Analysis of polarization mode dispersion in optical fiber communications and photonic microwave filters

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Analysis of Polarization Mode Dispersion in Optical Fiber Communications and Photonic Microwave Filters

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School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in fulfillment of the requirements for the degree of
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Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research and has not been submitted for a higher degree to any other University or Institution.

June 22, 2006
Date

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Abstract

The birefringence of real-life optical fibers causes pulse broadening in fiber-optic communication systems. This phenomenon is known as polarization mode dispersion (PMD). PMD has recently emerged as one of the critical concerns in achieving high-performance optical fiber transmission systems. A major part of the thesis is devoted to the study of PMD, together with chromatic dispersion (CD), in high bit-rate optical fiber communication systems.

A new approach has been presented in this thesis for the description of PMD and determining its effect on pulse and continuous RF signals. This approach has the advantage of (a) clear description of the behavior of the state-of-polarization (SOP) and the PMD vector, and (b) easy calculation of the effect of PMD and CD on pulse and continuous signals. The work presented in this dissertation includes: (i) A novel analytical expression relating the output state of polarization and the PMD vector including polarization-dependent chromatic dispersion (PCD), in terms of the angle of precession of the output state of polarization around the PMD vector. A number of new expressions incorporating for the first time this angle of precession have been derived. First, a general relation to study the effect of differential group delay (DGD), PCD and chromatic dispersion on pulses of arbitrary shapes has been derived. From this general relation, we derive expressions for pulse broadening and power penalty for Gaussian pulses. Measured results are presented to support the expressions derived above. To study the effects of higher-order PMD, two new terms have been introduced: depolarization vector and depolarization-dependent chromatic dispersion. A novel analytical expression relating the PMD vector and depolarization vector including depolarization-dependent chromatic dispersion (DCD), in terms of the angle

of precession of the PMD vector around the depolarization vector, has been derived. Pulse broadening and power penalty induced by second-order PMD, DCD and CD are also derived. (ii) Two novel PMD compensators that use Hi-Bi chirped gratings. First, A PMD compensator using a double pass configuration has been investigated. This is followed by a PMD compensator using a pair of Hi-Bi chirped gratings. Both compensators are substantially free from chromatic dispersion. (iii) Analysis of the RF power fading due to CD and PMD including PCD. A novel technique for simultaneous monitoring of PMD and CD has been demonstrated. For many aspects of the work described above, supporting experimental results are also presented. (iv) While in general PMD causes impairment in an optical fiber system, it may also be useful, for example, in photonic microwave filters. The thesis presents two novel designs for tunable photonic microwave notch filters, using DGD elements, which are free from the limitation of optical coherence and phase noise. The first design uses a Hi-Bi fiber together with an adjustable DGD element; the second design makes use of a variable DGD element and a Hi-Bi coupler. For both designs, the filters transfer function is derived and verified through measurement.

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List of Acronyms

ASE	amplified spontaneous emission
BER	bit error rate
BERT	bit error rate tester
CD	chromatic dispersion
DCD	depolarization-dependent chromatic dispersion
DFB	distributed feedback
DGD	differential group delay
DLOB	dual loop optical buffer
DOP	degree of polarization
DSB	double side band
DRA	distributed Raman amplifier
EDFA	erbium-doped fiber amplifier
EOM	electro-optic modulator
FBG	fiber Bragg grating
FEC	forward error correction
FSR	free spectral range
FWHM	full-width half maximum
Hi-Bi	high-birefringence
MZ	Mach-Zehnder
NOLM	nonlinear optical loop mirror
NRZ	nonreturn-to-zero
OSA	optical spectrum analyzer
PBS	polarization beam splitter

PC	polarization controller
PCD	polarization-dependent chromatic dispersion
PD	photodetector
PDL	polarization dependent loss
PM	polarization maintaining
PMD	polarization mode dispersion
PPG	pulse pattern generator
PRBS	pseudo-random binary sequence
PSP	principal state of polarization
RF	radio frequency
RFSA	radio frequency spectrum analyzer
RMS	root mean square
RZ	return-to-zero
SCM	sub-carrier multiplexing
SMF	single mode fiber
SOP	state of polarization
SSB	single side band
TTL	transistor-transistor logic
WDM	wavelength-division multiplexing

Notation

A	power penalty parameter for polarization mode dispersion
B	power penalty parameter for chromatic dispersion, polarization-dependent chromatic dispersion and depolarization-dependent chromatic dispersion
c	speed of light in vacuum
D	dispersion parameter in ps/(km.nm)
$\hat{e}_{\Omega+}, \hat{e}_{\Omega-}$	PSP _s characterized by two orthogonal unit Jones vectors
$f(t)$	initial pulse amplitude
$F(\omega - \omega_0)$	Fourier transform of $f(t)$
M	a Muller matrix
Δn_{eff}	effective refractive index difference between the slow and fast modes
P	a skewsymmetric matrix P corresponding to $\vec{\Omega} \times$
$P(t)$	signal power at the detector
P_i	optical power output from laser
P_e	detector electrical power
$\vec{q}$	depolarization vector
$\hat{q}_+$	unit slow depolarization vector
$\hat{q}_-$	unit fast depolarization vector
$\vec{S}$	state of polarization represented in terms of Stokes vector
$\hat{S}$	unit state of polarization represented in terms of Stokes vector
$\vec{S}_i$	input state of polarization in Stokes space

$\vec{S}_o$	output state of polarization
$\hat{S}_o(\omega)$	a unit vector in the direction of the output state of polarization
T	2×2 Jones vector
T_0	full width at half maximum (FWHM) of the optical pulse
x, y, z	fiber coordinates; z is the direction of propagation; x and y are the transverse coordinates—i.e., those of Jones space
2α	angle between $\hat{\Omega}$ and $\hat{S}_o$
$2\alpha_0$	angle between $\hat{q}$ and $\hat{\Omega}$
β	propagation constant
β_1	first-order frequency derivative of propagation constant β
β_2	second-order frequency derivative of propagation constant β
$\Delta\beta$	birefringence in optical fibers
β_s	propagation constant of slow mode
β_f	propagation constant of fast mode
$\hat{B}$	equal to $\hat{\Omega} \times \hat{S}_o$
σ^2	RMS width of broadened pulse
σ_0^2	initial RMS pulse width
$\hat{B}_0$	the unit vector corresponding to $\hat{q} \times \hat{\Omega}_0$
Δv_g	group velocity difference between the orthogonal modes
$\Delta\tau$	differential group delay
$\Delta\tau_1$	first-order frequency derivative of DGD
$\vec{\Omega}(\omega)$	PMD-vector in Stokes space

$\hat{\Omega}_0$	unit vector for the first-order PMD vector at ω_0
$\hat{\Omega}_+$	slow unit PSP in terms of Stokes vector
$\hat{\Omega}_-$	fast unit PSP in terms of Stokes vector
$\Delta\omega$	deviation from the angular carrier frequency ω_0 of the light
$\bar{\Omega}_\omega$	second-order PMD vector
$\hat{\Omega}$	unit PMD vector
$\hat{\Omega}_\omega$	rotation speed of the PSPs with frequency

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Chapter 1

Introduction

1.1 Background and motivation

The demand for information bandwidth has been increasing rapidly in recent years to support data, video, multimedia and other novel applications. Optical-fiber communication can meet this demand because an optical fiber has a potential bandwidth of 30 THz. However, there are a number of factors, which limit the transmission capacity of the single mode fiber (SMF), which is employed commonly. The first of these factors is the fiber attenuation with a value less than 0.2 dB/km [1]. The attenuation can be countered by using Erbium-doped fiber amplifier (EDFA) [2] and distributed Raman amplifier (DRA) [3]; the latter has wide gain bandwidth. The second factor limiting the transmission capacity is fiber chromatic dispersion (CD) that refers to the frequency dependence of the group velocity and causes broadening of optical pulses transmitted through fiber. More recently, a third factor limiting the transmission capacity has been identified. As the transmission bit rate increases, a new type of fiber dispersion arises due to the difference between the group velocities of light in the two orthogonal polarizations of the HE_{11} mode in SMF [4]. This phenomenon is referred to as polarization mode dispersion (PMD) [5]. In an ideal fiber, the two HE_{11} modes are indistinguishable. However, any real single mode fiber (SMF) has imperfections such as core deformation, bends, twists, and inner stress. These imperfections cause birefringence, that is, the group velocity of the two orthogonal HE_{11} modes is also different. Moreover, PMD in a fiber network is

intrinsically dynamic due to random variation of birefringence in fibers due to random perturbations of environmental factors such as temperature and stress [6]. With the transmission capacity increasing to 10 Gbit/s over a single channel and even 40 Gbit/s systems being planned, the penalties due to PMD and CD can cause severe system degradation.

In this chapter, fundamental concepts related to PMD and CD are described. The chapter also includes a brief description of PMD compensation and monitoring of PMD and CD. This is followed by a brief description of photonic microwave filters that can be realized using differential group delay (DGD) elements. The chapter also includes major objectives of this research, a listing of major contributions of this thesis, and a description of the organization of the thesis.

1.2 Polarization mode dispersion and chromatic dispersion

Generally, optical signals in the fiber suffer from dispersion effects, such as chromatic dispersion (CD) and polarization mode dispersion (PMD). CD refers to the frequency dependence of the group velocity. PMD is caused by asymmetries and stress distribution in the fiber core, which lead to birefringence. Since the effects of higher-order PMD are compounded in the presence of chromatic dispersion, the combined effect of PMD and CD is an important consideration because of the presence of both PMD and CD in many systems.

1.2.1 Polarization mode dispersion

To understand how PMD arises in optical fibers, it is useful to start with the concept of polarization followed by uniform birefringence within a fiber span. Any single

mode fiber can be considered as a concatenation of such uniform birefringence elements.

1.2.1.1 Polarization of light

Light in optical fibers can be treated as an electromagnetic wave. The electric field in two orthogonally polarized waves through the same fiber propagating in the same direction can be expressed in complex notation as

$$\begin{cases} \vec{E}_x(z, t) = \hat{x}A_xE_0 \exp j(\omega t - \beta z + \phi_x) \\ \vec{E}_y(z, t) = \hat{y}A_yE_0 \exp j(\omega t - \beta z + \phi_y) \end{cases} \quad (1.1)$$

The relative phase difference between the two waves is $\Delta\phi = \phi_y - \phi_x$. Moreover, both waves are traveling in z direction. The resultant optical field is the vector sum of these two orthogonal fields:

$$\vec{E}(z, t) = \vec{E}_x(z, t) + \vec{E}_y(z, t) \quad (1.2)$$

1.2.1.1.1 Jones vector representation

R. C. Jones introduced a matrix approach in the resolution of problems related to light polarization. To describe a state of polarization (SOP), he introduced a vector, called Jones vector [7]:

$$\vec{E} = \begin{bmatrix} A_x e^{j\phi_x} \\ A_y e^{j\phi_y} \end{bmatrix} \quad (1.3)$$

For example, $[\cos\theta \quad \sin\theta]^T$ represents a linearly polarized wave with the phase difference $\Delta\phi = \phi_y - \phi_x = 0$. Circular polarization can be represented by

$\frac{1}{\sqrt{2}}[1, -j]^T$ for right circular polarization, and $\frac{1}{\sqrt{2}}[1, j]^T$ for left circular polarization, where $\Delta\phi = \pi/2$. Other more general elliptical states of polarization can be represented by $[\cos\theta \quad \sin\theta e^{j\Delta\phi}]^T$. It is useful to note that each polarization state $\vec{E}_1$ has an orthogonal polarization state $\vec{E}_2$ defined by

$$\vec{E}_1^+ \vec{E}_2 = 0 \quad (1.4)$$

where $\vec{E}_1^+$ represents the conjugate transpose of $\vec{E}_1$. The transmission properties of any linear birefringent element can be represented by a frequency-dependent 2×2 Jones vector T , which transforms the input polarization state $\vec{E}_{in}$ into $\vec{E}_{out}$ according to

$$\vec{E}_{out} = T\vec{E}_{in} \quad (1.5)$$

Under the assumption of no polarization dependent loss (PDL), the frequency dependent Jones matrix can be given by

$$T(\omega) = e^{\alpha(\omega)} \begin{bmatrix} u_1(\omega) & u_2(\omega) \\ -u_2^*(\omega) & u_1^*(\omega) \end{bmatrix} \quad (1.6)$$

where $\alpha(\omega)$, $u_1(\omega)$ and $u_2(\omega)$ are complex functions of the angular frequency of light.

1.2.1.1.2 Stokes vector representation

Jones representation is a complex representation. In this representation, there are two angles describing an arbitrary Jones vector. These angles can be interpreted as coordinates in a spherical coordinate system, and each polarization state can be

represented by a Stokes vector, $\vec{S} = [s_1 \ s_2 \ s_3]^T$ on the unity sphere, which is called Poincare sphere. The three Cartesian components are defined as [8][9]:

$$\begin{cases} s_1 = \frac{|E_x|^2 - |E_y|^2}{|E_0|^2} = \cos 2\theta \\ s_2 = \frac{2 \operatorname{Re}(E_x E_y^*)}{|E_0|^2} = \sin 2\theta \cos \phi \\ s_3 = \frac{2 \operatorname{Im}(E_x E_y^*)}{|E_0|^2} = \sin 2\theta \sin \phi \end{cases} \quad (1.7)$$

where $|E_0|^2 = |E_x|^2 + |E_y|^2$. A polarization state represented on Poincare sphere is

Fig. 1.1. A polarization state represented by a Stokes vector on Poincare sphere

shown in Fig 1.1. The right and left circular polarizations are on the north and south poles respectively. Linear polarization states are represented on the equator. Further,

the optical intensity $s_0 = \left(|E_x|^2 + |E_y|^2 \right) / |E_0|^2$, is a fourth Stokes parameter relevant for the description of un- or partly polarized light. For completely polarized light, $s_0^2 = s_1^2 + s_2^2 + s_3^2$. The degree of polarization (DOP) is defined as $\sqrt{s_1^2 + s_2^2 + s_3^2} / s_0$. Similar to Jones vector, the transmission properties of any linear birefringent element can be represented by a frequency-dependent 3×3 Mueller matrix, M , which transforms the input polarization state $\vec{S}_i$ into $\vec{S}_o$ according to

$$\vec{S}_o = M \vec{S}_i \quad (1.8)$$

1.2.1.2 Birefringence

In an ideal optical fiber, which is circularly symmetric, the two orthogonally polarized HE_{11} modes are degenerate. In a real fiber with circular asymmetry, this degeneracy is disturbed. This can be expressed as a difference in the local propagation constants for the two modes [4]:

$$\Delta\beta = \beta_s - \beta_f = \frac{\omega(n_s - n_f)}{c} = \frac{\omega\Delta n_{eff}}{c} \quad (1.9)$$

where $\omega = 2\pi c / \lambda$ is the angular frequency of light, c is the speed of light in vacuum, $\Delta n_{eff} = n_s - n_f$ is the effective refractive index difference between the slow and fast modes, λ is the wavelength of light in vacuum, and β_s and β_f are the propagation constants of the two orthogonal modes. The difference between β_s with β_f is referred to as birefringence. The value of Δn_{eff} is usually between 10^{-7} and 10^{-5} .

The birefringence indicated in Eq. (1.9) is usually accompanied by a difference in the local group velocity. This can limit the bandwidth of the fiber by broadening pulses.

The group velocity difference is described by differential group delay (DGD), $\Delta\tau$. As shown in Fig.1.2, this can lead to broadening or splitting of a pulse.

Fig. 1.2. Pulse broadening due to DGD

The value of $\Delta\tau$ can be obtained by taking the frequency derivative of the propagation constants of Eq. (1.9):

$$\Delta\tau = \frac{d(\beta_s - \beta_f)}{d\omega} L = \left(\frac{\Delta n_{eff}}{c} + \frac{\omega d\Delta n_{eff}}{cd\omega} \right) L \quad (1.10)$$

The quantity $\Delta\tau/L$ is referred to as the PMD of the fiber and is usually expressed in units of ps/km. Typical PMD values of fiber are 0.1 to 1.5 ps/km [10].

1.2.1.3 Definition of PMD

The spatial variation of the birefringence in single mode fibers complicates the analysis of PMD. The concept of principal states of polarization (PSP), proposed by Poole and Wagner [11], is very useful for understanding the effects of PMD in fibers. The concept of PSP means that there exists at every frequency an orthogonal pair of input PSPs that map into orthogonal output PSPs independent of frequency to first-order for every fiber. The differential time delay between the two PSPs is called DGD.

In the first-order approximation for PMD, the DGD and PSPs are assumed constant over the spectrum of the signal; a pulse with input SOP aligned with fast or slow input PSP will arrive at the fiber output without distortion. Moreover, an input pulse can be decomposed into pulses aligned with the two input PSPs.

Polarization mode dispersion is characterized by the PMD vector, $\vec{\Omega}(\omega)$, in Stokes space, around which an output state of polarization (SOP), $\hat{S}_o$, rotates when the frequency changes [12], i.e.

$$\frac{d\hat{S}_o(\omega)}{d\omega} = \vec{\Omega}(\omega) \times \hat{S}_o(\omega) \quad (1.11)$$

This phenomenon is shown in Fig.1.3. However in a real fiber, with random

Fig. 1.3. Output state of polarization rotates around the PMD vector when frequency changes

polarization mode coupling, both the magnitude and direction of PMD vector are dependent on frequency, and then the output state of polarization can not be described simply as a rotation around the PMD vector. The modulus of the PMD vector, $\Delta\tau = |\vec{\Omega}|$, is the DGD between the PSPs. Thus

$$\vec{\Omega}(\omega) = \Delta\tau \hat{\Omega}(\omega) \quad (1.12)$$

where the unit vector $\hat{\Omega}(\omega)$ points in the direction of the slow PSP. The second-order PMD is described by the derivative [13]

$$\frac{d\vec{\Omega}(\omega)}{d\omega} = \Delta\tau_1 \hat{\Omega}(\omega) + \Delta\tau \hat{\Omega}_\omega \quad (1.13)$$

where $\Delta\tau_1 = d\Delta\tau/d\omega$ is called polarization-dependent chromatic dispersion (PCD) and $\hat{\Omega}_\omega = d\hat{\Omega}(\omega)/d\omega$ is rotation of PSPs with frequency. Second-order PMD has two terms, the first term is parallel to the first-order PMD vector, and the second-term $\hat{\Omega}_\omega$ is perpendicular to the first-order PMD vector.

1.2.1.4 Statistical nature of PMD

The principal states model has been discussed in section 1.2.1.3. However, PMD is different from many other lightwave system properties in that the impairment due to PMD usually changes stochastically because of the statistical nature of local birefringence. It has been reported that for long fiber lengths the average differential group delay, $\langle\Delta\tau\rangle$, has a square root of length dependence, $\langle\Delta\tau\rangle \propto \sqrt{L}$ [14-16]. The probability density function for DGD follows a Maxwellian distribution [13][17-20]. Hence the probability density function of the normalized DGD, $\rho = \Delta\tau/\langle\Delta\tau\rangle$, is given by

$$f(\rho, \langle \Delta \tau \rangle) = \frac{32\rho^2}{\pi^2 \langle \Delta \tau \rangle^3} \exp\left(-\frac{4\rho^2}{\pi \langle \Delta \tau \rangle^2}\right) \quad (1.14)$$

These results can be obtained theoretically by modeling the PMD using the three-dimensional PMD vector, $\vec{\Omega}$, and allowing this vector to grow in a random-walk-like process along the fiber [21][22]. By integration of the Maxwellian probability density function, it can be found that the probability for the DGD to be larger than $2\langle \Delta \tau \rangle$ is about 1.7% and 4.2×10^{-5} for it to be larger than $3\langle \Delta \tau \rangle$. This means that the DGD can be larger than $3\langle \Delta \tau \rangle$ during 22 minutes per year.

1.2.2 Chromatic dispersion

When an electromagnetic wave interacts with bound electrons of a silica fiber, the medium response is dependent on the optical frequency ω . This, referred to as material dispersion, manifests through the frequency dependence of the refractive index $n(\omega)$ [23]. Mathematically, the effects of fiber dispersion are accounted for by expanding the mode-propagation constant β in a Taylor series about the centre frequency ω_0 [23][24].

$$\beta(\omega) = n(\omega) \frac{\omega}{c} = \beta_0 + \beta_1(\omega - \omega_0) + \frac{1}{2}\beta_2(\omega - \omega_0)^2 + \dots \quad (1.15)$$

where

$$\beta_m = \left. \frac{d^m \beta}{d\omega^m} \right|_{\omega=\omega_0} \quad (m=0, 1, 2, \dots) \quad (1.16)$$

The envelope of a pulse signal moves at the group velocity ($v_g = 1/\beta_1$), while the parameter β_2 is responsible for pulse broadening. β_2 also depends on the wavelength

of the optical signals. The more commonly used system parameter is called dispersion parameter D (ps/(nm.km)) which is 17 ps/(nm.km) at 1550 nm for SMF. The quantity D , related to β_2 , can be expressed as

$$D = \frac{d\beta_1}{d\lambda} = -\frac{2\pi c}{\lambda^2} \beta_2 \quad (1.17)$$

The negative values of D correspond to normal dispersion (β_2 is positive), and positive values of D correspond to anomalous dispersion (β_2 is negative). Both negative and positive dispersion cause pulse broadening at the output of fiber.

1.3 Impairments caused by polarization mode dispersion and chromatic dispersion in optical fiber systems

Digital lightwave systems depend on undistorted transmission of optical pulses through long lengths of fiber. Dispersion effects such as PMD and CD cause pulse spreading and distortion and thus can lead to system penalties. Dispersion induced pulse broadening is detrimental for optical communication systems since it interferes with detection process which is based upon the criterion that pulse energy remains confined in its own time slot. Dispersion induced pulse broadening increases with fiber length.

The power penalty for first-order PMD is given by [25]

$$\varepsilon(dB) \cong A \frac{\Delta\tau^2 \gamma (1-\gamma)}{T_0^2} \quad (1.18)$$

where $0 \leq \gamma \leq 1$ is the power split ratio between the two orthogonal components of the signal, T_0 is the full width at half maximum (FWHM) of the optical pulse, and A is a

dimensionless parameter that depends on the optical shape and receiver filter characteristics.

The PMD limit is the maximum value of PMD for which system performance is acceptable. A commonly quoted value for the PMD limit is $0.1 T_0$. It has also been reported that for a digital system to avoid incurring a power penalty of 1 dB or greater for a time of 30 min per year, the average differential delay time between the principal states must be less than $0.14 T_0$ [4]. Moreover, the pulse distortion caused by second-order PMD depends on the spectral width of the pulse. The pulse broadening and power penalty induced by second-order PMD has been discussed in section 2.3.

Chromatic dispersion also places a limit on the maximum distance a signal can be transmitted without regenerating the original digital signal. The CD limited distance can be estimated by determining the transmission distance at which a pulse has been broadened by one bit interval. The estimated CD limited distance L_D for a signal having non-return to zero (NRZ) intensity modulation can be obtained by [4]

$$L_D = \frac{c}{\lambda^2 D B^2} \quad (1.19)$$

where B is the bit rate, c is the speed of light, and λ is the wavelength of optical signal. The CD limited distance decreases as the square of the bit rate. For conventional signal mode fiber, L_D turns out to be about 70 km for a 10 Gbit/s data with 1550 nm optical signal wavelength. Therefore it is necessary to compensate the chromatic dispersion of SMF before 10 Gbit/s signal can be transmitted for more than 70 km.

1.4 PMD compensation and monitoring of CD and PMD

Similar to the compensation of CD, compensation of PMD also can improve system performance. However, compensation of CD and PMD first requires monitoring of CD and PMD.

1.4.1 PMD compensation

Polarization mode dispersion compensation can be accomplished either in the optical or electrical domains. One optical technique is to split the signal into two orthogonal polarizations and delay one of the polarization components relative to the other [26]. Electrical compensation, equivalent to delaying the optical power in one polarization, is performed after detection at the receiver using phase diversity [27] or electrical filtering [28] [29]. However, these techniques cannot be performed in-line along a large optical system. In chapter 3, two novel tunable PMD compensators using Hi-Bi chirped fiber gratings have been discussed.

1.4.2 CD and PMD monitoring

Since dispersion (PMD and CD) causes distortion in an optical communication system, it is important to monitor PMD and CD. For example, in a reconfigurable system, it is required to monitor the accumulated CD and feed proper control signal to either chromatic dispersion compensator or equalizer. For management and monitoring purposes, simple, sensitive, flexible chromatic dispersion monitoring techniques may prove quite beneficial. Moreover, PMD equalization and optical PMD compensation are powerful techniques that potentially allow upgrading the currently installed fiber

links to higher-bit-rates. PMD monitoring is the only way to manage the accumulated PMD that does not interrupt the data transmission in high-speed optical systems and networks. A novel PMD monitoring technique has been discussed in chapter four. Further, a novel and simple technique for online simultaneous monitoring of PMD and CD has also been discussed in the same chapter.

1.5 Photonic microwave filters using DGD elements

In general, PMD induces impairment in an optical communication system. However, PMD can be useful in other fields, such as microwave photonic filters. The ability for processing microwave and millimeter-wave signals directly within the optical fiber has attracted significant interest. Because photonic microwave filters have many advantages such as large time bandwidth product, insensitivity to electromagnetic interference, low loss and light weight, many filters and signal processing structures have been reported in the literature [30-34]. In chapter five, two novel tunable photonic microwave filters have been discussed.

1.6 Objectives

PMD and CD can cause severe system degradation in high bit rate systems such as those operating at 10 Gbit/s or 40 Gbits/s. Therefore it is important to estimate the shape distortion and power penalty caused by PMD and CD for different types of pulse signals. Further, limitations of first-order PMD compensation and need for the compensation of higher-order PMD effects, which become significant as the bit rates approach 40 Gbit/s or higher, have been pointed out in the literature [35-43]. Although the effect of PMD on pulses [44-54], together with power penalty induced

by PMD [25], [55-61], have been areas of active study, some desirable studies have not been conducted so far. For instance, a general expression relating PMD to pulses of any shape, and including the precession angle between output state of polarization (SOP) and PMD vector, is not available. Such an expression should also include the effects of polarization-dependent chromatic dispersion (PCD) and chromatic dispersion. Thus, the first major objective of the thesis is to investigate the effect of PMD and chromatic dispersion (CD) in high bit-rate optical fiber communication systems. A general equation, which incorporates the precession angle and the effects of DGD, PCD and chromatic dispersion, on a pulse of any shape, is sought. Such an equation can lead to expressions for shape distortion, pulse broadening, and power penalty. These expressions for shape distortion and power penalty should be helpful in estimating the degradation due to PMD, including higher-order PMD effects, in high bit rate optical communication systems employing different types of pulse shapes.

To maintain adequate performance in optical fiber systems, compensation of PMD is necessary. Because of the temporal drift of the PMD characteristics, it is desirable to have an adjustable PMD compensator in systems in operation. Tunable PMD compensators have been proposed using high birefringence (Hi-Bi) nonlinearly chirped FBG [62-64]. However, fabrication of nonlinearly chirped gratings requires nonlinearly chirped phase masks, which are costly and difficult to fabricate. Tunable PMD compensators using Hi-Bi linearly chirped FBG have also been reported [65-67]. Unfortunately, the linearly chirped FBG also induces a chirp in the PMD compensated signal. Also, tuning of the FBG can shift the centre wavelength for the reflected signal. The second major objective is to propose novel PMD compensators that use Hi-Bi chirped gratings, are substantially free from chromatic dispersion, and are suitable for multi-wavelength systems.

Subcarrier multiplexing (SCM) is a modulation scheme where multiple signals are multiplexed in the RF domain and transmitted on a single optical carrier. SCM has several important applications in optical systems, including cable television. The transmission of analog and digital SCM signals under the influence of PMD has been studied earlier [68-69]. However, the effect of higher-order PMD and chromatic dispersion on RF signals has not been reported. Besides, although RF signals have been used to monitor PMD or CD [70], the existing techniques cannot monitor PMD and CD at the same time. Thus the third major objective here is to analyse RF power fading including the effects of higher order PMD and CD. Application of the results of this study has also been suggested in a novel scheme for simultaneous monitoring of PMD and CD.

Generally, PMD induces impairment in an optical fiber communication system; however, PMD can be useful in photonic microwave filters. As mentioned in section 1.5, because of their advantages, many photonic microwave filters and signal processing structures have been reported [30-34]. The final objective of this work is to propose novel designs for tunable photonic microwave notch filters employing DGD elements.

1.7 Major contributions of the thesis

In this thesis, a new approach is presented for the mathematical description of polarization mode dispersion, and its effects on pulsed and continuous RF signals. This approach clearly brings out the behaviour of the SOP as well as the behaviour of PMD vector when first- and higher-order PMD effects are included. Further, this approach permits an easy calculation of the effects of PMD and CD on pulsed and

continuous RF signals. This is the general contribution of this thesis. The specific contributions of the thesis are listed below:

1. A novel analytical expression relating the output state of polarization and the PMD vector including polarization-dependent chromatic dispersion (PCD); this expression explicitly includes the angle of precession of the output state of polarization around the PMD vector. This expression is used to derive a number of new expressions, incorporating for the first time this angle of precession. First, a general relation to study the effect of DGD, PCD and chromatic dispersion on pulses of arbitrary shapes has been derived. From this general relation, we derive expressions for pulse broadening and power penalty for Gaussian pulses; also derived are the expressions for pulse broadening, the power penalty as well as the average power penalty. The power penalty results are verified experimentally.
2. A novel analytical expression for the higher-order PMD vector including second-order effects and depolarization-dependent chromatic dispersion (DCD) is presented. Based on this expression, we get the phase difference value between the two orthogonal depolarization vectors. For this case also, expressions for the pulse broadening and power penalty due to higher-order PMD (including second-order PMD and DCD) and chromatic dispersion are derived.
3. Two novel PMD compensation methods using a linearly chirped FBG in a double-pass configuration have been presented. For both schemes, measured results verify that the chromatic dispersion is reduced significantly in these PMD compensator configurations. Furthermore, both schemes offer tunability

over a large bandwidth and are fiber compatible. The double-pass configuration proposed here can compensate PMD over a large range. The two-FBG design presented here makes the DGD tuning range start from zero, is free from cross-talk and can compensate PMD over a range of wavelengths.

4. New expressions for power fading of RF signals due to higher-order PMD including PCD and chromatic dispersion are derived. These expressions explicitly include the angle of precession of the output SOP around the PMD vector. Based on these expressions, we propose a simple and novel pilot-tone-based technique for simultaneous monitoring of chromatic dispersion and PMD. Experiment results demonstrate the effectiveness of the proposed techniques; these results also agree very well with the theoretical analysis. Moreover, based on the power fading expression, we suggest a novel method to measure the DGD between the fast and slow axes of a Hi-Bi chirped grating. Further, a similar method can measure the PMD of a chirped grating easily.
5. Two continuously tunable photonic microwave notch filter configurations, which are free from chromatic dispersion, are proposed. The first uses a combination of a Hi-Bi fiber and a variable differential group delay (DGD) element; the other is based on a variable polarization beam splitter (PBS), a 3 dB Hi-Bi coupler, and a variable time delay line. Two orthogonally polarized light beams are generated in these schemes; hence these are free from the problem of optical coherence interference and the associated phase noise problem. Expressions for the filters transfer function are derived and measured results are presented to demonstrate the operation of the filters. Measured results agree well with the theory. These configurations are also free from the

problem of chromatic dispersion. The two schemes can operate over a wide range of optical carrier frequencies, restricted only by the bandwidth of the various optical components in the system. Further, these structures can be easily incorporated in optical fiber microwave transmission systems.

1.8 Organization of the Thesis

This thesis consists of six chapters.

In Chapter 1, fundamental concepts related to PMD and CD are described. The chapter also includes a brief description of PMD compensation and monitoring of PMD and CD. This is followed by a brief description of photonic microwave filters that can be realized using differential group delay (DGD) elements. The chapter also includes major objectives of this research, a listing of major contributions of this thesis, and a description of the organization of the thesis.

In Chapter 2, we first present an expression relating the frequency dependent output state of polarization, $\vec{S}_o(\omega)$, and the PMD vector $\vec{\Omega}$, including PCD. This relation is consistent with the established behaviour of $\vec{S}_o(\omega)$ with respect to $\vec{\Omega}$, and clearly describes the details of the behaviour of SOP with respect to the PMD vector including the effect of PCD. Based on this expression, we get the phase difference value between the two PSPs. Following this, we derive a general equation, which incorporates the precession angle, and the effects of DGD, PCD and chromatic dispersion, on a pulse of any shape. This new equation is used to obtain expressions for shape distortion, pulse broadening, and power penalty for the special case of Gaussian pulses. An expression for average power penalty has also been derived.

Next, we present an expression for the higher-order PMD vector including second-order effects, namely, depolarization and depolarization-dependent chromatic dispersion (DCD). Both of these are new terms and are explained in Section 2.4. Based on this expression, we derive the phase difference value between two orthogonal depolarization vectors. Further, the pulse broadening and power penalty induced by the higher-order PMD effects (including second-order PMD and DCD) and chromatic dispersion are also discussed. These expressions for shape distortion and power penalty should be helpful in estimating the degradation due to PMD, including higher order PMD effects, in high bit rate optical communication systems employing different types of pulse shapes.

Chapter 3 presents two novel PMD compensation methods using a linearly chirped FBG. First, we propose a new PMD compensation scheme using a Hi-Bi linearly chirped FBG in a double pass configuration. This scheme offers tunability over a large bandwidth and is fiber compatible as well as can compensate PMD over a large range. Measured results are presented to show that such a configuration achieves significantly reduced chromatic dispersion, together with PMD compensation. Next, we propose another new tunable polarization mode dispersion (PMD) compensator, which is substantially free from chromatic dispersion. The design uses a pair of high-birefringence linearly chirped fiber gratings. Each FBG can adjust the differential group delay (DGD) nonlinearly by using a cantilever beam. The compensator can be tuned to operate over a range of wavelengths and can yield DGD values starting from 0 ps. Another feature of this design is that, compared to the double pass configurations, it is free from cross-talk. Measured eye-diagrams for a 10 Gbit/s system show that this compensator can greatly reduce the chromatic dispersion induced by chirped grating.

In Chapter 4, we use the expression relating the output state of polarization and PMD vector, including polarization-dependent chromatic dispersion (PCD), as discussed in chapter two. Based on this expression we get the power fading of RF signals including higher-order PMD and chromatic dispersion, which is dependent on the angle of precession of output state of polarization around the PMD vector. From the expression for power fading, we get the average power penalty for chromatic dispersion and higher-order PMD. Moreover, a simple and novel pilot-tone-based technique is proposed for simultaneous monitoring of chromatic dispersion and PMD. Experiment results demonstrate the effectiveness of the proposed technique; these results also agree very well with the theoretical analysis. Further, based on the power fading expression, we propose a novel method to measure the DGD between the fast and slow axes of Hi-Bi chirped grating. A similar method can measure the PMD of chirped grating easily.

In Chapter 5, first we propose a novel configuration for a continuously tunable photonic microwave notch filter, which is free from chromatic dispersion. The filter uses a combination of a Hi-Bi fiber and a variable differential group delay (DGD) element. The proposed system can operate over a wide range of optical wavelengths. Measured results presented agree well with the theory. Next, we also demonstrate a tunable microwave-photonic filter based on a variable polarization beam splitter (PBS), a 3 dB Hi-Bi coupler, and a variable time delay line. This configuration also is free from the problems of optical coherence interference and chromatic dispersion. A notch rejection greater than 35 dB is obtained. It is shown that the notch frequency is continuously tunable by tuning the time delay line. The scheme can operate over a wide range of optical carrier frequencies, restricted only by the bandwidth of the various optical components in the system.

Chapter 6 summarizes the important conclusions of the thesis, and includes suggestions for further work. Some possible applications of the results in this thesis are also suggested.

Chapter 2

Effect of Polarization Mode Dispersion and Chromatic Dispersion on Pulse Signals

2.1 Introduction

In optical fiber communication systems, at high bit rates such as 10Gbit/s or 40Gbit/s, polarization mode dispersion (PMD) is considered the ultimate limiting factor for transmission of information [6]. Therefore, in the case of transmission of pulse signals, it is important to estimate the shape distortion and power penalty caused by PMD for different types of signals. Further, there is a great interest in techniques to compensate the effects of PMD. There have been several studies, which point out the limitations of the first-order PMD compensation and the need for the compensation of higher-order PMD effects [35-43]. Higher order PMD effects become significant as the bit rates approach or exceed 40 Gbit/s. Therefore, in this chapter, the effect of PMD on pulse signals is studied. Higher order PMD effects, as well as the effect of chromatic dispersion are particularly emphasized.

PMD is characterized by the PMD vector, $\vec{\Omega} = \Delta\tau\hat{\Omega}$, which has its magnitude as the differential group delay (DGD) $\Delta\tau$, and $\hat{\Omega}$ is the unit PMD vector directed along the slow principal state of polarization (PSP) [12]. It is well known that when frequency changes, in the absence of higher-order PMD, the output state of polarization (SOP) precesses around the PMD vector. The angle of precession depends on the angle

between the SOP and the PSP. However, a clear mathematical expression governing the details of precession, such as rate of rotation, does not appear to be available in the literature. Next, while the effect of PMD on signal pulses has been an active area of study [e.g., 44-54], a general expression relating PMD to pulses of any shape, and including the precession angle between the SOP and the PMD vector as well as polarization-dependent chromatic dispersion (PCD) and chromatic dispersion, has not been presented so far. Further, power penalty induced by PMD, including the effect of PCD on signal pulses has also been an active area of study [e.g., 55-60]. However, once again, the power penalty as a function of the precession angle, taking into account PMD as well as PCD and chromatic dispersion, has not been reported and experimentally verified. When the power penalty expression is available in terms of the precession angle, the average power penalty can be derived conveniently.

In this chapter, we first present a new expression relating the frequency dependent output state of polarization, $\vec{S}_o(\omega)$, and the PMD vector $\vec{\Omega}$, including PCD. This relation is consistent with the established behaviour of $\vec{S}_o(\omega)$ with respect to $\vec{\Omega}$, and clearly describes the details of the behaviour of SOP with respect to the PMD vector including the effect of PCD. Based on this expression, we get the phase difference value between the two PSPs. Following this, we derive a general equation, which incorporates the precession angle, and the effects of DGD, PCD and chromatic dispersion, on a pulse of any shape. This new equation is used to obtain expressions for shape distortion, pulse broadening, and power penalty for the special case of Gaussian pulses. An expression for average power penalty has also been derived.

Next, we present an expression for the higher-order PMD vector including second-order effects, namely, depolarization and depolarization-dependent chromatic

dispersion (DCD). Both of these are new terms and are explained in Section 2.4. Based on this expression, we derive the phase difference value between two orthogonal depolarization vectors. Further, the pulse broadening and power penalty induced by the higher-order PMD effects (including second-order PMD and DCD) and chromatic dispersion are also discussed.

These expressions for shape distortion and power penalty should be helpful in estimating the degradation due to PMD, including higher-order PMD effects, in high bit rate optical communication systems employing different types of pulse shapes.

2.2 First-order PMD

2.2.1 Relation between output SOP and PMD vector

In this section, we neglect the polarization dependent loss (PDL) and polarization-dependent chromatic dispersion (PCD). The frequency-dependent behaviour of the output SOP is given by [12]:

$$\frac{d\hat{S}_o(\omega)}{d\omega} = \vec{\Omega} \times \hat{S}_o(\omega) \quad (2.1)$$

When we neglect higher-order PMD, $\vec{\Omega}$ can be considered as a constant vector. Replacing $\vec{\Omega} \times$ by a skew symmetric matrix $\mathbf{P}$, and following some complex calculations [See Appendix I] [71-72], one gets:

$$\hat{S}_o(\omega) = \hat{\Omega} \cos 2\alpha + [\hat{S}_o(\omega_0) - \hat{\Omega} \cos 2\alpha] \cos \Delta \tau \Delta \omega + \hat{B} \sin 2\alpha \sin \Delta \tau \Delta \omega \quad (2.2)$$

where 2α is the angle between $\hat{\Omega}$ and $\hat{S}_o$, $\hat{B}$ is the unit vector $\hat{\Omega} \times \hat{S}_o$ and $\Delta \omega$ is the frequency deviation with respect to (w.r.t.) ω_0 . In Eq. (2.2), the three unit vectors

$\hat{\Omega}$, $(1/\sin 2\alpha)[\hat{S}_o(\omega_0) - \hat{\Omega} \cos 2\alpha]$ and $\hat{B}$ constitute a 3-D orthogonal coordinate system. Eq. (2.2) means that the output state of polarization precesses around the PMD vector with a precession angle 2α . This is shown in Fig.2.1. The precession angle 2α is independent of frequency, but depends upon the angle between the input SOP and $\vec{\Omega}$; 2α is zero when these two vectors have the same direction; when these vectors are inclined at 45° , 2α is 90° . Further, in the plane normal to $\vec{\Omega}$, projection of $\vec{S}_o(\omega)$ rotates by an angle $\Delta\tau\Delta\omega$.

Fig. 2.1. Relation between output state of polarization $\hat{S}_o(\omega)$ and unit PMD vector $\hat{\Omega}$

The abovementioned interpretation of Eq. (2.2) is verified through the following experiment. The experimental setup is shown in Fig. 2.2. The output from a tunable laser is connected through a quarter-wave plate to a first-order PMD emulator; the

Fig. 2.2. Experimental setup

emulator consists of different lengths of Panda high-birefringence (Hi-Bi) fiber. The results for the observed SOP are shown in Fig. 2.3 When we change the length of the

(a) Small DGD value

(b) Larger DGD value

(c) Rotating $\lambda/4$ waveplate and large DGD value

Fig. 2.3. Measured results for output SOP

Panda Hi-Bi fiber to realize different values of DGD, the speed of rotation of the SOP changes but the size of the SOP circle remains the same (Figs. 2.3 (a) and 2.3 (b)). If we rotate the quarter-wave plate, the angle 2α changes; as a result, the radius of the circle representing the output SOP changes but the speed of rotation of the SOP does not change (Fig. 2.3 (c)). Similarly, when we use a DFB laser whose linewidth is smaller than that for the tunable laser, i.e., $\Delta\omega$ reduces, with all other conditions the same, the speed of rotation is observed to be smaller with the circle size in the Poincare sphere unchanged.

2.2.2 Pulse broadening and power penalty due to first-order PMD

2.2.2.1 Derivation of results

Next, consider propagation of a pulse signal $f(t)$ on the fiber. The signal has its Fourier transform:

$$F(\omega - \omega_0) = \int_{-\infty}^{+\infty} f(t) \exp[-j(\omega - \omega_0)t] dt \quad (2.3)$$

The pulse, after propagation, can be decomposed into two components, $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$, one each along the two orthogonal PSPs, $\hat{e}_{\Omega+}$ and $\hat{e}_{\Omega-}$ respectively.

Using the fact that the two components differ in phase by $\Delta\tau\Delta\omega$ [11], the components $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$ are written as:

$$f_{\Omega+}(t) = \frac{\cos\alpha}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp(j\Delta\tau\Delta\omega/2) e^{j(\omega - \omega_0)t} d\omega \quad (2.4)$$

$$f_{\Omega-}(t) = \frac{\sin\alpha}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp(-j\Delta\tau\Delta\omega/2) e^{j(\omega - \omega_0)t} d\omega \quad (2.5)$$

Combining the components, one gets

$$\vec{f}(t) = f_{\Omega+}(t)\hat{e}_{\Omega+} + f_{\Omega-}(t)\hat{e}_{\Omega-} \quad (2.6)$$

Eqs (2.4)-(2.6) are general equations, applicable to a pulse of any shape. To the best of our knowledge, it is the first time such a general equation for the effect of PMD on pulse signals has been presented. When the input pulse is Gaussian, i.e., $f(t) = M \exp(-t^2/2T_0^2)$,

$$f_{\Omega+}(t) = M \cos \alpha \exp\left\{-\left[(\Delta\tau/2 + t)^2/2T_0^2\right]\right\} \quad (2.7)$$

$$f_{\Omega-}(t) = M \sin \alpha \exp\left\{-\left[(\Delta\tau/2 - t)^2/2T_0^2\right]\right\} \quad (2.8)$$

In the above equations, T_0 is the full width at half maximum (FWHM) of the Gaussian pulse and M is the magnitude factor of Gaussian pulse. The signal power at the detector is $P(t) = |\vec{f}(t)|^2 = \vec{f}(t) \cdot \vec{f}^*(t)$. Since $\hat{e}_{\Omega+}$ is orthogonal to $\hat{e}_{\Omega-}$,

$$\begin{aligned} P(t) &= |f_{\Omega+}(t)|^2 + |f_{\Omega-}(t)|^2 \\ &= M^2 \cos^2 \alpha \exp\left[-(\Delta\tau/2 + t)^2/T_0^2\right] + M^2 \sin^2 \alpha \exp\left[-(\Delta\tau/2 - t)^2/T_0^2\right] \end{aligned} \quad (2.9)$$

Eq. (2.9) indicates that the output waveform, affected by the first-order PMD, consists of two humps whose relative magnitudes are dependent on $\cos^2 \alpha$ and $\sin^2 \alpha$, respectively. The RMS pulse width of the optical pulse as a function of 2α is [48]:

$$\sigma^2 = \langle t^2 \rangle - \langle t \rangle^2 = \frac{\int_{-\infty}^{+\infty} t^2 P(t) dt}{P_0} - \left[\frac{\int_{-\infty}^{+\infty} t P(t) dt}{P_0} \right]^2 = \frac{T_0^2}{2} + \left(\frac{\Delta\tau}{2} \right)^2 \sin^2 2\alpha \quad (2.10)$$

Since the input pulse width is $\sigma_0 = T_0/\sqrt{2}$, the RMS width of the broadened pulse is

$$\sigma^2 = \sigma_0^2 + (\Delta\tau/2)^2 \sin^2 2\alpha \quad (2.11)$$

Assuming only small changes in the output pulse, the fractional RMS broadening is

$$\frac{\sigma - \sigma_0}{\sigma_0} = \frac{1}{2} (\Delta\tau^2 \sin^2 2\alpha / 4\sigma_0^2) = (\Delta\tau/2T_0)^2 \sin^2 2\alpha \quad (2.12)$$

The fractional RMS broadening and the corresponding power penalty reported in [25] (using the present notation) are:

$$\varepsilon(dB) = A(\Delta\tau/T_0)^2 \gamma(1-\gamma) \quad (2.13)$$

In Eq (2.13), γ is the ratio of optical power in one PSP relative to the entire power and the factor A is dependent on the pulse shape and modulation format as well as the receiver characteristics [6][25][59]. Our expression clearly shows that RMS pulse broadening depends on $\sin^2 2\alpha$, that is the angle between output SOP and PMD vector. Further, from Eqs (2.10)-(2.13), the power penalty can be expressed as

$$\varepsilon(dB) = A(\Delta\tau/2T_0)^2 \sin^2 2\alpha \quad (2.14)$$

Thus when the angle between the input SOP and $\vec{\Omega}$ changes, the power penalty, that is dependent on 2α , also changes. When the input signal excites all the power in one PSP, that is $2\alpha = 0$, the power penalty is 0. The maximum power penalty occurs when the input signal excites power equally in the two PSPs; this value is:

$$\varepsilon(dB) = A(\Delta\tau/2T_0)^2 \quad (2.15)$$

The distribution of $\sin(2\alpha)$, where 2α is the angle between two random polarization vectors (output state of polarization and the PMD vector) with a uniform distribution, is given by [73]. The $\sin(2\alpha)$ distribution f_x and $\cos(2\alpha)$ distribution f_y are given by [73]:

$$\begin{cases} f_x = \frac{x}{\sqrt{1-x^2}} & (0 \leq x \leq 1) \\ f_y = 1/2 & (-1 \leq y \leq 1) \end{cases} \quad (2.16)$$

Using Eqs. (2.15) and (2.16), the average power penalty is

$$\bar{\varepsilon}(\text{dB}) = A(\Delta\tau/2T_0)^2 \int_0^1 \sin^2 2\alpha \frac{\sin 2\alpha}{\sqrt{1-(\sin 2\alpha)^2}} d \sin 2\alpha = \frac{2}{3} A(\Delta\tau/2T_0)^2 \quad (2.17)$$

This means that when the angle between output state of polarization and the PMD vector varies uniformly, the penalty is about 2/3 of the worst-case penalty. The derivation of Eq. (2.17) based on the expression for output state of polarization and PMD vector is novel. Next, we undertake an experimental verification of the power penalty result in Eq. (2.17), which has not been provided earlier.

It should be pointed out that the new expression Eq. (2.14) is more basic compared to Eq. (2.13) [25], since the role of angle 2α is more basic compared to that of power split ratio. The pulse, after propagation in the optical fiber, can be decomposed into two components, $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$, one each along the two orthogonal PSPs, $\hat{e}_{\Omega+}$ and $\hat{e}_{\Omega-}$ respectively, where $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$ are proportional to $\cos(\alpha)$ and $\sin(\alpha)$, respectively. This step is fundamental in determining the effects of higher-order PMD and CD on a pulse. Further, measurement of the power split ratio is also based on the relation $[\sin(\alpha)]^2 = \gamma$.

2.2.2.2 Experimental verification

The experimental setup is shown in Fig. 2.4 [74]. Light of wavelength 1552.1 nm is modulated with a 10 Gbit/s return-to-zero (RZ) pseudorandom pattern of length $2^{15}-1$ using two X-cut LiNbO₃ Mach-Zehnder (MZ) modulators. The first MZ modulator generates a 10 Gbit/s pseudo-random binary sequence (PRBS) signal. The nonreturn-to-zero (NRZ) signal is converted to a 50% duty-cycle RZ signal by pulse carving with a second MZ modulator driven by the pulse pattern generator (PPG)

Fig. 2.4. Experimental setup

$\overline{Clock}$, which is a 10 GHz sinusoidal signal. The modulated signal passes through an optical amplifier. A filter comprising a uniform fiber Bragg grating (FBG) and a circulator gets rid of the ASE noise of the EDFA. The centre wavelength and bandwidth of the FBG are 1552.15 and 0.2 nm respectively. Then the signal passes through an attenuator and a polarization controller (PC), which can change the value of 2α , and a PMD emulator, which consists of two polarization beam splitters and a

Fig. 2.5. Measured pulse shape results for the set-up in Fig. 2.4 (x-axis: 50 ps /Div)

DGD element with Hi-Bi pigtail. Following this, the signal is detected by a photodetector (PD) and is observed on a sampling oscilloscope.

For a fixed data sequence, the measured pulse shapes for different values of 2α are shown in Fig. 2.5 (a)-(d). The corresponding eye-diagrams are shown in Fig. 2.6; these results show that when $2\alpha = 0$ or $2\alpha = \pi$, the eye-diagram is almost open and the output pulse has just one peak. However, when 2α has some other value, the

Fig. 2.6. Measured eye diagram results for the set-up in Fig. 2.4 (x-axis: 20 ps /Div)

output waveform, affected by the first-order PMD, consists of two humps whose magnitudes are dependent on $\cos^2 \alpha$ and $\sin^2 \alpha$ respectively. When $2\alpha = \pi/2$, the magnitudes of the humps are the same, as shown in Fig. 2.6 (b). The third waveform in Fig. 2.6 (b)-(d), is the signal value obtained by adding the two humps. For $2\alpha = 0$ or $2\alpha = \pi$, the third value is the waveform itself. It is clear that the magnitude of the third waveform is the lowest when $2\alpha = \pi/2$, and the largest, when $2\alpha = 0$ or $2\alpha = \pi$.

Thus, the power penalty is dependent on $\sin^2 \alpha$ and the experimental results support Eq. (2.14). The DGD value used for Figs. 2.5 and 2.6 is 67 ps.

The experimental setup for measuring average power penalty due to PMD (not including CD) is shown in Fig. 2.7. The signal from a laser whose output power can be tuned is fed to a MZ modulator with 10 Gbit/s $2^{15}-1$ PRBS NRZ data. The patterns with $N=15$ and $N=23$ have been adopted in international standards [75]. Generally, the larger the N , the more difficult is the synchronous clock recovery in BER tester. Therefore a larger value of N is likely to cause a higher BER. The output of the modulator was transmitted through a quarter-wave plate that is used to change the excitation ratio between the fast and the slow axis of the following first-order PMD emulator; the PMD emulator consists of two polarization beam splitters

Fig. 2.7. Experimental setup (without chromatic dispersion)

with Hi-Bi pigtails and a DGD element whose pigtail is also a Hi-Bi fiber. The output of the first-order PMD emulator is fed to a photo-detector that is connected to an error detector. The bit error rate tester is used to measure the power penalty due to PMD at a bit error rate of 10^{-9} . The power meter is used to ensure that the input power to the

Fig. 2.8. Power penalty vs. the angle between the output state of polarization and the PMD vector, with optimum decision threshold

photodetector is the same for different values of DGD. On the optical bench, by adjusting the waveplate, one can determine the slow and fast axes of the PMD emulator. This determines the orientation of the PSPs. Rotation of the waveplate with respect to this reference position gives 2α and can be easily read on the waveplate. Fig. 2.8 shows the measured power penalty for three different values of DGD (66ps, 54ps and 30ps) for different values of the angle between output SOP and PMD vector, when the decision threshold is optimized. There was no penalty at first, for high input laser power; with suitable reduction of the input power, the power penalty appears. A value of $A = 38$ fits the measured curves. So the calculated results plotted in Fig. 2.8 have been obtained from Eq. (2.14) using $\varepsilon(dB) = 38(\Delta\tau/2T_0)^2 [\sin(2\alpha)]^2$; other parameters used for calculations are shown in Table 2.1.

Table 2.1: Parameters and equation of calculated results in Fig.2.8

Fig.	$\Delta\tau$ (DGD)	A	T_0	Eq.
2.8	30 ps; 54 ps; 66 ps	38	100 ps	2.14 $\varepsilon(dB) = 38(\Delta\tau/2T_0)^2 [\sin(2\alpha)]^2$

Since A , in particular, needs to be determined by simulation and experiment [6], there does not appear to be available a straightforward method for estimating A . Our results (for example Fig. 2.8) show that, based on the experimental data, a value of A can be determined so that there is a good match between the theory and experiment.

The discrepancy between the measured and the calculated power penalties increases as DGD increases (near the peaks in Fig.2.8) This is due to the following reason.

Since $\varepsilon(dB) = A(\Delta\tau/2T_0)^2 \sin^2 2\alpha$, when 2α is close to 90° , the power penalty assumes the largest value for a given DGD. Similarly, as DGD increases, power penalty increases. Therefore, as DGD increases, the discrepancy between the measured and the calculated power penalties appears large, particularly near the peaks. If one considers the discrepancy as a percentage of the measured value, the error is quite uniform.

2.2.3 Pulse broadening and power penalty due to first-order PMD and CD

If we include the effect of chromatic dispersion, then the two components of the pulse given in Eq.(2.4)-(2.5) can be expressed as:

$$f_{\Omega+}(t) = \frac{\cos\alpha}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp \left[j(\Delta\tau\Delta\omega/2) - j(\beta_2 L \Delta\omega^2/2) \right] e^{j(\omega - \omega_0)t} d\omega \quad (2.18)$$

$$f_{\Omega-}(t) = \frac{\sin \alpha}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp \left[(-j \Delta \tau \Delta \omega / 2) - (j \beta_2 L \Delta \omega^2 / 2) \right] e^{j(\omega - \omega_0)t} d\omega \quad (2.19)$$

In the above equations, L is the length of single mode fiber and $\beta_2 = \left(\partial^2 \beta / \partial \omega^2 \right) \Big|_{\omega=\omega_0}$ (β is the propagation constant). For a Gaussian pulse ($f(t) = M \exp(-t^2/2T_0^2)$), following the steps in Eq. (2.6)-(2.8), the signal power at the detector can be expressed as

$$\begin{aligned} P(t) &= |f_{\Omega+}(t)|^2 + |f_{\Omega-}(t)|^2 \\ &= \frac{M^2 T_0^2 \cos^2 \alpha}{\sqrt{T_0^4 + (\beta_2 L)^2}} \exp \left\{ -\frac{(\Delta \tau / 2 + t)^2 T_0^2}{(T_0^4 + (\beta_2 L)^2)} \right\} \\ &\quad + \frac{M^2 T_0^2 \sin^2 \alpha}{\sqrt{T_0^4 + (\beta_2 L)^2}} \exp \left\{ -\frac{(\Delta \tau / 2 - t)^2 T_0^2}{(T_0^4 + (\beta_2 L)^2)} \right\} \end{aligned} \quad (2.20)$$

Following the procedure in section 2.2.2.1, the RMS pulse width is

$$\sigma^2 = \frac{T_0^2}{2} + \frac{(\beta_2 L)^2}{2T_0^2} + \left(\frac{\Delta \tau}{2T_0} \right)^2 \sin^2(2\alpha) \quad (2.21)$$

Similar to Eq. (2.14), the power penalty can be expressed as

$$\varepsilon(dB) = B \frac{(\beta_2 L)^2}{2T_0^2} + A \left(\frac{\Delta \tau}{2T_0} \right)^2 \sin^2(2\alpha) \quad (2.22)$$

where the factor A depends on the pulse shape and modulation format as well as the receiver bandwidth [6][25][59]; B is a parameter for power penalty due to chromatic dispersion and is also dependent on the pulse shape, modulation format and receiver bandwidth. The pulse shape (due to different modulation formats) affects the parameters A and B strongly. Usually, NRZ is more tolerant to residual CD and less tolerant against first-order polarization mode dispersion (PMD), that is, A is larger and B is smaller, compared with RZ. The reason is that NRZ has a lower spectrum occupancy. Next, compared to RZ, NRZ format has poorer margin in the time domain

and is distorted more by PMD. These effects have been recently verified through measurements [74] [76].

From Eq. (2.22), when the angle between the input SOP and $\vec{\Omega}$ changes randomly with a uniform probability, the average power penalty can be expressed as

$$\varepsilon(dB) = B \frac{(\beta_2 L)^2}{2T_0^2} + \frac{2A}{3} \left(\frac{\Delta\tau}{2T_0} \right)^2 \quad (2.23)$$

Eq. (2.23) is the expression for average power penalty for a Gaussian pulse including the effect of PMD and chromatic dispersion.

2.3 Relation between output SOP and PMD vector including PCD

The second-order PMD is the frequency derivative of $\vec{\Omega}$:

$$\frac{d\vec{\Omega}}{d\omega} = \frac{d\Delta\tau}{d\omega} \hat{\Omega} + \Delta\tau \frac{d\hat{\Omega}}{d\omega} \quad (2.24)$$

where the first term on the right-hand side is parallel to $\hat{\Omega}$ and is caused by PCD, and the second term, perpendicular to $\hat{\Omega}$, is the rotation of the PSPs with frequency. If we take the polarization-dependent chromatic dispersion into account and neglect the rotation of PSPs, then Eq. (2.24) can be expressed as [77]

$$\begin{aligned} \hat{S}_o(\omega) = & \hat{\Omega} \cos 2\alpha + \left[\hat{S}_o(\omega_0) - \hat{\Omega} \cos 2\alpha \right] \cos(\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2) \\ & + \hat{B} \sin 2\alpha \sin(\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2) \end{aligned} \quad (2.25)$$

where $\Delta\tau_1$ is PCD, defined as $\Delta\tau_1 = d\Delta\tau/d\omega$. In this case also, the output state of polarization precesses around the PMD vector with precession angle 2α . However, the projection of the output state of polarization in the plane normal to the PMD vector rotates by an angle $\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2$; this result is derived in the next section.

2.3.1 Determination of phase difference between the principal states of polarization (PSPs) in the presence of PCD

The phase difference $\Delta\phi(\omega)$ between the two PSPs plays an important role in determining the output pulse shape for a given input pulse. Let the SOP be represented as $\hat{S}_o(\omega)$ in Stokes notation and as $\hat{e}(\omega)$ in the Jones notation. We can express the SOP in terms of the two orthogonal principal states of polarization, $\hat{e}_{\Omega+}$ and $\hat{e}_{\Omega-}$ in Jones vector notation:

$$\hat{e}(\omega) = \cos\alpha e^{j\phi_+} \hat{e}_{\Omega+} + \sin\alpha e^{j\phi_-} \hat{e}_{\Omega-} \quad (2.26)$$

For the special case of $\alpha = \pi/4$, one gets $\hat{e}(\omega)|_{\alpha=\pi/4} = \hat{e}'(\omega)$:

$$\hat{e}'(\omega) = \cos\frac{\pi}{4} e^{j\phi_+} \hat{e}_{\Omega+} + \sin\frac{\pi}{4} e^{j\phi_-} \hat{e}_{\Omega-} \quad (2.27)$$

In the above equations, $\phi_{\pm}$ and $\phi'_{\pm}$ are the phase angles in the Jones notation for the two PSPs; also, 2α is independent of frequency. Since PSPs are a property of the fiber and are independent of α , the phase difference $\Delta\phi$ ($\Delta\phi = \phi_+ - \phi_-$) is equal to $\Delta\phi'$ ($\Delta\phi' = \phi'_+ - \phi'_-$). Thus, for calculating $\Delta\phi$, we can use $\alpha = \pi/4$ without loss of generality because $\Delta\phi$ is independent of α . We start with Eq. (2.25); for $\alpha = \pi/4$, one gets

$$\hat{S}'_o(\omega) = \hat{S}'_o(\omega_0) \cos(\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2) + \hat{B}' \sin(\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2) \quad (2.28)$$

In Eq. (2.28), $\hat{B}' = \hat{\Omega} \times \hat{S}'_o(\omega_0)$. For simplicity, we further assume that $\hat{\Omega} = [0, 0, 1]^T$ and $\hat{S}'_o(\omega_0) = [1, 0, 0]^T$. Then,

$$\hat{S}'_o(\omega) = \begin{bmatrix} \cos(\Delta\omega\Delta\tau - \Delta\tau_1\Delta\omega^2) \\ \sin(\Delta\omega\Delta\tau - \Delta\tau_1\Delta\omega^2) \\ 0 \end{bmatrix} \quad (2.29)$$

In the Jones notation, this SOP can be written as

$$\hat{e}'(\omega) = \begin{bmatrix} \cos(\Delta\tau\Delta\omega/2 - \Delta\tau_1\Delta\omega^2/2)e^{j0} \\ \sin(\Delta\tau\Delta\omega/2 - \Delta\tau_1\Delta\omega^2/2)e^{-j0} \end{bmatrix} \quad (2.30)$$

The Jones vector $\hat{e}_{\Omega+}$ corresponding to $\hat{\Omega}$ is

$$\text{For } \hat{e}_{\Omega+} = \begin{bmatrix} \cos(\pi/4)e^{(j\pi/4)} \\ \sin(\pi/4)e^{(-j\pi/4)} \end{bmatrix} \quad (2.31)$$

Next, to calculate ϕ_+ and ϕ_- in Eq. (2.26), we use Eqs. (2.27), (2.30), and (2.31), and use the following relations [78]:

$$\tan\phi_+ = \frac{\cos(\theta' + \theta_p)}{\cos(\theta' - \theta_p)} \tan(\delta' - \delta_p) \quad (2.32)$$

$$\tan\phi_- = -\frac{\sin(\theta' + \theta_p)}{\sin(\theta' - \theta_p)} \tan(\delta' - \delta_p) \quad (2.33)$$

In Eq. (2.32) and Eq. (2.33) $\theta' = \Delta\tau\Delta\omega/2 - \Delta\tau_1\Delta\omega^2/2$, $\theta_p = \pi/4$, $\delta' = 0$ and

$\delta_p = \pi/4$. From Eq. (2.32), $\phi_+ = \frac{\Delta\tau\Delta\omega}{2} - \frac{\Delta\tau_1\Delta\omega^2}{2} - \frac{\pi}{4}$ and from Eq.(2.33), one gets

$\phi_- = -\frac{\Delta\tau\Delta\omega}{2} + \frac{\Delta\tau_1\Delta\omega^2}{2} - \frac{\pi}{4}$. Hence $\Delta\phi = \Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2$. It means that the phase

difference between the two PSPs, including the effect of PCD, is $\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2$.

2.3.2 The effect of DGD, PCD and chromatic dispersion on pulse shape

Let us assume an input pulse signal $f(t)$, with intensity $P(t) = |f(t)|^2$. The signal has its Fourier transform, as shown in Eq. (2.3),

$$F(\omega - \omega_0) = \int_{-\infty}^{\infty} f(t) \exp[-j(\omega - \omega_0)t] dt. \quad (2.3)$$

The pulse, after propagation, can be decomposed into two components, $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$, one each along the two orthogonal PSPs, $\hat{e}_{\Omega+}$ and $\hat{e}_{\Omega-}$. If the chromatic dispersion and polarization-dependent chromatic dispersion are taken into account, then the two components differ in phase by $\Delta\phi = \Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2$. Thus the components $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$ are written as:

$$f_{\Omega+}(t) = \frac{\cos\alpha}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp\left[j \frac{\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2 - \beta_2 L \Delta\omega^2}{2}\right] e^{j(\omega - \omega_0)t} d\omega \quad (2.34)$$

$$f_{\Omega-}(t) = \frac{\sin\alpha}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp\left[-j \frac{\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2 + \beta_2 L \Delta\omega^2}{2}\right] e^{j(\omega - \omega_0)t} d\omega \quad (2.35)$$

where $\Delta\tau_1 = d\Delta\tau/d\omega$, L is the length of single mode fiber, and β_2 is the group velocity dispersion. When the input signal is a Gaussian pulse, $f(t) = M \exp\left(-\frac{t^2}{2T_0^2}\right)$,

with frequency spectrum $F(\omega - \omega_0) = \sqrt{2\pi} M T_0 \exp(-\Delta\omega^2 T_0^2/2)$, then

$$f_{\Omega+}(t) = \frac{M T_0 \cos\alpha}{\sqrt{T_0^2 - j(\beta_2 L + \Delta\tau_1)}} \exp\left\{-\frac{(\Delta\tau/2 + t)^2}{2[T_0^2 - j(\beta_2 L + \Delta\tau_1)]}\right\} \quad (2.36)$$

$$f_{\Omega-}(t) = \frac{T_0 \sin\alpha}{\sqrt{T_0^2 - j(\beta_2 L - \Delta\tau_1)}} \exp\left\{-\frac{(\Delta\tau/2 - t)^2}{2[T_0^2 - j(\beta_2 L - \Delta\tau_1)]}\right\} \quad (2.37)$$

Combining the components, one gets

$$\vec{f}(t) = f_{\Omega+}(t)\hat{e}_{\Omega+} + f_{\Omega-}(t)\hat{e}_{\Omega-} \quad (2.38)$$

The signal power at the detector is $P(t) = |\vec{f}(t)|^2 = \vec{f}(t) \cdot \vec{f}^*(t)$. Because $\hat{e}_{\Omega+}$ is orthogonal to $\hat{e}_{\Omega-}$,

$$\begin{aligned} P(t) &= |f_{\Omega+}(t)|^2 + |f_{\Omega-}(t)|^2 \\ &= \frac{M^2 T_0^2 \cos^2 \alpha}{\sqrt{T_0^4 + (\Delta\tau_1 + \beta_2 L)^2}} \exp \left[-\frac{(\Delta\tau/2 + t)^2 T_0^2}{(T_0^4 + (\Delta\tau_1 + \beta_2 L)^2)} \right] \\ &\quad + \frac{M^2 T_0^2 \sin^2 \alpha}{\sqrt{T_0^4 + (\beta_2 L - \Delta\tau_1)^2}} \exp \left[-\frac{(\Delta\tau/2 - t)^2 T_0^2}{(T_0^4 + (\beta_2 L - \Delta\tau_1)^2)} \right] \end{aligned} \quad (2.39)$$

Eq. (2.39) gives the output pulse shape for an input Gaussian pulse affected by chromatic dispersion, DGD and PCD. Following the procedure in section 2.2.2.1, the RMS pulse width is

$$\sigma^2 = \frac{T_0^2}{2} + \frac{(\Delta\tau_1)^2 + (\beta_2 L)^2}{2T_0^2} + \frac{\cos(2\alpha)\Delta\tau_1\beta_2 L}{T_0^2} + \left(\frac{\Delta\tau}{2T_0} \right)^2 \sin^2(2\alpha) \quad (2.40)$$

2.3.3 Power penalty due to pulse broadening

Similar to section 2.2.3 and based on Eq. (2.40), the power penalty can be expressed as

$$\varepsilon(dB) = B \left[\frac{(\Delta\tau_1)^2 + (\beta_2 L)^2}{2T_0^4} + \frac{\cos(2\alpha)\Delta\tau_1\beta_2 L}{T_0^4} \right] + A \left(\frac{\Delta\tau}{2T_0} \right)^2 \sin^2(2\alpha) \quad (2.41)$$

where the factor A depends on the pulse shape and modulation format as well as the receiver bandwidth [6][25][59]; B is also a parameter for power penalty due to

chromatic dispersion and PCD and is also dependent on the pulse shape, modulation format and receiver bandwidth.

From Eq. (2.41), it can be seen that when the angle between the input SOP and PMD vector $\vec{\Omega}$ changes, the power penalty that is dependent on 2α , also changes. When the input signal excites all the power in one PSP, that is $2\alpha = 0$, the power penalty can be expressed as

$$\varepsilon(dB) = B \frac{(\Delta\tau_1 + \beta_2 L)^2}{2T_0^4} \quad (2.42)$$

From Eq. (2.42), when the input state of polarization is aligned with the slow PSP ($2\alpha = 0$), the power penalty due to DGD can be gotten rid of; however, the effect of PCD and CD is maximum. But when $2\alpha = \pi$, which means that the input state of polarization is aligned with the fast PSP, then the power penalty is

$$\varepsilon(dB) = B \frac{(\Delta\tau_1 - \beta_2 L)^2}{2T_0^4} \quad (2.43)$$

Comparing Eq. (2.42) and (2.43), we can see that the overall power penalty can be reduced when the input state of polarization is aligned with the fast PSP rather than the slow PSP. The greater power penalty occurs when the input signal excites power equally in the two PSPs ($2\alpha = \pi/2$). In this case,

$$\varepsilon(dB) = B \frac{(\Delta\tau_1)^2 + (\beta_2 L)^2}{2T_0^4} + A \left(\frac{\Delta\tau}{2T_0} \right)^2 \quad (2.44)$$

Following the procedure used earlier, Eq. (2.41) yields the average power penalty as

$$\bar{\varepsilon}(dB) = B \frac{(\Delta\tau_1)^2 + (\beta_2 L)^2}{2T_0^4} + \frac{2A}{3} \left(\frac{\Delta\tau}{2T_0} \right)^2 \quad (2.45)$$

This means that when the angle between the output state of polarization and the PMD vector changes randomly with a uniform probability, the average power due to PCD and chromatic dispersion is the same as that for $2\alpha = \pi/2$; however, the average power penalty due to DGD is about $2/3$ of the worst case. To the best of our knowledge, this is the first time an expression for the power penalty and its average value due to DGD, PCD and chromatic dispersion has been derived.

2.3.4 Experimental results and discussion

In the following experimental setup, the PCD is not taking into account since a higher-order PMD emulator is not available. The experimental setup for measuring the average power penalty due to PMD is shown in Fig. 2.9. The signal from a laser whose output power can be tuned was fed to a MZ modulator with 10Gbit/s $2^{15} - 1$ PRBS NRZ data.

The output of the modulator was transmitted through a quarter-wave plate that was used to change the excitation ratio between the fast and the slow axis of the following first-order PMD emulator; the PMD emulator consists of two polarization beam splitters with Hi-Bi pigtails and a time delay line whose pigtail is also a Hi-Bi fiber. After the PMD emulator, the signal passes through an EDFA and a bandpass filter to overcome the ASE noise of the amplifier. Then after an attenuator, the signal passes through a 50 km long SMF, before passing through a 90:10 coupler. A power meter is used to ensure that the input power to the photodetector for bit error test is the same for different values of DGD as for the case of $2\alpha = 0$. Then the signal passes through a photodetector to monitor the bit error rate which is maintained at 10^{-9} .

Fig. 2.9. Experimental setup to observe power penalty due to PMD and chromatic dispersion

Fig. 2.10 shows the measured power penalty for three different values of DGD (66ps, 54ps and 30ps) and chromatic dispersion of 50 km single mode fiber for different values of the angle between output SOP and PMD vector. There was no penalty at first, for high input laser power; with suitable reduction of the input power, the power penalty appears. When there is no PMD, i.e., $\Delta\tau = 0$, the power penalty due to CD in the 50 km single mode fiber is about 1dB; moreover, a value of $A = 38$ fits the measured curves. So the calculated results plotted in Fig. 2.10 have been obtained from Eq. (2.22) or (2.41), using $\varepsilon(\text{dB}) = 1 + 38(\Delta\tau/2T_0)^2 [\sin(2\alpha)]^2$; the other parameters used for calculations are shown in Table 2.2. It is clear that the power penalties due to CD and PMD simply add together when higher-order PMD effects are absent.

Table 2.2: Parameters and equations of calculated results in Fig.2.10

Fig.	$\Delta\tau$ (DGD)	A	T_0	Eq.
2.10	30 ps; 54 ps; 66 ps	38	100 ps	2.22 or 2.41 $\varepsilon(dB) = 1 + 38(\Delta\tau/2T_0)^2 [\sin(2\alpha)]^2$

Fig. 2.10. Power penalty vs. the angle between the output state of polarization and the PMD vector, with optimum decision threshold for PMD and CD

2.4 Expression for higher-order PMD vector including depolarization-dependent chromatic dispersion

In section 2.3, rotation of the PSPs with frequency is neglected. In this section, we first derive an expression for the second-order PMD vector including the effect of the rotation of the PSPs with frequency. For this purpose, we introduce a novel vector

called depolarization vector. Next, we develop an expression for the PMD vector including the effect of frequency variation of the depolarization vector. Frequency variation of the depolarization vector is given a new name: depolarization-dependent chromatic dispersion (DCD). These expressions are then used to determine the effect on pulse shape due to these higher-order effects, and the corresponding power penalty. These higher-order effects assume importance when bit rates go to 40 Gbit/s and higher.

2.4.1 Expression for PMD vector including depolarization vector

Following the earlier work on PMD [6], we characterize the first- and second-order PMD vectors, $\vec{\Omega}$ and $\vec{\Omega}_\omega$, in the Stokes space as:

$$\vec{\Omega}_\omega = \frac{d\vec{\Omega}}{d\omega} = \frac{d\Delta\tau}{d\omega} \hat{\Omega} + \Delta\tau \frac{d\hat{\Omega}}{d\omega} = \frac{d\Delta\tau}{d\omega} \hat{\Omega} + \Delta\tau \hat{\Omega}_\omega \quad (2.46)$$

It is well known that $\vec{\Omega}_\omega$ and $\vec{\Omega}$ are orthogonal to each other [6]. Therefore, we can characterize $\hat{\Omega}_\omega$ using a new vector $\vec{q}$, with magnitude Δq , and the vector $\hat{\Omega}$:

$$\frac{d\hat{\Omega}}{d\omega} = \vec{q} \times \hat{\Omega} \quad (2.47)$$

$$\frac{d\vec{\Omega}}{d\omega} = \Delta\tau_1 \hat{\Omega} + \Delta\tau \vec{q} \times \hat{\Omega} \quad (2.48)$$

where $\Delta\tau_1 = d\Delta\tau/d\omega$ is the PCD. We call the new vector $\vec{q}$ the depolarization vector.

The magnitude of $\vec{q}$, Δq , gives the rate of rotation of PSPs, i.e., $\Delta q = \left| \hat{\Omega}_\omega \right| = \left| \frac{d\hat{\Omega}}{d\omega} \right|$.

Corresponding to the two orthogonal PSPs, $\vec{\Omega}_+$ and $\vec{\Omega}_-$, there also exist two orthogonal depolarization vectors, $\vec{q}_+$ and $\vec{q}_-$. When the third-order PMD is

neglected, $\vec{q}$ and $\Delta\tau_1$ are constant with respect to frequency. Also, one can write $DGD = \Delta\tau + \Delta\tau_1\Delta\omega$, where $\Delta\tau$ is the magnitude of the first-order PMD, $\Delta\omega = \omega - \omega_0$, and ω_0 is the optical carrier frequency. Thus, from Eq. (2.48) we can get:

$$\frac{d\vec{\Omega}}{d\omega} = \left(\frac{\Delta\tau_1}{\Delta\tau + \Delta\tau_1\Delta\omega} + \Delta q \hat{q} \times \right) \vec{\Omega} \quad (2.49)$$

Next, we interpret $\vec{q} \times$ as a skew-symmetric matrix R' , integrate Eq.(2.49), and simplify the exponential term $\exp[R'(\omega - \omega_0)]$ using the eigenvalues and the eigenvector matrix of R . After some complex manipulations [Appendix II], we get:

$$\vec{\Omega}(\omega) = (\Delta\tau_0 + \Delta\tau_1\Delta\omega) \left\{ \begin{aligned} &\hat{q}(\hat{q} \cdot \hat{\Omega}_0 + [\hat{\Omega}_0 - \hat{q}(\hat{q} \cdot \hat{\Omega}_0)] \cos(\Delta q \Delta\omega)) \\ &+ \hat{B}_0 \sin 2\alpha_0 \sin(\Delta q \Delta\omega) \end{aligned} \right\} \quad (2.50)$$

Or

$$\vec{\Omega}(\omega) = (\Delta\tau_0 + \Delta\tau_1\Delta\omega) \left\{ \begin{aligned} &\hat{q} \cos 2\alpha_0 + [\hat{\Omega}_0 - \hat{q} \cos 2\alpha_0] \cos(\Delta q \Delta\omega) \\ &+ \hat{B}_0 \sin 2\alpha_0 \sin(\Delta q \Delta\omega) \end{aligned} \right\} \quad (2.51)$$

In Eq. (2.51), $\hat{\Omega}_0$ is the unit vector for the first-order PMD vector at carrier frequency ω_0 that is $\hat{\Omega}_0 = \hat{\Omega}(\omega_0)$, $\hat{B}_0$ is the unit vector corresponding to $\hat{q} \times \hat{\Omega}_0$, and $2\alpha_0$ is the angle between $\hat{\Omega}_0$ and $\vec{q}$. This is an expression for the frequency-dependent PMD vector taking into account the second-order PMD, but ignoring the other higher-order PMDs. Analogous to Eq. (2.2) for first-order PMD, in Eq. (2.51), the three unit vectors $\hat{q}$, $\frac{1}{\sin 2\alpha_0}(\hat{\Omega}_0 - \hat{q} \cos 2\alpha_0)$, and $\hat{B}_0$ constitute a 3-D orthogonal coordinate system; these vectors are shown in Fig.2.11. Eq. (2.51) means that the PMD vector precesses around $\hat{q}$ with a precession angle $2\alpha_0$, which is independent of frequency.

Also, the magnitude of PMD vector change as $(\Delta q + \Delta q_1 \Delta \omega)$. The projection of the PMD vector in the plane normal to $\hat{q}$ rotates by an angle $\Delta q(\omega - \omega_0)$.

Fig. 2.11. Relation between PMD vector $\vec{\Omega}(\omega)$ and $\hat{q}$

If we neglect the PCD, that is the magnitude of PMD vector is considered constant, then Eq. (2.51) simplifies to:

$$\vec{\Omega}(\omega) = \Delta \tau \left\{ \hat{q} \cos 2\alpha_0 + \left[\hat{\Omega}_0 - \hat{q} \cos 2\alpha_0 \right] \cdot \cos(\Delta q \Delta \omega) + \hat{B}_0 \sin 2\alpha_0 \cdot \sin(\Delta q \Delta \omega) \right\} \quad (2.52)$$

In Eq. (2.52), the three unit vectors $\hat{q}$, $\frac{1}{\sin 2\alpha_0} (\hat{\Omega}_0 - \hat{q} \cos 2\alpha_0)$, and $\hat{B}_0$ constitute a 3-D orthogonal coordinate system. Eq. (2.52) means that the PMD vector precesses around the $\hat{q}$ with a precession angle $2\alpha_0$, which is independent of frequency. We can see that the projection of the PMD vector in the plane normal to $\hat{q}$ rotates by an angle $\Delta q(\omega - \omega_0)$. The situation is represented in Fig.2.12.

In the following discussion, we will neglect the change of DGD value, i.e. the magnitude of the PMD vector is considered constant.

Fig. 2.12. Relation between PMD vector $\vec{\Omega}(\omega)$ and $\hat{q}$, when DGD ($\Delta\tau$) is constant

Further, analogous to section 2.3, the frequency derivative of the depolarization vector

$\vec{q} = \Delta q \hat{q}$ is:

$$\frac{d\vec{q}}{d\omega} = \frac{d\Delta q}{d\omega} \hat{q} + \Delta q \frac{d\hat{q}}{d\omega} \quad (2.53)$$

where the first component, which is parallel to $\hat{q}$, is given the name depolarization-dependent chromatic dispersion (DCD) and is defined as $\Delta q_1 = d\Delta q/d\omega$; the second term, perpendicular to $\hat{q}$, is the rate of rotation of the depolarization vector with frequency. If we take the depolarization-dependent chromatic dispersion (DCD) into account and neglect the rotation of the depolarization vector, then Eq. (2.53) can be expressed as

$$\vec{\Omega}(\omega) = (\Delta\tau_0 + \Delta\tau_1\Delta\omega) \left\{ \begin{aligned} &\hat{q} \cos 2\alpha_0 + [\hat{\Omega}_0 - \hat{q} \cos 2\alpha_0] \cos(\Delta q\Delta\omega - \Delta q_1\Delta\omega^2) \\ &+ \hat{B}_0 \sin 2\alpha_0 \sin(\Delta q\Delta\omega - \Delta q_1\Delta\omega^2) \end{aligned} \right\} \quad (2.54)$$

In this case also, PMD vector precesses around the depolarization vector with precession angle $2\alpha_0$. However, the projection of the PMD vector in the plane normal to the depolarization vector rotates by an angle $\Delta q\Delta\omega - \Delta q_1\Delta\omega^2$. It is useful to mention that all the terms related to PMD value such as DGD, depolarization, PCD, DCD are randomly varying quantities [79].

2.4.2 Determination of phase difference between $\hat{q}_+$ and $\hat{q}_-$

As mentioned in section 2.4.1, there exist two orthogonal depolarization vectors, $\hat{q}_+$ and $\hat{q}_-$. The phase difference $\Delta\varphi(\omega)$ between the two orthogonal depolarization vectors plays an important role in determining the output pulse shape for a given input pulse in the presence of higher-order PMD effects. Let the PMD vector be represented as $\hat{\Omega}(\omega)$ in Stokes notation and as $\hat{e}_{\Omega+}(\omega)$ in the Jones notation. Proceeding in a manner similar to section 2.3.1, the phase difference between the two orthogonal depolarization vectors is equal to $\Delta\varphi = \Delta q\Delta\omega - \Delta q_1\Delta\omega^2$ (see Appendix III)

2.4.3 The effect of Higher-order PMD on pulse shape and power penalty

2.4.3.1 Higher-order PMD effect on pulse shape without PCD, DCD, and CD

In this section, we first neglect the PCD, DCD and CD, for estimating the effect of higher-order PMD on pulse shape. In such a case, the depolarization vector is constant. The pulse, after propagation, can be decomposed into two components, $f_{q+}(t)$ and $f_{q-}(t)$, one each along the two orthogonal depolarization vectors, $\hat{e}_{q+}$ and $\hat{e}_{q-}$.

$$\vec{f}(t) = f_{q+}(t)\hat{e}_{q+} + f_{q-}(t)\hat{e}_{q-} \quad (2.55)$$

From section 2.4.2 (Appendix III), we know that the PSPs can be expressed in term of two orthogonal depolarization vectors. Using the result of section 2.4.2 that $\hat{q}_+$ and $\hat{q}_-$ differ in phase by $\Delta\varphi = \Delta q\Delta\omega$ (with the assumption of no DCD):

$$\hat{e}_{\Omega+} = \cos\alpha_0 \exp(j\Delta q\Delta\omega/2)\hat{e}_{q+} + \sin\alpha_0 \exp(-j\Delta q\Delta\omega/2)\hat{e}_{q-} \quad (2.56)$$

$$\hat{e}_{\Omega-} = \sin\alpha_0 \exp[j(\pi/2 - q\Delta\omega/2)]\hat{e}_{q+} + \cos\alpha_0 \exp[-j(\pi/2 - q\Delta\omega/2)]\hat{e}_{q-} \quad (2.57)$$

In Eqs. (2.56) and (2.57) α_0 is the angle between the PMD vector and the depolarization vector. To be able to express an output state of polarization (SOP) in the term of $\hat{q}_+$ and $\hat{q}_-$, we first expand the output SOP in the terms of the two orthogonal PSPs. Thus in Jones notation,

$$\hat{e} = \cos\alpha \exp(j\Delta\tau\Delta\omega/2)\hat{e}_{\Omega+} + \sin\alpha \exp(-j\Delta\tau\Delta\omega/2)\hat{e}_{\Omega-} \quad (2.58)$$

when we take the second-order PMD into account but neglect PCD, DCD, and CD. In Eq. (2.58), α is the angel between SOP and $\hat{\Omega}_+$, and the phase difference between $\hat{\Omega}_+$ and $\hat{\Omega}_-$ is $\Delta\tau\Delta\omega$.

For a Gaussian pulse, the signal power in the detector $P(t) = |\vec{f}(t)|^2 = \vec{f}(t) \cdot \vec{f}^*(t)$ can be expressed as: (see Appendix IV)

$$\begin{aligned}
 P(t) &= |f_{q+}(t)|^2 + |f_{q-}(t)|^2 \\
 &= M^2 \cos^2 \alpha \cdot \cos^2 \alpha_0 \exp \left[-\frac{(\Delta q/2 + \Delta \tau/2 + t)^2}{T_0^2} \right] + M^2 \sin^2 \alpha \cdot \sin^2 \alpha_0 \exp \left[-\frac{(\Delta q/2 + \Delta \tau/2 - t)^2}{T_0^2} \right] \\
 &\quad + M^2 \cos^2 \alpha \cdot \sin^2 \alpha_0 \exp \left[-\frac{(t + \Delta \tau/2 - \Delta q/2)^2}{T_0^2} \right] + M^2 \sin^2 \alpha \cdot \cos^2 \alpha_0 \exp \left[-\frac{(t + \Delta q/2 - t/2)^2}{T_0^2} \right]
 \end{aligned} \quad (2.59)$$

Eq. (2.59) shows that the output signal may be not Gaussian. The pulse shape depends on $\Delta \tau$ and Δq which are the magnitude of PMD and depolarization vector, respectively. Using Eq. (2.59), the RMS pulse width is

$$\begin{aligned}
 \sigma^2 &= \langle t^2 \rangle - \langle t \rangle^2 = \frac{\int_{-\infty}^{+\infty} t^2 P(t) dt}{P_0} - \left[\frac{\int_{-\infty}^{+\infty} t P(t) dt}{P_0} \right]^2 \\
 &= \frac{T_0^2}{2} + \left(\frac{\Delta \tau}{2} \right)^2 \sin^2 2\alpha + \left(\frac{\Delta q}{2} \right)^2 \sin^2 2\alpha_0
 \end{aligned} \quad (2.60)$$

Eq. (2.60) shows that the pulse broadening due to rotation of PSP is dependent on the angle between the PMD vector and the depolarization vector, and on the rate of rotation of PSPs with frequency.

2.4.3.2 Higher-order PMD effect on pulse shape including PCD, DCD and CD

The polarization-dependent chromatic dispersion will not affect the phase difference between the PSP and the two orthogonal depolarization vectors. If both the change of DGD and the chromatic dispersion are taken into account, then for a Gaussian pulse, the signal power in the detector $P(t)$ is (See Appendix V):

$$\begin{aligned}
 P(t) &= |f_{q+}(t)|^2 + |f_{q-}(t)|^2 \\
 &= \frac{T_0^2 \cos^2 \alpha \cdot \cos^2 \alpha_0}{\sqrt{T_0^4 + (\beta_2 L + \Delta \tau_1 + \Delta q_1)^2}} \exp \left\{ -\frac{T_0^2 (t + \Delta q/2 + \Delta \tau/2)^2}{[T_0^4 + (\Delta q_1 + \Delta \tau_1 + \beta_2 L)^2]} \right\} \\
 &+ \frac{T_0^2 \sin^2 \alpha \cdot \sin^2 \alpha_0}{\sqrt{T_0^4 + (\beta_2 L - \Delta \tau_1 - \Delta q_1)^2}} \exp \left\{ -\frac{T_0^2 (t - \Delta q/2 - \Delta \tau/2)^2}{[T_0^4 + (\beta_2 L - \Delta \tau_1 - \Delta q_1)^2]} \right\} \\
 &+ \frac{T_0^2 \cos^2 \alpha \cdot \sin^2 \alpha_0}{\sqrt{T_0^4 + (\beta_2 L + \Delta \tau_1 - \Delta q_1)^2}} \exp \left\{ -\frac{T_0^2 (t + \Delta \tau/2 - \Delta q/2)^2}{[T_0^4 + (\beta_2 L + \Delta \tau_1 - \Delta q_1)^2]} \right\} \\
 &+ \frac{T_0^2 \sin^2 \alpha \cdot \cos^2 \alpha_0}{\sqrt{T_0^4 + (\beta_2 L - \Delta \tau_1 + \Delta q_1)^2}} \exp \left\{ -\frac{T_0^2 (t - \Delta \tau/2 + \Delta q/2)^2}{[T_0^4 + (\beta_2 L - \Delta \tau_1 + \Delta q_1)^2]} \right\}
 \end{aligned} \quad (2.61)$$

Eq. (2.61) shows the effect of chromatic dispersion and higher-order PMD including DCD, PCD and depolarization vector on the output pulse. Using (2.61), the RMS pulse width is:

$$\begin{aligned}
 \sigma^2 &= \langle t^2 \rangle - \langle t \rangle^2 = \frac{\int_{-\infty}^{+\infty} t^2 P(t) dt}{P_0} - \left[\frac{\int_{-\infty}^{+\infty} t P(t) dt}{P_0} \right]^2 \\
 &= \frac{T_0^2}{2} + \frac{(\Delta \tau_1)^2 + (\beta_2 L)^2 + (\Delta q_1)^2 + 2 \cos 2\alpha \cdot \cos 2\alpha_0 \Delta \tau_1 \Delta q_1}{2T_0^2} \\
 &+ \frac{\cos(\alpha + \alpha_0) \cdot \cos(\alpha - \alpha_0) (\Delta \tau_1 + \Delta q_1) \beta_2 L}{T_0^2} \\
 &+ \frac{\sin(\alpha + \alpha_0) \cdot \sin(\alpha - \alpha_0) (\Delta \tau_1 - \Delta q_1) \beta_2 L}{T_0^2} \\
 &+ \left(\frac{\Delta \tau}{2} \right)^2 \sin^2 2\alpha + \left(\frac{\Delta q}{2} \right)^2 \sin^2 2\alpha_0
 \end{aligned} \quad (2.62)$$

The input RMS pulse width is $\sigma_0 = \frac{T_0}{\sqrt{2}}$. Significantly, Eq (2.62) means that PCD,

DCD and CD do not act on the pulse independently. Also, the RMS pulse width is dependent on the angles 2α and $2\alpha_0$ (see Fig. 2.1 and Fig 2.11).

2.4.3.3 Power penalty due to higher-order PMD and chromatic dispersion

Following the procedure in section 2.3.3 and according to Eq. (2.62), the power penalty is

$$\begin{aligned} \varepsilon(dB) = B & \left[\frac{(\Delta\tau_1)^2 + (\beta_2 L)^2 + (\Delta q_1)^2}{2T_0^4} + \frac{\cos(\alpha + \alpha_0) \cdot \cos(\alpha_0 - \alpha)(\Delta\tau_1 + \Delta q_1)\beta_2 L}{T_0^4} \right. \\ & \left. + \frac{\sin(\alpha + \alpha_0) \cdot \sin(\alpha_0 - \alpha)(\Delta\tau_1 - \Delta q_1)\beta_2 L}{T_0^4} + \frac{\cos 2\alpha \cdot \cos 2\alpha_0 \Delta\tau_1 \Delta q_1}{T_0^4} \right] \\ & + A \left[\left(\frac{\Delta\tau}{2T_0} \right)^2 \sin^2(2\alpha) + \left(\frac{\Delta q}{2T_0} \right)^2 \sin^2(2\alpha_0) \right] \end{aligned} \quad (2.63)$$

When the angle between the input SOP and $\vec{\Omega}(\omega_0)$ and the angle between PMD vector and depolarization vector $\vec{q}$ change, the power penalty, which is dependent on 2α and $2\alpha_0$, also changes. If $2\alpha = \pi/2$ and $2\alpha_0 = \pi/2$, the power penalty is:

$$\varepsilon(dB) = B \frac{(\Delta q_1)^2 + (\Delta\tau_1)^2 + (\beta_2 L)^2}{2T_0^4} + A(\Delta\tau/2T_0)^2 + A(\Delta q/2T_0)^2 \quad (2.64)$$

If the PSP coincides with the depolarization vector, that is $2\alpha_0 = 0$, and when the input state of polarization is aligned with the slow PSP, then the power penalty can be expressed as

$$\varepsilon(dB) = B \frac{(\beta_2 L + \Delta\tau_1 + \Delta q_1)^2}{2T_0^4} \quad (2.65)$$

Eq. (2.65) shows that in this case the power penalty for DGD and depolarization can be gotten rid of; however, the effect of PCD, CD and DCD is maximum. But when $2\alpha = \pi$ and $2\alpha_0 = \pi$, which means that the input state of polarization is aligned with the fast PSP and the slow PSP (PMD vector) is aligned with the fast depolarization vector, then the power penalty is

$$\varepsilon(dB) = B \frac{(\beta_2 L - \Delta\tau_1 - \Delta q_1)^2}{2T_0^4} \quad (2.66)$$

Comparing Eq. (2.65) and (2.66), we can see that the power penalty can be reduced when the input state of polarization is aligned with the fast PSP and the slow PSP is aligned with the fast depolarization vector.

2.5 Conclusion

We first present an expression relating the frequency dependent output state of polarization, $\vec{S}_o(\omega)$, and the PMD vector $\vec{\Omega}$, including the effect of PCD; this relation is consistent with the established behaviour of $\vec{S}_o(\omega)$ with respect to $\vec{\Omega}$, and clearly describes the details of the behaviour of SOP with respect to the PMD vector. Based on this expression, we get the phase difference value between the two PSPs. Following this, we derive a general equation, which incorporates the precession angle, and the effects of DGD, PCD and chromatic dispersion, on a pulse of any shape. This new equation is used to calculate the shape distortion for the special case of Gaussian pulses; also derived are the expressions for pulse broadening, the power penalty as well as the average power penalty. The power penalty results are verified experimentally.

Next, we present an expression for the higher-order PMD vector including second-order effects and depolarization-dependent chromatic dispersion (DCD). Based on this expression, we get the phase difference value between the two orthogonal depolarization vectors. For this case also, expressions for the pulse broadening and power penalty due to higher-order PMD (including second-order PMD and DCD) and chromatic dispersion are derived.

It is believed that such a derivation of power penalty in terms of the precession angle between the SOP and the PMD vector, including higher-order PMD effects and chromatic dispersion, and experimental verification of the power penalty expressions, have not been reported earlier. When the power penalty expression is available in terms of the precession angle, the average power penalty can be derived conveniently. These expressions for shape distortion and power penalty should be helpful in estimating the degradation due to PMD, including the higher order effects, in high bit rate optical communication systems employing different types of pulse shapes.

Chapter 3

PMD Compensation

3.1 Introduction

As is well known, at high bit rates such as 10Gbit/s or 40Gbit/s, optical fiber transmission systems are highly susceptible to performance degradation due to optical fiber related effects, such as chromatic dispersion (CD), and polarization mode dispersion (PMD). By managing fiber dispersion, the high bit rate capacity can be realized to some extent. However, PMD has been identified as one of the major remaining sources of performance limitation in long-distance optically amplified systems. This situation has occurred since many fiber segments that were deployed throughout the 1980s have a PMD parameter value that is over 100 times greater than the better quality fiber deployed today. So it is required to compensate, if possible, PMD for the old embedded fiber.

In general, PMD is caused by different group velocity of the signal in the two principal states of polarization (PSPs). A convenient way to describe PMD in a given system is the PMD vector, $\vec{\Omega}(\omega)$, whose direction is parallel to the slow PSP and whose magnitude is the differential group delay (DGD) in the system at frequency ω . As already mentioned, to maintain adequate performance in optical fiber communication systems, compensation of PMD is necessary. Different techniques for compensation or mitigation of PMD have been proposed. These are divided into two main categories, passive mitigation and active mitigation.

Passive methods for PMD compensation include the use of novel modulation formats [76][80-81] and forward error correction (FEC) [82][83]. However, in some cases, passive techniques are inadequate and active PMD mitigation techniques must be employed. The active techniques include optical PMD compensation [26][84-87], and electrical equalization [28][29]. As it has been proved that electrical equalization is less efficient than optical compensation, and hard to realize at 40 Gbit/s and beyond, only optical compensators are discussed in this chapter. Since the birefringence of a fiber varies randomly along the fiber length, PMD is a statistical or random quantity. For a Gaussian pulse, the pulse broadening due to first-order PMD as described in section 2.2.2.1, can be expressed by:

$$\sigma^2 = \langle t^2 \rangle - \langle t \rangle^2 = \frac{T_0^2}{2} + \left(\frac{\Delta\tau}{2} \right)^2 \sin^2 2\alpha \quad (3.1)$$

where 2α is the angle between the output state of polarization (SOP) and the PMD vector (in Stokes notation) and $\Delta\tau$ is DGD. Since the input pulse width is $\sigma_0 = T_0/\sqrt{2}$, the RMS width of the broadened pulse is rewritten as

$$\sigma^2 = \sigma_0^2 + (\Delta\tau/2)^2 \sin^2 2\alpha \quad (3.2)$$

Eq. (3.2) clearly shows that there are two ways to compensate the first-order PMD. One is to make 2α equal to zero, that is to align the signal SOP to a PSP using

Fig. 3.1. PMD compensation by transmitting in a principal state of polarization

polarization controller (PC) at the fiber input; this is shown in Fig.3.1 [59][88-89]. The second method is to nullify the PMD vector of the system at the fiber output using a PMD compensating element. This scheme requires an adjustable DGD element to match the DGD of the fiber and a polarization controller to adjust the direction of the compensator's PMD vector; this is shown in Fig.3.2.

Fig. 3.2. PMD compensation by using an adjustable DGD element

Because of the temporal drift of the PMD characteristics, it will be beneficial to have an adjustable PMD compensator in systems which are in operation. A number of PMD compensators using a tunable DGD element have been proposed in the literature [35][38][71]. Tunable PMD compensators have also been proposed using high birefringence (Hi-Bi) nonlinearly chirped FBG [62-64]. However, fabrication of nonlinearly chirped FBG requires nonlinearly chirped phase masks, which are costly and difficult to fabricate. Further, the nonlinearly chirped FBG induces chromatic dispersion in the PMD-compensated signal [90]. Tunable PMD compensators using a high-birefringence linearly chirped FBG also have been reported [65-67][91]. Unfortunately, the linearly chirped FBG also induces a chirp in the PMD compensated signal. This chirp may limit the distance of transmission after the compensator due to

the fiber chromatic dispersion. Recently, a distributed PMD compensation technique has been introduced [92]; in such a case, it is important to realize a PMD compensator that is free from chromatic dispersion.

A tunable PMD compensator, using a Hi-Bi nonlinearly chirped FBG in a double-pass configuration, has been proposed earlier [90]; the double pass design significantly reduces the chromatic dispersion induced by the grating and also doubles the DGD tuning range. In this chapter, we propose a new PMD compensation scheme using a Hi-Bi linearly chirped FBG in a double pass configuration [93]. As mentioned earlier, fabrication of a linearly chirped FBG is easier compared to that of a nonlinearly chirped FBG. In our implementation, the FBG is mounted on a cantilever beam in such a way that the bandwidth of the reflected signal can be tuned with little as a very small shift of the centre wavelength. This eliminates the possibility that the signal wavelength may lie at the edge of the effective reflection band of the grating [94]; thus, our scheme is likely to be more suitable for PMD compensation in a multi-wavelength system. Measured results are presented to show that such a configuration achieves significantly reduced chromatic dispersion, together with PMD compensation.

Next, we propose another new tunable polarization mode dispersion (PMD) compensator, which is also substantially free from chromatic dispersion. The design uses a pair of high-birefringence linearly chirped fiber gratings [95]. Each FBG can adjust the differential group delay (DGD) nonlinearly by using a cantilever beam. The compensator can be tuned to operate over a range of wavelengths and can yield DGD values starting from 0 ps. Another feature of this design is that, compared to the double pass configurations, it is free from cross-talk. Measured eye-diagrams for a 10

Gbit/s system show that this compensator can greatly reduce the chromatic dispersion induced by chirped grating.

3.2 Hi-Bi chirped grating

A fiber grating consists of a periodic modulation of index of refraction in the core of the fiber [96][97]. In the most general case, the index perturbation $\delta n(z)$ takes the form of a phase and amplitude-modulated waveform [96]:

$$\delta n(z) = \delta n_0(z) \left\{ 1 + m \cos \left[\frac{2\pi z}{\Lambda(z)} + \phi(z) \right] \right\} \quad (3.3)$$

where $\delta n_0(z)$ is the maximum index perturbation spatially averaged over a grating period, m is the fringe visibility of the index change, $\Lambda(z)$ is the local grating period including grating chirp, $\phi(z)$ is the phase change in the grating and z is the distance along the fiber axis. For a uniform grating, the strongest interaction or mode coupling occurs at the Bragg wavelength, λ_B , which is the free space centre wavelength of the input light that can be back reflected from the grating. The Bragg wavelength is given by

$$\lambda_B = 2n_{eff} \Lambda \quad (3.4)$$

where n_{eff} is the effective refractive index of the fiber core and it is a spatial average of $n_{eff}(z)$.

A chirped grating has a variable pitch $\Lambda(z)$ and it reflects different wavelengths at different points along its length. Unlike a uniform grating, which satisfies the Bragg condition for a specific wavelength, a chirped grating satisfies the Bragg condition for

a range of wavelength. The Bragg condition at an axial position z of a chirped FBG is given by:

$$\lambda_B(z) = 2n_{eff}(z)\Lambda(z) \quad (3.5)$$

where $\Lambda(z)$ is the local grating period at position z . Since the Bragg wavelength varies along the grating, different frequency components of an incident optical pulse are reflected at different points and have different time delays depending on where the Bragg condition is satisfied locally.

Further, a Hi-Bi chirped grating, chirped or uniform, is equivalent to two virtually identical gratings for the two orthogonal polarization axes: fast axis and slow axis. The wavelength difference for the two gratings is determined by the difference of effective refractive index between the fast and slow axes or the beat length of the Hi-Bi or polarization maintaining (PM) fiber. The reflectivity measurement scheme for a Hi-Bi chirped grating is shown in Fig.3.3. A tunable laser source (ANDO AQ4321D) feeds a circulator. Before the light is reflected by the Hi-Bi chirped grating, a polarization controller (PC) is used to adjust the input light SOP to align to the fast or

Fig. 3.3. Reflectivity measurement for Hi-Bi chirped grating

slow axis of the Hi-Bi grating. The reflected signal is coupled to an optical spectrum analyzer (OSA) (ANDO AQ6317B).

The measured reflectivity of a Hi-Bi chirped grating fabricated in NTU is shown in Fig.3.4. It clearly shows that the Hi-Bi chirped grating is equivalent to two virtually identical gratings for the two orthogonal polarization axes. The centre wavelength for the two identical gratings is shifted by about 0.4 nm.

Fig. 3.4. Measured reflectivity for fast and slow axis of Hi-Bi chirped grating

Fig.3.5 shows the schematic reflection spectrum and group delay of the Hi-Bi chirped grating. As shown in Fig.3.5 (a), the DGD between the two orthogonal states of polarization is $\Delta\tau$. When we tune the Hi-Bi linearly chirped FBG, the bandwidth of the grating broadens or narrows, with a fixed centre wavelength, as shown in

Fig.3.5 (a) Schematic reflectivity (left axis) and group delay (right axis) for Hi-Bi chirped grating (without tuning)

Fig.3.5 (b) Schematic reflectivity (left axis) and group delay (right axis) for Hi-Bi chirped grating (broadening)

Fig.3.5 (c) Schematic reflectivity (left axis) and group delay (right axis) for Hi-Bi chirped grating (narrow)

Fig. 3.5. Schematic reflectivity and group delay for Hi-Bi chirped grating

Fig. 3.5(b) and Fig. 3.5(c) respectively; the value of DGD between the two orthogonal states of polarization will be reduced or increased accordingly. Fig. 3.6 shows the

Fig. 3.6. The schematic diagram of the cantilever beam

schematic diagram of the arrangement [94] for tuning the grating. We glue the Hi-Bi chirped FBG in a slanted orientation onto the vertical side of the beam.

3.3 PMD compensation technique using a double pass configuration

3.3.1 Double pass configuration

As mentioned above, in a Hi-Bi chirped FBG, the high birefringence of the fiber gives two almost identical gratings for the two orthogonal polarization axes. The wavelength difference for the two gratings is determined by the difference of effective refractive index between the fast and slow axes. For a given signal wavelength which lies within the common bandwidth of the two gratings, after the signal is reflected by the Hi-Bi chirped grating, there occurs DGD between the two parts of the signal travelling along the fast and slow axes because of the position difference of reflection for the two polarization axes.

Fig. 3.7 shows our double-pass PMD compensator. The input 10 Gbit/s NRZ signal, using a pseudorandom binary sequence (PRBS) of length $2^{15} - 1$, is generated by a pulse pattern generator (Anritsu MP 1763C). This signal modulates an optical signal in a Mach-Zehnder modulator. The modulated optical signal passes through a first-order PMD emulator, which consists of an adjustable DGD element and a polarization controller, and generates a Maxwellian distribution of DGD with different mean values. After the emulator, the signal passes through a polarization controller (Thorlabs PLC 003) which is operated manually. Using the polarization controller, we can align the slow PSP (fast PSP) with the slow axis (fast axis) of the Hi-Bi FBG. Next, after passing through

Fig. 3.7. Double pass PMD compensator

a polarization maintaining circulator, the signal is reflected from the Hi-Bi linearly chirped grating from the short wavelength port so that the slow axis polarization state has a smaller time delay than the fast axis polarization state [97]. Next, the reflected optical signal passes through the polarization maintaining circulator and a polarization rotator. The latter rotates the polarization state by 90° and thus switches the propagation axes for the second reflection, which occurs from the long wavelength port. This time, the fast axis (originally slow axis) polarization state will have a smaller time delay. Consequently, the double pass scheme can double the time delay between the two PSPs. More significantly, the sign of chromatic dispersion depends on the port of incidence of the Hi-Bi linearly chirped FBG. Thus, any chromatic dispersion induced during the first reflection is reduced greatly by the second reflection. Here, we mount the FBG on a cantilever beam such that the centre wavelength of reflection spectrum remains fixed [94]. When the bandwidth of the Hi-Bi chirped grating changes, keeping the centre wavelength fixed, the DGD between

the two parts of the signal along the fast and slow axes also changes. When the grating period is changed nonlinearly, the DGD also changes nonlinearly.

3.3.2 Experimental results and discussion

The Hi-Bi linearly chirped FBG is written by exposing a hydrogen-loaded high-birefringence fiber to 244 nm laser beam through a linearly chirped phase mask. The length of the grating is 9.2 cm. The 3-dB bandwidth for each axis is 1.704 nm, and the centre wavelengths are 1554.356 nm and 1554.756 nm for the fast and the slow axis, respectively. The reflectivity of the original grating is about 99.9% within the effective bandwidth. Fig. 3.8(a) and 3.8(b) show the measured reflection spectrum for beam displacements of 6 mm at the long wavelength port, and -1 mm at the short wavelength port, respectively. The bandwidth of the grating for the two orthogonal

Fig. 3.8. (a) Reflection spectrum for beam displacements of 0 and 6 mm.

Fig. 3.8. (b) Reflection spectrum for beam displacements of 0 and -1 mm.

states of polarization can be broadened greatly to 5.72 nm (displacement = 9 mm) and high reflectivity can be maintained. It is worthwhile to note that the insertion loss changes by about 1.1 dB for maximum tuning. The power penalty due to the increased insertion loss can be overcome by increasing the laser power or the gain of the amplifier in the fiber link.

Measured group delay results are shown in Fig. 3.9. There exists a small ripple in the group delay values. As observed in [67], a small ripple in the group delay does not sensitively affect the eye-opening penalty. Since the slopes for the different ports of incidence have opposite signs, the double pass configuration has much lower chromatic dispersion compared to that for the single pass configuration. Figure 3.10 shows the time delay between the fast and slow axes after only one reflection (single pass), and after two reflections (double pass) for an optical wavelength of 1554.55 nm. The time delay between the two polarization states ranges from 65-450 ps for single-

pass; for double pass, the time delay tuning range is 130-904 ps. Fig. 3.11 shows the measured eye-diagrams for a bit-rate of 10 Gbit/s PRBS.

Fig. 3.9. Group delay for fast and slow axes (without tuning)

Fig. 3.10. DGD between the fast and slow axis

The eye-diagrams are measured (a) just after the transmitter, (b) just after the emulator with a DGD of 100ps, (c) after double-pass, with the grating tuned to compensate a DGD of 100ps, and (d) after single-pass, with the grating tuned to compensate a DGD of 100ps. While the chromatic dispersion is more than 650 ps/nm for single pass, it reduces to less than 80 ps/nm after the double pass configuration. As seen in Fig. 3.11(c) and Fig.3.11 (d), the eye opening for the double pass configuration is much better than that for single pass. Thus, the performance of the double-pass configuration is clearly better than that for the single-pass one.

Fig. 3.11. Measured eye-diagrams

3.4 PMD compensation technique using a pair of high-birefringence linearly chirped FBGs

In this section, we propose a new, chromatic dispersion highly reduced, tunable PMD compensator utilizing a pair of Hi-Bi linearly chirped FBGs. Once again, using a special arrangement, the reflection spectrum of each FBG can be nonlinearly tuned without shift in the centre wavelength [94], making it suitable for WDM systems. Another feature of the present design is that it can provide DGD values starting from

0 ps. Finally, compared with the double pass configurations in [90] and section 3.3, in which cross-talk may occur when tuning the Hi-Bi chirped FBG, this design is free from cross-talk. Eye-diagrams are presented to show the effectiveness of the proposed configuration.

3.4.1 Configuration and principle of operation

Fig. 3.12 shows our proposed two-FBG compensator. The input 10 Gbit/s NRZ signal, using a pseudorandom bit sequence of length $2^{15} - 1$ modulating a laser source, enters the first-order PMD emulator, which generates a Maxwellian distribution of DGD. Next, the incoming signal is reflected from the first Hi-Bi linearly chirped grating from the short wavelength port. With a polarization controller, we can make the slow

Fig. 3.12. PMD compensator using a pair of linearly chirped FBGs

PSP (fast PSP) align with the slow axis (fast axis) of the FBG; the slow axis polarization state has a smaller time delay than the fast axis polarization state when the signal enters the chirped FBG from the short wavelength port. After reflection from the first FBG, the optical signal is reflected from the long wavelength port of the

second Hi-Bi linearly chirped FBG, with a fixed displacement ($f=9\text{mm}$) of the free end of the beam; this time, the slow PSP has a longer time delay. Consequently, with suitable tuning, the 2-FBG scheme can yield time delay values beginning with zero; this is important for PMD compensation. Moreover, when the signal is reflected from the long (short) wavelength port, the dispersion is negative (positive). Thus, any chromatic dispersion induced during the first reflection is reduced greatly by the second reflection. The set-up uses polarization-maintaining circulators.

3.4.2 Experimental results

The Hi-Bi linearly chirped FBG was written by exposing a hydrogen-loaded high-birefringence fiber to 244 nm laser beam through a linearly chirped phase mask. The

Fig. 3.13. Reflection spectrum for displacement at 0 mm and 4 mm

length of the grating is 12.2 cm. Fig. 3.13 shows the measured spectrum for displacements of $f = 0$ and 4 mm for the FBG. Without tuning, the 3-dB bandwidth for each axis is 1.14 nm, and the centre wavelengths are 1543.89 nm and 1544.28 nm for the fast and the slow axis, respectively. With tuning, the bandwidth of the FBG, and hence that of the compensator, can be broadened to 5.34 nm (for $f = 9$ mm), maintaining high reflectivity. However, it is worthwhile to note that insertion loss changes by about 0.8 dB for maximum tuning. The power penalty of the increased insertion loss can be overcome by increasing the laser power or the gain of the amplifier in the fiber link. Fig. 3.14 shows the time delay between the fast and slow axes after reflection from one FBG (upper curve), and after reflection from two FBGs (lower curve) for an optical wavelength of 1554.1 nm. The curves show that the time delay between the two polarization states ranges from 72-446 ps for single-reflection, and from 0-374 ps for double-reflection. This way, one does not need an additional

Fig. 3.14. DGD vs. displacement for reflection from one FBG and two FBGs

Hi-Bi fiber to cover the zero DGD range.

Figs. 3.15 (a-e) present the eye-diagrams for the following conditions, with 60 ps DGD emulation: (a) Before the emulator. (b) After the first circulator; a polarization maintaining (PM) fiber is inserted before the first FBG to nullify the DGD due to the first FBG ($f = 9$ mm). The diagram shows the effect of PMD and the chromatic dispersion due to the first FBG. (c) After the second circulator, with both FBGs having $f = 9$ mm, so that only the chromatic dispersion is substantially removed. While the chromatic dispersion is more than 800 ps/nm.km for the first FBG, it reduces to less than 100ps/nm.km after the second FBG. The diagram shows the effect

Fig. 3.15. Measured eye diagrams showing the effect of one FBG and two FBGs for different conditions. (a) Before the emulator in Fig.3.11. (b) & (d) After the first circulator. (c) & (e) After the second circulator. (See text for details)

of PMD alone. (d) Same as in (b), with the FBG tuned to compensate the PMD. The diagram shows the effect of chromatic dispersion due to the first FBG. (e) Same as in (c), with the first FBG tuned to compensate the PMD; in this condition, both PMD and chromatic dispersion are substantially removed. Fig 3.15(e) clearly shows the improvement caused by the proposed two-FBG configuration. It may be noted that in the two-FBG scheme proposed here, a PM fiber is not required in practice; a PM fiber is inserted before the first FBG only when the eye diagrams are obtained to show the effect of a single FBG (Figs. 3.15(b) and 3.15(d)).

It has been mentioned in the literature that for a digital system to avoid incurring a power penalty of 1 dB or greater for a fractional time of 30 min per year, the average differential delay time between the principal states must be less than 0.14 of the bit period. It means that in a 10 Gbit/s system, the PMD should be less than about 14 ps, and less than about 3.5 ps in a 40 Gbit/s system. Therefore, in a 40 Gbit/s system, the PMD compensator is required to have greater accuracy and stability. Unfortunately, we do not have access to a 40 Gbit/s system to test our compensators. However, we believe that the two PMD compensators can be extended to 40 Gbit/s, if one uses an electrically controlled micrometer to tune the Hi-Bi FBG instead of the manually controlled micrometer.

3.5 Conclusion

Two novel PMD compensation methods using a linearly chirped FBG in a double-pass configuration have been presented. For both schemes, measured results verify that the chromatic dispersion is reduced significantly in these PMD compensator configurations. Furthermore, both schemes offer tunability over a large bandwidth and are fiber compatible. The double-pass configuration proposed here can compensate PMD over a large range. The two-FBG design presented here makes the DGD tuning range start from zero, is free from cross-talk and can compensate PMD over a range of wavelengths.

Chapter 4

Power Fading due to Polarization Mode

Dispersion and Chromatic Dispersion for Radio

Frequency Signals and Its Application for PMD

and CD Monitoring

4.1 Introduction

Polarization mode dispersion (PMD) is considered as the ultimate limiting factor for ultrahigh bit-rate transmission [11]. PMD arises from the birefringence in optical fibers due to some amount of asymmetry caused by imperfections in the manufacturing process or mechanical stress on the fiber after manufacture. Further, sub-carrier multiplexing (SCM) is a modulation scheme where multiple signals are multiplexed in the RF domain and transmitted on a single optical carrier. An important advantage of SCM is that microwave devices are more mature than optical devices: the stability of microwave oscillators and the frequency selectivity of microwave filters are much better than their optical counterparts. SCM has several important applications in optical systems, including cable television. Because of the simple and low-cost implementation, SCM has also been proposed to transmit multichannel signals for high-capacity local area networks [68]. Transmission of analog and digital SCM signals over fiber will be severely affected by PMD [68][69][98]. The statistics of PMD induced power fading for double sideband and single sideband SCM signals has been reported in [99]; however, the average power

fading is dependent on the precession angle, which is between the input state of polarization and the PMD vector including the polarization-dependent chromatic dispersion (PCD) effect, has not been reported. Moreover, the power fading including higher-order PMD and chromatic dispersion also has not been reported.

As discussed in section 1.4, monitoring of chromatic dispersion (CD) and PMD is important in the context of compensation of CD and PMD. RF signals have been used for the proposing of monitoring quite effectively. Several techniques for chromatic dispersion monitoring have been reported [70] [100-103]. These include using eye opening [100], detecting the intensity modulation induced by a phase modulation [101], extracting the bit-rate frequency component (clock) from photodetected data [102], inserting a sub-carrier at the transmitter [70] [103], or using sideband optical filter and clock phase shift detection [104]. The pilot-tone based CD monitoring techniques offer good sensitivity, dynamic range, and reconfigurability. Further, with regard to PMD monitoring, the PMD monitoring techniques using DOP monitoring and eye monitoring have been reported in [105] [106]. However, among the many reported techniques, the ones based on radio frequency (RF) power measurement are relatively simple. A method that tracks the effects of PMD by simply measuring the RF power has been demonstrated [107]; it is simple and is also applicable to wavelength division-multiplexing systems [108]. However, the RF tone is affected not only by PMD, but also by chromatic dispersion. A PMD monitoring technique that avoids the effects of chromatic dispersion by using SSB subcarrier transmission and tracking the chromatic dispersion insensitive RF power of the vestigial sideband has also been reported [109-112]. But the above-mentioned techniques cannot monitor the chromatic dispersion and PMD at the same time. Since the magnitude of PMD (i.e., differential group delay (DGD)) depends on the angle between the state-of-

polarization (SOP) and the PMD vector, it is also important to monitor the effect of this angle on power penalty due to DGD.

In this chapter, we use the expression relating the output state of polarization and PMD vector, including polarization-dependent chromatic dispersion (PCD), as discussed in chapter two. Based on this expression, we get the power fading including higher-order PMD and chromatic dispersion, which is dependent on the angle of precession of output state of polarization around the PMD vector. From the expression for power fading, we get the average power penalty for chromatic dispersion and higher-order PMD. Moreover, a simple and novel pilot-tone-based technique has been proposed for simultaneous monitoring of chromatic dispersion and PMD. Experimental results demonstrate the effectiveness of the proposed technique; these results also agree very well with the theoretical analysis.

Further, based on the power fading expression, we propose a novel method to measure the DGD between the fast and slow axes of Hi-Bi chirped grating. A similar method can measure the PMD of chirped grating easily.

4.2 Effect of first-order PMD on an RF signal

4.2.1 RF power fading for the case of double sideband (DSB) modulation

Polarization mode dispersion is characterized by the PMD-vector, $\vec{\Omega}$, in the Stokes space, around which an output state of polarization (SOP), $\hat{S}_o(\omega)$, precesses when the carrier frequency is changed, i.e. [12],

$$\frac{d\hat{S}_o(\omega)}{d\omega} = \vec{\Omega} \times \hat{S}_o(\omega) \quad (4.1)$$

The modulus of the PMD vector, $\Delta\tau = |\Omega|$, is the differential group delay (DGD); DGD is the time delay between the two principal states of polarization (PSPs) which are the two eigen-polarizations [11]. The PMD vector $\vec{\Omega}$ can be expressed as $\vec{\Omega} = \Delta\tau\hat{\Omega}$, where $\hat{\Omega}$ is the unit PMD vector.

In this section, we assume that the polarization dependent loss (PDL) and chromatic dispersion are negligible. Following the discussion in section 2.2.1, the output SOP can be described as

$$\hat{S}_o(\omega) = \hat{\Omega}\cos 2\alpha + [\hat{S}_o(\omega_0) - \hat{\Omega}\cos 2\alpha] \cos \Delta\tau\Delta\omega + \hat{B}\sin 2\alpha \sin \Delta\tau\Delta\omega \quad (4.2)$$

where 2α is the angle between $\hat{\Omega}$ and $\hat{S}_o$, $\hat{B}$ is the unit vector of $\hat{\Omega} \times \hat{S}_o$. In Eq. (2.2), the three unit vectors $\hat{\Omega}$, $(1/\sin 2\alpha)(\hat{S}_o(\omega_0) - \hat{\Omega}\cos 2\alpha)$ and $\hat{B}$ constitute a 3-D orthogonal coordinate system. Eq. (4.2) implies that the output state of polarization precesses around the PMD vector with a precession angle 2α , which is independent of frequency. The projection of the output state of polarization in the plane normal to the PMD vector rotates by an angle $\Delta\tau\Delta\omega$. Based on Eq. (4.2), we can express the SOP $\hat{e}(\omega)$ in terms of the two orthogonal principal states of polarization, $\hat{e}_{\Omega+}$ and $\hat{e}_{\Omega-}$ in Jones notation:

$$\hat{e}(\omega) = \cos\alpha e^{j\phi_+} \hat{e}_{\Omega+} + \sin\alpha e^{j\phi_-} \hat{e}_{\Omega-} \quad (4.3)$$

The phase difference between the two orthogonal states of polarization is $\Delta\tau\Delta\omega$ [11],

So Eq. (4.3) can be expressed as

$$\hat{e}(\omega) = \cos\alpha e^{\frac{j\Delta\tau\Delta\omega}{2}} \hat{e}_{\Omega+} + \sin\alpha e^{\frac{-j\Delta\tau\Delta\omega}{2}} \hat{e}_{\Omega-} \quad (4.4)$$

Next, the input RF voltage to the external modulator can be represented as

$$V(t) = 1 + m \cos \omega_m t \quad (4.5)$$

Considering a linear intensity modulation, the output optical power of the modulator can be expressed as:

$$P_i(t) = P_i(1 + m \cos \omega_m t) \quad (4.6)$$

where P_i is the optical power output from the laser, m is the modulation index, which equals to the peak intensity change divided by the average intensity [113] and ω_m is the modulating angular frequency. The corresponding electric field of the light wave can be expressed as:

$$f(t) = \sqrt{P_i} \sqrt{1 + m \cos(\omega_m t)} \cos(\omega_0 t) \quad (4.7)$$

where ω_0 is the angular frequency of the optical carrier. It is assumed that the modulator has a Hi-Bi pigtail and the polarization of the output light from the laser is optimally aligned to the modulator. Under small signal condition ($m \ll 1$), Eq. (4.7) can be approximated as:

$$\begin{aligned} f(t) &= \sqrt{P_i} \left(1 + \frac{1}{2} m \cos \omega_m t \right) \cos \omega_0 t \\ &= \sqrt{P_i} \left[\cos \omega_0 t + \frac{m}{4} \cos(\omega_0 + \omega_m)t + \frac{m}{4} \cos(\omega_0 - \omega_m)t \right] \end{aligned} \quad (4.8)$$

From Eq. (4.8), it is obvious that, under small signal condition, an intensity-modulated signal can be approximated to a typical amplitude modulated signal with double sidebands separated by $2\omega_m$. Next, consider the propagation of the optical signal $f(t)$ on the fiber. The signal has its Fourier transform:

$$F(\omega - \omega_0) = \int_{-\infty}^{+\infty} f(t) \exp[-j(\omega - \omega_0)t] dt \quad (4.9)$$

The signal, after propagation, can be decomposed into two components, $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$, one each along the two orthogonal PSPs, $\hat{\Omega}_+$ and $\hat{\Omega}_-$. The two components $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$ are written as:

$$f_{\Omega+}(t) = \frac{M \cos \alpha}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp[j(\Delta \tau \Delta \omega / 2)] \exp[j(\omega - \omega_0)t] d\omega \quad (4.10)$$

$$f_{\Omega-}(t) = \frac{M \sin \alpha}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp[-j(\Delta \tau \Delta \omega / 2)] \exp[j(\omega - \omega_0)t] d\omega \quad (4.11)$$

where M is signal amplitude factor; $\Delta \tau$ is the difference of the time delays between the two orthogonally polarized signal components. Combining the components, we get

$$\vec{f}(t) = f_{\Omega+}(t) \hat{\Omega}_+ + f_{\Omega-}(t) \hat{\Omega}_- \quad (4.12)$$

Eqs. (4.10)-(4.12) are general equations, applicable to an optical signal of any shape.

When the input signal is an RF signal with double sideband (DSB) modulation,

$$f_{\Omega+}(t) = \cos(\alpha) M \sqrt{P_i} \left\{ \cos(\omega_0 t) + \frac{m}{4} \cos[(\omega_0 + \omega_m)t + (\Delta \tau \omega_m / 2)] \right. \\ \left. + \frac{m}{4} \cos[(\omega_0 - \omega_m)t - (\Delta \tau \omega_m / 2)] \right\} \quad (4.13)$$

$$f_{\Omega-}(t) = \sin(\alpha) M \sqrt{P_i} \left\{ \cos(\omega_0 t) + \frac{m}{4} \cos[(\omega_0 + \omega_m)t - (\Delta \tau \omega_m / 2)] \right. \\ \left. + \frac{m}{4} \cos[(\omega_0 - \omega_m)t + (\Delta \tau \omega_m / 2)] \right\} \quad (4.14)$$

Since $\hat{\Omega}_+$ and $\hat{\Omega}_-$ are orthogonal, the optical signal power at the detector is

$$P(t) = |\vec{f}(t)|^2 = \vec{f}(t) \cdot \vec{f}^*(t) = |f_{\Omega+}(t)|^2 + |f_{\Omega-}(t)|^2 \quad (4.15)$$

By assuming the responsivity of the photodiode to be R , we get the RF output current as:

$$I(t) = (\cos \alpha)^2 R M^2 P_i \cdot \{ m \cos[\omega_m(t + \Delta \tau / 2)] \} \\ + (\sin \alpha)^2 R M^2 P_i \cdot \{ m \cos[\omega_m(t - \Delta \tau / 2)] \} \quad (4.16)$$

The detector electrical power $P_e(t) = |I(t)|^2 R_L$, where R_L is the load resistance; its power spectrum can be derived using Fourier transform as:

$$P_e(\omega_m) \propto \left\{ 1 - (\sin 2\alpha)^2 \cdot [\sin(\Delta\tau\omega_m/2)]^2 \right\} \quad (4.17)$$

Eq. (4.17) shows the output RF power under the influence of first-order PMD not only depends on the DGD but also is dependent on the angle between the output state of polarization and the PMD vector. Moreover, the distribution of $\sin(2\alpha)$ is given by [73]:

$$f_x = \frac{x}{\sqrt{1-x^2}} \quad (0 \leq x \leq 1) \quad (4.18)$$

So the average output power can be expressed by

$$P_e(f_m) \propto \left\{ 1 - \frac{2}{3} [\sin(\pi f_m \Delta\tau)]^2 \right\} \quad (4.19)$$

where $f_m = \omega_m/2\pi$ is the frequency of modulated RF signal. Eq. (4.19) clearly shows that the average power fading induced by PMD is about 2/3 of the worst case; the worst case occurs when the input signal equally excites components along the two PSPs ($2\alpha = \pi/2$).

4.2.2 RF power fading for the case of single sideband (SSB) modulation

Eqs. (4.10)-(4.12) are general equations, applicable to an optical signal of any shape. When the input signal is an RF signal and with single sideband (SSB) modulation, then the two components $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$ are described as:

$$f_{\Omega+}(t) = \cos(\alpha) M \sqrt{I_i} \left[\cos(\omega_0 t) + \frac{m}{4} \cos[(\omega_0 + \omega_m)t + (\Delta\tau\omega_m/2)] \right] \quad (4.20)$$

$$f_{\Omega-}(t) = \sin(\alpha) M \sqrt{I_i} \left[\cos(\omega_0 t) + \frac{m}{4} \cos[(\omega_0 + \omega_m)t - (\Delta\tau\omega_m/2)] \right] \quad (4.21)$$

Similar to discussion in section 4.2.1, by assuming the responsivity of the photodiode to be R , we get the RF output current as:

$$I(t) = (\cos \alpha)^2 R M^2 P_i \cdot \left\{ (m/2) \cos[\omega_m(t + \Delta\tau/2)] \right\} \\ + (\sin \alpha)^2 R M^2 P_i \cdot \left\{ (m/2) \cos[\omega_m(t - \Delta\tau/2)] \right\} \quad (4.22)$$

The detector electrical power $P_e(t) = |I(t)|^2 R_L$, where R_L is the load resistance, its power spectrum is given as:

$$P_e(\omega_m) \propto \left\{ 1 - (\sin 2\alpha)^2 \cdot [\sin(\Delta\tau\omega_m/2)]^2 \right\} \quad (4.23)$$

Comparison of Eq. (4.17) and Eq. (4.23) clearly shows that the power fading due to first-order PMD is the same for DSB and SSB modulation.

4.2.3 Experimental setup for monitoring the effect of first-order PMD

The experimental setup for RF power fading due to PMD is shown in Fig. 4.1. The laser source is modulated by a 10 GHz RF signal in an intensity modulator with a Hi-Bi Panda pigtail fiber. Then the signal passes through a $\lambda/4$ waveplate fixed on a fiber

Fig. 4.1. Experimental setup

bench. After transmission through the first-order PMD emulator, which consists of two polarization beam splitters (PBS) with Hi-Bi pigtails and a time delay element, which is also a Hi-Bi pigtail fiber, the signal pass through a 90:10 coupler. Thus a small portion of the signal goes to a power meter which helps in ensuring same power level for different setting of the emulator. Then the RF signal is detected by a 12.5 GHz photodetector (PD) and is measured using an RF spectrum analyzer (RFSa). The measured output RF power levels with DGD values for 20 ps, 33 ps and 39 ps, are shown in Fig. 4.2.

Fig. 4.2. Measured RF power for different values of DGD with different values of the angle between input state of polarization and PMD vector

It clearly shows that the RF power fading is dependent on the angle between the output state of polarization and PMD vector. The measured power fading experimental results agree well with the theoretical results based on Eq. (4.17), which are also plotted in Fig. 4.2.

4.3 Effect of CD and PMD, including PCD, on an RF signal

4.3.1 The case of double sideband (DSB) modulation

As discussed in section 2.3 and [76], the output state of polarization can be expressed as

$$\begin{aligned} \hat{S}_o(\omega) = & \hat{\Omega} \cos 2\alpha + \left[\hat{S}_o(\omega_0) - \hat{\Omega} \cos 2\alpha \right] \cos(\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2) \\ & + \hat{B} \sin 2\alpha \sin(\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2) \end{aligned} \quad (4.24)$$

Here $\Delta\tau_1$ is PCD, which is defined as $\Delta\tau_1 = d\Delta\tau/d\omega$, 2α is the angle between $\hat{\Omega}$ and $\hat{S}_o(\omega_0)$, $\hat{B}$ is the unit vector $\hat{\Omega} \times \hat{S}_o(\omega_0)$. Eq. (4.24) means that the output state of polarization precesses around the PMD vector with a precession angle 2α , which is independent of frequency. The projection of the output state of polarization in the plane normal to the PMD vector rotates by an angle $\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2$. Based on Eq. (4.24) and similar to the discussion in section 2.3, we can express the SOP in terms of the two orthogonal principal states of polarization, $\hat{e}_{\Omega+}$ and $\hat{e}_{\Omega-}$ in Jones notation:

$$\hat{e}(\omega) = \cos\alpha e^{j\phi_+} \hat{e}_{\Omega+} + \sin\alpha e^{j\phi_-} \hat{e}_{\Omega-} \quad (4.25)$$

As shown in chapter two, in this case the phase difference between the two orthogonal states of polarization is $\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2$. So Eq. (4.25) can be expressed as

$$\hat{e}(\omega) = \cos\alpha e^{\frac{j(\Delta\tau\Delta\omega - \Delta\tau_1\Delta\omega^2)}{2}} \hat{e}_{\Omega+} + \sin\alpha e^{\frac{j(-\Delta\tau\Delta\omega + \Delta\tau_1\Delta\omega^2)}{2}} \hat{e}_{\Omega-} \quad (4.26)$$

Similar to Eqs. (4.5)-(4.9), the signal, after propagation, can be decomposed into two components, $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$, one each along the two orthogonal axes, $\hat{\Omega}_+$ and $\hat{\Omega}_-$.

Also, the propagation constant $\beta(\omega)$ can be expanded as a Taylor series around $\beta(\omega_0)$ as

$$\beta(\omega) = \beta_0 + \beta_1(\omega - \omega_0) + \frac{1}{2}\beta_2(\omega - \omega_0)^2 + \dots \quad (4.27)$$

where

$$\beta_m = \left(\partial^m \beta / \partial \omega^m \right) \Big|_{\omega=\omega_0} \quad (4.28)$$

The phase change due to the chromatic dispersion can be expressed as

$$\Phi_{\Omega\pm} = (\beta_2 L \omega_m^2 / 2) \quad (4.29)$$

Based on Eq. (4.26) and (4.29), the two components $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$ are written as:

$$f_{\Omega+}(t) = \frac{M \sin \alpha}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp \left\{ j \left[\left(\Delta \tau \Delta \omega / 2 \right) - \left(\Delta \tau_1 \Delta \omega^2 / 2 \right) \right] \right\} e^{j(\omega - \omega_0)t} d\omega \quad (4.30)$$

$$f_{\Omega-}(t) = \frac{M \sin \alpha}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp \left\{ -j \left[\left(\Delta \tau \Delta \omega / 2 \right) - \left(\Delta \tau_1 \Delta \omega^2 / 2 \right) \right] \right\} e^{j(\omega - \omega_0)t} d\omega \quad (4.31)$$

where M is signal amplitude factor; $\Delta \tau$ is the DGD, $\Delta \omega$ corresponds to ω_m and $\Delta \tau_1$ is PCD. Combining the components, one gets

$$\vec{f}(t) = f_{\Omega+}(t) \hat{\Omega}_+ + f_{\Omega-}(t) \hat{\Omega}_- \quad (4.32)$$

Eqs. (4.30)-(4.32) are general equations, applicable to an optical signal of any shape.

When the input signal is an RF signal with DSB modulation

$$f_{\Omega+}(t) = \cos(\alpha) M \sqrt{I_i} \left\{ \begin{aligned} &\cos(\omega_0 t) + \frac{m}{4} \cos \left[\left(\omega_0 + \omega_m \right) t + \left(\Delta \tau \omega_m / 2 \right) \right. \\ &\quad \left. - \left(\Delta \tau_1 \omega_m^2 / 2 \right) - \left(\beta_2 L \omega_m^2 / 2 \right) \right] \\ &+ \frac{m}{4} \cos \left[\left(\omega_0 - \omega_m \right) t - \left(\Delta \tau \omega_m / 2 \right) \right. \\ &\quad \left. - \left(\Delta \tau_1 \omega_m^2 / 2 \right) - \left(\beta_2 L \omega_m^2 / 2 \right) \right] \end{aligned} \right\} \quad (4.33)$$

$$f_{\Omega-}(t) = \sin(\alpha) M \sqrt{I_i} \left\{ \begin{aligned} &\cos(\omega_0 t) + \frac{m}{4} \cos \left[\left(\omega_0 + \omega_m \right) t - \left(\Delta \tau \omega_m / 2 \right) \right. \\ &\quad \left. + \left(\Delta \tau_1 \omega_m^2 / 2 \right) - \left(\beta_2 L \omega_m^2 / 2 \right) \right] \\ &+ \frac{m}{4} \cos \left[\left(\omega_0 - \omega_m \right) t + \left(\Delta \tau \omega_m / 2 \right) \right. \\ &\quad \left. + \left(\Delta \tau_1 \omega_m^2 / 2 \right) - \left(\beta_2 L \omega_m^2 / 2 \right) \right] \end{aligned} \right\} \quad (4.34)$$

Since $\hat{\Omega}_+$ and $\hat{\Omega}_-$ are orthogonal, the optical signal power at the detector is

$$P(t) = |\vec{f}(t)|^2 = \vec{f}(t) \cdot \vec{f}^*(t) = |f_{\Omega+}(t)|^2 + |f_{\Omega-}(t)|^2 \quad (4.35)$$

By assuming the responsivity of the photodiode to be R , we get the RF output current as:

$$\begin{aligned} I(t) = & (\cos \alpha)^2 R M^2 P_i \cdot \left\{ m \cos \left[\left(\Delta \tau_1 \omega_m^2 / 2 \right) + \left(\beta_2 L \omega_m^2 / 2 \right) \right] \cdot \cos \left[\omega_m (t + \Delta \tau / 2) \right] \right\} \\ & + (\sin \alpha)^2 R M^2 P_i \cdot \left\{ m \cos \left[\left(\beta_2 L \omega_m^2 / 2 \right) - \left(\Delta \tau_1 \omega_m^2 / 2 \right) \right] \cdot \cos \left[\omega_m (t - \Delta \tau / 2) \right] \right\} \end{aligned} \quad (4.36)$$

The detector electrical power $P_e(t) = |I(t)|^2 R_L$, where R_L is the load resistance; its power spectrum can be written as:

$$P_e(\omega_m) \propto \left\{ \begin{aligned} & \left[\cos(\beta_2 L \omega_m^2 / 2) \cdot \cos(\Delta \tau_1 \omega_m^2 / 2) \right]^2 \\ & + (\cos 2\alpha)^2 \cdot \left[\sin(\beta_2 L \omega_m^2 / 2) \cdot \sin(\Delta \tau_1 \omega_m^2 / 2) \right]^2 \\ & - (1/2) \cos(2\alpha) \cdot \sin(\Delta \tau_1 \omega_m^2) \cdot \sin(\beta_2 L \omega_m^2) \\ & - \left[(\sin 2\alpha)^2 \cdot \cos \left[(\Delta \tau_1 \omega_m^2 / 2) + (\beta_2 L \omega_m^2 / 2) \right] \right. \\ & \left. \times \cos \left[(\beta_2 L \omega_m^2 / 2) - (\Delta \tau_1 \omega_m^2 / 2) \right] \cdot \left[\sin(\Delta \tau \omega_m / 2) \right]^2 \right] \end{aligned} \right\} \quad (4.37)$$

If the PCD is neglected, then Eq. (4.38) can be expressed as

$$P_e(\omega_m) \propto \left[\cos(\beta_2 L \omega_m^2 / 2) \right]^2 \left\{ 1 - (\sin 2\alpha)^2 \cdot \left[\sin(\Delta \tau \omega_m / 2) \right]^2 \right\} \quad (4.38)$$

Eq. (4.38) is the RF output power affected by first-order PMD and chromatic dispersion. The maximum output power occurs when $2\alpha = 0$ or $2\alpha = \pi$, which means that the signal is aligned with one of the two principal states of polarization. Moreover, the chromatic dispersion parameter D can be expressed as $D = -2\pi c \beta_2 / \lambda^2$ and $\omega_m = 2\pi f_m$, where f_m is the frequency of the modulating RF signal. Then, the maximum output power is

$$P_e \propto \left[\cos(\beta_2 L \omega_m^2 / 2) \right]^2 = \left[\cos(\pi f_m^2 \lambda^2 D l / c) \right]^2 \quad (4.39)$$

Comparison of Eq. (4.37) and Eq. (4.38) clearly shows that PCD and chromatic dispersion is dependent on the effect of output RF signal power.

In the presence of PCD, for the special case $2\alpha = 0$, which means that input state of polarization is aligned with the slow PSP, the output RF signal power is

$$P_e(\omega_m) \propto \left\{ \begin{aligned} & \left[\cos(\beta_2 L \omega_m^2 / 2) \cdot \cos(\Delta \tau_1 \omega_m^2 / 2) \right]^2 + \left[\sin(\beta_2 L \omega_m^2 / 2) \cdot \sin(\Delta \tau_1 \omega_m^2 / 2) \right]^2 \\ & - (1/2) \sin(\Delta \tau_1 \omega_m^2) \cdot \sin(\beta_2 L \omega_m^2) \end{aligned} \right\} \quad (4.40)$$

In this case the output power is not the maximum. However, if $2\alpha = \pi$, which means that the input state of polarization is aligned with the fast PSP, the output is the maximum. In this case the maximum output power can be shown as:

$$P_e(\omega_m) \propto \left\{ \begin{aligned} & \left[\cos(\beta_2 L \omega_m^2 / 2) \cdot \cos(\Delta \tau_1 \omega_m^2 / 2) \right]^2 + \left[\sin(\beta_2 L \omega_m^2 / 2) \cdot \sin(\Delta \tau_1 \omega_m^2 / 2) \right]^2 \\ & + (1/2) \sin(\Delta \tau_1 \omega_m^2) \cdot \sin(\beta_2 L \omega_m^2) \end{aligned} \right\} \quad (4.41)$$

Comparison of Eq. (4.40) with Eq. (4.41) clearly shows that the RF power fading can be reduced greatly when the input state of polarization is align with fast PSP rather than the slow PSP.

Further, the distribution of $\sin(2\alpha)(f_x)$ and $\cos(2\alpha)(f_y)$ are given by [73]:

$$\begin{cases} f_x = \frac{x}{\sqrt{1-x^2}} & (0 \leq x \leq 1) \\ f_y = 1/2 & (-1 \leq y \leq 1) \end{cases} \quad (4.42)$$

Using Eq. (4.37) and Eq. (4.42), the average output power can be described as

$$P_e(\omega_m) \propto \left\{ \begin{aligned} & \left[\cos(\beta_2 L \omega_m^2 / 2) \cdot \cos(\Delta \tau_1 \omega_m^2 / 2) \right]^2 + \frac{2}{3} \left[\sin(\beta_2 L \omega_m^2 / 2) \cdot \sin(\Delta \tau_1 \omega_m^2 / 2) \right]^2 \\ & - \frac{2}{3} \left\{ \cos \left[(\Delta \tau_1 \omega_m^2 / 2) + (\beta_2 L \omega_m^2 / 2) \right] \right. \\ & \left. \times \cos \left[(\beta_2 L \omega_m^2 / 2) - (\Delta \tau_1 \omega_m^2 / 2) \right] \cdot \left[\sin(\Delta \tau \omega_m / 2) \right]^2 \right\} \end{aligned} \right\} \quad (4.43)$$

When we do not take the PCD into account, then the average output power can be expressed as

$$P_e(\omega_m) \propto \left[\cos(\pi f_m^2 \lambda^2 D l / c) \right]^2 \left\{ 1 - \frac{2}{3} \left[\sin(\pi \Delta \tau f_m) \right]^2 \right\} \quad (4.44)$$

To the best of our knowledge, it is the first time to get the average power penalty induced by first-order PMD and CD.

It should be pointed out that an expression similar to Eq. (4.38) has been reported in Reza et al [114]. However, the expression given in [114] only considers the first-order PMD. The expressions presented here include the effects of higher-order PMD and

CD on power fading and we also present expressions for the average value of detected power. Moreover, our expressions explicitly include the angle 2α ; the advantages of such an approach have been mentioned in Section 2.2.2.1. It should also be pointed out that the cosine variation for CD in our expressions is consistent with the results reported in literature [e.g., 70].

4.3.2 The case of single sideband (SSB) modulation

When the input signal is an RF signal and with single sideband (SSB) modulation, the two components $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$ can be written as

$$f_{\Omega+}(t) = \cos(\alpha) M \sqrt{P_i} \left\{ \cos(\omega_0 t) + \frac{m}{4} \cos \left[(\omega_0 + \omega_m)t + (\Delta\tau\omega_m/2) - (\Delta\tau_1\omega_m^2/2) - (\beta_2 L\omega_m^2/2) \right] \right\} \quad (4.45)$$

$$f_{\Omega-}(t) = \sin(\alpha) M \sqrt{P_i} \left\{ \cos(\omega_0 t) + \frac{m}{4} \cos \left[(\omega_0 + \omega_m)t - (\Delta\tau\omega_m/2) + (\Delta\tau_1\omega_m^2/2) - (\beta_2 L\omega_m^2/2) \right] \right\} \quad (4.46)$$

By assuming the responsivity of the photodiode to be R , we get the RF output current as:

$$I(t) = (\cos \alpha)^2 R M^2 P_i \cdot \left\{ (m/2) \cdot \cos \left[\omega_m(t + \Delta\tau/2) - (\Delta\tau_1\omega_m^2/2) - (\beta_2 L\omega_m^2/2) \right] \right\} \\ + (\sin \alpha)^2 R M^2 P_i \cdot \left\{ (m/2) \cdot \cos \left[\omega_m(t - \Delta\tau/2) - (\beta_2 L\omega_m^2/2) + (\Delta\tau_1\omega_m^2/2) \right] \right\} \quad (4.47)$$

The detector electrical power $P_e(t) = |I(t)|^2 R_L$, where R_L is the load resistance; its power spectrum is:

$$P_e(\omega_m) \propto \left\{ 1 - (\sin 2\alpha)^2 \cdot \left\{ \sin \left[(\Delta\tau\omega_m/2) - (\Delta\tau_1\omega_m^2/2) \right] \right\}^2 \right\} \quad (4.48)$$

Eq. (4.48) shows that the SSB is beneficial in reducing the overall PMD induced RF power fading. The SSB modulation not only gets rid of the effect of chromatic dispersion but also reduces the effect of first-order PMD effect. Further, based on the

distribution of $\sin(2\alpha)$, as shown in Eq. (4.42), the average output power can be expressed as

$$P_e(f_m) \propto \left\{ 1 - \frac{2}{3} \left[\sin(\pi f_m \Delta \tau - 2\pi^2 f_m^2 \Delta \tau_1) \right]^2 \right\} \quad (4.49)$$

To the best of our knowledge, it is the first time to get an expression for the average output power for the case of SSB modulation. Eq. (4.49) means that the average power fading due to PMD and PCD is reduced for the case of SSB modulation compared with power fading due to first-order PMD only.

4.3.3 A new technique for simultaneous monitoring of PMD and CD

For CD monitoring, based on Eq. (4.17) and Eq. (4.38), it is seen that we can easily measure the power fading due to chromatic dispersion by comparing the output RF power for DSB modulation with that for SSB modulation, because DSB modulation involves the power fading due to both chromatic dispersion and PMD, whereas, SSB modulation only involves power fading due to PMD. The CD monitoring technique based on the above observation is shown in Fig. 4.3. Light signal at the wavelength of 1558.1 nm is modulated with a 10.709 Gbit/s NRZ pseudorandom pattern of length $2^{15} - 1$ and 10 GHz RF signal using a Mach-Zehnder (MZ) modulator with a Hi-Bi pigtail. The 10.709 Gbit/s NRZ is the signal pulse transmission in optical fiber communication system, as well as the 10 GHz RF signal to monitor PMD and CD, which is important for PMD and CD compensation. The modulated signal passes through a $\lambda/4$ waveplate which is fixed on a fiber bench. Then the signal passes through the first-order PMD emulator. The emulator consists of two polarization beam splitters (PBS) and a time delay element (both the two PBS and the time delay line is with Hi-Bi pigtail). Then the signal passes through an EDFA. A bandpass filter

removes the ASE noise of the EDFA. Then the signal passes through a 20 km long single mode fiber. After this, a small part of the signal is tapped for monitoring. The monitoring signal is divided into two paths by a 3 dB coupler. One part is directly detected by a photodetector (PD) and is observed on an RFSA; the other part passes through a uniform FBG which is used to filter out the upper side band of the modulated RF signal. The resulting SSB signal is detected by a photodetector and measured by an RFSA. The attenuator in the DSB path is to adjust

Fig. 4.3. Experimental setup for CD monitoring

the optical power of the optical signal to be the same as that for SSB before it is detected by the photodetector. The measured SSB optical spectrum at the output of the FBG is shown in Fig. 4.4. Optical SSB is very clearly evident, with the power in the lower sideband larger than that in the upper sideband by ~21dB. The measured normalized RF output power for DSB and SSB modulation, for DGD values 20 ps, 33 ps, and 39 ps, is shown in Fig. 4.5, Fig. 4.6, and Fig. 4.7 respectively. The calculated

curves in these figures are based on Eq. (4.49) ($\Delta\tau_1 = 0$) and Eq. (4.38) for normalized output RF power for SSB and DSB, respectively. The parameters used for calculations are shown in Table 4.1.

Table 4.1: Parameters and equations for calculated results in Figs.4.5, 4.6 and 4.7.

Figs.	4.5; 4.6; 4.7
λ (Wavelength)	1558.1 nm
$f_m; c$	10 GHz; 2×10^8
DL	340 ps/nm
$\Delta\tau$ (DGD)	20 ps; 33 ps; 39 ps
DSB	Eq. (4.38) $P_e(\omega_m) \propto \left[\cos(\pi f_m^2 \lambda^2 DL / c) \right]^2 \left\{ 1 - (\sin 2\alpha)^2 \cdot [\sin(\pi f_m \Delta\tau)]^2 \right\}$
SSB	Eq. (4.48) ($\Delta\tau_1 = 0$) $P_e(\omega_m) \propto \left\{ 1 - (\sin 2\alpha)^2 \cdot [\sin(\pi f_m \Delta\tau - 2\pi^2 f_m^2 \Delta\tau_1)]^2 \right\}$

The measured power fading due to CD in the 20 km single mode fiber is about 3.7 dB which shows that the power fading due to CD is independent of the DGD and the angle between SOP and PMD vector. The calculated power fading for a 20 km SMF ($D=17\text{ps}/(\text{nm.km})$) using Eq. (4.39) is also about 3.7 dB. This result is based on the parameters shown in Table 4.1. The measured results agree well with the calculated results for different values of DGD and different angle between input SOP and PMD vector.

Fig. 4.4. Measured result for SSB

Fig. 4.5. Measured output RF signal power for DSB and SSB modulation (with DGD=20 ps)

Fig. 4.6. Measured output RF signal power for DSB and SSB modulation (with DGD=33 ps)

Fig. 4.7. Measured output RF signal power for DSB and SSB modulation (with DGD=39 ps)

For simultaneous monitoring of PMD and CD, based on the expressions in section 4.3.1 and 4.3.2, a simple and novel inband pilot-tone-based technique for simultaneous monitoring of chromatic dispersion and PMD is proposed in this section. The effectiveness of the technique is demonstrated through experimental results.

The experimental setup for monitoring both PMD and CD is shown in Fig. 4.8 [115]. Light of wavelength 1552.41 nm is modulated with a 10.709 Gbit/s NRZ pseudorandom

Fig. 4.8. Experimental setup for both PMD and CD monitoring

pattern of length $2^{15} - 1$ and an inband pilot-tone at 10 GHz using Mach-Zehnder (MZ) modulator with a Hi-Bi pigtail. The power-spectral density corresponding to the data is quite small at the pilot-tone frequency as shown in Fig.4.9. The modulated signal passes through a $\lambda/4$ waveplate, a first-order PMD emulator and an EDFA. A bandpass filter removes the ASE noise of the EDFA. Then the signal passes through a single mode fiber; fiber of different lengths permits emulation of different values of chromatic dispersion. Following this, a small part of the signal is tapped for

monitoring. The tapped signal is divided into two parts by a 3 dB coupler. One part is directly detected by a photodetector (PD) and is observed on a spectrum analyzer, RFSA3, to measure the power of the 10 GHz tone; the other part passes through a uniform grating which filters out the upper sideband of the modulated RF signal. The output of the grating is further sub-divided into two parts using a 3 dB coupler. The first part, an SSB signal, is detected by a PD and the pilot tone power is measured by RFSA2. The second part passes through a circulator and another uniform FBG that reflects the lower sideband, so that the optical carrier and lower sideband are

Fig. 4.9. RF spectrum of RF tone and NRZ format

separated in the ports 2 and 3 of the circulator, respectively. The centre wavelength separation between the two uniform FBGs is 0.16 nm and the bandwidth of the two FBGs is about 0.08 and 0.1 nm respectively with a 99.9% reflection coefficient. Next, the reflected sideband passes through a polarization controller before recombining with the optical carrier through a 3 dB coupler. The recombined RF signal goes to a

PD and its power is measured by RFSA1. The two attenuators in the RFSA2 and RFSA3 branches ensure that the optical power incident on all the PDs are identical in the absence of RF modulation. By adjusting the polarization for the lower sideband, the SOP of the sideband and the carrier can be made identical; thus the RF power fading induced by PMD can be compensated [110][111]. So when we compare the RF power of SSB before and after PMD compensation, i.e., power in RFSA2 and RFSA1, respectively, the power penalty induced by PMD can be easily measured. Similarly, comparison of powers in RFSA3 and RFSA2 yields the power fading due to CD.

The difference between the measured output RF powers of RFSA1 and RFSA2, with

Fig. 4.10. Power fading due to PMD (RFSA1 – RFSA2) vs. angle between input SOP and PMD vector

DGD values of 20 ps, 33 ps and 39 ps and 44 ps, respectively, is shown in Fig. 4.10. The continuous line curves are calculated from Eq. (4.23). The parameters used for calculations are described in Table 4.2.

Table 4.2: Parameters and equation for calculated results in Fig.4.10

Fig.	$\Delta\tau$ (DGD)	f_m	Eq.
4.10	20 ps; 33 ps; 39 ps; 44 ps	10 GHz	(4.23) $P_e(\omega_m) \propto \left\{1 - (\sin 2\alpha)^2 \cdot [\sin(\pi f_m \Delta\tau)]^2\right\}$

It is clearly seen that the RF output power fading is dependent on the angle between the state of polarization and the PMD vector. Since the PMD monitoring technique is based on SSB, it is independent of the CD effect.

Fig. 4.11. Power fading due to different values of chromatic dispersion (RFSA2 – RFSA3) vs. the angle between input SOP and PMD vector.

Similarly, the power fading for different CD values and different angles between the state of polarization and PMD vector are shown in Fig. 4.11. The measured power fading results agree well with the theoretical analysis (continuous line curves based on Eq. (4.39)). Fig.4.12 shows the CD induced power fading, with DGD as a parameter.

The continuous line curves are calculated using Eq. (4.38) for the relative output power in the presence of both PMD and CD. The results are measured with 0, 20, 33, 39, and 44 ps DGD, and 10 km, 13 km, 20 km, 23 and 30 km long SMF to provide different values of CD. The calculated results match well with measured results, which verifies that the effect of CD is independent of the value of DGD. Together, Fig. 4.11 and Fig. 4.12 show that in this monitoring technique CD induced power fading is independent of the DGD and the angle between SOP and PMD vector.

Fig. 4.12. CD induced RF power fading (RFSA3) for different DGD values when

$$2\alpha = \pi/2$$

It should be pointed out that the RF spectrum analyzers shown in Fig.4.1, Fig.4.3 and Fig.4.8 can be replaced by electronic bandpass filters (BPF) followed by RF power meter in real systems; then the cost will be highly reduced. Further, the input SOP is not known in real systems. There are two ways of getting over this problem. One is to use an electronic polarization controller to find the maximum power fading; this can ensure that $2\alpha = \pi/2$, regardless of the input SOP. Another relatively simple

way is to add a polarization scrambler in the system; this enables measurement of the average value of RF power fading, which is again independent of the input SOP.

4.4 A New Method for the Measurement of DGD for chirped Fiber Gratings

From Eq. (4.38) in the previous section, it can be observed that the RF power fading due to PMD depends on the angle between the input state of polarization and the PMD vector. Based on this observation, we propose a novel method to measure the DGD between the fast and slow axes of a Hi-Bi chirped grating. Further, we show that a similar method can also measure the DGD of a chirped grating fabricated using a single mode fiber. Measured values of DGD are important in assessing the extent of PMD in an optical communication system using such gratings.

From Eq. (4.38), for $2\alpha = 0$, one gets

$$P_e(\omega_m) \propto \left[\cos(\beta_2 L \omega_m^2 / 2) \right]^2 \quad (4.50)$$

and for $2\alpha = \pi/2$, one gets

$$P_e(\omega_m) \propto \left[\cos(\beta_2 L \omega_m^2 / 2) \right]^2 \left\{ 1 - \left[\sin(\Delta\tau \omega_m / 2) \right]^2 \right\} \quad (4.51)$$

Thus, while there is no effect of DGD for $2\alpha = 0$, there is maximum effect of DGD for $2\alpha = \pi/2$. Comparison of Eqs. (4.50) and (4.51) shows that one can determine the value of $\Delta\tau$, i.e., DGD, by measuring the ratio of the RF power for these two cases.

As mentioned in section 3.2, a Hi-Bi chirped grating actually is two almost identical gratings for the two orthogonal polarization axes: fast axis and slow axis. The wavelength difference for the two gratings is determined by the difference of effective

refractive index between the fast and slow axes or the beat length of the PM fiber. In general, a signal can be decomposed into two parts: each part polarised along one of the two orthogonal polarization axes. Because of the index difference along the two axes, the two parts will exhibit DGD. In a similar manner, even for a chirped grating based on a single mode fiber, whether intrinsic or due to the fabrication of a chirped grating, the birefringence in the single mode fiber induces PMD [116]. A chirped grating has been reported to have a 28 ps DGD with a birefringence of 10^{-5} [116].

The experimental setup used for the measurement of DGD for a Hi-Bi chirped grating is shown in Fig. 4.13. A 1.5 GHz radio frequency (RF) signal drives an electro-optic modulator (EOM) with a Hi-Bi fiber pigtail, which modulates the output of a laser with a 300 kHz line-width. The output of the EOM is launched into a $\lambda/4$ wave-plate that is fixed in a fiber bench, and then coupled through a polarization maintaining circulator into the Hi-Bi chirped grating under test. The reflected signal from the Hi-Bi chirped grating is fed to a 10 GHz photo-detector, which feeds an RF spectrum analyzer.

Fig. 4.13. Experimental setup for measuring DGD of a Hi-Bi chirped grating

Fig. 4.14. Measured DGD values for a Hi-Bi chirped FBG; dashed line represents the results measured using Agilent 86038A optical dispersion analyzer

By measuring the ratio of the RF powers for $2\alpha = 0$ and $2\alpha = \pi/2$, corresponding to Eqs. (4.50) and (4.51), one easily gets the DGD values at different optical wavelengths. The results for the DGD values are shown in Fig. 4.14; the dashed line in the Fig.4.14 shows the DGD values measured by Agilent 86038A optical dispersion analyzer. For most of the optical wavelengths, the match between the two sets of DGD values is good. The maximum difference between the two results is about 15% but the typical difference is less than 8%. There appear three possible ways to improve the accuracy. One is to optimize the RF frequency to improve the sensitivity of our method. The second one is to use a smoothing filter, when using the Agilent 86036A, with a higher smoothing parameter so as to reduce the ripples in the measured results. Finally, one can measure more points to smoothen the curve obtained using our method.

Using a similar method, one can also measure the DGD for a single mode fiber chirped grating. The experimental setup is shown in Fig. 4.15.

Fig. 4.15. Experimental setup for measurement DGD of a chirped grating

The operating wavelength for the laser is 1564.4 nm, which is the centre wavelength of the chirped FBG. The frequency of the RF signal used is 12 GHz. The maximum and minimum RF powers, corresponding to $2\alpha = 0$ and $2\alpha = \pi/2$, respectively, measured using the RF spectrum analyzer, are -40.65 dBm, and -43.69 dBm respectively. So the power fading due to DGD of the linearly chirped grating is about 3.04 dB. Therefore, the DGD value for the chirped FBG is calculated to be about 20.92 ps.

4.5 Conclusion

In this chapter, we derive expressions for power fading of RF signals due to higher-order PMD including PCD and chromatic dispersion. These expressions explicitly include the angle of precession of the output SOP around the PMD vector. Based on these expressions, a simple and novel pilot-tone-based technique has been proposed for simultaneous monitoring of chromatic dispersion and PMD. Experimental results demonstrate the effectiveness of the proposed techniques; these results also agree very well with the theoretical analysis.

Moreover, based on the power fading expression, we suggest a novel method to measure the DGD between the fast and slow axes of a Hi-Bi chirped grating. Further, a similar method can measure the PMD of a chirped grating easily.

Chapter 5

Tunable Photonic Microwave Filters Using DGD Elements

In general, polarization mode dispersion (PMD) induces impairment in an optical communication system as discussed in [6][59]. PMD is caused by differential group delay (DGD) for a signal in two principal states of polarization (PSP) in an optical fiber. However, PMD can also be useful in another field, such as photonic microwave filters. The ability to process microwave and millimeter-wave signals directly within the optical fiber has attracted significant interest. In this chapter, we propose and experimentally demonstrate two novel and simple photonic microwave filters based on DGD elements. The first one uses a Hi-Bi fiber, together with a tunable DGD element. The second one is based on a variable polarization beam splitter (PBS), a Hi-Bi coupler, and a variable time delay line. Both configurations are free from the problems of optically coherent interference and chromatic dispersion. Application of one of these filters is also demonstrated for harmonic suppression in a radio-over-fiber system.

5.1 Introduction

The ability to process microwave and millimeter-wave signals directly within the optical fiber has attracted significant interest. All-optical processing of microwave and millimeter-wave signals provides advantages such as large time-bandwidth products, insensitivity to electromagnetic interference, and lightweight. A very useful

component in radio frequency (RF) systems is a high-resolution tunable notch filter [117]. Further, a number of different types of photonic microwave filters have been reported in the literature [117-129]. The simple and most cost-effective technique for realizing photonic signal processor filters is the optical delay structure [117-120]. To overcome the optical coherence problem, in this application, the coherence time of the light source is kept smaller than the minimum delay time of the filter [117-120], or optical beams are detected and combined electrically to avoid interference problems associated with the optical coherence of the laser [121-123]. To avoid optical coherence problem, a laser array is also used [124-129]; this leads to rather complicated systems. Incoherent summing has also been achieved by using a Hi-Bi fiber [130]; however, such a filter cannot be tuned easily and needs a long Hi-Bi fiber for small free spectral range (FSR) values. Photonic microwave filters are periodic in spectrum since they sample the input signal at a time rate given by τ . Thus the spectral period or FSR is given by $1/\tau$. A step tunable design with a tunable laser and uniform Hi-Bi gratings is reported in [131]. A continuously tunable design using a Hi-Bi chirped grating has been reported in [132]; unfortunately, the grating causes a very large chromatic dispersion, which will affect the system performance; the operating wavelength range is also limited due to the FBG bandwidth; moreover, phase noise will be introduced by the crosstalk when the chirped grating is tuned.

In conventional intensity modulation systems, the optical carrier is modulated to generate an optical field with carrier and sidebands. In the presence of chromatic dispersion (CD), the two sidebands differ in phase, leading to power fading at the photodetector; the chromatic dispersion will make the magnitude of frequency response of photonic microwave filters uneven. For example, a 3 dB power

degradation can occur for a fiber length of 6 km for a 20GHz signal [133-134]. So it is important to overcome the effect of CD in photonic microwave filters.

In this chapter, first we propose a novel configuration for a continuously tunable photonic microwave notch filter, which is free from chromatic dispersion. The filter uses a combination of a Hi-Bi fiber and a variable differential group delay (DGD) element. Two orthogonally polarized light beams are generated in this scheme. The beams travel along the fast and slow axes of the Hi-Bi fiber that induces a fixed DGD; next, the two beams pass through the adjustable DGD element, acquiring differential time delay, which is tunable. Direct detection of the output optical power gives incoherent summing of the two optical beams. The proposed system can operate over a wide range of optical wavelengths. Measured results presented here agree well with the theory. The use of the DGD element and Hi-Bi fiber results in an output that is not sensitive to environment.

Next, we also demonstrate a tunable microwave-photonic filter based on a variable polarization beam splitter (PBS), a 3 dB Hi-Bi coupler, and a variable time delay line. This configuration also is free from the problems of optical coherence interference and chromatic dispersion. A notch rejection greater than 35 dB is obtained. It is shown that the notch frequency is continuously tunable by tuning the time delay line. The scheme can operate over a wide range of optical carrier frequencies, restricted only by the bandwidth of the various optical components in the system.

5.2 Novel tunable notch filter using a Hi-Bi fiber and a DGD element

In this section, we propose and experimentally demonstrate a novel and simple photonic microwave notch filter using a Hi-Bi fiber that gives a fixed differential

group delay (DGD), together with a DGD element that gives a tunable DGD. This configuration overcomes the problems of optical coherence interference and chromatic dispersion, which may occur in schemes using fiber delay lines or fiber gratings. Also presented is theoretical analysis for the performance of the microwave filter using the present configuration. There is good agreement between experimental results and theoretical analysis [135].

5.2.1 System configuration and principle of operation

The filter configuration using Hi-Bi fiber and DGD element is shown in Fig. 5.1. The radio frequency (RF) signal (40 MHz to 40 GHz) drives an electro-optic modulator (EOM) with a Hi-Bi fiber pigtail, which modulates the output of a laser with a 100 kHz line-width. The output of the EOM is launched into a $\lambda/4$ waveplate that is fixed in a fiber bench, and then coupled into the Hi-Bi fiber. The output of the

Fig. 5.1. Filter configuration with Hi-Bi fiber and tunable DGD element

Hi-Bi fiber is fed to a polarization controller to align the principal axes of the Hi-Bi fiber with those of the electronically adjustable differential group delay (DGD)

element. If the tunable DGD element has a Hi-Bi fiber pigtail, then the polarization controller in Fig.5.1 is not required. The adjustable DGD element, based on a crystal, operates over a wavelength range of 1550 ± 50 nm. The DGD value can be changed using a parallel 6-bit transistor-transistor logic (TTL) control command, with a resolution of 1.36 ps. The DGD value can range between 0 and 45 ps. Due to the birefringence of the Hi-Bi fiber and the DGD element, light propagating in the two orthogonal states of polarization will undergo differential time delay. The total DGD consists of two parts: one is the fixed DGD induced by the Hi-Bi fiber, another is the tunable DGD due to the tunable DGD element. The output of the DGD element passes to the 45 Gb/s photodetector and the vector network analyzer. The two orthogonal parts of the signal combine in the photo-detector; this leads to the filter action at RF. By tuning the DGD element, the FSR of the microwave filter can be tuned. Moreover, the two orthogonally polarized signals can overcome the interference problems of the laser source [130-132], so that one does not need multiple photodetectors to overcome the coherent interference problems of the laser source [122-123].

5.2.2 Theoretical analysis

Depending on the polarisation state at the output of the $\lambda/4$ waveplate, the relative magnitude of the signal excited in the two orthogonal polarizations along the principal axes in the Hi-Bi fiber will vary. Let us assume that the polarisation at the output of the $\lambda/4$ waveplate is linear, and it makes an angle α with the direction $\hat{\Omega}_+$, where the directions of the principal axes of the Hi-Bi fiber are denoted as $\hat{\Omega}_+$ and $\hat{\Omega}_-$. When α is $\pi/4$, the two orthogonal polarizations are equally excited, leading to a maximum notch depth in the filter response [130].

The following analysis is similar to that in [126]. However, in our case, multiple signal components are generated due to orthogonal polarization and not due to multiple optical sources. Let the input RF voltage to the external modulator be represented as

$$V(t) = 1 + m \cos \omega_m t \quad (5.1)$$

Considering a linear intensity modulation, the output optical power of the modulator can be expressed as:

$$P_i(t) = P_i(1 + m \cos \omega_m t) \quad (5.2)$$

where P_i is the optical power output from the laser, m is the modulation index, which equals the peak intensity change divided by the average intensity [136], and ω_m is the modulating angular frequency. The corresponding electric field of the light wave can be expressed as:

$$f(t) = \sqrt{P_i} \sqrt{1 + m \cos(\omega_m t)} \cos(\omega_0 t) \quad (5.3)$$

where ω_0 is the angular frequency of the optical carrier signal. It is assumed that the modulated has a Hi-Bi pigtail and the polarization of the output light from the laser is optimally aligned to the modulator. Under small signal condition ($m \ll 1$), Eq. (5.3) can be approximated as:

$$\begin{aligned} f(t) &= \sqrt{P_i} \left(1 + \frac{1}{2} m \cos \omega_m t \right) \cos \omega_0 t \\ &= \sqrt{P_i} \left[\cos \omega_0 t + \frac{m}{4} \cos(\omega_0 + \omega_m)t + \frac{m}{4} \cos(\omega_0 - \omega_m)t \right] \end{aligned} \quad (5.4)$$

From Eq. (5.4), it is obvious that, under small signal condition, an intensity-modulated signal can be approximated to a typical amplitude modulated signal with double sidebands separated by $2\omega_m$. Next, consider propagation of the optical signal $f(t)$ on the fiber. The signal has its Fourier transform:

$$F(\omega - \omega_0) = \int_{-\infty}^{+\infty} f(t) \exp[-j(\omega - \omega_0)t] dt \quad (5.5)$$

The signal, after propagation, can be decomposed into two components, $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$, one each along the two orthogonal axes, $\hat{\Omega}_+$ and $\hat{\Omega}_-$. Also, the propagation constant $\beta(\omega)$ can be expanded as a Taylor series around the carrier frequency ω_0 as

$$\beta(\omega) = \beta_0 + \beta_1(\omega - \omega_0) + \frac{1}{2}\beta_2(\omega - \omega_0)^2 + \dots \quad (5.6)$$

where

$$\beta_m = \left(\frac{\partial^m \beta}{\partial \omega^m} \right) \bigg|_{\omega=\omega_0} \quad (5.7)$$

Using the fact that the two orthogonal components differ in phase by $\Delta\tau\Delta\omega$ [11], the components $f_{\Omega+}(t)$ and $f_{\Omega-}(t)$ are written as:

$$f_{\Omega+}(t) = M \frac{\cos \alpha}{2\pi} \int_{-\infty}^{+\infty} \left\{ F(\omega - \omega_0) \exp\left(j \frac{\Delta\tau\Delta\omega}{2}\right) \times \exp\left(-j\beta_1 L \Delta\omega - j \frac{\beta_2 \Delta\omega^2 L}{2}\right) e^{j(\omega - \omega_0)t} d\omega \right\} \quad (5.8)$$

$$f_{\Omega-}(t) = M \frac{\sin \alpha}{2\pi} \int_{-\infty}^{+\infty} \left\{ F(\omega - \omega_0) \exp\left(-j \frac{\Delta\tau\Delta\omega}{2}\right) \times \exp\left(-j\beta_1 L \Delta\omega - j \frac{\beta_2 \Delta\omega^2 L}{2}\right) e^{j(\omega - \omega_0)t} d\omega \right\} \quad (5.9)$$

where M is signal amplitude factor; $\Delta\tau$ is the difference of the time delays between the two orthogonally polarized light beams and $\Delta\omega$ corresponds to ω_m . Combining the components, one gets

$$\vec{f}(t) = f_{\Omega+}(t) \hat{\Omega}_+ + f_{\Omega-}(t) \hat{\Omega}_- \quad (5.10)$$

Eqs. (5.8)-(5.10) are general equations, applicable to an optical signal of any shape.

Because the Hi-Bi fiber length in our configuration is less than 50m, the chromatic dispersion of the Hi-Bi fiber can be neglected. Further, to get a high rejection for the notch frequency, as mentioned earlier, α is $\pi/4$. In this situation, the two orthogonal components are equally excited, and one gets:

$$f_{\Omega_+}(t) = \frac{\sqrt{2}}{2} M \sqrt{P_i} \left[\begin{aligned} &\cos(\omega_0 t) + \frac{m}{4} \cos[\omega_0 t + \omega_m(t + \Delta\tau/2 - \beta_1 L)] \\ &+ \frac{m}{4} \cos[\omega_0 t - \omega_m(t + \Delta\tau/2 - \beta_1 L)] \end{aligned} \right] \quad (5.11)$$

$$f_{\Omega_-}(t) = \frac{\sqrt{2}}{2} M \sqrt{P_i} \left[\begin{aligned} &\cos(\omega_0 t) + \frac{m}{4} \cos[\omega_0 t + \omega_m(t - \Delta\tau/2 - \beta_1 L)] \\ &+ \frac{m}{4} \cos[\omega_0 t - \omega_m(t - \Delta\tau/2 - \beta_1 L)] \end{aligned} \right] \quad (5.12)$$

Since $\hat{\Omega}_+$ and $\hat{\Omega}_-$ are orthogonal, the optical signal power at the detector is

$$P(t) = |\vec{f}(t)|^2 = \vec{f}(t) \cdot \vec{f}^*(t) = |f_{\Omega_+}(t)|^2 + |f_{\Omega_-}(t)|^2 \quad (5.13)$$

Eq. (5.13) shows that the optical power incident on the photodetector consists of two components, which have orthogonal states of polarization. So, one gets rid of the coherence problem induced by the laser source. By assuming the responsivity of the photodiode to be R , we get the RF output current as:

$$I(t) = \frac{1}{2} R M^2 P_i \cdot \left\{ \begin{aligned} &\frac{1}{2} + \frac{m}{2} \cos[\omega_m(t + \Delta\tau/2 - \beta_1 L)] \\ &+ \frac{m^2}{8} [\cos(\omega_m(t + \Delta\tau/2 - \beta_1 L))]^2 \end{aligned} \right\} \\ + \frac{1}{2} R M^2 P_i \cdot \left\{ \begin{aligned} &\frac{1}{2} + \frac{m}{2} \cos[\omega_m(t - \Delta\tau/2 - \beta_1 L)] \\ &+ \frac{m^2}{8} [\cos(\omega_m(t - \Delta\tau/2 - \beta_1 L))]^2 \end{aligned} \right\} \quad (5.14)$$

Under small signal conditions, the m^2 term in Eq. (5.14) can be neglected. Next, comparing the RF output current in Eq. (5.14) with the RF input voltage in Eq. (5.1), the transfer function for the microwave filter can be obtained as:

$$|H(\omega_m)| = \frac{R}{4} M^2 \left| \exp\left(\frac{j\Delta\tau\omega_m}{2}\right) + \exp\left(-\frac{j\Delta\tau\omega_m}{2}\right) \right| = \frac{R}{2} M^2 |\cos(\Delta\tau\omega_m/2)| \quad (5.15)$$

From Eq. (5.15), the notch frequencies are given by:

$$f = \left(k + \frac{1}{2}\right) \frac{1}{\Delta\tau}, \quad k = 0, 1, 2, \dots \quad (5.16)$$

where $\Delta\tau$, the difference of the time delays between the two orthogonally polarized light beams, consists of two parts: one part is the fixed DGD induced by the Hi-Bi fiber and the other part is the variable DGD of the tunable DGD element. When one tunes the DGD element, the FSR will vary according to Eq. (5.16).

5.2.3 Experimental results and discussion

First, we use only the Hi-Bi fiber. When the polarization of the output light from the $\lambda/4$ wave -plate is so aligned that it equally excites the two orthogonal polarizations in the Hi-Bi fiber, maximum notch depth occurs. This is shown in Fig. 5.2. The measured DGD value for the Hi-Bi fiber is 66.8 ps. The measured FSR is 14.785 GHz; this agrees well with Eq. (5.16). Next, we put an adjustable DGD element following the Hi-Bi fiber with a polarization controller to make the fast and slow axes of Hi-Bi fiber align with the two principal axes of the DGD element. In this case, the FSR of the filter can be continuously tuned by changing the DGD values of the DGD element. Fig. 5.3 demonstrates filter tunability achieved in this manner. As we change the DGD value of the DGD element to 30 ps, the curve in Fig. 5.3 shows that the FSR reduces to 10.435 GHz. This result also agrees well with Eq. (5.16).

Fig. 5.2. Measured frequency response of the filter for Hi-Bi fiber alone

Fig. 5.3. Measured frequency response of the filter with Hi-Bi fiber and the DGD element

A notch rejection greater than 40 dB is obtained; the stability of the results demonstrates that the filter overcomes the limitation of optical coherence of the laser. Besides, the proposed filter is free from the problem of chromatic dispersion. Fig. 5. 4 presents the variation in FSR as the DGD value obtained from the adjustable DGD element varies from 0 to 45 ps. The figure also includes the FSR values calculated using Eq. (5.16). The experimental results agree well with the calculated values.

Fig. 5.4. Comparison between measured (triangle dots) and calculated (solid line) FSR values, as a function of DGD values

5.2.4 Application for harmonic suppression in a radio-over-fiber (RoF) system

The experimental implementation for the filter described in Fig.5.1 in an RoF system is shown in Fig.5.5. A 1550 nm laser source (ANDO AQ8201-3A) with a narrow

linewidth of 5 MHz is employed. Light is applied to the intensity modulator with Hi-Bi pigtails in both input and output port. The filter shown in Fig.5.1 is connected to

Fig. 5.5. Experimental setup

the output of the intensity modulator driven by an RF signal (2.885 GHz). The filter

Fig. 5.6. Measured normalized filter response

output is applied to a 20 km standard single mode fiber which has a chromatic dispersion of 17 ps/(km.nm) at 1550 nm. After photodetection, the signal is observed on an RF spectrum analyser (RFSA). It should be noted that without the filter, due to

the chromatic dispersion of the fiber, the second harmonic of the modulated RF signal will be generated.

The measured normalized filter response with an FSR of 11.54 GHz is shown in Fig. 5.6, with the first notch filter rejection frequency at 5.77 GHz. The measured output RF signal without the filter is shown in Fig.5.7. It clearly shows that the intensity modulation by the RF signal generates the second-order harmonic at 5.77 GHz in the RoF system due to the chromatic dispersion of the RoF link (single mode fiber). However, as shown in Fig.5.8, the second-order harmonic can be suppressed significantly by using the microwave filter.

Fig. 5.7. Measured RFSA output power without filter

Fig. 5.8. Measured RFSA output power with filter

5.3 Novel tunable notch filter using a variable polarization beam splitter and a Hi-Bi coupler

In this section, we describe a tunable microwave-photonic filter based on a variable polarization beam splitter (PBS), a 3 dB Hi-Bi coupler, and a variable time delay line. The configuration is free from the problems of optical coherence interference and chromatic dispersion. An expression for the filter transfer function is derived and measured results are presented to demonstrate the filter operation. A notch rejection greater than 35 dB is obtained. It is shown that the notch frequency is continuously tunable by tuning the time delay line. The scheme can operate over a wide range of optical carrier frequencies, restricted only by the bandwidth of the various optical components in the system [137].

5.3.1 System configuration and principle of operation

The filter configuration is shown in Fig. 5.9. An RF signal drives an electro-optic modulator (EOM), which modulates the output of a 1550 nm DFB laser with 100kHz

Fig. 5.9. Experimental setup

linewidth. The output of the EOM, which has a Hi-Bi pigtail, is launched into a variable polarization beam splitter (PBS), which also has Hi-Bi Pigtails. A quarter-wave plate is inserted in port 1 of the PBS; this enables one to change the power-split ratio in the two output arms of the PBS that have orthogonal states of polarization. One of the output arms of the PBS incorporates a time delay line with a Hi-Bi pigtail and is connected to a 3 dB Hi-Bi coupler through a fiber of length L_1 . The time delay line can provide a large tunable time delay. The other output arm of the PBS is connected to the 3dB Hi-Bi coupler through a fiber of length L_2 . The time delay line (OZ Optics Ltd.) has an insertion loss of about 2.5 dB. The quarter-wave plate and the variable PBS can ensure that equal optical power is incident on the input ports of the coupler; this leads to a strong notch in the filter response. The Hi-Bi coupler combines the two input signals that are orthogonally polarised. One output port of the Hi-Bi

coupler is connected to a photo-detector and a vector network analyser. The other output port of the Hi-Bi coupler is connected to an optical spectrum analyser (OSA). While the OSA is not essential to the operation of the filter, it helps in monitoring incoherent operation of the filter. The differential time delay between the two branches of the resulting filter consists of two parts: one caused by the length difference between L_1 and L_2 , and the other caused by the time delay $\Delta\tau_d$ provided by the time delay line.

Following an analysis similar to that in section 5.2.2, under small signal condition ($m \ll 1$), the output signal of EOM can be approximated as:

$$\begin{aligned} f(t) &= \sqrt{P_i} \left(1 + \frac{1}{2} m \cos \omega_m t \right) \cos \omega_0 t \\ &= \sqrt{P_i} \left[\cos \omega_0 t + \frac{m}{4} \cos(\omega_0 + \omega_m)t + \frac{m}{4} \cos(\omega_0 - \omega_m)t \right] \end{aligned} \quad (5.17)$$

where m is the modulation index, ω_m is the modulation frequency. The Fourier transform of the signal is written as:

$$F(\omega - \omega_0) = \int_{-\infty}^{+\infty} f(t) \exp[-j(\omega - \omega_0)t] dt \quad (5.18)$$

Assuming equal power in the two output arms of the PBS, the signals in the two arms can be expressed as:

$$f_{\Omega+}(t) = M\sqrt{P_i} \left\{ \begin{aligned} &\cos(\omega_0 t) + \frac{m}{4} \cos[\omega_0 t + \omega_m(t - \beta_1 L_1 - \Delta\tau_d)] \\ &+ \frac{m}{4} \cos[\omega_0 t - \omega_m(t + \beta_1 L_1 + \Delta\tau_d)] \end{aligned} \right\} \quad (5.19)$$

$$f_{\Omega-}(t) = M\sqrt{P_i} \left\{ \begin{aligned} &\cos(\omega_0 t) + \frac{m}{4} \cos[\omega_0 t + \omega_m(t - \beta_1 L_2)] \\ &+ \frac{m}{4} \cos[\omega_0 t - \omega_m(t + \beta_1 L_2)] \end{aligned} \right\} \quad (5.20)$$

where $\Delta\tau_d$ is the time delay introduced by the time delay line and $\beta_1 = \left. \frac{d\beta}{d\omega} \right|_{\omega=\omega_0}$.

Since the signals in the two arms are orthogonally polarized, the optical signal power at the detector is

$$P(t) = |\vec{f}(t)|^2 = \vec{f}(t) \cdot \vec{f}^*(t) = |f_{\Omega+}(t)|^2 + |f_{\Omega-}(t)|^2 \quad (5.21)$$

Eq. (5.21) shows that the optical power incident on the photodetector consists of two components with orthogonal states of polarization, thus getting rid of the coherence problem induced by the laser source. The two orthogonal components are summed up in the photodetector incoherently. Therefore, any fluctuations in optical phase due to environmental perturbations do not affect the stability of the filter operation. This can be verified by monitoring the optical power with the help of an optical spectrum analyzer (OSA).

By assuming the responsivity of the photodiode to be R , we get the RF output current as:

$$\begin{aligned} I(t) = & RM^2 P_i \cdot \left\{ \frac{1}{2} + \frac{m}{2} \cos[\omega_m(t - \beta_1 L_1 - \Delta\tau_d)] \right. \\ & \left. + \frac{m^2}{8} [\cos(\omega_m(t - \beta_1 L_1 - \Delta\tau_d))]^2 \right\} \\ & + RM^2 P_i \cdot \left\{ \frac{1}{2} + \frac{m}{2} \cos[\omega_m(t - \beta_1 L_2)] \right. \\ & \left. + \frac{m^2}{8} [\cos(\omega_m(t - \beta_1 L_2))]^2 \right\} \end{aligned} \quad (5.22)$$

Under small signal conditions, the m^2 terms in Eq. (5.22) can be neglected. Next, comparing the RF output current in Eq. (5.22) with the RF input voltage in Eq. (5.17) one gets the transfer function of the microwave filter:

$$|H(\omega_m)| = RM^2 \left| \exp[-j(\beta_1 L_1 + \Delta\tau_d)\omega_m] + \exp[-j(\beta_1 L_2)\omega_m] \right| \quad (5.23)$$

As explained earlier, the time delay between the two arms of the PBS is

$$\Delta\tau = \Delta\tau_0 + \Delta\tau_d \quad (5.24)$$

where $\Delta\tau_0 = \beta_1 L_1 - \beta_1 L_2$ is the fixed time delay. Further, the notch frequencies are given by:

$$f = \left(k + \frac{1}{2}\right) \frac{1}{\Delta\tau}, \quad k = 0, 1, 2, \dots \quad (5.25)$$

When one changes the value of time delay $\Delta\tau_d$, the FSR varies according to Eq. (5.25).

5.3.2 Experimental results and discussion

Case 1: The length difference between L_1 and L_2 is about 2.347 m and $\Delta\tau_d$ is zero.

So the fixed time delay value $\beta_1 L_1 - \beta_1 L_2$ is about 11735 ps. As shown in Fig. 5.10, a

Fig. 5.10. Measured and calculated filter response for Case 1 ($\Delta\tau_d = 0$)

filter with an FSR of 0.08526 GHz and notch depths more than 35 dB is obtained. Fig.5.10. also includes the results calculated using Eq. (5.23). The calculated and measured results match very well. Moreover, the optical power monitored by the OSA that is connected to the other output port of the 3 dB coupler is very stable, which means that the operation is free from the problem of optical coherence.

Fig. 5.11. Measured and calculated notch filter response for Case 2 ($\Delta\tau_d = 110$ ps)

Case 2: L_1 and L_2 same as in Case 1; $\Delta\tau_d$ about 110 ps. Fig. 5.11 shows the measured and calculated results. The measured FSR is about 0.08375 GHz. The difference between the measured and calculated FSR is less than 4%.

Case 3: The length difference between L_1 and L_2 is about (2.347-1.6) m and $\Delta\tau_d$ is zero. So the calculated fixed time delay $\beta_1 L_1 - \beta_1 L_2$ reduces to 3735 ps. As shown in Fig. 5.12, a filter with an FSR of 0.269 GHz is obtained. The measured results match very well the calculated results.

Fig. 5.12. Measured and calculated notch filter response for Case 3 ($\Delta\tau_d = 0$)

Fig. 5.13. Measured and calculated notch filter response for Case 4 ($\Delta\tau_d = 310$ ps)

Case 4: L_1 and L_2 same as in Case 3; $\Delta\tau_d$ about 310 ps. As shown in Fig. 5.13, the measured FSR is about 0.2473 GHz. The measured and calculated results match very well, once again.

One of the advantages for the design shown in Fig. 5.9 is that it can easily give small FSR values using relatively short fiber lengths. For comparison, consider the design based on Hi-Bi fiber [130], as shown in Fig. 5.14.

Fig. 5.14. Experimental setup for long PM fiber

The radio frequency (RF) signal (40 MHz to 40 GHz) drives an electro-optic modulator (EOM) with Hi-Bi fiber pigtail, which modulates the output of a laser with a 100 kHz linewidth. The output of the EOM is launched into a $\lambda/4$ wave-plate that is fixed in a fiber bench, and then coupled into the 490 m Panda Hi-Bi fiber. The output of the Hi-Bi fiber is fed to the 10 Gb/s photodetector and the vector network analyzer. The measured FSR as shown in Fig. 5.15, is 1.524 GHz.

Fig. 5.15. Measured FSR for 490 m PM Panda fiber

Compared with Fig. 5.10 and Fig. 5.12, Fig. 5.15 shows that it is difficult to realize small FSR values with birefringence elements such as Hi-Bi fiber or the design discussed in section 5.2. The present scheme can yield small FSR values without needing long fiber lengths and can operate over a wide range of optical carrier frequencies.

5.4 Conclusion

We propose two continuously tunable photonic microwave notch filter configurations, which are free from chromatic dispersion. The first uses a combination of a Hi-Bi fiber and a variable differential group delay (DGD) element; the other is based on a variable polarization beam splitter (PBS), a 3 dB Hi-Bi coupler, and a variable time delay line. Two orthogonally polarized light beams are generated in these schemes; hence these are free from the problem of optical coherence interference and the

associated phase noise problem. Expressions for the filters transfer function are derived and measured results are presented to demonstrate the operation of the filters. Measured results agree well with the theory. These configurations are also free from the problem of chromatic dispersion. The two schemes can operate over a wide range of optical carrier frequencies, restricted only by the bandwidth of the various optical components in the system. Further, these structures can be easily incorporated in optical fiber microwave transmission systems. As an example, application of one of these filters is demonstrated for harmonic suppression in a radio-over-fiber system.

Chapter 6

Conclusions and Further Work

In this chapter, conclusions based on the work presented in this thesis are drawn. In addition, topics for further study of polarization mode dispersion (PMD) and microwave photonics have also been suggested.

6.1 Conclusions

PMD and chromatic dispersion (CD) have emerged as two serious problems that put a limit on achievement of high bit rates (i.e ≥ 10 Gbit/s) in long-distance optical fiber transmission systems. To overcome the limit in transmission capacity imposed by PMD and CD, several critical issues have been explored. These are: (i) effect of polarization mode dispersion and chromatic dispersion on pulse signals; (ii) polarization mode dispersion compensation using high-birefringence (Hi-Bi) linearly chirped fiber Bragg grating (FBG); (iii) power fading due to PMD and CD for radio frequency (RF) signals and its application for PMD and CD monitoring. Furthermore, PMD can also be useful, for example in photonics microwave filters. New configurations for microwave photonic filters using differential group delay (DGD) elements have been proposed in this thesis.

In this thesis, a new approach has been proposed for the description of PMD and its effects on pulse and continuous RF signals. This approach clearly shows the behaviour of the state-of-polarization (SOP) and the PMD vector. It has been

demonstrated that this approach permits an easy calculation of the effect of PMD and CD on pulse and continuous RF signals.

First, the thesis presents an expression relating the frequency dependent output state of polarization, $\vec{S}_o(\omega)$, and the PMD vector $\vec{\Omega}$, including polarization-dependent chromatic dispersion (PCD). This relation is consistent with the established behaviour of $\vec{S}_o(\omega)$ with respect to $\vec{\Omega}$, and clearly describes the details of the behaviour of SOP with respect to the PMD vector including the effect of PCD. Based on this expression, we get the phase difference value between the two principal states of polarization (PSPs). Following this, we derive a general equation, which incorporates the precession angle, and the effects of DGD, PCD and chromatic dispersion, on a pulse of arbitrary shape. This new equation is used to obtain expressions for shape distortion, pulse broadening, and power penalty for the special case of Gaussian pulses. An expression for average power penalty has also been derived.

Next, the thesis presents an expression for the higher-order PMD vector including second-order effects, namely, depolarization and depolarization-dependent chromatic dispersion (DCD). Both of these are new terms and are explained in Section 2.4. Based on this expression, we derive the phase difference value between two orthogonal depolarization vectors. Further, the pulse broadening and power penalty induced by the higher-order PMD effects (including second-order PMD and DCD) and chromatic dispersion are also discussed.

These expressions for shape distortion and power penalty should be helpful in estimating the degradation due to PMD, including CD and higher-order PMD effects,

in high bit rate optical communication systems employing different types of pulse shapes.

We propose a new PMD compensation scheme using a Hi-Bi linearly chirped FBG in a double pass configuration, which is free from chromatic dispersion. Fabrication of a linearly chirped FBG is easier compared to that of a nonlinearly chirped FBG. In our implementation, the FBG is mounted on a cantilever beam in such a way that the bandwidth of the reflected signal can be tuned with little or a very small shift of the centre wavelength. This eliminates the possibility that the signal wavelength may lie at the edge of the effective reflection band of the grating; thus, our scheme is likely to be more suitable for PMD compensation in a multi-wavelength system. Measured results are presented to show that such a configuration achieves significantly reduced chromatic dispersion, together with PMD compensation.

Next, another new tunable polarization mode dispersion (PMD) compensator has been proposed, which is also substantially free from chromatic dispersion. The design uses a pair of high-birefringence linearly chirped fiber gratings. Each FBG can adjust the differential group delay (DGD) nonlinearly by using a cantilever beam. The compensator can be tuned to operate over a range of wavelengths, can yield DGD values starting from 0 ps and free from cross-talk. Measured eye-diagrams for a 10 Gbit/s system show that this compensator can greatly reduce the chromatic dispersion induced by chirped grating.

An expression for power fading of continuous RF signals, including higher-order PMD and chromatic dispersion, has been derived. This expression includes the angle of precession of the output state of polarization around the PMD vector. From the expression for power fading, we get the average power penalty for chromatic

dispersion and higher-order PMD. Moreover, a simple and novel pilot-tone-based technique for simultaneous monitoring of chromatic dispersion and PMD has been proposed. Experiment results demonstrate the effectiveness of the proposed technique; these results also agree very well with the theoretical analysis. Further, based on the power fading expression, we propose a novel method to measure the DGD between the fast and slow axes of Hi-Bi chirped gratings. A similar method can measure the PMD of chirped gratings easily.

A continuously tunable photonic microwave notch filter configuration has been proposed. The configuration is free from chromatic dispersion. The filter uses a combination of a Hi-Bi fiber and a variable DGD element. The proposed system can operate over a wide range of optical wavelengths. We also present a theoretical analysis for the transfer function of the microwave filter using the proposed configuration. Measured results presented here agree well with the theory. The use of the DGD element and Hi-Bi fiber results in an output that is not sensitive to environment. Moreover, we demonstrate a tunable microwave-photonic filter based on a variable polarization beam splitter (PBS), a 3 dB Hi-Bi coupler, and a variable time delay line. The configuration is free from the problems of optical coherence interference and chromatic dispersion. An expression for the filter transfer function is derived and measured results are presented to demonstrate the filter operation. A notch rejection greater than 35 dB is obtained. It is shown that the notch frequency is continuously tunable by tuning the time delay line. The scheme can operate over a wide range of optical carrier frequencies, restricted only by the bandwidth of the various optical components in the system.

6.2 Further work

In chapters 2 and 4, the effects of PMD and CD on pulse and continuous RF signals have been discussed in the context of optical fiber communication. There is another promising technology to provide future interactive multimedia services; this technology is millimeter-wave radio access networks. The millimeter-wave frequency band offers large transmission bandwidth and also overcomes spectral congestion at lower frequencies. In addition, millimeter-wave radio systems enable efficient frequency reuse due to the limited propagation distances at these frequencies. Optical fiber with its low loss, large bandwidth, and immunity to electromagnetic interference characteristics serves as an ideal transmission medium for the distribution of the millimeter-wave wireless access systems with optical backbone. The radio signals are generated at a central location before distribution via fiber to a number of remote antenna base stations for wireless distribution. However, chromatic dispersion as well as PMD will induced power penalties in millimeter-wave transmission and fiber-wireless systems [138-140]. It should be useful to study the higher-order PMD and CD effects on millimeter and fiber-wireless systems.

Optical packet-switched networks will be very important for future ultra fast optical networks. In such packet-switched networks, direct optical signal processing such as optical packet storage will be required without electrical conversion. Optical buffers using parametric amplification in fiber and nonlinear optical loop mirror (NOLM) as an optical switch to erase the buffer have been reported in [141]. Also a dual loop optical buffer (DLOB) based on 3×3 collinear fiber coupler has also been reported in [142-143]. With the increasing number of circulations in such buffers, the PMD and CD will strongly affect the pulse waveform. It is important to study the effect of PMD

and CD in optical buffers. With the bit rate increasing (more than 10Gbit/s), the optical buffer based on fiber loop must also include CD and PMD compensation; otherwise we can not observe the pulse waveform after several circulations. The study of the effect of PMD and CD on the pulse shape in an optical buffer is important for PMD and CD compensation.

Even though PMD is often looked at as a property of a passive fiber, when the optical network becomes more agile with fast restoration and provisioning, and with the emergence of burst and packet switching, dependence of PMD on rapidly changing optical power in adjacent channels should be considered. In multi-channel long haul systems, high optical power may cause non-linear polarization rotation or cross-talk [144-146]. Therefore, it is important to study the effects of PMD and CD in high power multi-channel systems.

In chapter 5, photonic microwave notch filters have been discussed. These are two-taps filters. For greater design flexibility, multi-tap filters are more attractive. For this purpose, one can consider tunable multi-wavelength laser sources or a spectral-sliced broadband source instead of the single wavelength laser sources considered here.

Appendix I

Derivation of relation between output SOP and PMD vector (Eq. 2.2)

Based on Eq. (2.1) and letting $\vec{\Omega} = [\Omega_1, \Omega_2, \Omega_3]^T$ in Stokes notation, the cross product ($\vec{\Omega} \times$) can be regarded as an operator represented by a skew-symmetric matrix P [147]. Then,

$$P = \vec{\Omega} \times = \begin{bmatrix} 0 & -\Omega_3 & \Omega_2 \\ \Omega_3 & 0 & -\Omega_1 \\ -\Omega_2 & \Omega_1 & 0 \end{bmatrix} \quad (\text{A1.1})$$

Using Eq. (2.1) and Eq. (A1.1), we have a new equation $\frac{\partial \hat{S}_o(\omega)}{\partial \omega} = P \hat{S}_o(\omega)$, whose solution is

$$\hat{S}_o(\omega) = \exp\{P(\omega - \omega_0)\} \hat{S}_o(\omega_0) \quad (\text{A1.2})$$

where $\hat{S}_0 = S_o(\omega_0) = [s_{01}, s_{02}, s_{03}]^T$ and $\Delta\tau = |\vec{\Omega}|$, It is easy to obtain three eigenvalues of P which are $(0, i\Delta\tau, -i\Delta\tau)$. Furthermore, we have the matrix of eigenvectors

$$T = \begin{bmatrix} \Omega_3 & \Omega_1\Omega_3 - j\Omega_2\Delta\tau & \Omega_1\Omega_3 + j\Omega_2\Delta\tau \\ \Omega_2 & \Omega_1\Omega_2 + j\Omega_3\Delta\tau & \Omega_1\Omega_2 - j\Omega_3\Delta\tau \\ \Omega_1 & \Omega_1^2 - \Delta\tau^2 & \Omega_1^2 - \Delta\tau^2 \end{bmatrix} \quad (\text{A1.3})$$

Then,

$$\exp\{[P(\omega - \omega_0)]\} = T \begin{bmatrix} 1 & 0 & 0 \\ 0 & e^{j\Delta\tau(\omega - \omega_0)} & 0 \\ 0 & 0 & e^{-j\Delta\tau(\omega - \omega_0)} \end{bmatrix} T^{-1} \quad (\text{A1.4})$$

Eq. (A1.4) leads to

$$\exp\{P(\omega - \omega_0)\}$$

$$= \frac{1}{\Delta\tau^2} \begin{bmatrix} \Omega_1^2 + \frac{\Omega_1^2\Omega_3^2 + \Omega_2^2\Delta\tau^2}{\Delta\tau^2 - \Omega_3^2} \cos\Delta\tau\Delta\omega & \Omega_1\Omega_2(1 - \cos\Delta\tau\Delta\omega) - \Omega_3\Delta\tau\sin\Delta\tau\Delta\omega & \Omega_1\Omega_3(1 - \cos\Delta\tau\Delta\omega) + \Omega_2\Delta\tau\sin\Delta\tau\Delta\omega \\ \Omega_1\Omega_2(1 - \cos\Delta\tau\Delta\omega) + \Omega_3\Delta\tau\sin\Delta\tau\Delta\omega & \Omega_2^2 + \frac{\Omega_2^2\Omega_3^2 + \Omega_1^2\Delta\tau^2}{\Delta\tau^2 - \Omega_3^2} \cos\Delta\tau\Delta\omega & \Omega_2\Omega_3(1 - \cos\Delta\tau\Delta\omega) - \Omega_1\Delta\tau\sin\Delta\tau\Delta\omega \\ \Omega_1\Omega_3(1 - \cos\Delta\tau\Delta\omega) - \Omega_2\Delta\tau\sin\Delta\tau\Delta\omega & \Omega_2\Omega_3(1 - \cos\Delta\tau\Delta\omega) + \Omega_1\Delta\tau\sin\Delta\tau\Delta\omega & \Omega_3^2 + (\Delta\tau^2 - \Omega_3^2) \cos\Delta\tau\Delta\omega \end{bmatrix} \quad (\text{A1.5})$$

Then,

$$\hat{S}_o(\omega) \stackrel{\text{def}}{=} \begin{bmatrix} s_1(\omega) \\ s_2(\omega) \\ s_3(\omega) \end{bmatrix} =$$

$$= \begin{bmatrix} \frac{\Omega_1}{\Delta\tau^2}(\Omega_1s_{01} + \Omega_2s_{02} + \Omega_3s_{03}) + \left[s_{01} - \frac{\Omega_1}{\Delta\tau^2}(\Omega_1s_{01} + \Omega_2s_{02} + \Omega_3s_{03}) \right] \cos\Delta\tau\Delta\omega + \frac{1}{\Delta\tau}(\Omega_2s_{03} - \Omega_3s_{02}) \sin\Delta\tau\Delta\omega \\ \frac{\Omega_2}{\Delta\tau^2}(\Omega_1s_{01} + \Omega_2s_{02} + \Omega_3s_{03}) + \left[s_{02} - \frac{\Omega_2}{\Delta\tau^2}(\Omega_1s_{01} + \Omega_2s_{02} + \Omega_3s_{03}) \right] \cos\Delta\tau\Delta\omega + \frac{1}{\Delta\tau}(\Omega_3s_{01} - \Omega_1s_{03}) \sin\Delta\tau\Delta\omega \\ \frac{\Omega_3}{\Delta\tau^2}(\Omega_1s_{01} + \Omega_2s_{02} + \Omega_3s_{03}) + \left[s_{03} - \frac{\Omega_3}{\Delta\tau^2}(\Omega_1s_{01} + \Omega_2s_{02} + \Omega_3s_{03}) \right] \cos\Delta\tau\Delta\omega + \frac{1}{\Delta\tau}(\Omega_1s_{02} - \Omega_2s_{01}) \sin\Delta\tau\Delta\omega \end{bmatrix} \quad (\text{A1.6})$$

The above equation can be rearranged as

$$\hat{S}_o(\omega) = \hat{\Omega}(\hat{\Omega} \cdot \hat{S}_o(\omega_0)) + \left[\hat{S}_o(\omega_0) - \hat{\Omega}(\hat{\Omega} \cdot \hat{S}_o(\omega_0)) \right] \cos\Delta\tau\Delta\omega$$

$$+ \hat{\Omega} \times \hat{S}_o(\omega_0) \sin\Delta\tau\Delta\omega \quad (\text{A1.7})$$

Or

$$\hat{S}_o(\omega) = \hat{\Omega} \cos 2\alpha + \left[\hat{S}_o(\omega_0) - \hat{\Omega} \cos 2\alpha \right] \cos\Delta\tau\Delta\omega + \hat{B} \sin 2\alpha \sin\Delta\tau\Delta\omega \quad (\text{A1.8})$$

Eq. (A1.8) is the desired relation.

Appendix II

Expression for PMD vector including depolarization vector (Eq. 2.50)

From Eq. (2.49):

$$\frac{\partial \vec{\Omega}}{\partial \omega} = \left(\frac{\Delta \tau_1}{\Delta \tau + \Delta \tau_1 \Delta \omega} + \vec{q} \times \right) \vec{\Omega} \quad (\text{A2.1})$$

From Eq. (A2.1), we know that $\frac{d|\vec{\Omega}|}{d\omega} = \left(\frac{\Delta \tau_1}{\Delta \tau + \Delta \tau_1 \Delta \omega} \right) |\vec{\Omega}|$. Also $\frac{d\hat{\Omega}}{d\omega} = \vec{q} \times \hat{\Omega}$. Further,

the PMD vector can be expressed as $\vec{\Omega} = |\vec{\Omega}| \hat{\Omega}$. The solution for the magnitude of PMD vector can be expressed as:

$$|\vec{\Omega}| = (\Delta \tau_0 + \Delta \tau_1 \Delta \omega) \quad (\text{A2.2})$$

Next, we derive the solution for the frequency derivation of $\hat{\Omega}$.

If $\vec{q} = [q_1, q_2, q_3]^T$, Then

$$R' = \vec{q} \times = \begin{bmatrix} 0 & -q_3 & q_2 \\ q_3 & 0 & -q_1 \\ -q_2 & q_1 & 0 \end{bmatrix} \quad (\text{A2.3})$$

We have a new equation $\frac{\partial \hat{\Omega}(\omega)}{\partial \omega} = R' \hat{\Omega}(\omega)$, whose solution is

$$\hat{\Omega}(\omega) = \exp \{ R' (\omega - \omega_0) \} \hat{\Omega}(\omega_0) \quad (\text{A2.4})$$

where $\hat{\Omega}_0 = \hat{\Omega}(\omega_0) = [\Omega_{01}, \Omega_{02}, \Omega_{03}]^T$. Also, let $\Delta q = |\vec{q}|$. It is easy to obtain three eigenvalues of R' which are $(0, j\Delta q, -j\Delta q)$. Furthermore, we have the matrix of eigenvectors

$$G = \begin{bmatrix} q_3 & q_1 q_3 - j q_2 q & q_1 q_3 + j q_2 q \\ q_2 & q_1 q_2 + j q_3 q & q_1 q_2 - j q_3 q \\ q_1 & q_1^2 - q^2 & q_1^2 - q^2 \end{bmatrix} \quad (\text{A2.5})$$

Then,

$$\exp[R'(\omega - \omega_0)] = G \begin{bmatrix} 1 & 0 & 0 \\ 0 & e^{jq(\omega - \omega_0)} & 0 \\ 0 & 0 & e^{-jq(\omega - \omega_0)} \end{bmatrix} G^{-1} \quad (\text{A2.6})$$

Eq. (A2.6) leads to

$$\begin{aligned} & \exp[R'(\omega - \omega_0)] \\ &= \frac{1}{\Delta q^2} \begin{bmatrix} q_1^2 + \frac{q_1^2 q_3^2 + q_2^2 \Delta q^2}{\Delta q^2 - q_3^2} \cos(\Delta q \Delta \omega) & q_1 q_2 (1 - \cos \Delta q \Delta \omega) - q_3 q \sin \Delta q \Delta \omega & q_1 q_3 (1 - \cos \Delta q \Delta \omega) + q_2 \Delta q \sin \Delta q \Delta \omega \\ q_1 q_2 (1 - \cos \Delta q \Delta \omega) + q_3 \Delta q \sin \Delta q \Delta \omega & q_2^2 + \frac{q_2^2 q_3^2 + q_1^2 \Delta q^2}{\Delta q^2 - q_3^2} \cos \Delta q \Delta \omega & q_2 q_3 (1 - \cos \Delta q \Delta \omega) - q_1 \Delta q \sin \Delta q \Delta \omega \\ q_1 q_3 (1 - \cos \Delta q \Delta \omega) - q_2 \Delta q \sin \Delta q \Delta \omega & q_2 q_3 (1 - \cos \Delta q \Delta \omega) + q_1 \Delta q \sin \Delta q \Delta \omega & q_3^2 + (\Delta q^2 - q_3^2) \cos \Delta q \Delta \omega \end{bmatrix} \end{aligned} \quad (\text{A2.7})$$

Then,

$$\begin{aligned} \hat{\Omega}(\omega) &\stackrel{\text{def}}{=} \begin{bmatrix} \Omega_1(\omega) \\ \Omega_2(\omega) \\ \Omega_3(\omega) \end{bmatrix} = \\ &= \begin{bmatrix} \frac{q_1}{\Delta q^2} (q_1 \Omega_{01} + q_2 \Omega_{02} + q_3 \Omega_{03}) + \left[\Omega_{01} - \frac{q_1}{\Delta q^2} (q_1 \Omega_{01} + q_2 \Omega_{02} + q_3 \Omega_{03}) \right] \cos \Delta q \Delta \omega + \frac{1}{\Delta q} (q_2 \Omega_{03} - q_3 \Omega_{02}) \sin \Delta q \Delta \omega \\ \frac{q_2}{\Delta q^2} (q_1 \Omega_{01} + q_2 \Omega_{02} + q_3 \Omega_{03}) + \left[\Omega_{02} - \frac{q_2}{\Delta q^2} (q_1 \Omega_{01} + q_2 \Omega_{02} + q_3 \Omega_{03}) \right] \cos \Delta q \Delta \omega + \frac{1}{\Delta q} (q_3 \Omega_{01} - q_1 \Omega_{03}) \sin \Delta q \Delta \omega \\ \frac{q_3}{\Delta q^2} (q_1 \Omega_{01} + q_2 \Omega_{02} + q_3 \Omega_{03}) + \left[\Omega_{03} - \frac{q_3}{\Delta q^2} (q_1 \Omega_{01} + q_2 \Omega_{02} + q_3 \Omega_{03}) \right] \cos \Delta q \Delta \omega + \frac{1}{\Delta q} (q_1 \Omega_{02} - q_2 \Omega_{01}) \sin \Delta q \Delta \omega \end{bmatrix} \end{aligned} \quad (\text{A2.8})$$

The above equation can be rearranged as

$$\hat{\Omega}(\omega) = \hat{q} [\hat{q} \cdot \hat{\Omega}(\omega_0)] + \left\{ \hat{\Omega}(\omega_0) - \hat{q} [\hat{q} \cdot \hat{\Omega}(\omega_0)] \right\} \cos q \Delta \omega + \hat{q} \times \hat{\Omega}(\omega_0) \sin q \Delta \omega \quad (\text{A2.9})$$

$$\text{Or } \hat{\Omega}(\omega) = \hat{q} \cos 2\alpha_0 + [\hat{\Omega}(\omega_0) - \hat{q} \cos 2\alpha_0] \cos \Delta q \Delta \omega + \hat{B}_0 \sin 2\alpha_0 \sin \Delta q \Delta \omega \quad (\text{A2.10})$$

Form Eqs. (A2.2), (A2.9) and (A2.10), we know that the solution of equation (A2.1) can be expressed as:

$$\vec{\Omega}(\omega) = (\Delta \tau_0 + \Delta \tau_1 \Delta \omega) \left\{ \begin{aligned} &\hat{q}(\hat{q} \cdot \hat{\Omega}_0) + [\hat{\Omega}_0 - \hat{q}(\hat{q} \cdot \hat{\Omega}_0)] \cos(\Delta q \Delta \omega) \\ &+ \hat{B}_0 \sin 2\alpha_0 \sin(\Delta q \Delta \omega) \end{aligned} \right\} \quad (\text{A2.11})$$

Or

$$\vec{\Omega}(\omega) = (\Delta \tau_0 + \Delta \tau_1 \Delta \omega) \left\{ \begin{aligned} &\hat{q} \cos 2\alpha_0 + [\hat{\Omega}_0 - \hat{q} \cos 2\alpha_0] \cos(\Delta q \Delta \omega) \\ &+ \hat{B}_0 \sin 2\alpha_0 \sin(\Delta q \Delta \omega) \end{aligned} \right\} \quad (\text{A2.12})$$

Eqs. (A2.11) and (A2.12) are the desired relations.

Appendix III

Phase difference between $\hat{q}_+$ and $\hat{q}_-$

Let the PMD vector be represented as $\hat{\Omega}(\omega)$ in Stokes notation and as $\hat{e}_{\Omega+}(\omega)$ in the Jones notation. We can express the PMD vector in terms of the two orthogonal vectors $\hat{q}_+$ and $\hat{q}_-$, $\hat{e}_{q+}$ and $\hat{e}_{q-}$ in Jones vector notation:

$$\hat{e}_{\Omega}(\omega) = \cos\alpha_0 e^{j\varphi_+} \hat{e}_{q+} + \sin\alpha_0 e^{j\varphi_-} \hat{e}_{q-} \quad (\text{A3.1})$$

For the special case of $\alpha_0 = \pi/4$, one gets $\hat{e}_{\Omega}(\omega)|_{\alpha=\pi/4} = \hat{e}'_{\Omega}(\omega)$:

$$\hat{e}'_{\Omega}(\omega) = \cos\frac{\pi}{4} e^{j\varphi'_+} \hat{e}_{q+} + \sin\frac{\pi}{4} e^{j\varphi'_-} \hat{e}_{q-} \quad (\text{A3.2})$$

In the above equations, $\varphi_{\pm}$ and $\varphi'_{\pm}$ are the phase angles in the Jones notation for the two orthogonal depolarization vectors $\hat{q}_+$ and $\hat{q}_-$; also, $2\alpha_0$ is independent of frequency. Since depolarization vectors are a property of the fiber, the phase difference $\Delta\varphi$ ($\Delta\varphi = \varphi_+ - \varphi_-$) is equal to $\Delta\varphi'$ ($\Delta\varphi' = \varphi'_+ - \varphi'_-$). Next, for calculating $\Delta\varphi$, we can use $\alpha = \pi/4$ without loss of generality because $\Delta\varphi$ is independent of α_0 .

We start with Eq. (2.54) for $\vec{\Omega}$; for $\alpha_0 = \pi/4$, one write for $\hat{\Omega}'(\omega)$

$$\hat{\Omega}'(\omega) = \hat{\Omega}'(\omega_0) \cos(\Delta q \Delta \omega - \Delta q_1 \Delta \omega^2) + \hat{B}'_0 \sin(\Delta q \Delta \omega - \Delta q \Delta \omega^2) \quad (\text{A3.3})$$

In Eq. (A3.3), $\hat{B}'_0 = \hat{q} \times \hat{\Omega}'(\omega_0)$. For simplicity, we further assume that $\hat{q} = [0, 0, 1]^T$

and $\hat{\Omega}'(\omega_0) = [1, 0, 0]^T$. Then,

$$\hat{\Omega}'(\omega) = \begin{bmatrix} \cos(\Delta\omega\Delta q - \Delta q_1\Delta\omega^2) \\ \sin(\Delta\omega\Delta q - \Delta q_1\Delta\omega^2) \\ 0 \end{bmatrix} \quad (\text{A3.4})$$

In the Jones notation, the PMD vector can be written as

$$\hat{e}'_{\Omega}(\omega) = \begin{bmatrix} \cos(\Delta q\Delta\omega/2 - \Delta q_1\Delta\omega^2/2)e^{j0} \\ \sin(\Delta q\Delta\omega/2 - \Delta q_1\Delta\omega^2/2)e^{-j0} \end{bmatrix} \quad (\text{A3.5})$$

The Jones vector $\hat{e}_{q+}$ corresponding to $\hat{q}_+$ is

$$\hat{e}_{q+} = \begin{bmatrix} \cos(\pi/4)e^{(j\pi/4)} \\ \sin(\pi/4)e^{(-j\pi/4)} \end{bmatrix} \quad (\text{A3.6})$$

Next, to calculate ϕ_+ and ϕ_- in Eq. (A3.1), we use Eqs. (A3.2), (A3.5), and (A3.6), and use the following relations [78]:

$$\tan\phi_+ = \frac{\cos(\theta' + \theta_p)}{\cos(\theta' - \theta_p)} \tan(\delta' - \delta_p) \quad (\text{A3.7})$$

$$\tan\phi_- = -\frac{\sin(\theta' + \theta_p)}{\sin(\theta' - \theta_p)} \tan(\delta' - \delta_p) \quad (\text{A3.8})$$

In Eq. (A3.7) and Eq. (A3.8) $\theta' = \Delta q\Delta\omega/2 - \Delta q_1\Delta\omega^2/2$, $\theta_p = \pi/4$, $\delta' = 0$ and

$\delta_p = \pi/4$. From Eq. (A3.7), $\phi_+ = \frac{\Delta q\Delta\omega}{2} - \frac{\Delta q_1\Delta\omega^2}{2} - \frac{\pi}{4}$ and from Eq.(A3.8), one gets

$\phi_- = -\frac{\Delta q\Delta\omega}{2} + \frac{\Delta q_1\Delta\omega^2}{2} - \frac{\pi}{4}$. Hence $\Delta\phi = \Delta q\Delta\omega - \Delta q_1\Delta\omega^2$. It means that the phase

difference between the two orthogonal vectors $\hat{q}_+$ and $\hat{q}_-$, including DCD is

$\Delta q\Delta\omega - \Delta q_1\Delta\omega^2$.

Appendix IV

Pulse shape due to higher-order PMD effect without PCD, DCD,

and CD

Based on Eqs (2.56)-(2.58), when we expand the PSPs in term of $\hat{q}_+$ and $\hat{q}_-$, the output state of polarization can be expressed in terms of the two orthogonal depolarization vectors as

$$\begin{aligned}\hat{e} = & \cos \alpha \cdot \cos \alpha_0 \exp \left[j \left(\Delta q \Delta \omega / 2 + \Delta \tau \Delta \omega / 2 \right) \right] \hat{e}_{q+} \\ & + \cos \alpha \cdot \sin \alpha_0 \exp \left[j \left(\Delta \tau \Delta \omega / 2 - \Delta q \Delta \omega / 2 \right) \right] \hat{e}_{q-} \\ & + \sin \alpha \cdot \sin \alpha_0 \exp \left[j \left(\pi / 2 - \Delta \tau \Delta \omega / 2 - \Delta q \Delta \omega / 2 \right) \right] \hat{e}_{q+} \\ & + \sin \alpha \cdot \cos \alpha_0 \exp \left[-j \left(\pi / 2 + \Delta \tau \Delta \omega / 2 - \Delta q \Delta \omega / 2 \right) \right] \hat{e}_{q-}\end{aligned}\quad (\text{A4.1})$$

Form Eq. (A4.1), we know that

$$\begin{aligned}f_{q+}(t) = & \frac{\cos \alpha \cdot \cos \alpha_0}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp \left[j \left(\Delta \tau \Delta \omega / 2 + \Delta q \Delta \omega / 2 \right) \right] \exp \left[j(\omega - \omega_0)t \right] d\omega \\ & + \frac{\sin \alpha \cdot \sin \alpha_0}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp \left[j \left(\pi / 2 - \Delta \tau \Delta \omega / 2 - \Delta q \Delta \omega / 2 \right) \right] \exp \left[j(\omega - \omega_0)t \right] d\omega\end{aligned}\quad (\text{A4.2})$$

$$\begin{aligned}f_{q-}(t) = & \frac{\cos \alpha \cdot \sin \alpha_0}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp \left[j \left(\Delta \tau \Delta \omega / 2 - \Delta q \Delta \omega / 2 \right) \right] \exp \left[j(\omega - \omega_0)t \right] d\omega \\ & + \frac{\sin \alpha \cdot \cos \alpha_0}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp \left[-j \left(\pi / 2 + \Delta \tau \Delta \omega / 2 - \Delta q \Delta \omega / 2 \right) \right] \exp \left[j(\omega - \omega_0)t \right] d\omega\end{aligned}\quad (\text{A4.3})$$

Combining the components, one gets

$$\vec{f}(t) = f_{q+}(t) \hat{e}_{q+} + f_{q-}(t) \hat{e}_{q-} \quad (\text{A4.4})$$

Eqs. (A4.2)-(A4.4) are general equations, applicable to a pulse of any shape. To the best of our knowledge, it is the first time such a general equation has been presented for the second-order PMD effect on pulse-shape. Again, when the input signal is a

Gaussian pulse $f(t) = M \exp \left(-\frac{t^2}{2T_0^2} \right)$, then

$$\begin{aligned}
 f_{q+}(t) = & M \cos \alpha \cdot \cos \alpha_0 \exp \left\{ -\frac{(\Delta \tau / 2 + \Delta q / 2 + t)^2}{2T_0^2} \right\} \\
 & + jM \sin \alpha \cdot \sin \alpha_0 \exp \left\{ -\frac{(t - \Delta \tau / 2 - \Delta q / 2)^2}{2T_0^2} \right\}
 \end{aligned} \tag{A4.5}$$

$$\begin{aligned}
 f_{q-}(t) = & M \cos \alpha \cdot \sin \alpha_0 \exp \left\{ -\frac{(t + \Delta \tau / 2 - \Delta q / 2)^2}{2T_0^2} \right\} \\
 & - jM \sin \alpha \cdot \cos \alpha_0 \exp \left\{ -\frac{(t - \Delta \tau / 2 + \Delta q / 2)^2}{2T_0^2} \right\}
 \end{aligned} \tag{A4.6}$$

The signal power in the detector is $P(t) = |\vec{f}(t)|^2 = \vec{f}(t) \cdot \vec{f}^*(t)$. Because $\hat{e}_{q+}$ is orthogonal to $\hat{e}_{q-}$,

$$\begin{aligned}
 P(t) = & |f_{q+}(t)|^2 + |f_{q-}(t)|^2 \\
 = & M^2 \cos^2 \alpha \cdot \cos^2 \alpha_0 \exp \left[-\frac{(\Delta q / 2 + \Delta \tau / 2 + t)^2}{T_0^2} \right] + M^2 \sin^2 \alpha \cdot \sin^2 \alpha_0 \exp \left[-\frac{(\Delta q / 2 + \Delta \tau / 2 - t)^2}{T_0^2} \right] \\
 & + M^2 \cos^2 \alpha \cdot \sin^2 \alpha_0 \exp \left[-\frac{(t + \Delta \tau / 2 - \Delta q / 2)^2}{T_0^2} \right] + M^2 \sin^2 \alpha \cdot \cos^2 \alpha_0 \exp \left[-\frac{(t + \Delta q / 2 - t / 2)^2}{T_0^2} \right]
 \end{aligned} \tag{A4.7}$$

Appendix V

Pulse shape due to higher-order PMD effect including PCD, DCD, and CD

Based on Eqs (2.56)-(2.58) and Eq. (A4.1), if both the change of DGD and the chromatic dispersion are taken into account, then $f_{q+}(t)$ and $f_{q-}(t)$ can be expressed as

$$\begin{aligned}
 f_{q+}(t) = & \frac{\cos \alpha \cdot \cos \alpha_0}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp \left[j \left(\frac{\Delta \tau \Delta \omega - \Delta \tau_1 \Delta \omega^2}{2} + \frac{\Delta q \Delta \omega - \Delta q_1 \Delta \omega^2 - \beta_2 L \Delta \omega^2}{2} \right) \right] \exp[j(\omega - \omega_0)t] d\omega \\
 & + \frac{\sin \alpha \cdot \sin \alpha_0}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp \left\{ j \left(\frac{\pi - \Delta \tau \Delta \omega + \Delta \tau_1 \Delta \omega^2}{2} + \frac{-\Delta q \Delta \omega + \Delta q_1 \Delta \omega^2 - \beta_2 L \Delta \omega^2}{2} \right) \right\} \exp[j(\omega - \omega_0)t] d\omega
 \end{aligned}
 \tag{A5.1}$$

$$\begin{aligned}
 f_{q-}(t) = & \frac{\cos \alpha \cdot \sin \alpha_0}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp \left[j \left(\frac{\Delta \tau \Delta \omega - \Delta \tau_1 \Delta \omega^2}{2} + \frac{-\Delta q \Delta \omega + \Delta q_1 \Delta \omega^2 - \beta_2 L \Delta \omega^2}{2} \right) \right] \exp[j(\omega - \omega_0)t] d\omega \\
 & + \frac{\sin \alpha \cdot \cos \alpha_0}{2\pi} \int_{-\infty}^{+\infty} |F(\omega - \omega_0)| \exp \left[-j \left(\frac{\pi + \Delta \tau \Delta \omega - \Delta \tau_1 \Delta \omega^2}{2} + \frac{-\Delta q \Delta \omega + \Delta q_1 \Delta \omega^2 + \beta_2 L \Delta \omega^2}{2} \right) \right] \exp[j(\omega - \omega_0)t] d\omega
 \end{aligned}
 \tag{A5.2}$$

Combining the components and using Eq. (A4.4), when the input signal is a Gaussian

pulse $f(t) = \exp\left(-\frac{t^2}{2T_0^2}\right)$ with frequency spectrum

$$F(\omega - \omega_0) = \sqrt{2\pi}T_0 \exp\left(-\frac{\Delta\omega^2 T_0^2}{2}\right), \text{ then}$$

$$\begin{aligned} f_{q+}(t) = & \frac{T_0 \cos \alpha \cdot \cos \alpha_0}{\sqrt{T_0^2 - j(\beta_2 L + \Delta\tau_1 + \Delta q_1)}} \exp\left\{-\frac{(\Delta\tau/2 + \Delta q/2 + t)^2}{2[T_0^2 - j(\beta_2 L + \Delta\tau_1 + \Delta q_1)]}\right\} \\ & + \frac{jT_0 \sin \alpha \cdot \sin \alpha_0}{\sqrt{T_0^2 - j(\beta_2 L - \Delta\tau_1 - \Delta q_1)}} \exp\left\{-\frac{(t - \Delta\tau/2 - \Delta q/2)^2}{2[T_0^2 - j(\beta_2 L - \Delta\tau_1 - \Delta q_1)]}\right\} \end{aligned} \quad (\text{A5.3})$$

$$\begin{aligned} f_{q-}(t) = & \frac{T_0 \cos \alpha \cdot \sin \alpha_0}{\sqrt{T_0^2 - j(\beta_2 L + \Delta\tau_1 - \Delta q_1)}} \left[\exp\left\{-\frac{(t + \Delta\tau/2 - \Delta q/2)^2}{2[T_0^2 - j(\beta_2 L + \Delta\tau_1 - \Delta q_1)]}\right\} \right] \\ & - j \frac{T_0 \sin \alpha \cdot \cos \alpha_0}{\sqrt{T_0^2 - j(\beta_2 L - \Delta\tau_1 + \Delta q_1)}} \exp\left\{-\frac{(t - \Delta\tau/2 + \Delta q/2)^2}{2[T_0^2 - j(\beta_2 L - \Delta\tau_1 + \Delta q_1)]}\right\} \end{aligned} \quad (\text{A5.4})$$

The signal power at the detector is $P(t) = |\vec{f}(t)|^2 = \vec{f}(t) \cdot \vec{f}^*(t)$. Because $\hat{e}_{q+}$ is orthogonal to $\hat{e}_{q-}$,

$$\begin{aligned} P(t) = & |f_{q+}(t)|^2 + |f_{q-}(t)|^2 \\ = & \frac{T_0^2 \cos^2 \alpha \cdot \cos^2 \alpha_0}{\sqrt{T_0^4 + (\beta_2 L + \Delta\tau_1 + \Delta q_1)^2}} \exp\left\{-\frac{T_0^2 (t + \Delta q/2 + \Delta\tau/2)^2}{[T_0^4 + (\Delta q_1 + \Delta\tau_1 + \beta_2 L)^2]}\right\} \\ & + \frac{T_0^2 \sin^2 \alpha \cdot \sin^2 \alpha_0}{\sqrt{T_0^4 + (\beta_2 L - \Delta\tau_1 - \Delta q_1)^2}} \exp\left\{-\frac{T_0^2 (t - \Delta q/2 - \Delta\tau/2)^2}{[T_0^4 + (\beta_2 L - \Delta\tau_1 - \Delta q_1)^2]}\right\} \\ & + \frac{T_0^2 \cos^2 \alpha \cdot \sin^2 \alpha_0}{\sqrt{T_0^4 + (\beta_2 L + \Delta\tau_1 - \Delta q_1)^2}} \exp\left\{-\frac{T_0^2 (t + \Delta\tau/2 - \Delta q/2)^2}{[T_0^4 + (\beta_2 L + \Delta\tau_1 - \Delta q_1)^2]}\right\} \\ & + \frac{T_0^2 \sin^2 \alpha \cdot \cos^2 \alpha_0}{\sqrt{T_0^4 + (\beta_2 L - \Delta\tau_1 + \Delta q_1)^2}} \exp\left\{-\frac{T_0^2 (t - \Delta\tau/2 + \Delta q/2)^2}{[T_0^4 + (\beta_2 L - \Delta\tau_1 + \Delta q_1)^2]}\right\} \end{aligned} \quad (\text{A5.5})$$

Eq. (A5.5) shows the effect of chromatic dispersion and higher-order PMD including DCD, PCD, and depolarization vector on output pulse.

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