



BSI Standards Publication

Fibre optic communication system design guidelines

Part 16: Coherent receivers and transmitters with
high-speed digital signal processing

National foreword

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TECHNICAL REPORT

**Fibre optic communication system design guidelines –
Part 16: Coherent receivers and transmitters with high-speed digital signal
processing**

INTERNATIONAL
ELECTROTECHNICAL
COMMISSION

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INTERNATIONAL ELECTROTECHNICAL COMMISSION

FIBRE OPTIC COMMUNICATION SYSTEM DESIGN GUIDELINES –

**Part 16: Coherent receivers and transmitters
with high-speed digital signal processing**

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The language used for the development of this Technical Report is English.

This document was drafted in accordance with ISO/IEC Directives, Part 2, and developed in accordance with ISO/IEC Directives, Part 1 and ISO/IEC Directives, IEC Supplement, available at www.iec.ch/members_experts/refdocs. The main document types developed by IEC are described in greater detail at www.iec.ch/standardsdev/publications.

A list of all parts in the IEC 61282 series, published under the general title *Fibre optic communication system design guidelines*, can be found on the IEC website.

Future documents in this series will carry the new general title as cited above. Titles of existing documents in this series will be updated at the time of the next edition.

The committee has decided that the contents of this document will remain unchanged until the stability date indicated on the IEC website under "<http://webstore.iec.ch>" in the data related to the specific document. At this date, the document will be

- reconfirmed,
- withdrawn,
- replaced by a revised edition, or
- amended.

INTRODUCTION

Coherent optical receivers are widely used in long-haul fibre optic communication systems, especially in systems that transmit optical carriers at data rates of 100 Gbit/s and higher. While the principle of coherent detection is very similar to that of super-heterodyne (or homodyne) detection in radio and microwave receivers, its implementation is significantly more challenging. The main reason is that optical frequencies are substantially higher than radio frequencies, so it becomes more difficult to match the local oscillator frequency in the coherent receiver to the frequency of the transmitted optical signal. Furthermore, optical signals tend to be highly polarized, which means that the amplitude of a coherently received signal can be substantially reduced or even vanish if the polarization state of the local oscillator light does not match the polarization state of the received optical signal. This polarization matching is particularly difficult to achieve in fibre optic communications systems, which usually do not preserve the launched state of polarization of the transmitted signal. To overcome these problems, modern coherent receivers typically consist of four parallel coherent optical mixers that provide phase and polarization diversity, and they rely on high-speed digital signal processors to retrieve the transmitted data from the four received electrical signals.

This rather complex coherent receiver architecture is further justified by the fact that it allows the receiver to mitigate various types of signal impairments introduced in the fibre optic link (or in the receiver itself) simply by means of additional electronic signal processing. Most notably, it is possible to substantially reduce the signal distortions caused by polarization-mode dispersion (PMD) or uncompensated chromatic dispersion (CD) in the fibre link, without requiring additional optical PMD or CD compensators. For this reason, coherent optical communication systems generally allow signal transmission at much higher data rates than communication systems using direct-detection receivers. Furthermore, coherent detection with subsequent digital signal processing facilitates the decoding of complex vector-modulated signals, such as quadrature-amplitude modulated signals (QAM) and polarization-multiplexed (PM) signals, and thereby the transmission of higher data rates.

Aside from fibre optic communications systems, coherent optical receivers are also used in various test and measurement instrumentation. Most notable examples are optical modulation analysers (OMAs) and high-resolution optical spectrum analysers (HR-OSAs). Optical modulation analysers are high-performance optical reference receivers and are used to assess the signal quality of complex vector-modulated optical signals. They are typically composed of a carefully calibrated coherent receiver and a high-speed real-time digitizing oscilloscope to record the coherently received signals, which are then analysed with the help of a software-based digital signal processor.

High-resolution optical spectrum analysers are often used to analyse narrowband features of the optical spectrum of a modulated signal, such as a residual optical carrier or other spectral lines. In contrast to OMAs, they typically employ a low-speed coherent receiver, so as to utilize the frequency-selectivity of coherent detection. The key component in these instruments is a continuously tuneable local oscillator, which is scanned over the frequency range of the signal to be analysed while the total power of the coherently received signal is recorded. The spectral resolution of these instruments can be of the order of a few MHz. Other examples of coherent optical test instruments include in-service PMD analysers and in-band optical signal-to-noise ratio analysers for polarization-multiplexed signals.

FIBRE OPTIC COMMUNICATION SYSTEM DESIGN GUIDELINES –

Part 16: Coherent receivers and transmitters with high-speed digital signal processing

1 Scope

This part of IEC 61282 is a technical report on coherent optical receiver and transmitter technologies that are employed in fibre optic communication systems as well as in optical test and measurement equipment. This document describes the principle of operation and functional capabilities of coherent optical receivers as well as the operation of optical transmitters used to generate complex vector-modulated signals. It is intended to serve as a technical foundation for other IEC documents and standards related to coherent optical transmission techniques.

2 Normative references

The following documents are referred to in the text in such a way that some or all of their content constitutes requirements of this document. For dated references, only the edition cited applies. For undated references, the latest edition of the referenced document (including any amendments) applies.

IEC 60050-731, *International Electrotechnical Vocabulary (IEV) – Part 731: Optical fibre communication* (available at www.electropedia.org)

IEC TR 61931, *Fibre optic – Terminology*

3 Terms, definitions, and abbreviated terms

3.1 Terms and definitions

For the purposes of this document, the terms and definitions given in IEC 60050-731 and IEC TR 61931 apply.

ISO and IEC maintain terminological databases for use in standardization at the following addresses:

- IEC Electropedia: available at <http://www.electropedia.org/>
- ISO Online browsing platform: available at <http://www.iso.org/obp>

3.2 Abbreviated terms

ABC	adaptive bias control
ADC	analogue-to-digital converter
AM	amplitude modulation
ASIC	application-specific integrated circuit
ASK	amplitude-shift keying
BER	bit-error ratio
BPSK	binary phase-shift keying
CD	chromatic dispersion
CFP	C form-factor pluggable
CMA	constant modulus algorithm

CMRR	common-mode-rejection ratio
COSA	coherent optical subassembly
CW	continuous wave
DAC	digital-to-analogue converter
DD-LMS	decision-directed least-mean-square
DGD	differential group delay
DSP	digital signal processor
DWDM	dense wavelength-division multiplexing
EDFA	erbium-doped fibre amplifier
ENOB	effective number of bits
EVM	error-vector magnitude
ER	extinction ratio
EQL	equalizer filter
FEC	forward error correction
FFT	fast Fourier transform
FIR	finite impulse response
FSE	fractionally spaced equalizer
FWM	four-wave mixing
GS	geometrically shaped
GVD	group velocity dispersion
HB-CDM	high-bandwidth coherent driver modulator
HR-OSA	high-resolution optical spectrum analyser
LO	local oscillator
LPF	low-pass filter
MAP	maximum a posteriori
MMA	multi-modulus algorithm
MMI	multi-mode interference
MZM	Mach-Zehnder modulator
NLFT	nonlinear Fourier transform
NLPN	nonlinear phase noise
I	in-phase component
IA	implementation agreement
ICR	intradyne coherent receiver
IC-TROSA	integrated coherent transmit-receive optical subassembly
iFFT	inverse fast Fourier transform
Im	imaginary part
InP	indium phosphide
ITLA	integrable tuneable laser assembly
OFDM	orthogonal frequency-division multiplexing
OMA	optical modulation analyser
OOK	on-off keying
OSA	optical spectrum analyser
OSNR	optical signal-to-noise ratio

OTN	optical transport network
PAM	pulse-amplitude modulation
PBC	polarization beam combiner
PBS	polarization beam splitter
PC	polarization controller
PD	photodiodes, photo detectors
PDL	polarization-dependent loss
PLC	planar lightwave circuit
PMD	polarization-mode dispersion
PM	polarization multiplexed
PM-QPSK	polarization multiplexed QPSK
PR	polarization rotator
PRBS	pseudo-random binary sequence
PS-QPSK	polarization switched QPSK
PSK	phase-shift keying
Q	quadrature component
QAM	quadrature amplitude modulation
QASK	quaternary amplitude-shift keying
QPSK	quaternary phase-shift keying
RAM	random access memory
RDE	radius-directed equalizer
Re	real part
RF	radio frequency
RMS	root-mean-square
ROADM	reconfigurable optical add-drop multiplexer
RRC	root-raised-cosine
SiPh	silicon photonics
SNR	signal-to-noise ratio
SPM	self-phase modulation
SSB	single-sideband
TDE	time-domain equalizer
TEF	transversal electrical filter
TIA	transimpedance amplifier
TROSA	transmit-receive optical subassembly
VCO	voltage-controlled oscillator
WDM	wavelength-division multiplexing
X	X-polarized component
XI	in-phase component of X-polarized signal
XPM	cross-phase modulation
XQ	quadrature component of X-polarized signal
Y	Y-polarized component
YI	in-phase component of Y-polarized signal
YQ	quadrature component of Y-polarized signal

4 Background

The development of coherent optical receivers started in the late 1980s and was driven by the desire to improve the sensitivity of optical receivers in order to increase the reach of fibre-optic communication systems [1]¹. In addition, coherent receivers were expected to facilitate decoding of phase-shift-keyed signals (PSKs) and to serve as narrow-band frequency selectors in wavelength-division multiplexed systems. However, the interest in coherent receivers diminished rapidly after practical Erbium-doped fibre amplifiers (EDFAs) became available just a few years later, because these optical amplifiers enabled repeater-less transmission of optical signals over very long distances without the need for coherent detection. Besides, coherent receivers at that time were still not mature enough to be deployed in practical fibre optic communication systems, because coherent detection was very sensitive to the polarization state of the received signals and, in addition, required transmitter and local-oscillator lasers with very high phase stability [2].

The interest in coherent optical receiver technology revived in the late 2000s, when high-speed analogue-to-digital converters (ADCs) and high-speed digital signal processors (DSPs) became available. These high-speed ADCs sample the received electrical signals at or above the symbol rate of the transmitted signals and deliver the digitized signals to a DSP, which then processes the sampled electrical signals in real time to retrieve the transmitted digital or analogue information [3]. It turned out that the combination of linear signal detection in coherent receivers and signal processing in high-speed DSPs not only solved the technical problems of early-generation coherent receivers, but also enabled electronic mitigation of linear (and even nonlinear) signal distortions encountered in the optical communication link [4]. Most importantly, coherent reception with subsequent digital signal processing allows adaptive mitigation of undesired impairments caused by polarization-mode dispersion (PMD) and uncompensated chromatic dispersion (CD) in the transmission fibre, either of which can severely distort the waveforms of optical signals at data rates of 10 Gbit/s or higher [6]. Modern coherent receivers are often capable of accommodating large amounts of PMD and CD and, as a result, allow signal transmission at much higher modulation rates and/or over much longer distances than direct-detection receivers [5]. Moreover, the DSP used in coherent receivers can even mitigate signal distortions resulting from non-ideal optical and electronic components in the coherent receiver itself, such as signal skew, amplitude imbalance, and other deviations from the desired frequency response, as described in more detail in Clause 5 of this document.

Coherent receivers also facilitate the decoding of complex vector-modulated signals, such as M -ary PSK or quadrature-amplitude modulation (M -QAM), which can carry several times the information of binary on-off-keyed (OOK) signals of the same symbol rate [7]. Moreover, modern coherent receivers are capable of decoding polarization-multiplexed (PM) signals, which can carry up to twice the information of corresponding single-polarized signals [7], [8]. In fact, the first commercially deployed fibre optic communication systems with coherent detection transmitted polarization-multiplexed quaternary phase-shift-keyed (PM-QPSK) signals at symbol rates of 10 GBd, yielding a total transmission rate of 40 Gbit/s per optical carrier [9]. Just a few years later, a second generation of coherent receivers was introduced allowing transmission of single-carrier PM-QPSK signals at 100 Gbit/s [10]. Subsequent development of transmitters with linear modulation response then enabled transmission of polarization-multiplexed 16QAM signals at 200 Gbit/s and even 400 Gbit/s [11]. Moreover, transmission of 32QAM and 64QAM signals under practical conditions has been demonstrated, and the feasibility of transmitting of even higher-order QAM signals is under investigation [13], [14], [15].

The generation of higher-order optical QAM signals usually requires linear modulators and driver amplifiers as well as careful equalization of non-ideal transmitter characteristics introduced by the optical and electronic transmitter components, including frequency roll-off, skew, and nonlinear response. This equalization can be achieved by employing digital signal processing in the transmitter to facilitate adjustable pre-distortion of the electrical signals applied to the optical modulator [2]. Therefore, as soon as fast high-speed digital-to-analogue

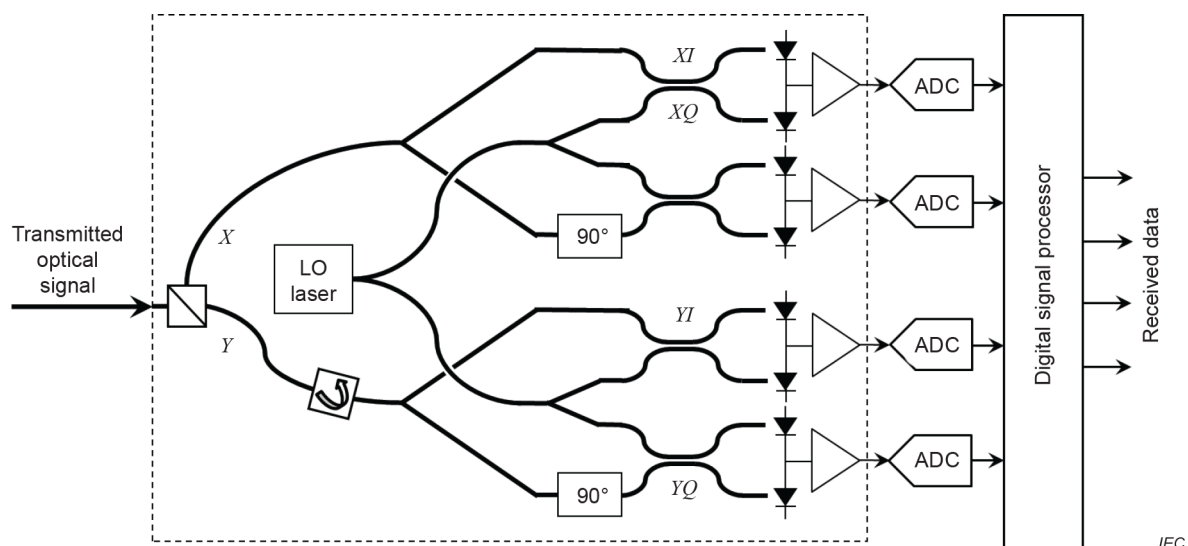
¹ Numbers in square brackets refer to the Bibliography.

converters (DACs) became commercially available, high-speed DSPs were also introduced in optical transmitters. In addition to equalization, the transmitter DSPs are frequently used to spectrally shape the electrical drive signals applied to the optical modulator in order to generate optical signals with the desired optical spectrum [12]. Transmitter DSPs also facilitate electronic pre-compensation of chromatic dispersion in the transmission fibre [6], [15]. Clause 8 describes the generation of optical signals with complex modulation formats and the functions of transmitter DSPs.

5 Coherent transmission of vector-modulated signals

5.1 Typical receiver architecture

As mentioned in Clause 4, modern coherent receivers for fibre optic communication systems usually employ four parallel coherent mixers that are configured in such a way that they detect the optical amplitude, phase, and polarization state of the received signal. Figure 1 displays a typical arrangement of these four mixers, which are labelled XI , XQ , YI , and YQ [16]. The mixers thus generate four analogue electrical signals, which are then converted to digital signals via high-speed analogue to digital converters (ADCs) and further processed in a high-speed digital signal processor (DSP).



Key

ADC Analogue-to-digital converter

LO Local oscillator

X X -polarized signal

Y Y -polarized signal

XI In-phase component of X -polarized signal

XQ Quadrature component of X -polarized signal

YI In-phase component of Y -polarized signal

YQ Quadrature component of Y -polarized signal

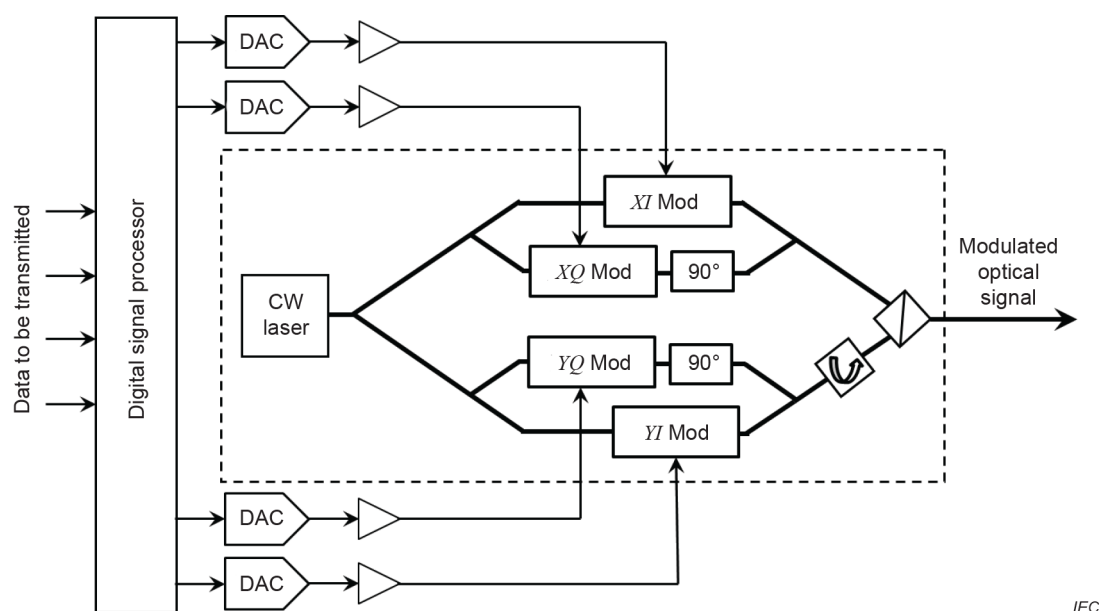
Figure 1 – Coherent optical receiver

The operation of these mixers and their capabilities will be discussed in more detail in Clause 6. It is important to note that the arrangement shown in Figure 1 makes the coherent receiver insensitive to time-varying fluctuations in the optical phase and polarization state of the received signal, which is a highly desirable feature for practical applications in fibre optic communication systems [1]. Moreover, it also allows the receiver to decode complex vector-modulated signals in which amplitude, phase, and polarization state are independently modulated [8], [17]. As described in more detail in 5.3, these vector-modulated signals can carry several times more

information than conventional on-off-keyed (OOK) signals transmitted at the same symbol rate. Hence, transmission of information at a given data rate with vector-modulated signals can be accomplished at much lower symbol rates, and hence with much lower electrical and optical bandwidth, than transmission with conventional OOK signals, which is also a highly desirable feature. However, decoding of these vector-modulated signals in the coherent receiver is a fairly complex process, which is usually accomplished by employing a high-speed DSP, as discussed in more detail in Clause 7.

5.2 Typical transmitter architecture

Coherent transmission systems frequently employ optical transmitters that are capable of generating complex vector-modulated signals. An example of such a transmitter is displayed in Figure 2, where the transmitter comprises four parallel modulators, which are labelled XI , XQ , YI , and YQ . These four modulators are typically employed in a nested Mach-Zehnder configuration in such a way that they allow independent modulation of the optical amplitude, phase, and polarization state of the transmitted signal, as discussed in more detail in Clause 8 [18].



Key

90° Optical phase shifter

CW Continuous wave

DAC Digital-to-analogue converter

Mod Optical modulator

XI In-phase component of X -polarized signal

XQ Quadrature component of X -polarized signal

YI In-phase component of Y -polarized signal

YQ Quadrature component of Y -polarized signal

Figure 2 – Optical transmitter for coherent transmission

The labels XI , XQ , YI , and YQ (or IX , QX , IY , and QY) are widely used in the industry to characterize the various signal components in vector-modulated signals, both for the optical transmitter and the optical receiver [16], [18]. However, it is very important to note that the signal components generated in the transmitter by the XI , XQ , YI , and YQ modulators are in general different from the signal components generated by the XI , XQ , YI , and YQ mixers in the coherent receiver. In fact, the received signal components in XI , XQ , YI , and YQ are related to the transmitted signal components in XI , XQ , YI , and YQ by a complex and often frequency-dependent function (see 7.1). Therefore, a DSP is often employed to de-convolute the transmitted signals from the four received signal components.

NOTE The use of the same labels for the transmitter signal components and the receiver signal components can lead to considerable confusion, because there usually is no simple relationship between the transmitter and receiver components. For example, if a transmitter generates signals using only the XI and YI modulators (e.g. to generate polarization-multiplexed binary phase-shift-keyed signals), it is sometimes believed that the coherent receiver would only need XI and YI mixers. As discussed in Clause 6, this is only the case for homodyne receivers with a phase-locked oscillator laser. However, modern intradyne receivers need all four XI , XQ , YI , and YQ mixers to recover the transmitted signal components from the received signal.

Frequently, DSPs are also employed in the transmitter in conjunction with high-speed digital-to-analogue converters (DACs). As discussed in more detail in Clause 9, the DSP is used here to map the information data into the transmitted signal components and to pre-distort the digital drive signals for the modulators.

Unlike conventional direct-detection receivers, a coherent receiver requires a continuous-wave (CW) laser to serve as a local oscillator (LO), with which the received signal components are mixed, as shown in Figure 1. This laser typically emits unmodulated light at a single (but tuneable) optical frequency, just like the transmitter lasers in conventional non-coherent communication systems [19]. Coherent detection, however, requires transmitter and LO lasers with relatively low optical phase noise. It should be noted that the phase noise requirements for LO lasers can be different from those for the transmitter lasers, as discussed in more detail in 6.3.3, 6.4.5, and 6.4.6.

5.3 Vector-modulated signals

5.3.1 Mathematical description

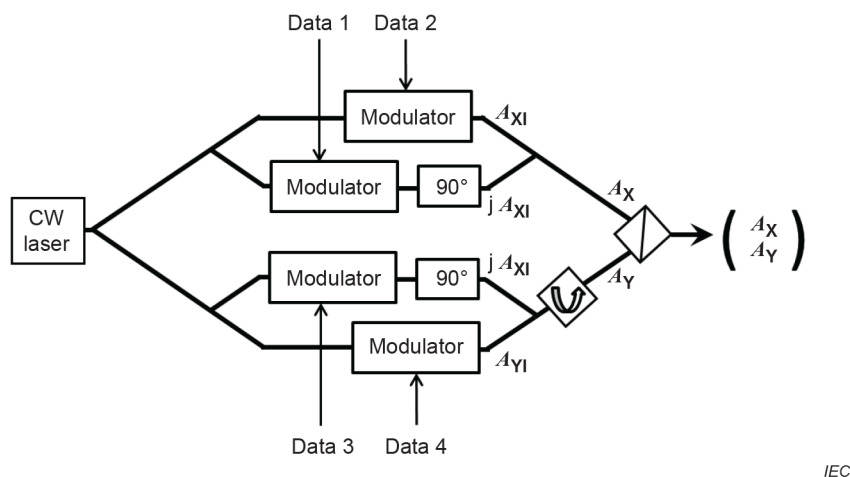
In this document, the term vector-modulated signals is used for optical signals that carry information data in their amplitude, phase, and/or polarization state [8], [17]. These signals can be described by a two-dimensional complex vector $\vec{A}_s(t)$ of the general form shown in Formula (1).

$$\vec{A}_s(t) = \sqrt{P_S} \begin{pmatrix} A_X(t) \\ A_Y(t) \end{pmatrix} e^{j\Phi_s(t)} = \sqrt{P_S} \begin{pmatrix} A_{XI}(t) + jA_{XQ}(t) \\ A_{YI}(t) + jA_{YQ}(t) \end{pmatrix} e^{j[\omega_s t + \varphi_s(t) + \hat{\varphi}_s(t)]} \quad (1)$$

where

P_S	is the average optical power of the signal;
$A_X(t)$ and $A_Y(t)$	are normalized complex amplitudes with $ A_X(t) ^2 + A_Y(t) ^2 = 1$, which describe the modulated signal components in two mutually orthogonal polarization states;
$A_{XI}(t)$, $A_{XQ}(t)$, $A_{YI}(t)$, and $A_{YQ}(t)$	are real-valued amplitudes, which represent the in-phase and quadrature signal components of $A_X(t)$ and $A_Y(t)$, respectively;
$\Phi_s(t) = \omega_s t + \varphi_s(t) + \hat{\varphi}_s(t)$	is a time-varying common phase;
$\omega_s = 2\pi f_s$	is the angular frequency of the modulated optical carrier;
$\varphi_s(t)$	represents optical phase variations;
$\hat{\varphi}_s(t)$	is a random variable representing laser phase noise from the transmitter laser;
e	is Euler's number, approximately equal to 2,71828; and
$j = \sqrt{-1}$	is the imaginary unit.

The four signal components A_{X_I} , A_{X_Q} , A_{Y_I} , and A_{Y_Q} usually carry different information (i.e. different payload data). As briefly discussed in 5.2, they are typically generated by four parallel amplitude modulators, which are arranged in a nested Mach-Zehnder structure. This arrangement includes 90° optical phase shifters to generate the quadrature components $A_{X_Q}(t)$ and $A_{Y_Q}(t)$ as well as a polarization rotator to generate the orthogonally polarized $A_Y(t)$ signal component, as shown schematically in Figure 3 and described in more detail in Clause 8.



Key

- 90° Optical phase shifter
 CW Continuous wave

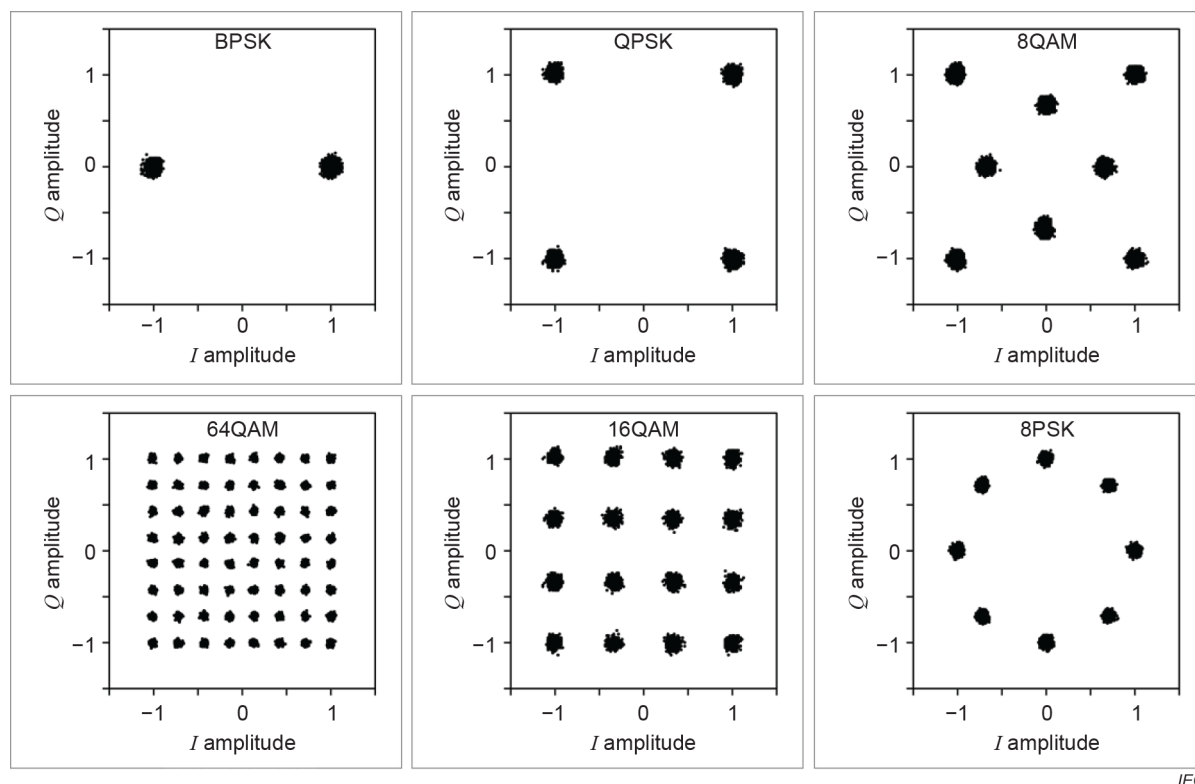
Figure 3 – Generation of vector modulated signals

The orthogonal polarization states represented by A_X and A_Y are often a pair of linearly polarized states when the signal is generated in the optical transmitter. However, this pair of orthogonal polarization states is generally transformed in the transmission fibre into another pair of orthogonal polarization states before the signal arrives at the receiver.

5.3.2 Binary amplitude and phase modulation

Conventional non-coherent communication systems typically transmit binary on-off-keyed (OOK) signals in a single polarization state [1]. In these signals, only $A_{X_I}(t)$ is modulated (i.e. toggled between 0 and 1) while $A_{X_Q} \equiv 0$ and $A_Y \equiv 0$ in Formula (1). Single-polarized binary OOK signals thus carry one bit of information per transmitted symbol and can be readily detected non-coherently with a single photodetector. OOK signals are sometimes generated either by an optical amplitude modulator (e.g. of the type discussed in 8.2), or alternatively, by a simpler intensity modulator.

Single-polarized binary phase-shift-keyed (BPSK) signals are typically generated by an optical amplitude modulator, which toggles $A_{X_I}(t)$ between -1 and 1 while keeping $A_{X_Q} \equiv 0$ and $A_Y \equiv 0$, as shown in the constellation diagram of Figure 4. Single-polarized BPSK signals can carry only one bit of information per transmitted symbol, but their detection requires a receiver that is sensitive to the phase of the received signal [20]. Single-polarized BPSK signals can be detected non-coherently with a properly tuned differential delay interferometer and a single photodetector. In principle, it is also possible to detect these signals with a coherent receiver having only a single mixer, as described in 6.2. However, in practical applications it is far more convenient and easier to detect these signals with a coherent receiver of the type shown in Figure 1, which has four parallel mixers, providing phase- and polarization-diversity, so that the receiver can tolerate phase and frequency variations in the transmission and reference lasers as well as time-varying polarization transformations in the transmission fibre.



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Key

BPSK Binary phase-shift keying

 I In-phase component

PSK Phase-shift keying

 Q Quadrature component

QAM Quadrature-amplitude modulation

QPSK Quaternary phase-shift keying

Figure 4 – Examples of modulation formats for coherent communication**5.3.3 Quadrature amplitude modulation**

The number of information bits carried by each transmitted symbol can be increased by also modulating the quadrature component A_{XQ} in addition to and independently of the in-phase component A_{XI} . Single-polarized signals using quaternary phase-shift-keying (QPSK), for example, carry two bits of information per transmitted symbol, since A_{XI} and A_{XQ} are both toggled independently between -1 and $+1$, as shown in Figure 4 [8]. QPSK signals and higher order quadrature amplitude modulated signals can be generated by two parallel amplitude modulators, as discussed in 8.3.

Even higher bit rates can be transmitted by using multi-level phase-shift keying (PSK) or quadrature amplitude modulation (QAM), such as 8PSK or 8QAM, which carry three bits per symbol. Moreover, 16QAM signals carry four bits per symbol and 64QAM signals even six bits per symbol. Single-polarized QPSK signals and multi-level PSK or QAM signals can be detected with coherent receivers that employ at least two parallel mixers with phase diversity, as discussed in 6.3. However, in practice it is far more convenient to detect these signals with a phase- and polarization-diversity coherent receiver, because of time-varying polarization transformations in the transmission fibre.

5.3.4 Polarization multiplexing

For any given modulation format, the number of bits per transmitted symbol can be increased further by transmitting polarization-multiplexed (PM) or dual-polarization (DP) signals, in which both polarized components, $A_X(t)$ and A_Y , are independently modulated. Polarization-multiplexed QPSK (PM-QPSK) signals, for example, carry four bits of information per symbol [8]. Polarization-multiplexed signals can readily be detected with phase- and polarization-diversity coherent receivers, which typically employ four parallel mixers, as described in 6.4. Coherent receivers thus facilitate the use of polarization multiplexing and higher order modulation formats, which allow transmission at higher data rates without requiring higher symbol rates. This is important because the electrical bandwidth requirements for the various electrical and optical components in the optical transmitter and receiver predominantly depend on the symbol rate.

5.3.5 Higher-dimensional coding and constellation shaping

It should be noted that higher-order QAM and PSK signals are more sensitive to optical noise than binary ASK or OOK signals and, hence, usually do not have the same transmission reach as binary signals [8]. In general, the more bits of information encoded in a signal, the more sensitive the signal is to optical noise and to other linear and nonlinear distortions. To make the transmission of higher-order QAM and PSK signals more robust to such impairments, new modulation techniques have been introduced in optical communication systems, which include multi-dimensional coding as well as geometrically and probabilistically shaped modulation formats [21], [22], [23].

In four-dimensional (4D) coding, the two polarization tributaries are not independently modulated, as described above, but instead are varied in such a way that the constellation points of the combined optical signal are distributed in a four-dimensional constellation space (whereas conventional polarization-multiplexed signals can be represented in a three-dimensional constellation space) [21]. A simple example of a 4D modulation format is polarization-switched QPSK (sometimes referred to as PS-QPSK), where only one polarization tributary is transmitted at any given time. Thus, when a QPSK symbol is transmitted in the X polarization in time slot n , then no signal is transmitted in the Y polarization, so that $A_{XC}(t_n) \neq 0$ and $A_{YC}(t_n) \equiv 0$; whereas, in another time slot m , a QPSK symbol can be transmitted in the Y polarization and no signal is transmitted in the X polarization, so that $A_{XC}(t_m) \equiv 0$ and $A_{YC}(t_m) \neq 0$. In PS-QPSK signals, the alternating occurrence of symbols and zeros in the two polarization tributaries is digitally modulated (i.e. non-periodic), so that an additional bit of information can be encoded in each transmitted QPSK symbol, yielding a total information rate of 3 bits per symbol (versus 4 bits per symbol for PM-QPSK). Hence, PS-QPSK signals carry less information than PM-QPSK signals, but when transmitted with the same average optical power, they are more resilient to optical noise than PM-QPSK signals [21]. This concept has been extended to modulation formats of higher cardinality, in which case the two polarization tributaries alternately transmit orthogonal subsets of constellation points, or even to eight-dimensional (8D) coding [21], [24].

In geometrically shaped (GS) modulation formats, the constellation points in either polarization tributary are arranged in such a way that the smallest Euclidian distance between them is as large as possible; whereas in probabilistically shaped modulation formats, the digital information is mapped into the constellation space in such a way that the frequency of occurrence of each constellation point is not equiprobable, as in conventional modulation formats, but is distributed, on average, according to a predetermined probability function, so that symbols carrying the innermost constellation points are more frequently transmitted than symbols carrying the outermost constellation points. Hence, for a given number of constellation points, probabilistic shaping reduces the information rate that can be transmitted, which is not the case for geometric shaping. To carry the same information rate as a conventional signal with equiprobable constellation points, a probabilistically shaped signal generally requires a modulation format of higher cardinality. For example, a probabilistically shaped 64QAM signal could transmit the same information rate as conventional 16QAM or 32QAM signals, and a probabilistically shaped 1024QAM signal the same information rate as a conventional 256QAM, but with higher tolerance to optical noise and with reduced optical bandwidth [25]. The higher

noise tolerance of probabilistically shaped signals arises from the fact that the innermost constellation points are more frequently transmitted than the outermost constellation points. Hence, at the same average optical signal power, the symbols of a probabilistically shaped signal carry more optical power than the same symbols of a conventional signal of the same modulation format and cardinality but with equiprobable constellation points. In terms of optical amplitude, therefore, the smallest Euclidian distance between the constellation points of a probabilistically shaped signal is effectively larger than that of a conventional signal of the same modulation format and cardinality.

The above described techniques of higher-dimensional coding, geometrical shaping, and probabilistic shaping are not mutually exclusive, but can in fact be employed simultaneously [22]. However, a more detailed description of these new coding schemes is beyond the scope of this document. Application of these advanced coding techniques requires additional signal processing at the transmitter and receiver, but can be accomplished with the same coherent receivers and optical transmitters as used for conventional vector-modulated formats (see 6.4 and 8.4).

Likewise, the same type of coherent receivers can be used for multi-carrier modulation formats, which are signals that are composed of independently modulated sub-carriers, like signals that employ orthogonal frequency-division multiplexing (OFDM) [26].

6 Coherent receiver architectures and functional capabilities

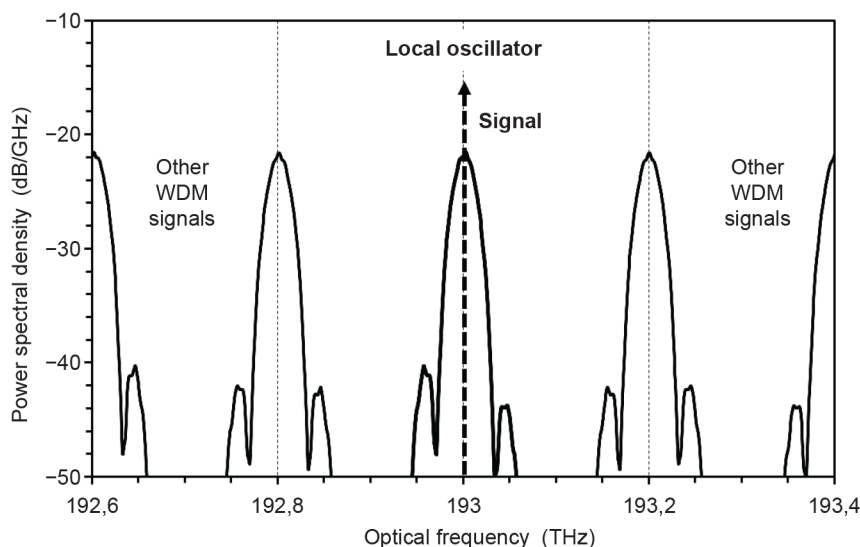
6.1 Basic principle of coherent detection

6.1.1 General

A coherent optical receiver mixes the transmitted optical signal with an optical reference signal from an unmodulated local-oscillator (LO) laser, which generates a beat signal that is proportional to the product of the optical amplitudes of the two signals [1], [2]. The two optical signals are usually mixed in one or more high-speed photodiodes (PDs), utilizing the nonlinear (i.e. quadratic) response of the photodetectors to generate the desired beat signal. The resulting photocurrents are proportional to the time-varying amplitude and phase of the transmitted optical signal, as described in more detail in 6.2, 6.3, and 6.4. In fact, the operation of a coherent receiver is very similar to that of super-heterodyne (or super-homodyne) receivers, which are often used in radio and microwave transmission systems to increase receiver sensitivity and to enhance the frequency selectivity of the receiver.

6.1.2 Homodyne and heterodyne detection

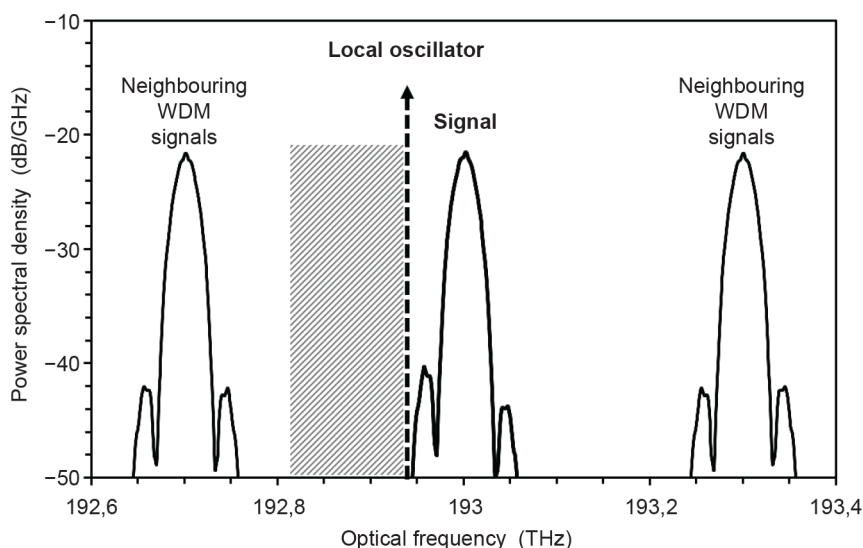
Coherent receivers can be operated as homodyne, heterodyne or intradyne detectors [2], [20]. In the case of homodyne detection, the reference signal has exactly the same optical frequency as the carrier frequency of the transmitted optical signal, as shown schematically in Figure 5 [26].



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Figure 5 – Signal and local oscillator frequencies for homodyne detection

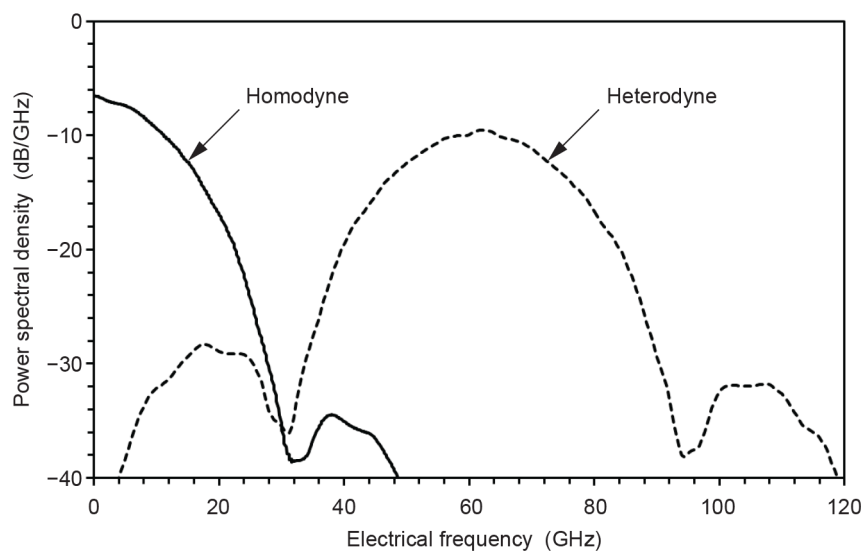
In the case of heterodyne detection, the reference frequency and the carrier frequency are offset by an amount that is slightly larger than the spectral width of the transmitted signal, as shown in Figure 6 [26]. Intradyne detection is very similar to homodyne detection, but it allows for relatively small frequency offsets between the reference signal and the carrier frequency of the transmitted optical signal [2]. Coherent fibre optic communication systems as well as optical modulation analysers typically use intradyne detection [9].



IEC

Figure 6 – Signal and local oscillator frequencies for heterodyne detection

In either case, mixing of the reference signal with the transmitted signal in the PD generates a photocurrent in the electrical baseband, having a single-sided spectrum, as shown schematically in Figure 7 for homodyne and heterodyne detection [26].



IEC

Figure 7 – Electrical spectra of homodyne and heterodyne beat signals

In the case of heterodyne detection, the carrier frequency of the transmitted signal is translated to an intermediate electrical frequency $|f_r - f_c|$, where f_r denotes the reference frequency and f_c the carrier frequency of the transmitted signal. If the two optical frequencies are sufficiently different, the entire (two-sided) spectrum of the transmitted optical signal can be translated to the single-sided electrical spectrum, as shown in the example of Figure 7, thus preserving the information contained in the upper and lower modulation sidebands of the transmitted signal.

6.1.3 Intradyne detection

In the case of homodyne detection, the reference frequency is exactly equal to the carrier frequency of the transmitted signal, that is $f_r = f_c$; whereas for intradyne detection, the reference frequency is only approximately equal to the transmitted carrier frequency, i.e. $f_r \approx f_c$ [2]. In both cases, the upper and lower modulation sidebands of the transmitted signal are folded into the single sideband of the electrical spectrum, which means that each frequency component of the resulting electrical spectrum is composed of two mixing products, one of which results from mixing of the LO laser with an optical frequency component in the lower modulation sideband of the transmitted signal at frequency f_l , and the other one from mixing of the LO laser with an optical frequency component in the upper modulation sideband of the signal at frequency f_u , with $f_r - f_l = f_u - f_r$.

In general, this spectral folding of the mixing products does not preserve all phase information contained in the transmitted optical signal. Hence, the decoding capabilities of a single coherent mixer are limited. To avoid these restrictions, modern coherent communication systems employ multiple intradyne mixers in the receiver, which generate mixing products at different optical phases between the LO laser output and the transmitted signal, as discussed in more detail in 6.3. This dual-mixer architecture is commonly referred to as a phase diversity receiver [27]. It should be noted that heterodyne receivers do not have these limitations but require photodiodes with twice the bandwidth of those used in intradyne receivers [26].

6.1.4 Polarization dependence

Heterodyne as well as homodyne mixers are sensitive to the polarization state of the transmitted signal. This effect causes fading of the mixing product when LO laser light and the received signal are in different polarization states. Such polarization fading can be avoided by employing additional mixers to generate mixing products between the LO laser and copies of the transmitted signal in different polarization states, which is known as polarization diversity [28]. Polarization-diversity receivers are discussed in more detail in 6.4 and 6.5.

It should be noted that heterodyne receivers typically require at least twice the electrical bandwidth of homodyne receivers, as can be clearly seen from Figure 7. While the bandwidth requirements for homodyne receivers are similar to those for direct-detection receivers (typically less than the symbol rate of the transmitted signal), heterodyne receivers need to have an electrical bandwidth of at least the entire width of the optical spectrum. Typically, the bandwidth of heterodyne receivers is substantially wider than twice the symbol rate of the transmitted signal, so as to avoid undesired signals distortion due to spectral folding of the mixing products [30]. In the example of Figure 7, the 3 dB bandwidth of the heterodyne receiver is about 3 times that of the homodyne receiver.

6.1.5 Frequency dependence

Coherent receivers in general are highly frequency dependent, just like radio or microwave homo- or heterodyne receivers. They hence need to be tuned to the frequency of the DWDM channel to be detected. The optical bandwidth of a coherent receiver is typically only as wide as the optical bandwidth of the transmitted signal [10].

Coherent receivers can thus serve as wavelength demultiplexers in DWDM systems. This feature is exploited, for example, in colourless drop ports of reconfigurable optical add-drop multiplexers (ROADMs) to select the wavelength channel of the signal to be received [30], which is simply accomplished by tuning the frequency of the LO laser to the desired WDM channel. To avoid interference from neighbouring DWDM signals, the channel spacing of the DWDM signals has to be larger than twice the bandwidth of the coherent receiver. Thus, coherent receivers based on homodyne detection allow much smaller WDM channel spacing than those using heterodyne detection, because the latter require unoccupied optical spectrum of the same width as the selected signal on the other side of the reference signal, as indicated by the hashed area in Figure 6. Hence, with heterodyne receivers, the number of useable wavelength channels is typically smaller than with homodyne receivers, thus limiting the transmission capacity of fibre optic DWDM communication systems [8].

6.1.6 Phase and polarization diversity

Early versions of coherent receivers employed only a single optical mixer to generate the desired beat signal. These simple receivers were very sensitive to polarization fluctuations in the transmitted signal and, in the case of homodyne receivers, required phase locking of the optical reference signal [1], [2]. These limitations made it quite difficult to use these receivers in commercial fibre-optic communication systems, where the fibre link usually does not preserve the state of polarization of the transmitted signals. However, it was shown that these problems can be circumvented by employing multiple mixers in a coherent receiver, as described in more detail in 6.3 and 6.4 [20].

Modern coherent receivers typically use intradyne detection and comprise four balanced coherent mixers to generate a set of four complementary beat signals, which carry information on the amplitude, phase, and polarization state of the transmitted signal [9]. These receivers are commonly referred to as phase- and polarization-diversity receivers [29]. It is important to note that these receivers facilitate the detection of complex vector-modulated signals, such as multi-level phase-shift keyed (PSK) or quadrature-amplitude modulated (QAM) signals, as well as the decoding of polarization-multiplexed signals, which can carry twice the information than conventional single-polarized signals without requiring additional optical bandwidth, as discussed in 5.3 [32]. Thus, coherent receivers have enabled a rapid increase in the transmission capacity of fibre optic communication systems [8].

However, a considerable amount of signal processing is needed to extract the transmitted information from the four beat signals, which is usually accomplished with the help of high-speed digital signal processors (DSPs), as discussed in more detail in Clause 7. The various types of coherent receivers are described in more detail in 6.2 to 6.4, beginning with single coherent mixers and ending with modern quadruple coherent mixers.

6.2 Single coherent mixer with balanced detection

6.2.1 Principle of operation

A single coherent mixer typically employs a symmetric 2×2 optical waveguide coupler (i.e. a 3 dB coupler), which combines the transmitted signal $\vec{A}_s(t)$ with the reference signal $\vec{A}_r(t)$, as shown schematically in Figure 8. The two output signals of the coupler are of the form $[\vec{A}_s(t) - j\vec{A}_r(t)]/\sqrt{2}$ and $[\vec{A}_r(t) - j\vec{A}_s(t)]/\sqrt{2}$, where $\vec{A}_s(t)$ is of the general form of Formula (1), and $\vec{A}_r(t)$ describes the optical amplitude of the reference signal whose polarization state is assumed to be fixed, as described by Formula (2).

$$\vec{A}_r(t) = \sqrt{P_{LO}} \begin{pmatrix} 1 \\ 0 \end{pmatrix} e^{j\Phi_r(t)} = \sqrt{P_{LO}} \begin{pmatrix} 1 \\ 0 \end{pmatrix} e^{j[\omega_r t + \varphi_r(t) + \hat{\varphi}_r(t)]} \quad (2)$$

where

P_{LO} is the optical power of the reference signal,

$\omega_r = 2\pi f_r$ the angular frequency of the reference signal,

$\varphi_r(t)$ is the optical phase of the reference signal, and

$\hat{\varphi}_r(t)$ is a random variable representing laser phase noise from the transmitter laser.

Subsequent mixing of the two output signals of the coupler in two separate high-speed photodiodes (PDs) generates the two photocurrents $i_1(t)$ and $i_2(t)$ given by Formula (3) [33].

$$\begin{aligned} i_1(t) &= R_1 \left\{ |\vec{A}_r(t)|^2 + |\vec{A}_s(t)|^2 + 2 \operatorname{Im} [\vec{A}_r(t) \cdot \vec{A}_s^*(t)] \right\} + \hat{i}_1(t) \\ i_2(t) &= R_2 \left\{ |\vec{A}_r(t)|^2 + |\vec{A}_s(t)|^2 - 2 \operatorname{Im} [\vec{A}_r(t) \cdot \vec{A}_s^*(t)] \right\} + \hat{i}_2(t) \end{aligned} \quad (3)$$

where

R_1 and R_2 represent the responsivities of the photodiodes, in units of A/W,

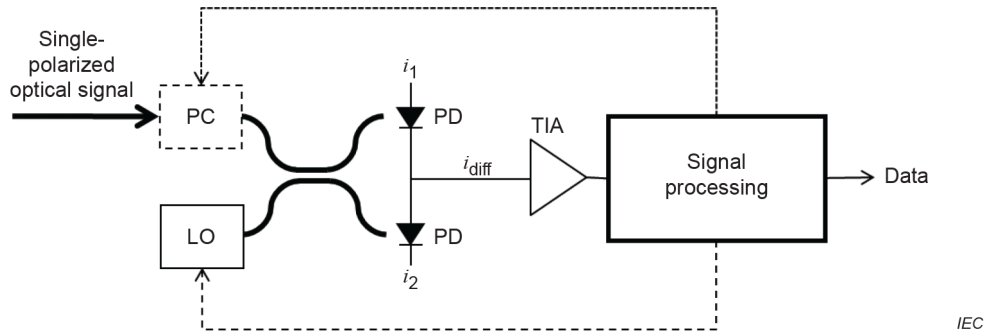
$\hat{i}_1(t)$ and $\hat{i}_2(t)$ are random variables that represent receiver noise,

$()^*$ denotes the complex conjugate, and

$() \cdot ()$ indicates the scalar vector product.

Note that R_1 and R_2 include here not only the photo-detection efficiency of the PDs but also the various optical splitting losses in the 3 dB coupler as well as optical insertion losses and attenuation in the receiver, so that the photocurrents are directly related to the input power of the received and reference signals [29].

NOTE In practice, the responsivity of each PD can be significantly different for light entering the signal input port of the mixer versus light entering the LO input port of the mixer, because of asymmetric insertion losses and additional optical elements (like optical attenuators and monitor PDs) in the signal input port. For the sake of simplicity, this difference is ignored in Formula (3).



Key

LO	Local oscillator laser
PC	Polarization controller
PD	Photodiode
TIA	Transimpedance amplifier

Figure 8 – Single balanced mixer for coherent reception

The first two terms on the right side of Formula (3) are undesired by-products of the mixing process, which result from non-coherent detection (also known as direct detection) of the reference signal and the received signal, whereas the third term represents the desired coherent beat signal. The random variables $\hat{i}_1(t)$ and $\hat{i}_2(t)$ represent receiver noise generated in the photo-detection process, including thermal and shot noise as well as input-referred amplifier noise [2], [20], [29], [34].

6.2.2 Common-mode rejection

With a perfectly balanced mixer, i.e. with $R_1 = R_2$, one can eliminate the undesired direct-detection products by subtracting the two photocurrents from each other to obtain the differential current $i_{\text{diff}}(t)$, as described by Formula (4) [20].

$$i_{\text{diff}}(t) = i_1(t) - i_2(t) = 2R_1 \operatorname{Im} \left[\vec{A}_r(t) \cdot \vec{A}_s^*(t) \right] + \hat{i}_{\text{tot}}(t) \quad (4)$$

This differential signal includes only the coherent beat signal and the receiver noise. The suppression of the direct detection products is known as common-mode rejection, since the direct-detection products have the same sign and magnitude in both photocurrents (common mode), whereas the coherent beat signals have opposite sign (differential mode). In practice, the common-mode rejection is limited, because of small receiver imbalance, $R_1 \neq R_2$, and differential time delays between the two photocurrents, which is known as skew. The amount of common-mode suppression in a balanced mixer is characterized by a common-mode-rejection ratio (CMRR), R_{CMR} , which is usually expressed in dB and calculated from Formula (5) [33].

$$R_{\text{CMR}} = 20 \log_{10} \left[\frac{i_1(t) - i_2(t)}{i_1(t) + i_2(t)} \right] \quad (5)$$

The CMRR can be measured by injecting light into only one of the two input ports of the mixer. In general, the signal input and the reference input yield different CMRR values, because of variations in the PD responsivities. High CMRR of the received signal is especially important for applications in colour-less ROADM ports, as discussed in 6.1.5, where the receiver sees a multitude of DWDM signals, of which only one is selected for coherent reception. In this case, all DWDM signals contribute to the direct-detection products in the photocurrents $i_1(t)$ and $i_2(t)$, as described by Formula (6).

$$\begin{aligned}
 i_1(t) &= R_1 \left[\left| \vec{A}_r(t) \right|^2 + \left| \vec{A}_s^{\lambda_1}(t) \right|^2 + \left| \vec{A}_s^{\lambda_2}(t) \right|^2 + \dots + \left| \vec{A}_s^{\lambda_n}(t) \right|^2 + 2 \operatorname{Im} \left\{ \vec{A}_r(t) \cdot \vec{A}_s^{*\lambda_1}(t) \right\} \right] + \hat{i}_1(t) \\
 i_2(t) &= R_2 \left[\left| \vec{A}_r(t) \right|^2 + \left| \vec{A}_s^{\lambda_1}(t) \right|^2 + \left| \vec{A}_s^{\lambda_2}(t) \right|^2 + \dots + \left| \vec{A}_s^{\lambda_n}(t) \right|^2 - 2 \operatorname{Im} \left\{ \vec{A}_r(t) \cdot \vec{A}_s^{*\lambda_1}(t) \right\} \right] + \hat{i}_2(t)
 \end{aligned} \tag{6}$$

where

n is the total number of DWDM signals,

$\vec{A}_s^{\lambda_m}(t)$ is the optical amplitude of the DWDM signal at wavelength λ_m , $m = 1 \dots n$, and

λ_1 is the wavelength to be coherently received.

The subtraction of the two photocurrents can be accomplished by connecting the two photodiodes in series and tapping the differential current at midpoint between the photodiodes, as shown in Figure 8. In practice, a differential transimpedance amplifier is often used to subtract the two photocurrents, as shown in Figure 9. In both cases, the losses and path lengths in the two mixer arms should be well matched, since there are usually no means for compensation of receiver imbalance or skew between the signals.

Imbalance between the two mixer arms degrades the CMRR at all frequencies, while skew introduces frequency dependence in the CMRR, degrading it more at higher frequencies than at lower frequencies. For this reason, the CMRR is often specified for different frequencies. Typical limits for the CMRR are 16 dB at low frequencies and 10 dB at high frequencies. Receivers designed to operate at symbol rates of around 25 GBd, for example, should have less than 2 ps differential time delay (or skew) between the two photodetector signals $i_1(t)$ and $i_2(t)$, which corresponds to a path length difference of less than 0,4 mm. These tight tolerances are very difficult to meet when using fibre-pigtailed optical couplers. Typically, coherent receivers for fibre-optic communication systems employ compact planar waveguide circuits for the optical splitter(s) and monolithically or hybrid integrated photodiodes that are directly connected to transimpedance amplifiers.

Furthermore, the signal-to-noise ratio (SNR) of balanced receivers is 3 dB higher than that of equivalent single-ended receivers using only one photodiode, because the amplitude of the beat signal in Formula (4) is twice as large as that in Formula (3), whereas the RMS amplitude of the total noise current in Formula (4) is only 41 % higher than that of the individual noise current in Formula (3). Assuming $\hat{i}_1(t) \approx \hat{i}_2(t)$, the total noise in the differential current is given by Formula (7).

$$\hat{i}_{\text{tot rms}} = \sqrt{\langle \hat{i}_1^2(t) \rangle + \langle \hat{i}_2^2(t) \rangle} \approx \sqrt{2} \hat{i}_{1\text{rms}} \tag{7}$$

where the brackets $\langle \dots \rangle$ indicate time averaging.

The optical power of the reference signal is typically in the range of 10 mW to 40 mW and thus much larger than the average power of the received signal, which can range from 0,01 mW to 1 mW, so that the reference signal amplifies the coherently received beat signal. As a result, coherent receivers usually have significantly higher sensitivity than direct detection receivers. In many cases, the power of the reference signal is high enough, so that the receiver noise is dominated by shot noise resulting from direct detection of the reference signal. The RMS amplitude of this shot noise, $\hat{i}_{\text{tot rms}}$, is given by Formula (8) [20].

$$\hat{i}_{\text{tot rms}} \approx \sqrt{2qRP_{\text{LO}}B} \quad (8)$$

where

$q \approx 1,60 \times 10^{-19} \text{ C}$ is the elementary charge,
 B is the electrical bandwidth of the photodiode (in Hz), and
 P_{LO} is the optical power of the reference signal from the LO laser (in W).

When the receiver noise is dominated by shot noise, the signal-to-noise ratio of the coherently received beat signal becomes independent of P_{LO} .

6.2.3 Polarization dependence

As discussed in 6.1.4, the magnitude of the coherent beat signal depends strongly on the polarization state of the received signal. It is maximal only when the received signal is in the same polarization state as the reference signal from the LO laser. The beat signal vanishes when the two signals are in orthogonal polarization states. This strong polarization dependence is very undesirable in fibre-optic communication systems, because the transmission fibres usually do not maintain the state of polarization of the transmitted signal [28], [32]. Hence, a single-polarized optical signal usually arrives at the receiver in an arbitrary polarization state, characterized by a polarization angle θ and a differential phase ψ , as described by Formula (9), where the phase parameters θ and ψ can both fluctuate randomly with time.

$$\vec{A}_s(t) = \sqrt{P_S} \begin{pmatrix} \cos(\theta) \\ \sin(\theta) e^{j\psi} \end{pmatrix} a_s(t) e^{j\Phi_s(t)} \quad (9)$$

The reference signal from the local oscillator laser is usually inserted into the coupler via polarization-maintaining fibre, so that it is always in a fixed and well-defined polarization state, as described by Formula (2).

For pure amplitude modulation with real-valued $a_s(t)$, Formula (4) becomes Formula (10).

$$i_{\text{diff}}(t) = R \sqrt{P_{\text{LO}} P_S} \cos(\theta) a_s(t) \sin[\Phi_r(t) - \Phi_s(t)] + \hat{i}_{\text{tot}}(t) \quad (10)$$

It is readily seen from Formula (10) that the amplitude of the coherent beat signal is proportional to $\cos(\theta)$ and, hence, completely vanishes when $\theta = \pm 90^\circ$. This effect is known as polarization-dependent signal fading. It can be avoided by placing an optical polarization controller (PC) in front of the coherent mixer, as indicated in Figure 8, and by continuously adjusting this controller to maximize the amplitude of the coherent beat signal [35]. Such automatic optical polarization controllers would add considerable complexity to the coherent receiver. Instead, as described in more detail in 6.4, modern coherent receivers employ four balanced coherent mixers which are operated in parallel to detect four different polarization components simultaneously [9]. The desired beat signal can then be reconstructed from these four signal components with the help of digital signal processing. It should also be noted that single coherent mixers cannot decode polarization-multiplexed signals, which are often used to transmit signals at line rates of 100 Gbit/s and above [32].

6.2.4 Homodyne detection

Single mixers can be operated either as homodyne or as heterodyne receivers. In the case of homodyne detection, the optical frequency f_r of the reference signal is exactly equal to the carrier frequency f_s of the transmitted optical signal, so that its optical phase is locked to that of the carrier frequency [1], [2]. With homodyne detection, the simple coherent receiver of Figure 8 can detect only signals that carry either pure amplitude (or intensity) modulation (AM) or, alternatively, pure binary phase modulation, but not quadrature amplitude modulation (QAM) or other combinations of amplitude and phase modulation. In the case of pure AM, having real-valued amplitude variations $a_s(t)$ but constant phase φ_s (e.g. $\varphi_s \equiv 0$), Formula (10) reduces to Formula (11).

$$i_{\text{diff}}(t) = R \sqrt{P_{\text{LO}} P_{\text{S}}} \cos(\theta) a_s(t) \sin[\varphi_r + \hat{\varphi}_{\text{tot}}(t)] + \hat{i}_{\text{tot}}(t) \quad (11)$$

where

$\hat{\varphi}_{\text{tot}}(t) = \hat{\varphi}_r(t) - \hat{\varphi}_s(t)$ is the combined phase noise of the transmitter and local oscillator lasers.

The beat signal of Formula (11) is proportional to the time-varying amplitude of the transmitted signal, as desired, but it is also proportional to $\sin(\varphi_r)$, so that its magnitude depends on the relative optical phase between the reference signal and the optical carrier of the transmitted signal. The beat signal is maximal only when $\varphi_r = 90^\circ$ (or alternatively -90°) and vanishes when $\varphi_r = 0^\circ$ or 180° . Thus, homodyne detection with a single balanced mixer requires active locking of the optical phase of the local oscillator phase, which is quite difficult to accomplish in practice. However, the single balanced mixer is capable of detecting arbitrary AM, in particular higher-order pulse-amplitude modulation (PAM).

The transmitter and local oscillator lasers used in a coherent transmission system should exhibit low phase noise, so that $|\hat{\varphi}_{\text{tot}}(t)|$ is sufficiently small and does not introduce significant additional amplitude variations in the beat signal, which would distort the received signal [36]. In general, coherent detection requires transmitter and local oscillator lasers with narrow linewidth [10]. The amount of acceptable phase noise depends on the type and cardinality of the modulation format as well as on the symbol rate of the transmitted signal [37]. In general, the higher the cardinality of the modulation format, the less phase noise from the transmitter and LO lasers can be tolerated. Moreover, the higher the symbol rate, the more phase noise can be tolerated from the transmitter laser, but not necessarily from the LO laser [38].

NOTE There are additional (and often stricter) requirements for the phase noise of the LO laser when chromatic dispersion is compensated in the coherent receiver, as described in more detail in 6.3.3 and 6.4.6. As a result, the tolerance to LO laser phase noise typically decreases with increasing symbol rate.

Similar considerations apply to purely phase modulated signals, which are of the form shown in Formula (12).

$$\vec{A}_s(t) = \sqrt{P_{\text{S}}} \begin{pmatrix} \cos(\theta) \\ \sin(\theta) e^{j\psi} \end{pmatrix} e^{j\varphi_s(t)} e^{j[\omega_s t + \hat{\varphi}_s(t)]} \quad (12)$$

In this case Formula (10) becomes Formula (13).

$$i_{\text{diff}}(t) = R \sqrt{P_{\text{LO}} P_{\text{S}}} \cos(\theta) \sin[\varphi_s(t) - \varphi_r + \hat{\varphi}_{\text{tot}}(t)] + \hat{i}_{\text{tot}}(t) \quad (13)$$

In contrast to the case of AM, the phase modulation cannot be arbitrary. A single balanced mixer is capable of detecting binary phase-shift keying (BPSK), where $\varphi_s(t)$ is toggled between only two phases, for example 0° and 180° (assuming $\varphi_r = 90^\circ$, as above), but it cannot decode quaternary phase-shift keying (QPSK), where $\varphi_s(t)$ is toggled between the four phases 0° , 90° , 180° and 270° , since $i_{\text{diff}}(t)$ would be zero when $\varphi_s(t)$ is 90° and 270° . This is a direct result of the spectral folding that occurs in single homodyne mixers (see 6.1.2).

6.2.5 Heterodyne detection

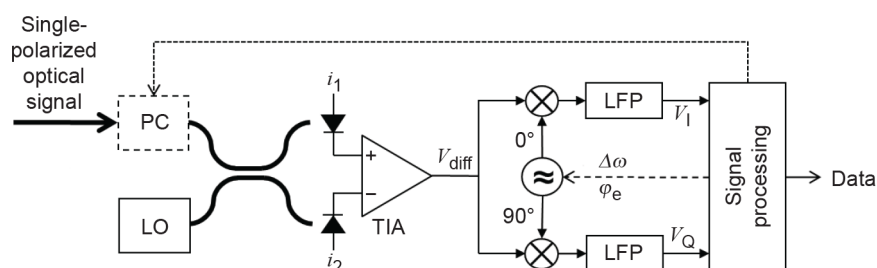
In the case of heterodyne detection with $\omega_s \neq \omega_r$, the transmitted signal can carry arbitrary AM and/or PM, and thus arbitrary QAM, so that the received differential photocurrent is given by Formula (14).

$$i_{\text{diff}}(t) = R \sqrt{P_{\text{LO}} P_{\text{S}}} \cos(\theta) a_s(t) \sin \left[\Delta\omega t + \varphi_r - \varphi_s(t) + \hat{\varphi}_{\text{tot}}(t) \right] + \hat{i}_{\text{tot}}(t) \quad (14)$$

where

$\Delta\omega = \omega_r - \omega_s$ denotes the frequency offset between the reference signal and the carrier frequency of the transmitted signal.

The phase and amplitude information encoded in the transmitted signal can be recovered, for example, by down-mixing the output signal of the transimpedance amplifier, $V_{\text{diff}}(t) = Z i_{\text{diff}}(t)$, simultaneously with the in-phase and quadrature outputs of a sinusoidal oscillator at frequency $\Delta\omega$, i.e. by down-mixing with the electrical signals $\sin(\Delta\omega t + \varphi_e)$ and $\cos(\Delta\omega t + \varphi_e)$, as shown schematically in Figure 9 [1].



IEC

Key

- LO Local oscillator laser
- LFP Low-pass filter
- PC Polarization controller
- TIA Transimpedance amplifier

Figure 9 – Balanced heterodyne mixer with electrical down-mixing

With $\varphi_e = \varphi_r - 90^\circ$ and after low-pass filtering, to reject all frequency components above $\Delta\omega$, one obtains the quadrature signals $V_I(t)$ and $V_Q(t)$ described by Formula (15).

$$\begin{aligned}
 V_I(t) &= \frac{Z R \sqrt{P_{LO} P_S} \cos(\theta)}{2} a_s(t) \sin[\varphi_s(t) - \hat{\varphi}_{\text{tot}}(t)] + \hat{V}_1(t) \\
 V_Q(t) &= \frac{Z R \sqrt{P_{LO} P_S} \cos(\theta)}{2} a_s(t) \cos[\varphi_s(t) - \hat{\varphi}_{\text{tot}}(t)] + \hat{V}_2(t)
 \end{aligned} \tag{15}$$

where

$\hat{V}_1(t)$ and $\hat{V}_2(t)$ represent the receiver noise after down-mixing and filtering, and Z is the transimpedance of the TIA (in units of V/A or Ω).

The two signals $V_I(t)$ and $V_Q(t)$ comprise all information on the amplitude and phase modulation in the transmitted signals, which can be seen by combining the two signals into a complex vector as shown in Formula (16).

$$V_I(t) + jV_Q(t) = \frac{Z R \sqrt{P_{LO} P_S} \cos(\theta)}{2} a_s(t) e^{j[\varphi_s(t) - \hat{\varphi}_{\text{tot}}(t)]} + \hat{V}_{\text{noise}}(t) e^{j\hat{\varphi}_{\text{noise}}(t)} \tag{16}$$

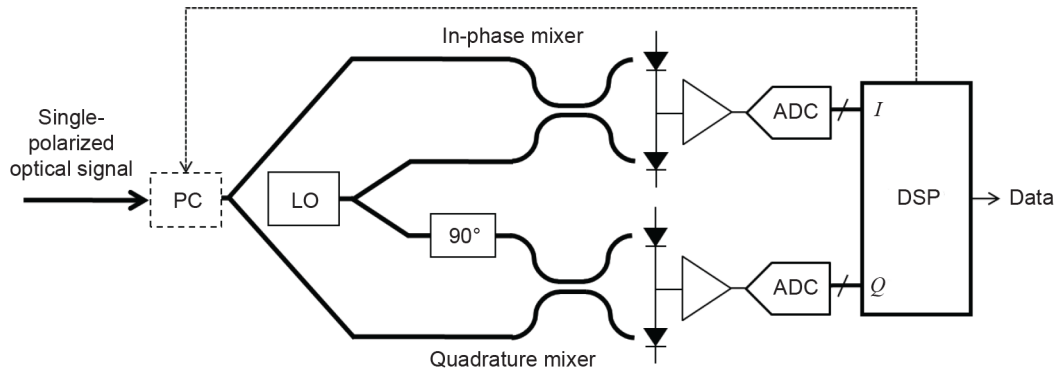
Aside from the additional noise terms on the rightmost side, Formula (16) resembles the complex phasor of the transmitted signal given in Formula (9). The down-mixing and filtering can be accomplished with the help of analogue electronics, as shown in Figure 9, or equivalently in a high-speed digital signal processor, after the analogue voltage $V_{\text{diff}}(t)$ is converted into a digital signal by a high-speed analogue-to-digital converter (ADC).

Hence, the simple heterodyne receiver of Figure 9 allows decoding of arbitrary M -QAM signals, without requiring phase-locking of the LO laser. Any drift in the frequency and phase of the transmitter and LO lasers can be compensated for by adaptive adjustment of the frequency $\Delta\omega$ and phase φ_e of the electrical oscillator in the receiver. However, heterodyne receivers require PDs and TIAs with at least twice the electrical bandwidth of a comparable homodyne receiver, which can limit the symbol rate of the transmitted signal [30]. In most applications, a polarization controller is required to avoid polarization-induced signal fading in the heterodyne receiver [35]. However, it should be noted that polarization-fading can also be avoided by encoding the information data into a dual-polarization signal, using for example Alamouti coding, in such a way that the coherently received beat signal does not depend on the polarization state of the transmitted signal at the receiver [39].

6.3 Dual coherent mixer with phase diversity

6.3.1 Principle of operation

The capabilities of coherent receivers can be greatly enhanced by employing multiple mixers in parallel, so as to recover more properties of the received optical signal. As a first step, in 6.3 a coherent receiver with phase diversity is considered. Such phase-diversity receivers typically employ two balanced mixers, an “in-phase mixer” and a “quadrature” mixer, which are operated in parallel [20], [29]. Either of the two mixers is similar to the single balanced mixer of Figure 8, but the quadrature mixer includes an additional optical phase shifter that introduces a 90° optical phase shift in the reference signal, as shown in Figure 10. This 90° phase shift can be generated, for example, by adding a suitable length of additional optical waveguide in the second mixer, so as to delay the reference signal by an amount that corresponds to a 90° shift in optical phase.



IEC

Key

- 90° Optical phase shifter
- ADC Analogue-to-digital converter
- DSP Digital signal processor
- LO Local oscillator laser
- PC Polarization controller

Figure 10 – Dual coherent mixer with phase diversity

The transmitted signal and the reference signal from the LO laser are both split into two equal copies, which are then fed symmetrically into two parallel mixers, as shown in Figure 10. Hence, the in-phase mixer combines the transmitted signal with the 0°-shifted copy of the reference signal, whereas the quadrature mixer combines the transmitted signal with the 90° phase-shifted copy of the reference signal, thus creating a coherent beat signal whose electrical phase is always shifted by 90° relative to the beat signal created by the in-phase mixer (phase quadrature). This combination of in-phase and quadrature mixer is often referred to as a 90° hybrid mixer, a term that was adopted from RF and microwave technologies [20].

It should be noted that the arrangement shown in Figure 10 is not the only way to implement a phase-diversity receiver. If the transmitted signal and the reference signal are combined in a three-branch equal-split directional coupler, for example, the three output ports of this coupler produce beat signals with relative phase shifts of 0°, 120°, and 240° [20], [29]. However, most practical applications employ balanced 90° hybrid mixers.

Assuming homodyne reception with $\Delta\omega = 0$ and $\varphi_r = 0$, as well as otherwise identical path lengths in the two mixers, the in-phase and quadrature mixers in Figure 10 generate two phase shifted photocurrents $i_I(t)$ and $i_Q(t)$, which are given in Formula (17) [1], [29].

$$\begin{aligned} i_I(t) &= R \sqrt{P_{LO} P_S} \cos(\theta) a_s(t) \sin \left[-\varphi_s(t) + \hat{\varphi}_{tot}(t) \right] + \hat{i}_{tot}(t) \\ i_Q(t) &= R \sqrt{P_{LO} P_S} \cos(\theta) a_s(t) \cos \left[-\varphi_s(t) + \hat{\varphi}_{tot}(t) \right] + \hat{i}_{tot}(t) \end{aligned} \quad (17)$$

These two photocurrents have the same form as the two down-mixed heterodyne signals of Formula (15). Therefore, a dual homodyne mixer with phase diversity provides the same functionality as a single heterodyne mixer. Hence, it allows detection of arbitrary M -QAM signals (including QPSK) without needing the wide electrical bandwidth of a heterodyne receiver. However, splitting of the reference and transmitted signals between the two mixers reduces the effective responsivity R by 50 %, compared with Formula (10), thus yielding substantially smaller photocurrents in each mixer than for the single mixer in Formula (10). The amplitude of each photocurrent can be raised to the signal level of a single mixer by increasing the power of the reference signal and the transmitted signal both by 3 dB or, alternatively, by increasing the

power of only the reference signal by 6 dB, where the latter would increase the signal-to-noise ratio (SNR) of the two signals in Formula (17).

When used with ASK or BPSK signals, the phase-diversity mixer of Figure 10 introduces more noise in the photocurrents than the single balanced mixer of Figure 8 [27], [29]. This can be readily seen in the case of homodyne detection with a phase-locked LO laser, as discussed in 6.2.4. If the reference phase φ_r is adjusted in such a way that only $i_I(t)$ carries the desired beat signal, then $i_Q(t)$ only carries receiver noise (e.g. shot noise), which is an undesired addition to the received signal. However, this disadvantage vanishes when quadrature modulated signals (such as QPSK or M -QAM) are received [27].

However, the phase-diversity mixer of Figure 10 is much more versatile than a single mixer, because it allows homodyne detection with small frequency offsets between the reference signal and the transmitted signal as well as compensation of I - Q skew and chromatic dispersion, as explained in 6.3.2 through 6.3.5.

6.3.2 Intradyne detection with frequency offset removal

In contrast to the single mixer discussed in 6.2, the phase-diversity mixer of Figure 10 allows homodyne detection without requiring optical phase locking of the LO laser to the optical carrier of the transmitted signal [29]. In fact, the LO laser does not even have to be operated at exactly the same frequency as the transmitter laser, because the two signals $i_I(t)$ and $i_Q(t)$ carry the information needed to remove undesired frequency and phase offsets [2]. This mode of operation is commonly referred to as intradyne reception.

To demonstrate this capability of the phase-diversity mixer, the parameters $\Delta\omega$ and φ_r are used to describe the frequency and phase offsets between the reference signal and the transmitted signal, so that the two photocurrents of Formula (17) become $\tilde{i}_I$ and $\tilde{i}_Q$ as described by Formula (18).

$$\begin{aligned}\tilde{i}_I(t) &= R \sqrt{P_{LO} P_S} \cos(\theta) a_s(t) \sin[\Delta\omega t + \varphi_r - \varphi_s(t) + \hat{\varphi}_{tot}(t)] + \hat{i}_{tot}(t) \\ \tilde{i}_Q(t) &= R \sqrt{P_{LO} P_S} \cos(\theta) a_s(t) \cos[\Delta\omega t + \varphi_r - \varphi_s(t) + \hat{\varphi}_{tot}(t)] + \hat{i}_{tot}(t)\end{aligned}\quad (18)$$

If the frequency offset $\Delta\omega$ and the phase offset φ_r are known or can be measured, the beat signals of Formula (18) can be transformed into the form of Formula (17) by applying the matrix multiplication described by Formula (19).

$$\begin{bmatrix} i_I(t) \\ i_Q(t) \end{bmatrix} = \begin{bmatrix} \sin(\Delta\omega t + \varphi_r) & \cos(\Delta\omega t + \varphi_r) \\ \cos(\Delta\omega t + \varphi_r) & -\sin(\Delta\omega t + \varphi_r) \end{bmatrix} \times \begin{bmatrix} \tilde{i}_I(t) \\ \tilde{i}_Q(t) \end{bmatrix}\quad (19)$$

This operation can be performed in real time, using either analogue multipliers or, alternatively, a high-speed digital signal processor (DSP) when the two beat signals are digitized via high-speed ADCs [40]. However, since $\Delta\omega$ and φ_r can fluctuate with time, especially when independent lasers are used for the transmitter and local oscillator, it is usually necessary to continuously adjust the parameters in Formula (19) to completely remove the frequency and phase offsets [40].

The frequency and phase offsets are usually not known a priori, but they can be estimated in the DSP using various methods [41], [42]. Relatively large frequency offsets, for example, can be estimated from a spectral analysis of the two photocurrents $\tilde{i}_I(t)$ and $\tilde{i}_Q(t)$ by Fourier transforming the complex signal $\tilde{i}_c(t) = \tilde{i}_I(t) + j\tilde{i}_Q(t)$, which yields a two-sided electrical spectrum $\tilde{S}_c(\omega)$ of the coherent beat signal.

Such a Fourier transform is often calculated in DSPs when chromatic dispersion is to be compensated, as discussed in more detail in 6.3.3. The frequency offset estimation can thus be performed in the same signal processing step. However, compensation of large frequency offsets should be performed before chromatic dispersion is compensated (see 7.2).

For most modulation formats, the transmitted signal spectrum is symmetric and centred at zero when $\Delta\omega = 0$. Hence, if a frequency offset $\Delta\omega$ is present, it can be estimated from the position of the central frequency of the spectrum, as shown schematically in Figure 11 [43]. Another method for coarse frequency offset estimation involves calculation of the time-varying phase changes in $\tilde{i}_c(t)$ [41].

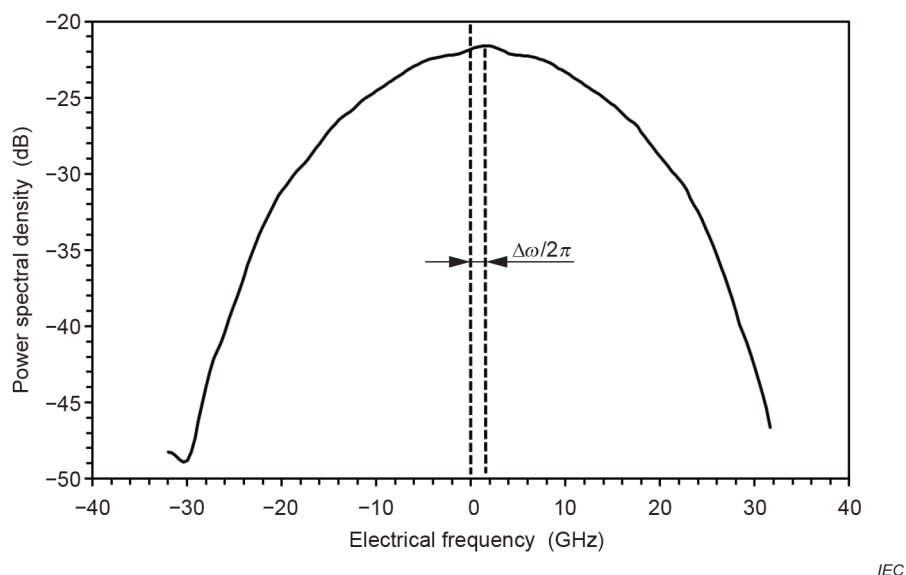


Figure 11 – Intradyn beat spectrum with 2 GHz frequency offset

Likewise, there are several methods to estimate time-varying phase offsets in $\tilde{i}_c(t)$ (including small frequency offsets), most of which strongly depend on the particular modulation format used for signal transmission [42]. Phase offset removal is usually referred to as carrier phase recovery [44]. An exemplary method for carrier phase recovery is described in 6.4.5.

With phase-diversity receivers, it is thus possible to use independently frequency-stabilized lasers for the transmitter and local oscillator, like integrable tuneable laser assemblies (ITLAs), which are frequently employed in coherent transmission systems [19]. These lasers typically have a frequency accuracy of better than $\pm 2,5$ GHz, so that the worst-case frequency difference between the transmitter and local oscillator lasers will not exceed 5 GHz. This offset is small compared with the total bandwidth of a 25 GBd optical signal.

6.3.3 Compensation of chromatic dispersion

Fibre optic communication links often exhibit considerable amounts of chromatic dispersion (CD), especially those employing dispersion-unshifted single-mode fibres, such as B-652, B-654, and B-657 fibres [45]. Chromatic dispersion causes optical signals at different frequencies to propagate at different speeds through the fibre optic link. When the amount of group velocity dispersion (GVD) is large enough, the frequency-dependent propagation speed can introduce significant differential time delays between the various frequency components of a modulated single-carrier optical signal, which in turn can cause severe waveform distortions in the transmitted signal [46]. Formula (20) shows how to calculate the CD-induced differential time delay Δt_{CD} between two frequency components of angular frequency difference $\Delta\omega_s$ [46].

$$\Delta t_{CD}(\Delta\omega_s) = \frac{2\pi D_L c \Delta\omega_s}{\omega_s^2} \quad (20)$$

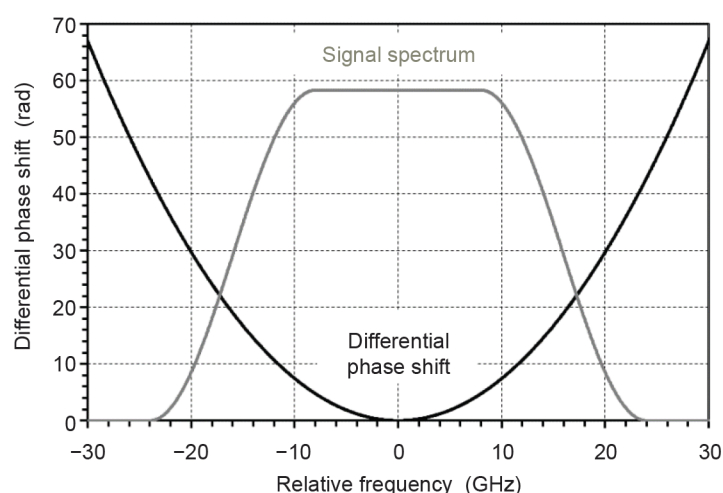
where

D_L denotes the total GVD which the signal has experienced (in units of s/m),

c is the speed of light in vacuum (approximately 299 792 458 m/s), and

ω_s is the angular carrier frequency of the signal (in units of rad/s).

When the frequency-dependent differential time delays Δt_{CD} are large enough, they introduce undesired differential phase shifts across the signal spectrum. This is shown in Figure 12 for the example of $D_L = 3\,000$ ps/nm, which is approximately the GVD in about 180 km of dispersion-unshifted B-652 fibre at 1,5 μm wavelength [45]. For comparison, the grey curve in Figure 12 shows the spectral shape of a 128 Gb/s PM-QPSK signal that has been filtered with a root-raised-cosine function having a roll-off factor 0,5 [12]. It is clearly seen from Figure 12 that the differential phase shifts at the edges of the optical spectrum increase rapidly with increasing bandwidth of the optical signal spectrum and, consequently, with increasing modulation rate of the transmitted signal.



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Figure 12 – Differential phase shifts introduced by 3 000 ps/nm GVD

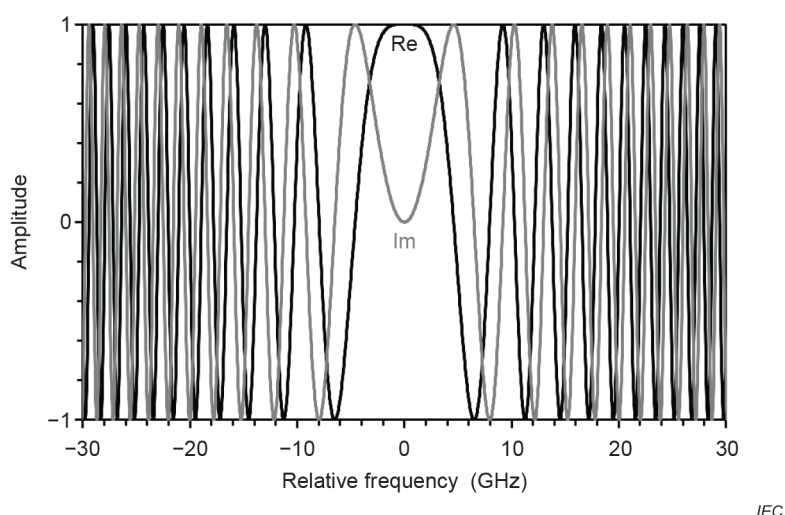
The effects of CD on a modulated signal can be described in the frequency domain by a complex transfer function $H(\omega_s + \Delta\omega_s)$, which is given in Formula (21) [48].

$$H(\omega_s + \Delta\omega_s) = \exp \left\{ j \frac{\pi D_L c \Delta\omega_s^2}{\omega_s^2} \right\} \quad (21)$$

where

$\Delta\omega_s$ is the angular frequency of a spectral component relative to the carrier frequency ω_s

The transfer function $H(\omega_s + \Delta\omega_s)$ is a rapidly oscillating function, whose oscillation frequency increases with increasing frequency offset $\Delta\omega_s$ as well as with increasing D_L [10]. These oscillations are shown in Figure 13 for the example of $D_L = 3\,000$ ps/nm around an optical frequency of 195 THz. The frequency of these oscillations increases rapidly with increasing frequency offset $\Delta\omega_s$.



Key

Re Real part

Im Imaginary part

Figure 13 – Transfer function for GVD of 3 ns/nm in frequency domain

Since the differential phase shifts Δt_{CD} are largest at the edges of the optical signal spectrum (see Figure 12), the CD-induced waveform distortions increase rapidly with increasing modulation frequency.

However, with a phase-diversity mixer it is possible to remove these undesired differential phase shifts by introducing similar phase shifts of opposite sign in the complex beat signal $i_c(t) = i_I(t) + j i_Q(t)$. This can be accomplished by multiplying the spectral components $S_c(\Delta\omega_s)$ of $i_c(t)$ with the transfer function $H^{-1}(\Delta\omega_s)$ given in Formula (22), which is the inverse function of Formula (21) [48].

$$H^{-1}(\Delta\omega_s) = \exp \left\{ -j \frac{\pi D_L c \Delta\omega_s^2}{\omega_s^2} \right\} \quad (22)$$

This procedure requires a Fourier transformation of $i_c(t)$ into the frequency domain and, after multiplication with $H^{-1}(\Delta\omega_s)$, an inverse a Fourier transformation back into the time domain [4]. Alternatively, one can remove the undesired effects of CD in the time domain, where the spectral transfer function $H^{-1}(\omega)$ corresponds to a convolution of the complex beat signal $i_c(t)$ with the transfer function $h^{-1}(t)$ given in Formula (23) [49].

$$h^{-1}(t) = \sqrt{\frac{\omega_s^2}{2\pi |D_L| c}} \exp \left\{ j \frac{\omega_s^2 t^2}{4\pi D_L c} \right\} \quad (23)$$

The convolution of $i_c(t)$ with $h^{-1}(t)$ is given by Formula (24).

$$\left[h^{-1} * i_c \right](t) \equiv \int_{-\infty}^{\infty} h^{-1}(\tau) i_c(t - \tau) d\tau \approx \int_{-\Delta t_{\max}}^{\Delta t_{\max}} h^{-1}(\tau) i_c(t - \tau) d\tau \quad (24)$$

where

$\left[h^{-1} * i_c \right](t)$ denotes the convolution of the function $h^{-1}(t)$ with the function $i_c(t)$.

The complex transfer function $h^{-1}(t)$ also exhibits rapid oscillations, whose frequency increases rapidly with time, as depicted in Figure 14, but decreases with increasing GVD D_L [50]. The integration range of the convolution in Formula (24) should be confined to the maximal differential time delays that are introduced in the optical spectrum of the transmitted signal, i.e. to the interval $-\Delta t_{\max}$ to $+\Delta t_{\max}$, with $\Delta t_{\max} = D_L (\lambda_{\max} - \lambda_{\min})/2$, where $\lambda_{\max}$ and $\lambda_{\min}$ denote the smallest and largest wavelengths at the ends of the signal spectrum [51].

Hence, the larger D_L , the longer $\Delta t_{\max}$ has to be, and the more transmitted symbols are involved in the convolution. This can be seen from Figure 14, which displays the real and imaginary parts of $h^{-1}(t)$ tailored to compensate 3 000 ps/nm and 10 000 ps/nm GVD in a 128 Gb/s PM-QPSK signal, which has the spectrum of Figure 12. In the case of 3 000 ps/nm GVD the correlation time is $2\Delta t_{\max} \approx 1\,200$ ps, whereas for 10 000 ps/nm it is $2\Delta t_{\max} \approx 4\,000$ ps. Thus, a GVD of 10 000 ps/nm correlates the amplitudes of about 128 symbols in the 128 Gb/s PM-QPSK signal.

Note that the number of correlated bits increases with the square of the symbol rate. In a similarly filtered 256 Gb/s PM-QPSK signal, therefore, a GVD of 10 000 ps/nm would correlate the amplitudes of about 512 symbols, which is four times the number of symbols correlated in a 128 Gb/s signal, because of a twofold increase in $2\Delta t_{\max}$ and a twofold increase in symbol rate.