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Signal Quality Monitoring for Optical Communications Networks

By Edward Anthony Flood

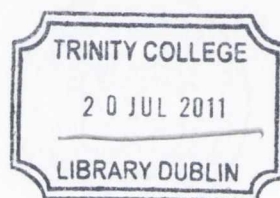
A thesis submitted for the degree of Doctor of Philosophy in the
University of Dublin.



Under the supervision of:
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November 2010



THESIS
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Declaration

This thesis has not been submitted as an exercise for a degree at this or any other university. The work contained herein is entirely my own except for the assistance of those mentioned in the acknowledgements. I agree that Trinity College library may lend or copy this thesis on request.

Abstract

The subject of this thesis is optical performance monitoring for transparent optical communications networks. To this end, a series of techniques for the characterisation of ultrashort telecommunications pulses were developed. Two phase-sensitive pulse monitoring techniques were employed to fully recover the electric field of 2ps pulses (i.e. the phase and spectrum) generated using a modelocked laser with varying levels of chirp. Both schemes involved the use of sinusoidal optical phase modulation followed by harmonic decomposition of the measured spectrum, which had been modified by the phase modulation. The scheme was validated through comparison with a standard SHG-FROG system and compared by way of simulation with Kang and Dorrer's sinusoidal phase modulation scheme.

A second scheme based on harmonic decomposition of a spectrum modified by sinusoidal phase modulation was investigated. This was similar in principle to our method mentioned above but with the added advantage of lower characterisation times as a result of only requiring one measurement of the modified spectrum. This scheme was validated through comparison with our initial sinusoidal phase modulation method. It was seen that a good agreement could be found between both of our sinusoidal phase modulation schemes but a comparison to similar schemes mentioned in the literature proved difficult as they could only recover the phase and intensity of highly chirped pulses with any reasonable accuracy due to signal to noise issues.

Two interferometric methods for characterisation of the in-band optical signal to noise ratio of communications signals were also investigated. The first of these involved a linear polariser in conjunction with a Michelson fibre interferometer. Measurements of the interferogram were taken for two orthogonal polarisation states and the in-band OSNR could be calculated from these. A measurement range from 10 to 25dB ± 0.5 dB

OSNR was achieved with such an arrangement for both OOK and DPSK signals at 10G with and without the addition of 680ps/nm dispersion. A second scheme was also investigated. This involved the use of a pair of Michelson fibre interferometers with different optical delays to recover the in-band OSNR without prior knowledge of the signal's coherence properties. A measurement range of up to $30\text{dB} \pm 0.5\text{dB}$ OSNR could be attained for amplitude modulated OOK signals and the scheme was seen to be free of influence from polarised noise, chromatic dispersion, signal extinction ratio degradation and data rates of up to 10G. The scheme was also tested for its resilience to phase modulation formats. The retrieval of the in-band OSNR from single and dual polarisation phase modulated QPSK signals proved more challenging than OOK signals as they are spectrally broader than intensity modulated signals and a measurement range up to approximately $22\text{dB} \pm 1\text{dB}$ could only be attained. However, no difference was seen in the results between the single and dual polarisation QPSK signals. The use of different optical delay values was investigated by way of simulation to find a range of delays that would allow measurement of 10G and 40G OOK and QPSK signals. It was seen that the use of up to six optical delays could enhance the measurement range significantly.

List of Publications and Presentations

- [1] E. Flood, W. Guo, D. Reid, M. Lynch, A. Bradley, L. Barry, and J. Donegan, *"In-band OSNR monitoring using a pair of Michelson fiber interferometers,"* Optics Express **18**, 3618-3625 (2010).
- [2] E. Flood, W. Guo, D. Reid, M. Lynch, A. Bradley, L. Barry, and J. Donegan, *"OSNR Monitoring Using Two Fiber Interferometers,"* in Optical Fiber Communication Conference, OSA Technical Digest (CD) (Optical Society of America, 2010), paper OWB3.
- [3] Edward A. Flood, W. H. Guo, A. L. Bradley, M. Lynch, D. Reid, L. P. Barry, J. F. Donegan, *"OSNR Monitoring Using Two Fibre Interferometers,"* Conference on Lasers and Electro-Optics, 16-MAY-10 - 21-MAY-10, San Jose, USA
- [4] E. Flood, W.H. Guo, D. Reid, M. Lynch, A.L Bradley, L.P. Barry and J.F. Donegan, *"Interferometer Based In-Band OSNR Monitoring of Single and Dual Polarisation QPSK signals",* European Conference on Optical Communications (ECOC), paper TH.9.C.6, 2010.
- [5] Wei-Hua Guo; O'Dowd, J.; Flood, E.; Quinlan, T.; Lynch, M.; Bradley, A.L.; Donegan, J.F.; Bondarczuk, K.; Maguire, P.J.; Barry, L.P.; , *"Suppression of Residual Single-Photon Absorption Relative to Two-Photon Absorption in High Finesse Planar Microcavities,"* Photonics Technology Letters, IEEE , vol.20, no.16, pp.1426-1428, Aug.15, 2008

- [6] J. O'Dowd, W. Guo, E. Flood, M. Lynch, A. Bradley, L. Barry, and J. Donegan, "*Polarization dependence of a GaAs-based two-photon absorption microcavity photodetector*," Optics Express **16**, 17682-17688 (2008).
- [7] O'Dowd, J.; Guo, W.H.; Lynch, M.; Flood, E.; Bradley, A.L.; Donegan, J.F.; , "*Description of polarisation dependence of two-photon absorption in silicon avalanche photodiodes*," Electronics Letters , **46** (12), p.854-856, (2010)

Acknowledgements

First of all, I would like to thank Professor John Donegan for allowing me to pursue this work under his supervision, his support and feedback have been invaluable.

Thanks to Dr. Weihua Guo for his willingness to share some of his astonishing encyclopaedia of knowledge, which has proved invaluable. His suggestions and input are what made most of this work possible.

For his suggestions, assistance, experimental wizardry and expertise in all things communications, I would like to thank Dr. Douglas Reid. I cannot count how many hours have been saved through Doug's input.

Many thanks to the Research Institute for Networks and Communications Engineering (RINCE) in DCU and in particular Professor Liam Barry for allowing me to perform most of this work in this institute and Dr. Frank Smyth for assistance in the lab.

I'd also like to thank Dr. John O'Dowd for his assistance in the early stages of this work, and the rest of the Semiconductor Photonics group; Louise Bradley, Mick Lynch, Vincent Weldon and all the folks who made the basement labs interesting over the years with various discussions regarding the nature of, and what was to be done with, the universe; Dave, Rob, Diarmuid, Pablo, Dave2, Ciaran and the late Tom Quinlan.

Outside of the SPG, I would like to thank my allies Ruairi, Tony, Matt, Mick, Dec, Susan, Murtagh, Joe, Amro. Those of you I've forgotten can voice your displeasure to me at time of your choosing via email or in person but it would take all day to mention everyone!

I would like to thank Dee for her understanding, support, patience, humour, kindness and love over the past months. It would be impossible to count up the hours of complaining she's had to listen to!

I would like to thank my parents Maura and Frank Flood for support, encouragement and understanding and my siblings, Orla, Niall and Francis, all of whom I haven't seen nearly enough of over the last few months.

Finally I'd like to dedicate this work to the memory of my late grandmother Brigid "Bridy" Hayden, who sadly passed away during the writing of this thesis. Her kind nature will be missed, as will her humour, which has had me in tears laughing at various points over the last 26 years. RIP.

List of Abbreviations

ASE	Amplified spontaneous emission
CW	Continuous wave
DCF	Dispersion compensation fibre
DPSK	Differential phase shift keying
EAM	Electroabsorption modulator
ECL	External cavity laser
EDFA	Erbium doped fibre amplifier
EOM	Electrooptic modulator
FPF	Fabry-Perot filter
FRAC	Fringe resolved autocorrelator
FROG	Frequency resolved optical gating
FWHM	Full width at half maximum
FWM	Four wave mixing
GVD	Group velocity dispersion
MZM	Mach-Zenhder modulator
NRZ	Non-return to zero
OCM	Optical channel monitor
OEO	Optical to electrical to optical
OOK	On-off keying
OPM	Optical performance monitoring
OSNR	Optical signal to noise ratio
PDL	Polarisation dependent loss
PMD	Polarisation mode dispersion
PRBS	Pseudo-random binary sequence

PSK	Phase shift keying
QAM	Quadrature amplitude modulation
QPSK	Quadrature phase shift keying
RF	Radio frequency
ROADM	Reconfigurable optical add-drop multiplexer
RZ	Return to zero
SBS	Stimulated Brillouin scattering
SRS	Stimulated Raman scattering
SSI	Spectral shearing interferometry
TBP	Time-bandwidth product
OTDM	Optical time division multiplexing
THG	Third harmonic generation
TIMI	Temporal information via intensity
TPA	Two-photon absorption
VOA	Variable optical attenuator
WDM	Wavelength division multiplexing
XPM	Cross phase modulation

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Chapter 1 Introduction

In order to facilitate the demand for optical bandwidth brought about by developments such as internet television, voice over internet protocols, videoconferencing and the increase in internet users over the last decade, it is necessary to further exploit the limits of the optical fibre. The most obvious way to facilitate increased data traffic is by increasing channel bitrates, the number of data channels and decreasing their spacing using various time and wavelength division multiplexed schemes [1,2,3]. Advanced quasi-multilevel and multilevel modulation formats which transmit more than one bit per symbol [4] may be used to this end as well as polarisation multiplexing [5,6]. The development of coherent detection has also made it possible to transmit data in which the phase, rather than the intensity of the signal is modulated [7]. To date, it has been possible to transmit data at 69.1Tb/s over 240km using 432 channels modulated with a 171Gb/s polarisation division multiplexed 16-QAM signal with a channel spacing of 25GHz in the C and extended L band [8]. However, as the spectral efficiencies (the amount of data within a given bandwidth in bit/s/Hz) of transmitted signals increase, the quality of the transmitted signal gains greater importance as the error margins decrease due to the physical and practical limitations of the technologies (the “window of operability”)[9].

The development of the transparent optical network [3] as a means of keeping this window open has replaced the periodic optical-electrical-optical (OEO) regeneration sites with optical amplifiers [10]. Such devices have inherent deleterious effects on the signal quality, they also by nature remove the signal quality monitoring that was inherently present as a result of the regeneration process. The lack of optical regeneration also means that impairments associated with transmitting data over long

distances become more prominent and the effects of nonlinear processes are no longer negligible. Impairments may be grouped into three categories [11],

Component based

Timing jitter

Amplifier noise

Amplifier distortion and transients

Linear impairments

Attenuation

Chromatic dispersion

Polarisation mode dispersion

Nonlinear impairments

Stimulated Raman scattering and stimulated Brillouin scattering

Four wave mixing

Self and cross phase modulation

Nonlinear crosstalk

1.1 Optical performance monitoring

In order to ensure an acceptable level of signal quality and to allow the tracking of errors and the active compensation of propagation impairments, the concept of optical performance monitoring (OPM) was introduced. Optical performance monitoring is a field that involves the use of optical measurement techniques to measure and overcome problems associated with the transmission of optical pulses, in other words, physical layer signal monitoring of the optical signal. Until recently, Optical layer monitoring involved optical channel monitoring (OCM)[12]. This involves measurement of the channel power, presence and (in some cases) wavelength in conjunction with alarms on individual components with the assumption that signal quality would be acceptable if all

components were working correctly. A more sophisticated approach was presented by Kilper et al in 2004 [13]. It was noted that there was no standard of measurements agreed for optical monitoring, unlike electrical monitoring where the parameters to be monitored have already been standardised. A series of measurements for a hypothetical OPM model is shown in Figure.1.1[14]

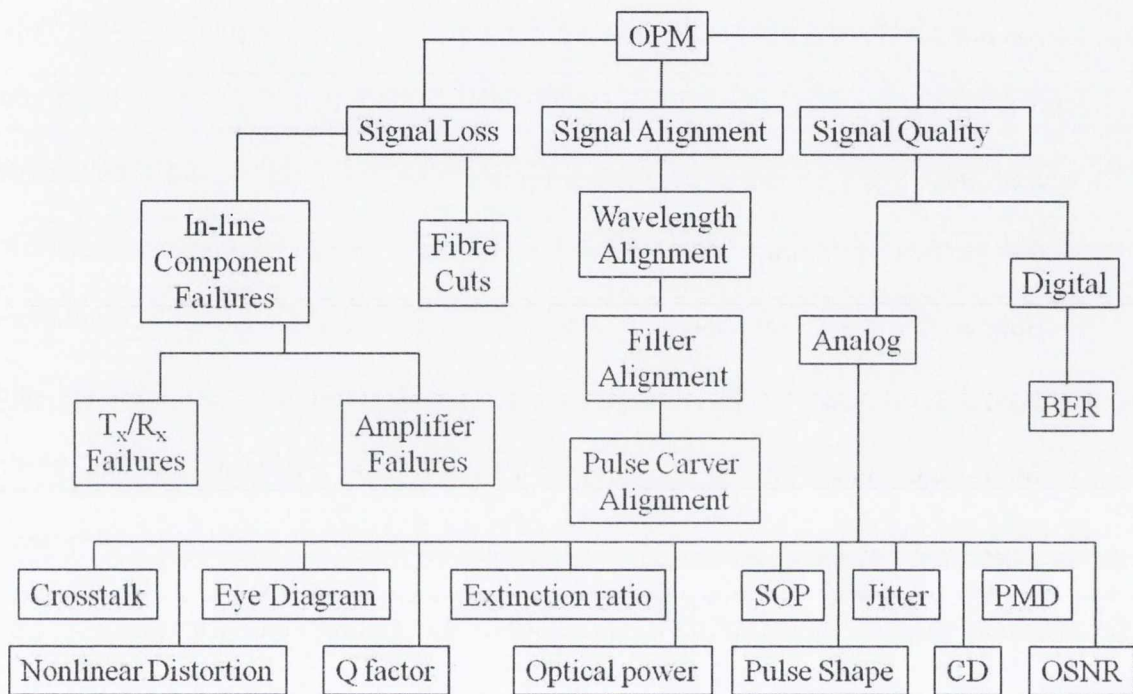


Figure.1.1 Hierarchy for a hypothetical OPM model, BER: bit-error rate, PMD, polarisation mode dispersion, OSNR: optical signal to noise ratio, SOP: state of polarisation, T_x/R_x : transmitter/receiver.

1.2 Implementation

OPM may be implemented using a nondisruptive dedicated monitor where the signal is tapped from a WDM fibre and the monitor is shared among the different wavelengths on the fibre. A second way involves a disruptive monitor in which one of multiple fibres may be switched to a monitor. This is undesirable due to the fact that the fibre is taken offline. It is imperative that a nondisruptive monitor functions at low optical powers and if possible, without reliance on nonlinear effects due to its input originating from the fibre tap.

To this end, a series of linear monitoring techniques relying on low optical input powers are examined in this thesis. Two techniques for the realisation of full electric field recovery (spectrum and phase) of the electric field of ultrashort pulses are examined. These techniques are based on the harmonic decomposition of a modified optical spectrum due to a sinusoidal optical phase modulation. The second series of measurements are based on monitoring of the in-band optical signal to noise ratio (OSNR) of intensity and phase modulated signals using two techniques based on Michelson fibre interferometers without prior knowledge of the signal's coherence properties. The first technique is based on a single fixed optical delay interferometer in conjunction with a linear polariser. The second uses a pair of Michelson interferometers with two fixed optical delays. Both measurements are immune to chromatic and polarisation mode dispersion and the latter is immune to the effects of polarised noise, signal depolarisation and can measure the in-band OSNR of polarisation multiplexed signals.

1.3 Outline of thesis

Chapter two: the expression for the electric field of an optical pulse is introduced. The phase and intensity information contained within the expression are explored and the parameters necessary to achieve full characterisation of the intensity and phase of ultrashort optical pulses are established. The generation of modulated signals and various commonly used modulation formats are investigated.

Chapter three outlines the more commonly used measurement techniques for the full characterisation of optical pulses using spectrographic, tomographic and interferometric means.

Chapter four introduces the optical signal to noise ratio and how it pertains to general signal health in a communications network. Included within this chapter is a description of existing OSNR measurement techniques and some issues relating to their implementation within transparent optical networks.

Chapter five introduces the phase sensitive monitoring of ultrashort optical pulses using the harmonic decomposition of a modified spectrum due to a sinusoidal optical phase modulation. The results of this method are compared to those obtained using existing methods.

Chapter six explores another harmonic decomposition based phase sensitive pulse monitoring technique which, while similar to that shown in chapter five, has the added advantage of a much faster characterisation time. The results obtained from this scheme are compared to those obtained using the scheme in chapter five and with those obtained using an existing method.

Chapter seven shows a variation on the interferometric in-band OSNR monitoring technique which combines an interferometric OSNR measurement with the existing polarisation nulling method and as with chapter seven, its ability to perform in the face of common network impairments is examined.

Chapter eight introduces another novel method for the measurement of the in-band OSNR of a communications signal using a pair of optical delay interferometers and its resilience to a series of network impairments and modulation formats is explored.

Chapter nine will summarise the findings of this work and their merits. Ways to improve the measurements and ideas for future investigations are explored.

Chapter 2 Ultrashort Optical Pulses

The characterisation of ultrashort pulses is a vital issue in optical communications networks. As mentioned in chapter 1, increasing bitrates lead to a demand for shorter pulse durations in optical networks, but at the cost of a greater sensitivity to transmission impairments. Monitoring technology must be such that full recovery of pulse information is achieved. Conventional monitoring techniques using fast photodiodes in conjunction with oscilloscopes are sufficient to temporally characterise pulses down to about 10 picoseconds in duration. The shortening of optical pulses beyond the point where their durations are less than that of the rise time of conventional photodetectors necessitates the use of more advanced techniques to characterize such pulses. Also, the advent of phase modulated signals in optical networks means that it is no longer sufficient to have a measure of the spectral characteristics of the pulses. It then becomes necessary to monitor the phase properties of the pulse.

In this section, a mathematical description of the electric field of a linearly polarised optical pulse is established. The parameters that are required to fully characterise the electric field of the pulse are discussed as a preface to chapter three.

2.1 Time domain representation

The expression for the electric field of a linearly polarised optical pulse as a function of time is given in equation (2.1). This is known as the scalar approximation as only one polarisation component is considered [15,16]

$$\varepsilon(t) = \text{Re}\left(\sqrt{I(t)} \exp\left(i\left[\omega_0 t - \phi(t)\right]\right)\right) \quad (2.1)$$

where ω_0 is the optical carrier frequency, $I(t)$ is the intensity and $\phi(t)$ is the phase of the signal. Figure 2.1 shows a graphical representation of an ultrashort pulse with a constant temporal phase $\phi(t)$.

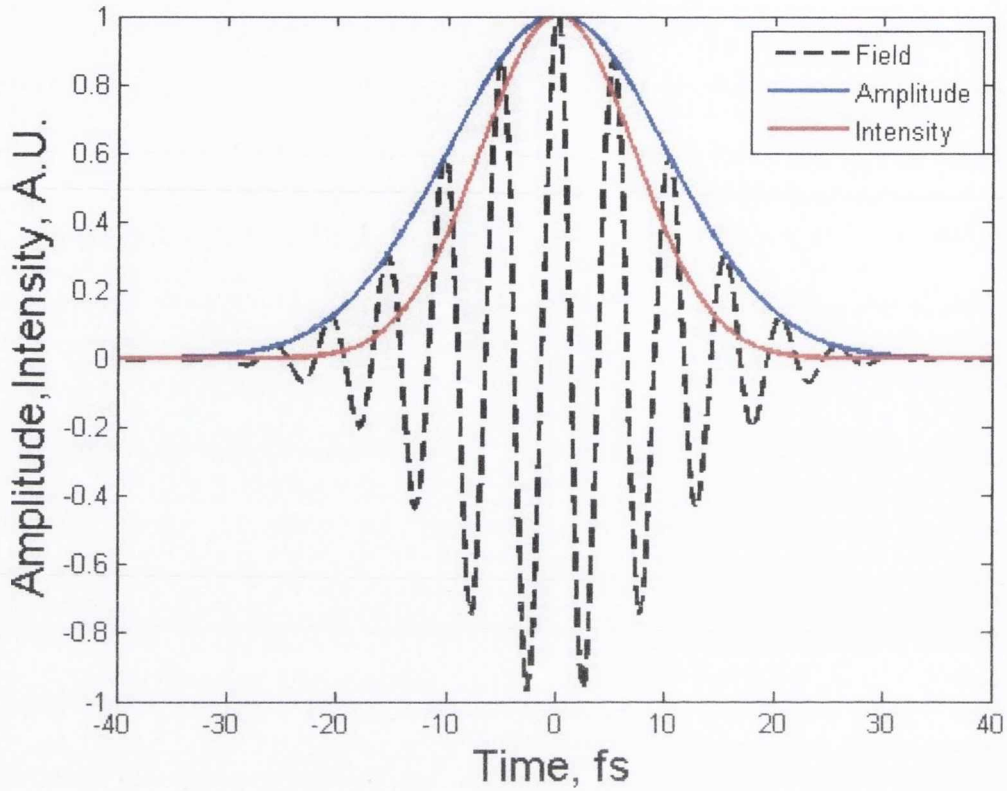


Figure 2.1 Electric field, amplitude and intensity of an ultrashort Gaussian optical pulse in the time domain with each quantity normalised. The intensity envelope is proportional to the magnitude of $E(t)^2$ and its width is $\sqrt{2}$ times that of $E(t)$.

If pulse is of a duration such that the oscillations in the optical carrier frequency are of much shorter duration than the variations in the $I(t)$ envelope function, the rapidly varying carrier wave $\exp(i\omega_0 t)$ may be removed from the calculations. This consideration applies to pulses which are picoseconds duration as the oscillations do not influence the overall envelope. When this is the case, the complex amplitude may then be described as equation (2.2)

$$E(t) = \sqrt{I(t)} \exp[-i\phi(t)] \quad (2.2)$$

The shape of the intensity may be defined as equation (2.3)

$$I(t) = |E(t)|^2 \quad (2.3)$$

Where we have disregarded some variables as the absolute magnitude of $I(t)$ is of little importance compared to its shape. The temporal phase $\phi(t)$ can be solved as follows

$$\phi(t) = -\arctan \left\{ \frac{\text{Im}[E(t)]}{\text{Re}[E(t)]} \right\} \quad (2.4)$$

or equivalently

$$\phi(t) = -\text{Im} \left\{ \ln[E(t)] \right\} \quad (2.5)$$

2.1.1 Pulse shape and length

The “length” of an optical pulse may be described by the full width at half maximum of its intensity distribution, τ_{FWHM} . Other quantities may be used to this end such as the RMS pulse length, the equivalent pulse width τ_e and the half width $1/e$ $\tau_{\text{HW}1/e}$ depending on the pulse’s intensity distribution as τ_{FWHM} will not adequately represent the length of a pulse with side bands whose peak intensity is below half that of the pulse’s peak intensity. For a Gaussian pulse distribution as generated by a modelocked laser which has not been subject to severe distortion τ_{FWHM} is sufficient to describe the pulse length.

2.2 Frequency domain representation

The field in the frequency domain is simply the Fourier transform of the time dependent field given as equation (2.6)

$$\tilde{\epsilon}(\omega) = \int_{-\infty}^{\infty} \epsilon(t) \exp(-i\omega t) dt \quad (2.6)$$

The $\sim$ above the function indicates that it is the Fourier transform. The inverse Fourier transform is given as

$$\varepsilon(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \tilde{\varepsilon}(\omega) \exp(i\omega t) d\omega \quad (2.7)$$

$\tilde{\varepsilon}(\omega)$ may then be written as follows

$$\tilde{\varepsilon}(\omega) = \sqrt{S(\omega)} \exp[-i\varphi(\omega)] \quad (2.8)$$

Where $S(\omega)$ is the spectrum of the pulse and is defined as the magnitude squared

Fourier transform of $\varepsilon(t)$, i.e. $S(\omega) = |\tilde{\varepsilon}(\omega)|^2$

$\varphi(\omega)$ is the spectral phase and can be defined in a similar fashion to $\phi(t)$

$$\varphi(\omega) = -\arctan\left(\frac{\text{Im}[\tilde{\varepsilon}(\omega)]}{\text{Re}[\tilde{\varepsilon}(\omega)]}\right) \quad (2.9)$$

And then we can write

$$\varphi(\omega) = -\text{Im}\{\ln[\tilde{\varepsilon}(\omega)]\} \quad (2.10)$$

2.2.1 Temporal phase and the instantaneous frequency.

The temporal phase provides information on the frequency evolution across the pulse and can be used to find the instantaneous angular frequency of the pulse at any point across its duration using equation (2.11)

$$\omega_{\text{inst}} = \omega_0 - \frac{d\phi(t)}{dt} \quad (2.11)$$

This may also be expressed in terms of frequency as shown in (2.12)

$$\nu_{\text{inst}} = \nu_0 - \frac{1}{2\pi} \frac{d\phi(t)}{dt} \quad (2.12)$$

The temporal phase may be described by any real function, usually a polynomial. It is common to expand $\phi(t)$ using a Taylor series as follows

$$\phi(t) = \phi_0 + t\phi_1 + t^2\phi_2 / 2... \quad (2.13)$$

2.2.2 Spectral phase and group delay

The spectral phase $\phi(\omega)$ provides information about the time vs frequency evolution of the pulse, which may be used to find the moment at which certain frequency components are present in the pulse. The group delay may be defined as the differential of the spectral phase of the pulse with respect to the angular frequency as shown in

$$t_g(\omega) = \frac{d\phi}{d\omega} \quad (2.14)$$

In other words the group delay provides information on the temporal delay imposed on certain spectral components of the pulse. The spectral phase may likewise be expanded using a Taylor series as follows

$$\phi(\omega) = \phi_0 + (\omega - \omega_0)\phi_1 + (\omega - \omega_0)^2 \phi_2 / 2... \quad (2.15)$$

2.2.3 Zeroth order phase

ϕ_0 and ϕ_0 , the zeroth order phase is the same in both time and frequency domains. The zeroth order phase is known as the “absolute phase”. However, ϕ_0 ϕ_0 are the relative phases of the carrier with respect to the envelope in the time and frequency domain respectively. It may be described simply as the phase of the carrier wave at the peak of the pulse envelope. This term is unimportant when the pulse is several cycles long as the slow variation of the carrier with respect to the envelope will ensure that a shift in phase of the carrier will have little effect on the overall envelope. Figure 2.2 shows the effect of a zero order phase shift of magnitude π , $\pi/2$ on a single cycle pulse relative to the envelope

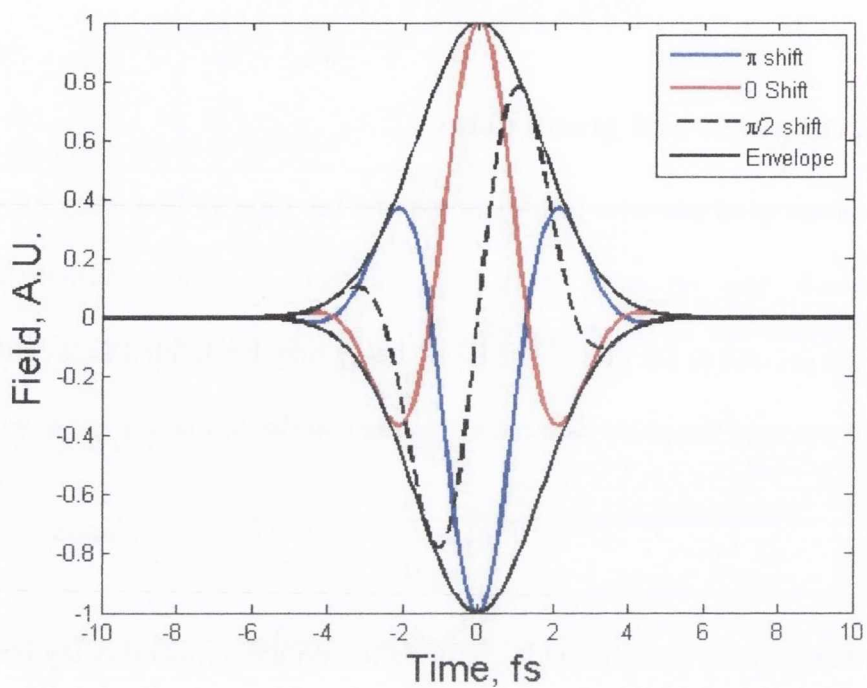


Figure 2.2 shift in ϕ_0 for a single cycle pulse

Figure 2.3 shows the effect of the same phase shift on pulse that is many cycles in duration and it may be appreciated that a shift of π has little influence on the overall envelope.

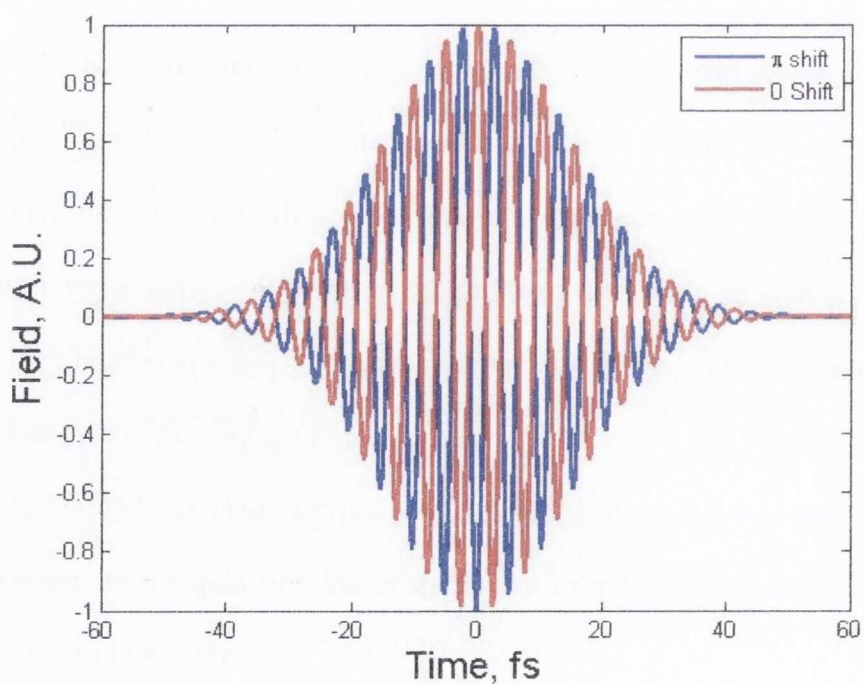


Figure 2.3 shift in ϕ_0 for a multiple-cycle pulse.

2.2.4 First order phase

Shifts in the first order phases signify a shift in time or frequency. The Fourier shift theorem states that $E(t-\tau)$ transforms to $\tilde{E}(\omega)\exp(-i\omega\tau)$. This also applies to the inverse Fourier transform so a linear term in the spectral phase ϕ_1 corresponds to a frequency shift and a linear term in φ_1 corresponds to a shift in time or a delay. ϕ_1 may be measured using interferometric methods and φ_1 can be measured using a spectrometer.

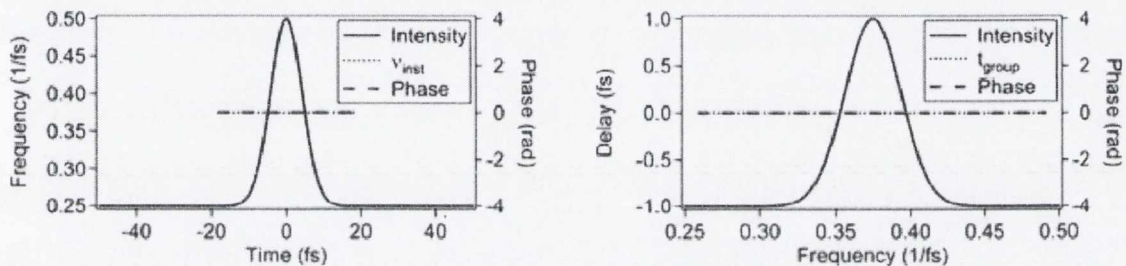


Figure 2.4 Temporal (left) and spectral (right) intensities of a pulse with constant first order phase

Figure 2.4 shows a Gaussian pulse with flat phase in spectrum and time. When the same pulse has a linear spectral phase it can be seen that the pulse is delayed in time as shown in Figure 2.5

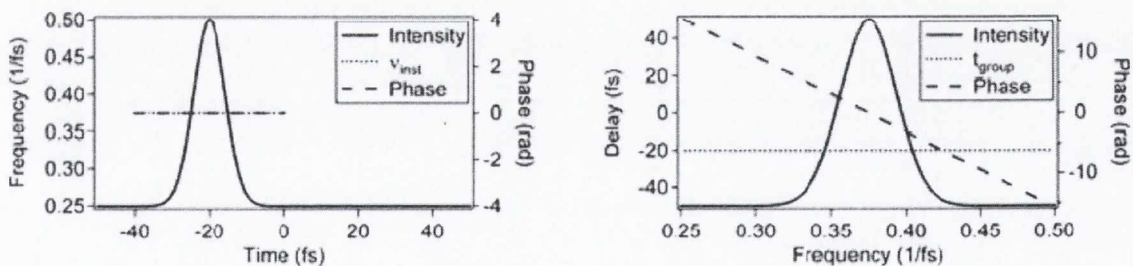


Figure 2.5 Temporal (left) and spectral (right) intensity of a pulse with a linear variation in spectral phase

Figure 2.6 shows the effect of a linear temporal phase which causes a shift in the spectrum

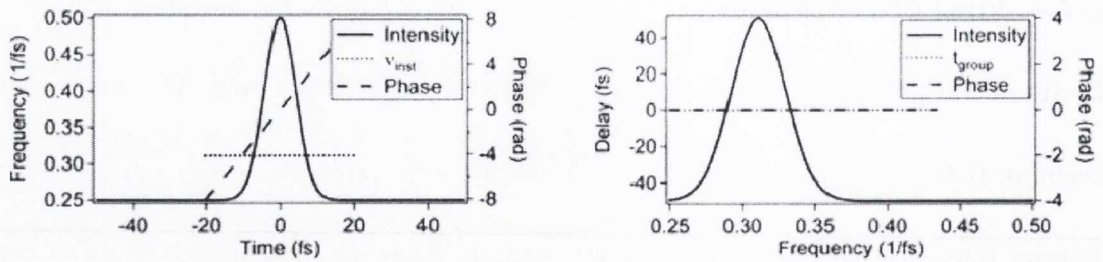


Figure 2.6 Temporal (left) and spectral (right) intensity of pulse with linear variation in temporal phase

2.2.5 Second order phase

A quadratic variation of $\phi(t)$ (a nonzero value of ϕ_2) represents a linear frequency shift across the pulse which is known as a “linear chirp”. This has the effect of spatially broadening the pulse. Linear chirp is caused by propagation of the wave through a dispersive medium (i.e. a medium with a frequency dependent refractive index). Figure 2.7 shows the intensity, phase and instantaneous frequency as a function of time for a linearly chirped Gaussian pulse. Figure 2.8 shows the electric field for the same pulse.

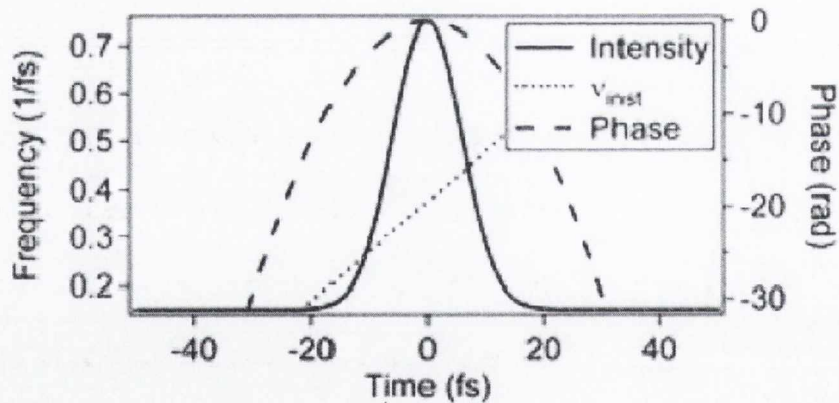


Figure 2.7 Temporal intensity, phase and instantaneous frequency of a pulse with a quadratic temporal phase

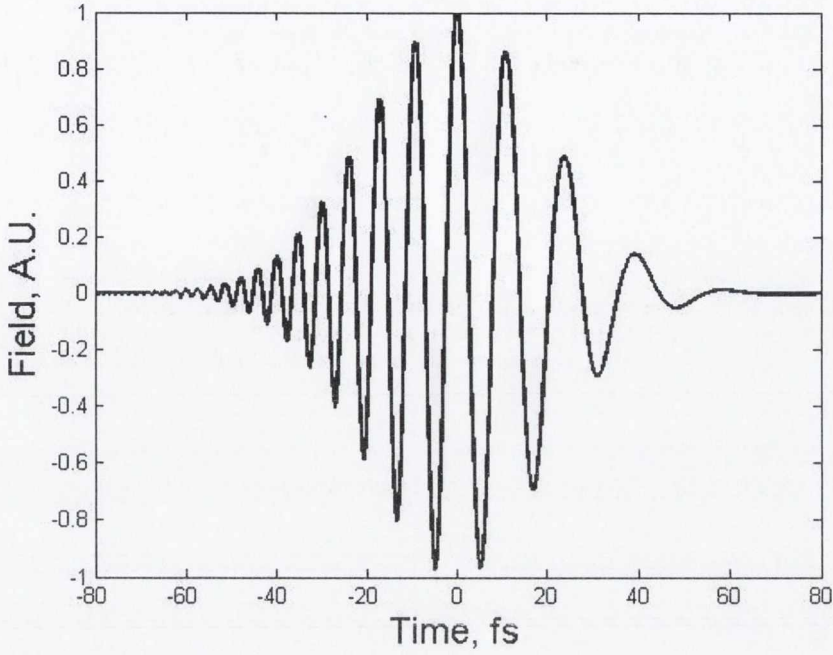


Figure 2.8 Electric field of a pulse with quadratic temporal phase

2.2.6 Third order phase

A nonzero ϕ_3 and higher order phase values manifest themselves as a nonlinear chirp across the duration of the pulse. Third and higher order phase distortions are considered in the frequency domain as the spectrum is easy to measure directly while the temporal intensity is not as will be seen in chapter three. The spectral phase can be used for full field retrieval but the temporal phase cannot. A nonzero $\phi_3(\omega)$ term results in a quadratic group delay as a function of frequency. The consequence of this is that the centre frequency of the pulse arrives at the detector first while frequencies either side arrive later. The different frequencies cause beats in intensity resulting in oscillations after the main pulse (when ϕ has a positive sign) and afterwards (when ϕ is negative). Third and higher order phase values have a severe effect on the intensity distribution of pulses and can be caused by higher order dispersions and nonlinear effects in optical fibres.

2.2.7 Time bandwidth product

This is the product of the temporal and spectral width and is a dimensionless quantity which gives an indication of the level of temporal or spectral phase distortions present in a pulse. The TBP is the product of the full width at half maximum (FWHM) of the spectral and temporal intensities, $\Delta\omega\Delta t/2\pi \geq K$, where K is a constant of the order of unity. This is the general relation between the functions $I(\omega)$ and $I(t)$ imposed by the uncertainty principle. Also, the coherence time τ_c is the reciprocal of the pulse bandwidth so the time-bandwidth product is the ratio of the pulse width to the coherence time. If the pulse spectral intensity is Gaussian, i.e. $I(\omega)=\exp[-(\omega-\omega_0)^2/\alpha]$, then its temporal intensity will also be Gaussian and given as $I(t)=\alpha\exp[-\alpha(t-t_0)^2]$. For this combination of intensity profiles, $K=(2\ln 2)/\pi=0.441$, which is the minimum TBP of a Gaussian pulse [17]. This value is difficult to attain practically due to distortions in even the most well behaved pulses [16]

2.3 Modulation of optical signals

Encoding of data onto a train of optical pulses is performed by manipulating the properties of the electric field of the pulses. In the electric field expression given in equation (2.16)

$$E(t) = A(t)\exp(i\omega t - \phi(t)) \quad (2.16)$$

$A(t)$ and $\phi(t)$, the amplitude and phase respectively, are the most commonly manipulated variables in optical communications networks.

2.3.1 Duty cycle, bit rate and symbol rate

For an intensity modulated signal the pulse duty cycle is the ratio of the signal temporal width Δt_{FWHM} to the bit period t_b . $D = \Delta t_{FWHM}/t_b$ which is basically the ratio of how long the signal is in the “on” state to the off state, t_b is the reciprocal of the data rate. For a

binary modulation format in which data is carried in either an “on” or “off” state, t_b is equal to the reciprocal of the bit rate. A multilevel modulation is one in which each symbol may take more than two values

2.3.2 Direct modulation of semiconductor lasers

One obvious way of generating an intensity modulated optical signal is by modulating the laser injection current [18][19]. The laser is biased near threshold and the modulation current is such that its maximum is above the lasing threshold and the minimum is below. An issue with this method is that electrons injected into a semiconductor contribute to the magnitude of the refractive index so the current modulation also leads to a modulation of the refractive index of the gain medium. This leads to an undesirable frequency modulation in the output optical field. This leads to an unwanted spreading of pulses as they propagate in a dispersive medium. For this reason there is a limit to the rate at which a semiconductor laser may be directly modulated, which is approximately 10Gb/s. Higher data rates have been achieved in a laboratory environment [20] but this was over quite short distances.

2.4 External modulation

For bitrates above 10Gb/s it is no longer possible to use direct modulation and the laser intensity must be modulated by external components, which fall into two different categories, namely electroabsorption modulators and electrooptic modulators.

2.4.1 Electroabsorption modulators

Electroabsorption modulators (EAM) utilise the Franz-Keldysh [21] effect by which the loss in a linear waveguide may be modulated by an applied electric field tilting the band structure allowing partial overlap between the valence and conduction band. This causes the probability of absorption of incident photons to increase greatly, thereby

corresponding to an “off state”. EAMs may also be based on the quantum confined Stark effect in which the energy levels of quantum wells shift in the presence of an electric field [22]. Such a device is attractive for several reasons, namely the possibility of monolithic integration with lasers and semiconductor optical amplifiers as they are only a few microns in length and have low power requirements. As with the case of the directly modulated laser, the output of an EAM is chirped but commercially available EAMs can be modulated at up to 40Gb/s with laboratory demonstration of up to 80Gb/s [23]

2.4.2 Electro-optic modulators

Another category of modulator utilises the electro optic effect, usually in Lithium Niobate (LiNbO_3) [24].

The material experiences a change in the refractive index as a result of an applied voltage. This changes the phase of an incident optical signal. Such an effect may be exploited for use in both phase and amplitude modulators.

2.4.2.1 Mach Zehnder amplitude and phase modulator

The Mach-Zehnder modulator works on the principle of interference of light. The incoming light is split into two and one or both arms are equipped with electro-optic phase modulators, which correspond to single or dual drive modulators respectively. The two fields interfere at the output and the transfer function is given as

$$T_E(V_1, V_2) = \frac{1}{2} \left[\exp(i\phi(V_1)) + \exp(i\phi(V_2) + i\psi) \right] \quad (2.17)$$

This then becomes

$$T_E(V_1, V_2) = \exp\left(\frac{i(\phi(V_1) + \phi(V_2) + \psi)}{2}\right) \cos\left[\frac{(\phi(V_1) - \phi(V_2))}{2} - \frac{\psi}{2}\right]$$

$\phi(V_{1,2})$ are the voltage dependent optical phases of the two interferometer arms. ψ is a constant phase shift between the arms brought about by applying a voltage known as the modulator bias. As the voltage difference between the two arms is scanned linearly, the output field intensity varies sinusoidally. The voltage difference required to effect a phase change of π between the two arms is known as $V\pi$, which is sufficient to switch the modulator from full transmission to full extinction. A dual drive Mach-Zehnder modulator may be used to cause a phase modulation [25,26] by manipulating V_1 and V_2 as can be seen from equation (2.18). By driving the two modulator arms of a dual drive mach Zehnder modulator equally but in opposite directions (i.e. $V_2 = -V_1$), the phase term vanishes and the signal is free of chirp. This is known as the “push-pull” modulation scheme [27].

2.5 Amplitude modulation

The most straightforward means of encoding data onto light pulses is by modulation of the signal amplitude while keeping ω_0 and $\phi(t)$ constant (in practice ϕ may change due to dispersion but this should not be an issue with conventional detection). For binary amplitude modulation $A(t)$ takes either one of two fixed values during each bit period, “1” when light is present above a certain power threshold and “0” when it is below that level. A benefit of binary intensity modulation schemes is that they are immune to the effects of phase changes in the received signal and require straightforward detection techniques. Two commonly used intensity modulation formats are highlighted below.

2.5.1 NRZ-OOK

Non return to zero (NRZ) on off keying (OOK) is the most basic of binary modulation formats. The optical signal takes either of its binary states over the entire bit period, so the signal power does not drop to zero between two consecutive “1” values. The generation of an NRZ signal may be by direct modulation or using an EAM (for short and intermediate distance) and a chirp-free MZM for long haul applications. Generation of an NRZ signal with an MZM may be achieved by biasing at 50% transmission (the quadrature point) and driving with a voltage of V_π .

2.5.2 RZ-OOK

Return to zero (RZ) on-off keying means that the signal power returns to zero within each bit period. As the temporal width of the pulses are ideally half that of a NRZ signal, the spectrum of an RZ signal is wider than that of an NRZ so its spectral efficiency is lower. However, the lower duty cycle of an RZ signal allows their use in optical time division multiplexed (OTDM) systems. An RZ signal may be generated by encoding a modulator with an RZ signal [28] or by carving pulses from an NRZ signal through use of an additional modulator (usually an EAM). The latter becomes necessary at data rates of 40Gb/s and above [29]. A range of duty cycles may be realised using this format with some of the lower duty cycle possibilities being candidates for use in optical time division multiplexed systems [2]. By way of example, it is possible to generate a signal with duty cycle of 50% of the bit period by sinusoidally driving an MZM at the data rate between the minimum and maximum transmission. By lowering the driving voltage it is possible to reduce the duty cycle to 36% but this is met with a corresponding insertion loss as the modulator is never actually driven to the point of maximum transmission. If an MZM is driven sinusoidally at half the data rate between the transmission minima the result is a pulse with duty cycle of 33% of the bit periods.

If an MZM is driven sinusoidally at half the data rate between its transmission maxima, the resultant pulses will have a 67% duty cycle with alternating phase (which is a “carrier suppressed return to zero” (CS-RZ) signal).

2.5.3 Eye diagram

The optical intensity eye diagram is a means of depicting the degradation of digital binary pulses due to pulse spreading, distortion and noise in an amplitude modulated signal. The stream of pulses is displayed on a storage oscilloscope and the horizontal scan is triggered to the bit rate of the signal. The oscilloscope records multiple sequences of zeros and ones which add up to the eye diagram. A clear transmission with a low bit error rate is signified by an open eye as shown in Figure 2.9

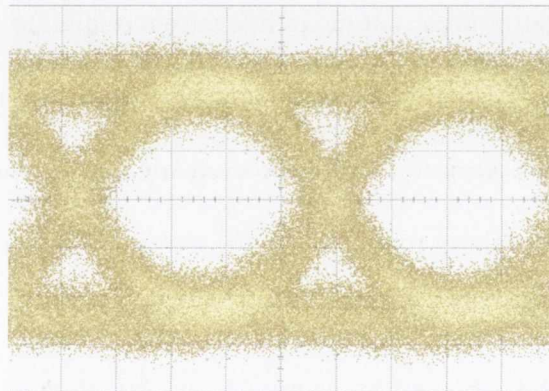


Figure 2.9 Eye diagram of a 10G NRZ-OOK PRBS (pseudo random binary sequence) signal

The eye diagram provides information on the optical field’s intensity although it is a qualitative measurement it can be used to infer the modulation format of a received signal and provide information about timing jitter, pulse rise and fall time and dispersion.

2.6 Phase modulation

As with the amplitude, it is possible to encode data onto the phase of the optical signal, $\phi(t)$. The generation of the phase modulated signals may be performed using either a Mach-Zehnder intensity modulator or a phase modulator. Detection of phase modulated

signals is more complicated than their intensity counterparts due to the requirement for coherent detection. Phase shift keying (PSK) is the term that refers to the modulation of the phase of the optical carrier. Some of the more common PSK formats are shown below.

2.6.1 DPSK

Differential binary phase shift keying (DPSK or DBPSK)[30] encodes information on a binary phase change of π between two adjacent bits. In this case a “1” is represented by a phase shift of π and a “0” is signified by the lack of a shift. Generation of a DPSK signal is achieved either by a straight line phase modulator or a Mach-Zehnder modulator. In the former case the intensity of the signal remains constant. In the case of a Mach-Zehnder modulator there will be an intensity modulation due to the fact that the signal is generated by biasing the modulator at zero transmission and driving it symmetrically. The latter is more commonly used for practical applications due to the ease of imposing an exact phase modulation [30] using such an arrangement. Like OOK modulation formats, PSK formats may also be RZ or NRZ. Detection of a phase modulated signal may be achieved by using a 1 bit optical delay interferometer, which converts the differential phase modulation into intensity modulation.

2.6.2 DQPSK

Differential quadrature phase shift keying (DQPSK) [31] is a multilevel phase modulation format (there is more than one bit per symbol) which utilises four phase shifts instead of two, namely 0, $\pm\pi/2$ and π . Generation of this format is somewhat more complicated than a DPSK signal. A pair of parallel Mach-Zehnder modulators may be used as phase modulators with a $\pi/2$ phase shifter after one of the modulators. The signals from the two modulators are then combined. Detection is performed by splitting

the signal into two paths and using a pair of delay interferometers in conjunction with balanced receivers. The delay in the interferometers is equal to the symbol duration, which is half the bit duration for a two level modulation format. This places greater emphasis on the tolerances in the delay interferometer.

2.6.3 Polarisation multiplexed QPSK

In order to further increase the spectral density further, it is possible to encode two independent data sets on orthogonal polarisations. This was first shown for a binary phase shift keyed signal in [5] and has since been expanded to QPSK [32]. Such formats have not yet seen practical application due to the complexity involved [7]

2.7 Transmission impairments

The transmission of optical pulses over long distances presents numerous obstacles which arise from various linear and nonlinear effects. Due to tighter tolerances and a lack of optical regeneration these may propagate along several amplifier spans and their effects become exaggerated. The large interaction length and high transmitted optical intensities mean that certain nonlinear effects may not be trivial in such networks.

2.7.1 Linear impairments

2.7.1.1 Loss

Fibre losses are caused by the attenuation of the optical power propagating in a fibre. If the power launched into a fibre is P_0 , the transmitted power P_T at a given length L from the input may be written as (2.19)

$$P_T = P_0 \exp(-\alpha L) \quad (2.19)$$

where α is the attenuation constant and is a measure of the losses from all sources. Fibre losses are wavelength dependent and the two major causes are material absorption and

Rayleigh scattering. Silica absorbs little light at communications wavelengths (~1550nm), but impurities can lead to significant losses due to material absorption. The loss spectrum of a single mode fibre [33] is shown in Figure 2.10.

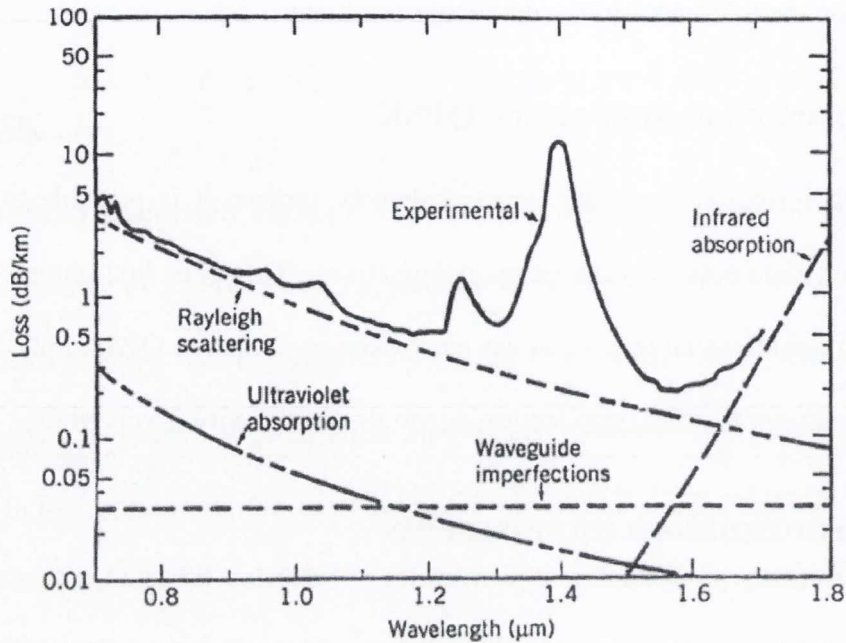


Figure 2.10. Loss spectrum of a singlemode Silica Fibre with the various contributions shown
Image taken from ref [10]

The main peak at 1.35μm is due to an overtone of an absorption peak at 2.73μm. Rayleigh scattering occurs due to density fluctuations in the fibre due to the manufacturing process, thereby causing fluctuations in the refractive index. A result of this is that light is scattered in all directions. The losses due to Rayleigh scattering scale with λ^{-4} as shown in equation (2.20)

$$\alpha_R = \frac{C_R}{\lambda^4} \tag{2.20}$$

α_R is attenuation constant due to Rayleigh scattering and C_R is a constant ranging from 0.7-0.9dB/(km-μm⁴) depending on the constituents of the fibre core. Typical loss values in commercial fibres are in the range of 0.2dB/km around 1.55μm.

2.7.1.2 Amplifier noise

The compensation of optical losses is achieved through optical amplification using Erbium doped fibre amplifiers (EDFA) through stimulated emission. However, spontaneous emission of photons may also occur. These spontaneous photons are amplified by the active medium of the amplifier. The level of noise added to the propagating signal by each amplifier may be quantified as the “noise figure” which is given as

$$F = \frac{\text{SNR}_{\text{input}}}{\text{SNR}_{\text{output}}} = \frac{S_i / N_i}{S_o / N_o} = \frac{N_o}{GN_i} = \frac{GN_i + N_{\text{amp}}}{GN_i} + 1 + \frac{N_{\text{amp}}}{GN_i} \quad (2.21)$$

SNR is the signal to noise ratio, S_i and N_i are the input signal and noise powers respectively while S_o and N_o are the output signal and noise powers respectively. G is the amplifier gain ($G=S_o/S_{in}$) and N_{amp} is the noise power added by the amplifier at the output.

2.7.1.3 Chromatic dispersion

Because amplification has reduced the need for regeneration in networks, issues arise due to the greater transmission distances afforded by this development. Due to the frequency dependence of the refractive index, $n(\omega)$, of optical fibres, different frequency components of an optical pulse will propagate with differing velocities $c/n(\omega)$. A result of this is both a broadening of the pulse temporal width and a frequency chirp across the pulse [34]. The effects of dispersion may be examined by expanding the mode propagation constant β using a Taylor series using equation (2.22).

$$\beta(\omega) = n(\omega) \frac{\omega}{c} = \beta_0 + \beta_1 (\omega - \omega_0) + \frac{1}{2} \beta_2 (\omega - \omega_0)^2 + \dots \quad (2.22)$$

where ω_0 is the pulse carrier frequency. β is given as

$$\beta_m = \left(\frac{d^m \beta}{d\omega^m} \right)_{\omega=\omega_0} \quad (m = 0, 1, 2, \dots) \quad (2.23)$$

β_1 and β_2 are given as

$$\beta_1 = \frac{1}{v_g} = \frac{n_g}{c} = \frac{1}{c} \left(n + \omega \frac{dn}{d\omega} \right) \quad (2.24)$$

$$\beta_2 = \frac{1}{c} \left(2 \frac{dn}{d\omega} + \omega \frac{d^2 n}{d\omega^2} \right) \quad (2.25)$$

n_g is the group index and v_g is the group velocity, which may be found as $\beta_1 = c/n_g$. β_2 represents the group velocity dispersion (GVD) and is responsible for pulse broadening. β_2 is the GVD parameter. The dispersion parameter, D is given as $d\beta_1/d\lambda$, which is related to β_2 and λ as follows

$$D = \frac{d\beta_1}{d\lambda} = \frac{-2\pi c}{\lambda^2} \beta_2 = -\frac{\lambda}{c} \frac{d^2 n}{d\lambda^2} \quad (2.26)$$

The dispersion parameter varies with wavelength in single mode fibres as shown in Figure 2.11

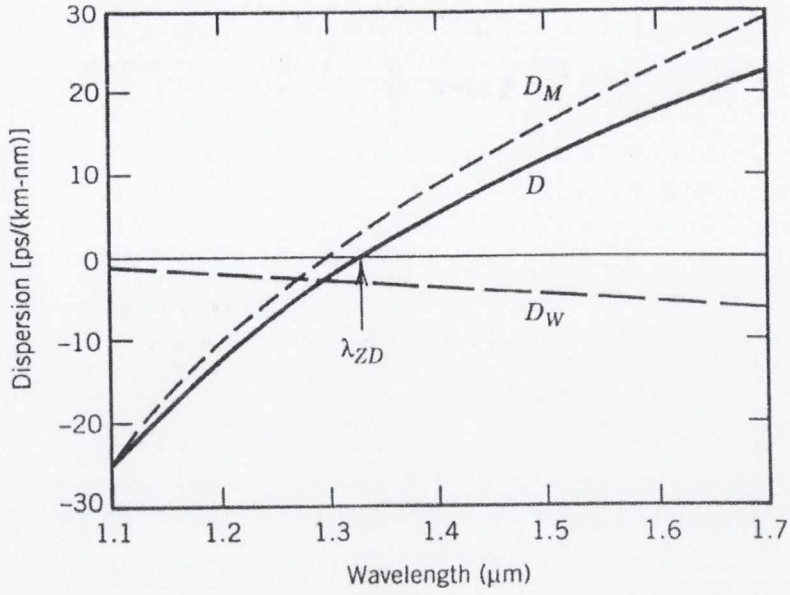


Figure 2.11 wavelength dependence of dispersion in single mode fibres. D_w is the contribution from waveguide dispersion, D_M is the contribution from material dispersion λ_{ZD} is their zero-dispersion wavelength

The zero crossing is described as the zero dispersion wavelength. However, at this wavelength, there will be a contribution from β_3 leading to a third order dispersion. λ_D may be shifted through the doping of single mode fibres. Also, due to the waveguiding process, a contribution to the total dispersion comes from the design of the waveguide meaning that the dispersion parameter may be tailored in this regard. This may find applications in dispersion compensated fibres, which can be used to mitigate the effects of chromatic dispersion.

2.7.1.4 Polarisation mode dispersion

A single mode fibre may in fact support two degenerate modes which are orthogonally polarised along what are called the ordinary and extraordinary axes. Due to applied stress or material factors, anisotropy in the refractive indices along the ordinary and extraordinary axes, the propagation constant β for each mode may vary; this is called “polarisation modal dispersion”. A consequence of this is that the two modes propagate at different group velocities leading to a spreading in the temporal width of the pulse.

The time delay τ between the two modes may be written as equation (2.27)

$$\tau = \left(\frac{d}{c} \Delta n + \frac{\omega d}{c} \frac{\partial \Delta n}{\partial \omega} \right) \quad (2.27)$$

where d is the length of fibre, c the speed of light and Δn is the birefringence given by $\Delta n = n_s - n_f$ where s and f refer to the refractive indices of the slow and fast axes of the fibre.

2.7.1.5 WDM crosstalk

The narrowing of inter-channel spacing can lead to interactions between two adjacent channels. A beating term between a channel's electric field $E(t)$ and the residual field of a neighbouring field $R(t)$ causes interference at square law detection such that

$$S \propto |E(t) + R(t)|^2 = E(t)^2 + R(t)^2 + 2\text{Re}\{E(t)R^*(t)\} \quad (2.28)$$

In the case where adjacent channels are copolarised the major degradation is brought about by the beating term $2\text{Re}\{E(t)R^*(t)\}$. This effect may be mitigated by a process known as “polarisation interleaving” where adjacent WDM channels are orthogonally polarised.

2.7.2 Nonlinear impairments

2.7.2.1 Stimulated inelastic scattering

Stimulated inelastic scattering may occur between an optical pump wave and the medium in which it is propagating. This leads to an energy transfer to the medium and a frequency-shifted Stokes wave when the incident power is above a certain level. Two types of stimulated inelastic scattering may manifest themselves in Silica fibres. Stimulated Raman scattering (SRS) involves the transfer of energy from incident “pump” photons to an optical phonon in Silica. Another photon is then created at a lower energy such that the total energy of the original photon is conserved, this photon

belongs to the scattered “Stokes wave”. Stimulated Brillouin scattering (SBS) involves the transfer of energy from an incident photon to an acoustic phonon along with the creation of a new photon at a lower frequency. Although the origin of the two processes is similar, SBS in single mode fibres only occurs in the backward direction while SRS may occur in both directions. For the case of SRS

$$\frac{dI_s}{dz} = g_R I_p I_s \quad (2.29)$$

where I_s is the intensity of the Stokes wave I_p is the pump intensity g_R is the Raman gain coefficient. The relation in equation (2.29) is similar for SBS only g_B replaces g_R where g_B is the Brillouin gain coefficient. Both SRS and SBS are threshold phenomena and only become appreciable over a certain intensity level $I_{th} \sim 16\alpha/g_R$ for Raman scattering and $I_{th} \sim 21\alpha/g_B$ for Brillouin scattering. Due to the energy differences in the phonons created, the Stokes waves will be downshifted in energy by different amounts and in the case of SRS scattering the shift will be in the order of 10^{13} Hz while the Stokes wave due to SBS will be shifted by less than 10^{10} Hz. Both effects have a parasitic effect on the intensity of the propagating signal but SRS may cause nonlinear crosstalk between channels, while SBS occurs at much lower optical powers ~ 1 mW

2.7.2.2 Self phase modulation and cross phase modulation

Self phase modulation (SPM) occurs as a result of an intensity-dependent refractive index in a medium. This manifests itself as a phase shift in the electric field of the propagating pulse the magnitude of which is proportional to the intensity distribution of the pulse. The greater intensity at the peak of the pulse propagates more slowly than the lower intensity wings. This causes significant spectral broadening and places a limit on the peak powers that may be used in WDM systems. In the case of cross phase modulation (XPM), if many WDM channels are propagating in a fibre (channels

1,2..N), channels 2-N may distort the index profile of the fibre such that channel 1 can experience a phase shift. As well as the spectral broadening that occurs as a result of these effects, the intensity-dependent index changes may cause severe phase noise issues with phase modulated signals.

2.7.2.3 Four wave mixing

When two or more channels with the same polarisation propagate in a fibre and satisfy the phase matching criterion then their interaction may generate new frequencies. This process is a third order nonlinear effect known as four-wave mixing (FWM) and the coincidence of the newly formed frequencies with existing DWDM channels can cause issues such as nonlinear crosstalk. This poses a problem even if the newly generated signals do not coincide with WDM channels as it degrades the signal power.

2.8 Summary

In this chapter the expression for the field of an optical pulse was introduced and the temporal and spectral phase and intensity properties were discussed. The manner in which data is encoded onto an optical signal was examined in the context of both direct and external modulation using both electroabsorption and electro-optic modulators. Several of the more prominent data modulation formats were outlined as was the manner in which they may be generated. Several linear and nonlinear impairments pertaining to the transmission of optical pulses in WDM networks were briefly discussed.

Chapter 3 Characterisation of optical pulses

In chapter 2 the properties of an optical pulse in the temporal and spectral domain were discussed. It was seen that the pulse field $E(t)$ can be uniquely determined if its spectrum and phase are known. In this chapter, measurements for the temporal and spectral characterisation of ultrashort pulses will be introduced. Following this, a series of phase sensitive measurement techniques will be discussed.

3.1 Time domain characterisation.

An obvious way to measure the temporal characteristics of an optical pulse is through use of a high bandwidth photodiode in conjunction with a fast sampling oscilloscope. This allows the measurement of the pulse shape, temporal intensity and eye diagram. However, the fastest commercially available detectors are only able to measure down to approximately 10ps temporal FWHM (u^2t XPDV), so for the moment it is necessary to use alternative means for characterisation of sub 10ps pulses.

3.1.1 Intensity autocorrelation

The *intensity autocorrelation* [35] is a measurement that seeks to represent the pulse's intensity as a function of time. The measurement involves splitting the pulse under test and using the copy of this pulse to sample itself. The pulse copy is subject to a variable delay before being recombined with the original in a nonlinear medium such as a second harmonic generation (SHG) medium [36], or a two-photon absorption microcavity photodetector [37] (TPA Autocorrelation). In the case of the more commonly used SHG autocorrelation, the crystal produces light at twice the frequency as the input pulses, the intensity of which is measured using a slow detector. The intensity of the SHG signal is related to the spatial overlap of the two pulses. An example of a SHG autocorrelator is shown in Figure 3.1

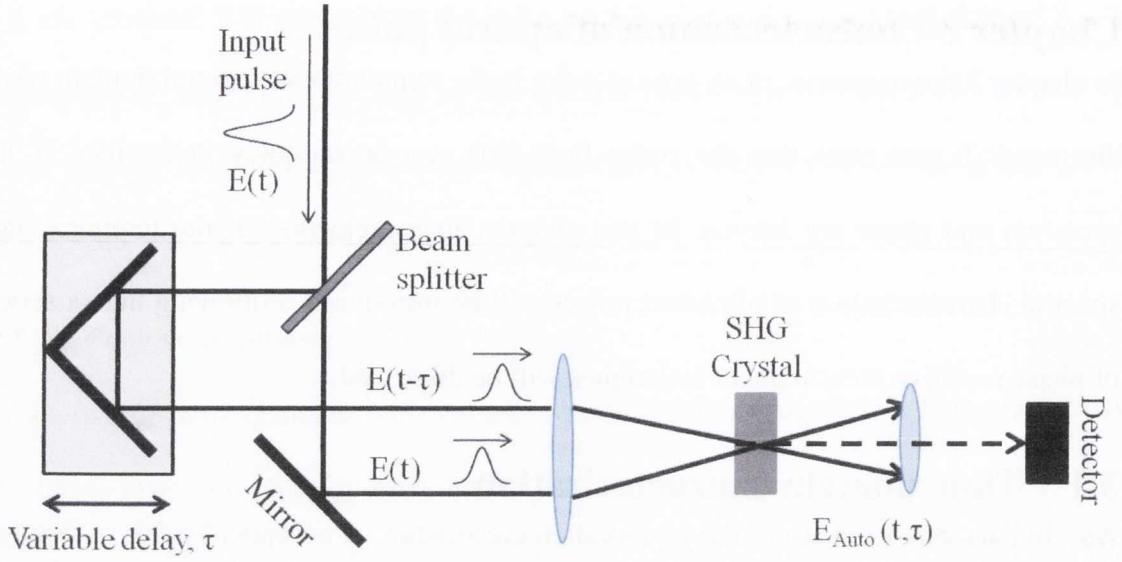


Figure 3.1 schematic for SHG autocorrelation measurement

The SHG field is given as

$$E_{\text{Auto}}^{\text{SHG}}(t, \tau) \propto E(t)E(t + \tau) \quad (3.1)$$

τ is the delay between the pulses. The intensity is proportional to the product of the two pulse intensities.

$$I_{\text{Auto}}^{\text{SHG}}(t, \tau) \propto I(t)I(t - \tau) \quad (3.2)$$

The intensity autocorrelation signal is measured using a slow detector and is given as

$$A^{(2)}(\tau) = \int_{-\infty}^{\infty} I(t)I(t - \tau)dt \quad (3.3)$$

This will yield information on the duration of the pulse due to the fact that there will be a change in the SHG signal as a function of delay if the pulses overlap spatially. However, because the autocorrelation is symmetrical, that is $A^{(2)}(\tau) = A^{(2)}(-\tau)$, it cannot distinguish a pulse from its mirror image and a large number of pulses may have the same autocorrelation function so the autocorrelation is only useful for providing an estimate of the FWHM of the intensity distribution and is only really useful with prior knowledge of the pulse shapes. Also, all phase information is lost in retrieving $A^{(2)}(\tau)$.

3.1.2 Fringe resolved autocorrelation

Fringe resolved autocorrelation combines quantities related to the autocorrelation and the spectrum in a single data trace [38]. Also known as interferometric autocorrelation, it involves measuring the second harmonic spectrum vs delay from a SHG crystal placed at the output of a Michelson interferometer, which is basically an autocorrelation measurement performed with collinear beams as shown in Figure 3.2.

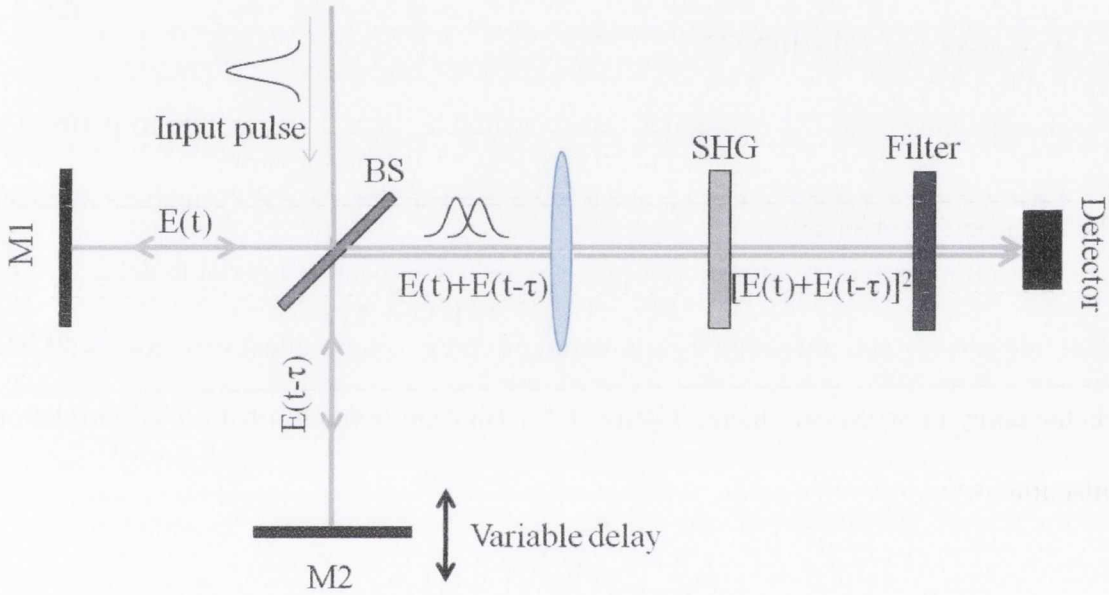


Figure 3.2 Schematic for FRAC measurement, M1 and M2 are mirrors, BS: beam splitter

The second harmonic light produced by the interaction of the two beams combines coherently with that produced by each individual beam. Interference occurs as a result of the coherent addition of the beams and interference fringes occur vs the delay. The measured signal is as follows

$$I_{\text{FRAC}} = \int_{-\infty}^{\infty} \left| [E(t)E(t-\tau)]^2 \right|^2 \quad (3.4)$$

which can be expanded to

$$I_{\text{FRAC}} = \int_{-\infty}^{\infty} \left| E(t)^2 + 2E(t)E(t-\tau) + E(t-\tau)^2 \right|^2 \quad (3.5)$$

The squared terms are due to the SHG signals of the individual pulses and their interference results in the cross term. The interference of these with each other and with the cross terms will yield additional information not present in the standard intensity autocorrelation. It is actually possible to retrieve the pulse intensity and phase using FRAC if it is combined with a measurement of the spectrum using an iterative algorithm. [39]

3.1.3 Cross-correlation

If a shorter pulse than the one under test is available, then it is possible to perform a cross-correlation measurement. The pulse to be measured is passed through a nonlinear medium and is “gated” by the shorter pulse, which is subject to a variable delay, τ . The SHG intensity is then measured as a function of delay using a linear detector much like an autocorrelation measurement. Figure 3.3 shows an example of a cross correlation apparatus.

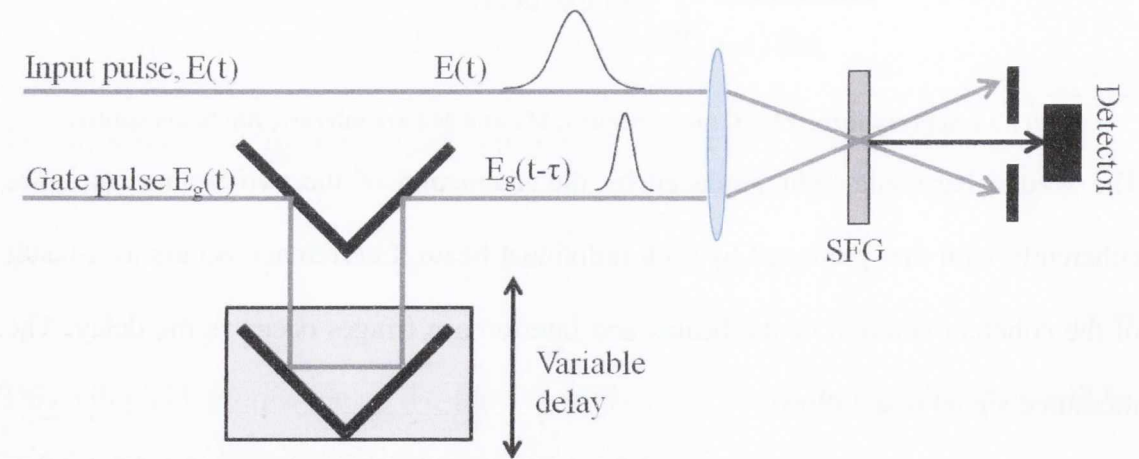


Figure 3.3 schematic for cross-correlation measurement

The cross-correlation, $C^{(2)}(\tau)$ is given as

$$C^{(2)}(\tau) = \int_{-\infty}^{\infty} I(t) I_g(t - \tau) dt \tag{3.6}$$

$I(t)$ is the intensity of the pulse under test while $I_g(t)$ is the intensity of the “gate pulse”. When the gate pulse is much shorter than the pulse under test, the cross correlation yields the temporal intensity precisely but no phase information is retrieved, although it is possible to retrieve the phase using an interferometric cross correlation [40].

3.2 Frequency domain characterisation.

The spectrum of a pulse is the magnitude squared Fourier transform of the time dependent electric field and is given as

$$S(\omega) = \left| \int_{-\infty}^{\infty} E(t) \exp(-i\omega t) dt \right|^2 \quad (3.7)$$

Measurement of the pulse spectrum is usually performed using a spectrometer, the most common of which involves the diffraction of a collimated beam off a grating followed by focusing it onto a camera [41]. It is also possible to use a narrow bandwidth tunable filter in conjunction with a low-speed photodetector. Knowledge of $S(\omega)$ is not alone sufficient to fully characterise $E(t)$.

3.3 Phase sensitive measurements

Attempts have been made at full pulse characterisation using an autocorrelation measurement in conjunction with a measurement of the spectrum. It was seen that the autocorrelation does not determine the intensity and the spectrum does not determine the field but a combination of both methods could be used to extract more information than the use of either technique in isolation [42]. This is known as the Temporal information via intensity (TIMI) method and was based on techniques used by image science and x-ray crystallography who noted that knowledge of the intensity and spectrum can be sufficient to yield the phase in either domain using the Gerchberg-Saxton algorithm [43]. This involves making an estimate for the phase and Fourier

transforming back and forth between the two domains and replacing the magnitude in the relevant domain with the measured quantity. $I(t)$ and $S(\omega)$ can retrieve the phase, however, TIMI measures the autocorrelation of the pulse, not $I(t)$. Refinements were made in order to find the intensity from the autocorrelation but this did not uniquely determine the pulse field.

3.4 Time frequency distributions

Simultaneous information on the temporal and spectral properties of an optical pulse can be used to fully characterise its electric field. A convenient way to represent this information is by using a time-frequency or “chronocyclic” function. Such a function allows the energy distribution of the signal to be described as a function of both frequency and time. The electric field of optical pulses may be represented by any number of time-frequency or “chronocyclic” distributions. Two of the more common ones will be discussed below, namely the Wigner-Ville distribution and the spectrogram/sonogram [44] as both are associated with the more commonly used characterisation techniques.

3.4.1 Spectrogram

The spectrogram is a time-frequency distribution which consists of spectrally resolving a series of time slices of an optical pulse. This may be described using equation (3.8)

$$S_g(\omega, \tau) \equiv \left| \int_{-\infty}^{\infty} E(t) g(t - \tau) \exp(-i\omega t) dt \right|^2 \quad (3.8)$$

where $g(t - \tau)$ is a variable delay gate function or “time nonstationary filter” which can be scanned across the temporal profile of the pulse in short slices. A time nonstationary filter is one which extracts information on the temporal resolution of the pulse distribution such as an autocorrelator. The gate function $g(t - \tau)$ may be realised in

several different ways in the optical domain and each is associated with a particular measurement technique, examples of which will be illustrated at the end of the section. The general scheme for resolving a spectrogram is shown in Figure 3.4.

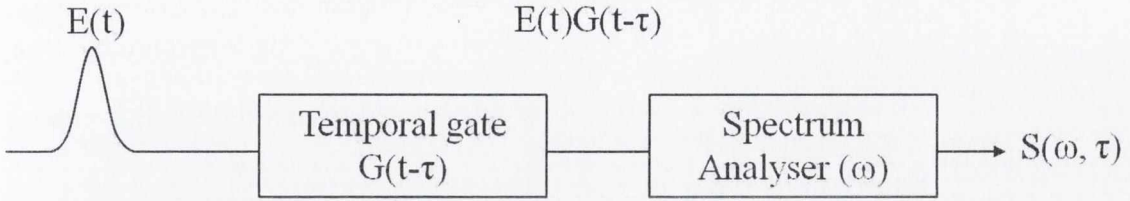


Figure 3.4 measurement of the spectrogram of a pulse train using a temporal gate and frequency resolving device.

The spectrogram is the set of all gated sections of $E(t)$ as τ is varied. Knowledge of the spectrogram of $E(t)$ is sufficient to completely determine $E(t)$.

3.4.2 Sonogram

A similar time-frequency distribution is known as the Sonogram. This differs slightly to the spectrogram in that the order of the elements is reversed and the pulse is gated in the frequency domain by a function $R(\omega-\Omega)$ and a series of spectral slices are resolved in the time domain. The general form of the sonogram is given by equation (3.9)

$$S(\Omega, T) = \left| \int_{-\infty}^{\infty} E(\omega) R(\omega - \Omega) \exp(-i\omega T) d\omega \right|^2 \quad (3.9)$$

Experimental realisation of a sonogram is performed using a scheme based on that shown in Figure 3.5

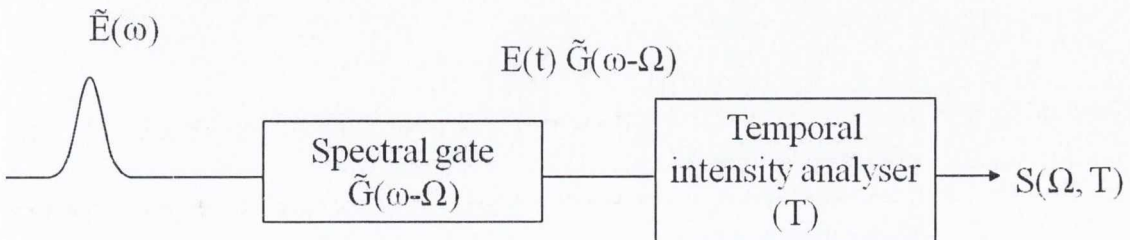


Figure 3.5. General form of a scheme used to measure the sonogram of an optical pulse train.

The spectrogram and sonogram are mathematically identical and the only difference is the manner in which they are generated.

3.4.3 Wigner-Ville distribution

Another chronocyclic representation scheme is the Wigner-Ville distribution, $W(\omega, t)$ [45]. The Wigner-Ville function for the pulse represented by the field $E(t)$ is shown in equation (3.10).

$$W(t, \omega) = \int E\left(t + \frac{t'}{2}\right) E^*\left(t - \frac{t'}{2}\right) \exp(-i\omega' t) dt' \quad (3.10)$$

Like the spectrogram, this gives simultaneous information on the time and frequency distribution of a pulse field in chronocyclic space. The temporal and spectral intensity are found through (3.11) and (3.12)

$$I(t) = |E_{in}(t)|^2 = \int W(\omega, t) d\omega \quad (3.11)$$

$$I(\omega) = |\tilde{E}_{in}(\omega)|^2 = \int W(\omega, t) dt \quad (3.12)$$

The group delay and chirp are equal to the first spectral and temporal moments of the Wigner-Ville function as shown in equations (3.13) and (3.14)

$$\frac{\delta\phi(t)}{\delta t} = \frac{\int \omega W(t, \omega) d\omega}{\int W(t, \omega) d\omega} \quad (3.13)$$

$$\frac{\delta\phi(\omega)}{\delta \omega} = \frac{\int t W(t, \omega) dt}{\int W(t, \omega) dt} \quad (3.14)$$

A schematic of the Wigner-Ville distribution of a linearly chirped pulse is shown in Figure 3.6.

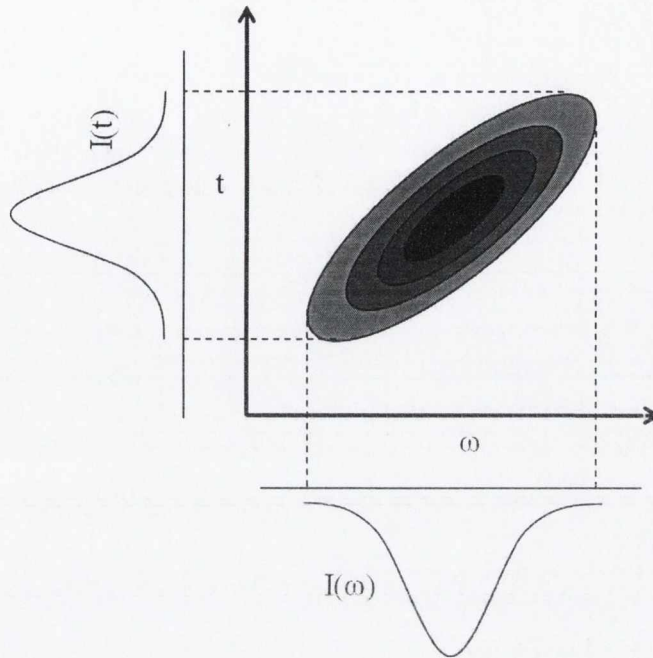


Figure 3.6 Schematic of Wigner-Ville distribution of linearly chirped pulse

The contours refer to the intensity (darker=higher intensity) which is projected for illustrative purposes onto the time and frequency axes.

3.5 Phase sensitive measurement techniques

Having discussed some of the more common time-frequency distributions for representing the field of an optical pulse, it is now time to examine how the electric field of ultrashort optical pulses may be measured. This section investigates some of the more common linear and nonlinear techniques to this end.

3.5.1 Frequency resolved optical gating

Frequency resolved optical gating or FROG is a means by which to measure the spectrogram of an optical pulse [46]. In the case of the commonly used second harmonic generation FROG (SHG-FROG), the gate function chosen to characterise the pulse is the pulse itself. A non collinear Michelson interferometer is used to split the

signal into two beams and impose a variable delay between both signals as shown in Figure 3.7. They are then recombined in a medium with a high second order susceptibility.

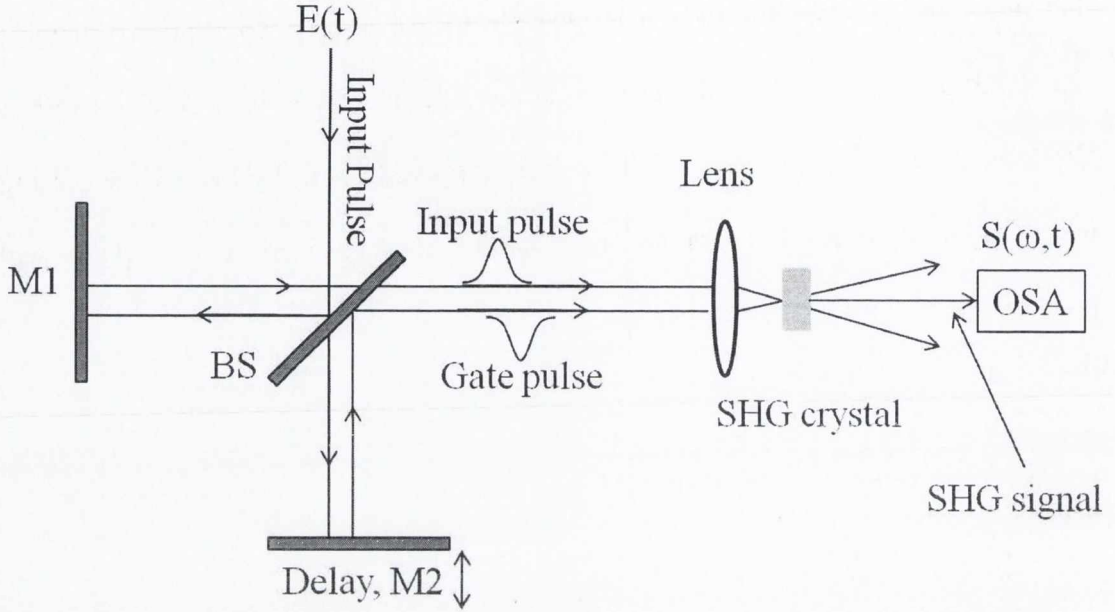


Figure 3.7. Basic FROG measurement scheme, M1 and M2 are mirrors, BS: beamsplitter, OSA: optical spectrum analyser.

The SHG signal is measured using a spectrometer for several delay values. It can be seen that the SHG-FROG scheme resembles a spectrally resolved autocorrelator. A consequence of which is that, like an autocorrelator, measurement of a pulse will yield the same result as one of its mirror image. This means that, in order to accurately characterise the temporal profile of the field there is a requirement for prior knowledge of the pulse shape. The measured SHG signal is proportional to $E(t)E(t-\tau)$ and its value is given as equation (3.15)

$$I_{\text{FROG}}^{\text{SHG}}(\omega, \tau) = \left| \int_{-\infty}^{\infty} E(t)E(t-\tau) \exp(-i\omega t) dt \right|^2 \quad (3.15)$$

It can be seen that this is also the expression for the spectrogram as shown in equation (3.8) . It is possible to use a large range of gate functions in the FROG measurement,

examples include two photon absorption (TPA-FROG) [47], polarisation gating (PG-FROG) [48], cross phase modulation (XPM-FROG) [49], third harmonic generation (THG-FROG) [50] and four wave mixing [51] (FWM-FROG). If a shorter pulse than that under test is available then it is possible to use a cross correlation scheme [52] (XFROG), an advantage in this case is that the gate pulse may be well characterised so $G(t-\tau)$ is known. The full pulse field may be retrieved from the FROG trace using an iterative phase retrieval algorithm. The FROG trace retrieved from a pulse with a positive linear chirp is shown in Figure 3.8.

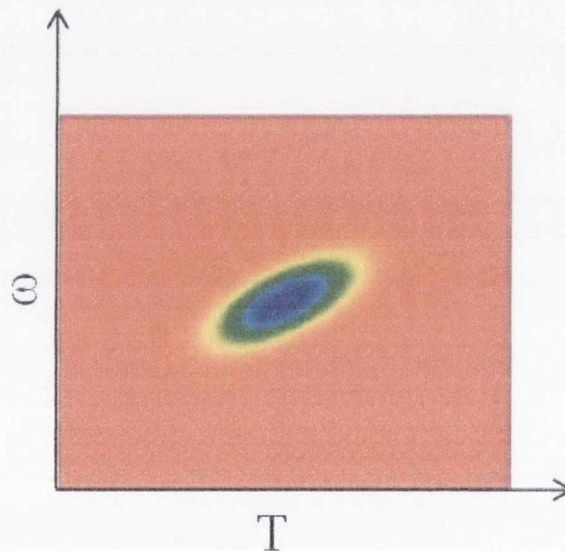


Figure 3.8 Example of FROG trace of a linearly chirped pulse,

3.5.2 Linear spectrograms

The use of nonlinear optical effects in the FROG technique makes it difficult to characterise pulses with low peak intensities and with low duty cycles. It has been shown that an electroabsorption modulator (EAM) can dispense with the requirement for SHG or TPA in the spectrogram measurement [53]. The EAM (as described in chapter 2) may be used as a time-nonstationary filter. A pulse described by $E(t)$ enters a modulator with a response function of $R(t)$ as shown in Figure 3.9. The modulator is

driven at the repetition frequency of the signal and for a series of delay values, τ between the modulation and pulse train a spectrum is recorded.

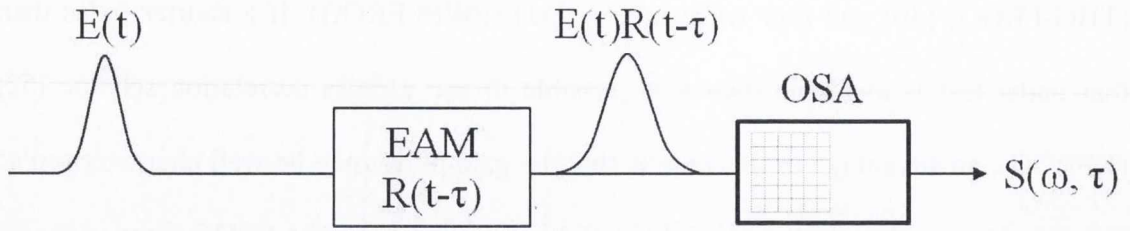


Figure 3.9 Schematic for retrieval of linear spectrogram using an electroabsorption modulator (EAM). OSA: optical spectrum analyser

The retrieved spectrogram is given as

$$\begin{aligned}
 S(\omega, \tau) &= \left| \int_{-\infty}^{\infty} E'(t, \tau) \exp(-i\omega t) dt \right|^2 \\
 &= \left| \int_{-\infty}^{\infty} E(t) R(t - \tau) \exp(-i\omega t) dt \right|^2
 \end{aligned} \tag{3.16}$$

And the extraction of the pulse, $E(t)$ and gate $R(t-\tau)$, from $S(\omega, t)$ is by a blind deconvolution [54], which is performed with the principle component generalised projection algorithm [55]. The pulse spectrum is obtained by either bypassing the modulator or switching it to a state of constant response function.

The gate pulse may be longer than the pulse to be measured, i.e. in [56] a 30ps gate characterised 900fs pulses). It is also possible to use electro-optic modulators to measure the linear spectrogram as shown in [57]

3.5.3 Tomographic pulse monitoring

Tomography is an imaging technique by which two and three dimensional objects are reconstructed by using one or two dimensional projections respectively. It is possible to implement the same principles to pulse characterisation using a method called Chronocyclic Tomography [58]. This is achieved by performing successive rotations of

the Wigner-Ville distribution of the pulse and measuring the resultant projection on the frequency axis (i.e. the spectrum) as indicated by Figure 3.10.

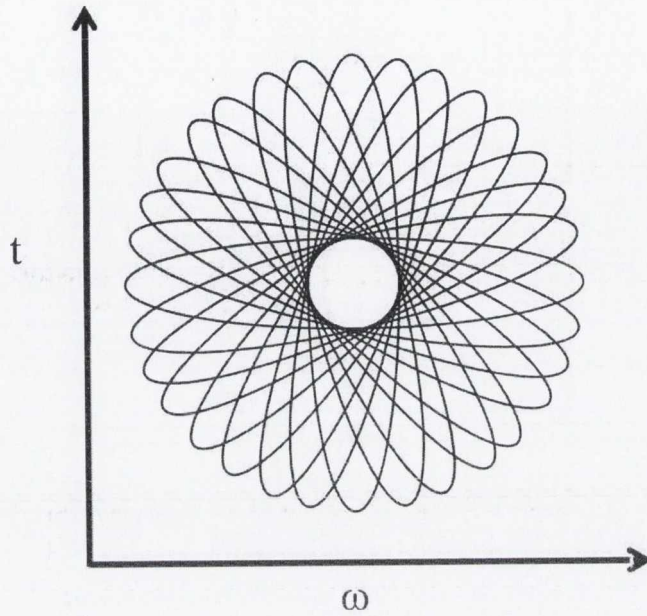


Figure 3.10 Schematic of the Chronocyclic Tomography method in which the Wigner-Ville function is rotated several times and the projections onto the frequency axis (spectra) are measured for each rotation.

The Wigner-Ville distribution is then found using a backprojection algorithm and the electric field of the pulse is found from this. The difficulty in producing large rotations of the Wigner-Ville distribution makes chronocyclic tomography difficult to realise experimentally as a large rotation requires a quadratic temporal and spectral phase modulation. Also, the large number of projections required to reconstruct the Wigner-Ville distribution of an optical field makes the measurement time consuming. As a result, this technique has not yet been implemented experimentally.

3.5.4 Time-frequency converter

If the pulse source is coherent then it is only necessary to measure two projections on the frequency axis. A time to frequency converter can be used to achieve this [59]. The temporal intensity is measured as the spectrum of the pulse that corresponds to the rotation of the W-V function by an angle of $\pi/2$ as shown in Figure 3.11.

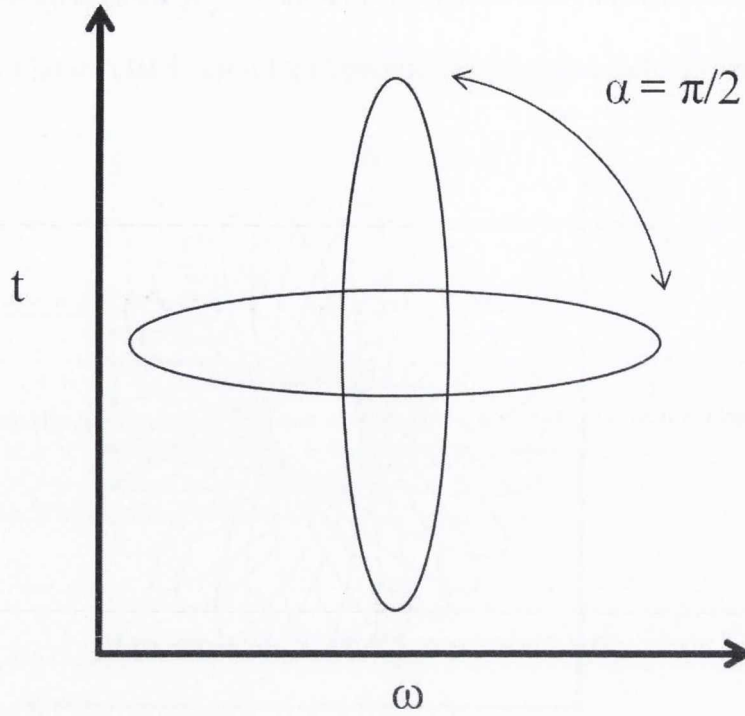


Figure 3.11 Schematic of the Rotation of a Wigner-Ville function caused by a time to frequency converter

A large optical bandwidth is required for this measurement and, even with knowledge of the spectral intensity of the pulse, it does not yield its electric field as no phase information can be retrieved this way.

3.5.5 Simplified chronocyclic tomography

It is possible to use the projections of two small angle rotations of the Wigner-Ville function of a pulse on the frequency axis to characterise its field [60]. With no rotation of the Wigner-Ville function the spectrum is given as

$$I(\omega) = \int W(t, \omega) dt \quad (3.17)$$

when the Wigner-Ville function is rotated by an angle α , its projection on the frequency axis is given as

$$I_{\alpha}(\omega) = \int W[t \cos(\alpha) + \omega \sin(\alpha), \omega \cos(\alpha) - t \sin(\alpha)] dt \quad (3.18)$$

For a rotation of the distribution by an angle α , such that $\alpha \ll 1$ radians the, fractional power spectrum is given as.

$$I_{\alpha}(\omega) = \int W(t + \omega\alpha, \omega - t\alpha) dt \quad (3.19)$$

This small rotation is depicted in Figure 3.12

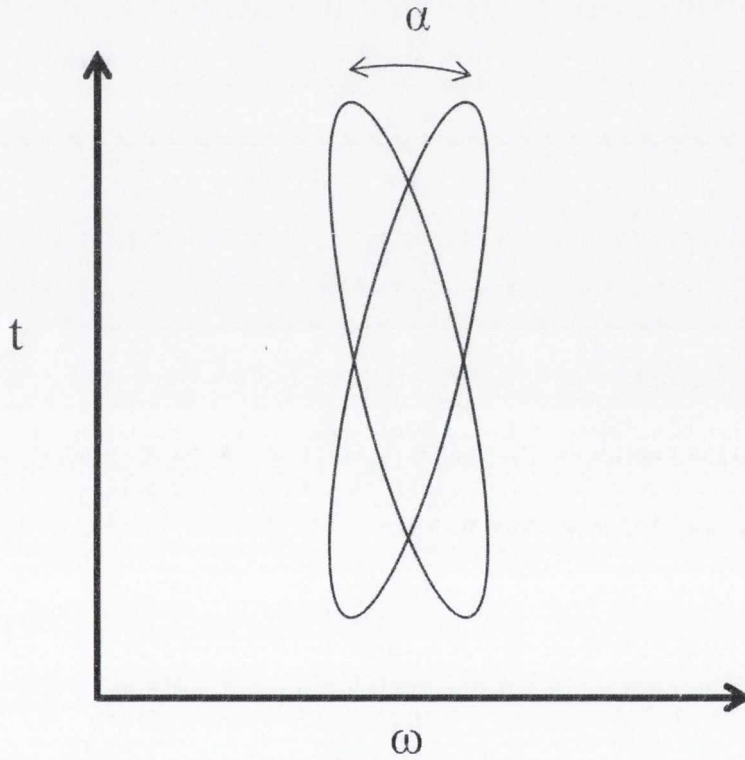


Figure 3.12 schematic of a shear of the Wigner-Ville function in chronocyclic space

Differentiating $I_{\alpha}(\omega)$ with respect to α yields equation

$$\frac{\partial I_{\alpha}}{\partial \alpha} = \int \left(\omega \frac{\partial W}{\partial t} - t \frac{\partial W}{\partial \omega} \right) dt = -\frac{\partial}{\partial \omega} \int t W dt \quad (3.20)$$

As discussed in equation, the group delay is given as

$$\frac{\delta\phi(\omega)}{\delta\omega} = \frac{\int tW(t, \omega) dt}{\int W(t, \omega) dt} \quad (3.21)$$

And, as in equation the spectrum is written as

$$I(\omega) = |\tilde{E}_{in}(\omega)|^2 = \int W(\omega, t) dt \quad (3.22)$$

Combining equations, the derivative of I_α with respect to α may then be written as

$$\frac{\delta I_\alpha}{\delta \alpha} = -\frac{\partial}{\partial \omega} \left(I \frac{\delta \phi}{\delta \omega} \right) \quad (3.23)$$

For a small finite difference in α

$$\frac{\delta I_\alpha}{\delta \alpha} = \frac{I_\alpha - I_{-\alpha}}{2\alpha} \quad (3.24)$$

So then

$$\frac{\partial}{\partial \omega} \left[I \frac{\partial \phi}{\partial \omega} \right] = \frac{I_{-\alpha}(\omega) - I_\alpha(\omega)}{2\alpha} \quad (3.25)$$

So the group delay may then be written as

$$\frac{\partial \phi}{\partial \omega} = \frac{1}{2\alpha I(\omega)} \int I_\alpha(\omega) - I_{-\alpha}(\omega) d\omega \quad (3.26)$$

The group delay may be obtained from the angular derivative of $I_\alpha(\omega)$ as it is rotated between the angles α and $-\alpha$ chronocyclic space. The spectrum is the arithmetic mean of their respective projections on the frequency axis.

$$I(\omega) = \frac{I_{\alpha}(\omega) + I_{-\alpha}(\omega)}{2} \quad (3.27)$$

So both the spectrum and spectral group delay may be found in this manner meaning that full pulse field recovery has been achieved. In order to effect a rotation of the Wigner-Ville function it is necessary to apply a quadratic temporal and spectral phase modulation, but a small rotation such as that discussed here may be approximated by a shear in the distribution and a shear requires only a quadratic temporal phase modulation. [60], This may be achieved by using a phase modulator driven by a sine wave at the pulse repetition rate. If the pulse duration is sufficiently short, and the peaks are aligned with the maxima and minima of the sinusoid, then the phase modulation experienced by the pulses will be approximately quadratic. Figure 3.13 shows the basic setup for such a measurement scheme and the alignment of the pulses with the sinusoidal phase modulation.

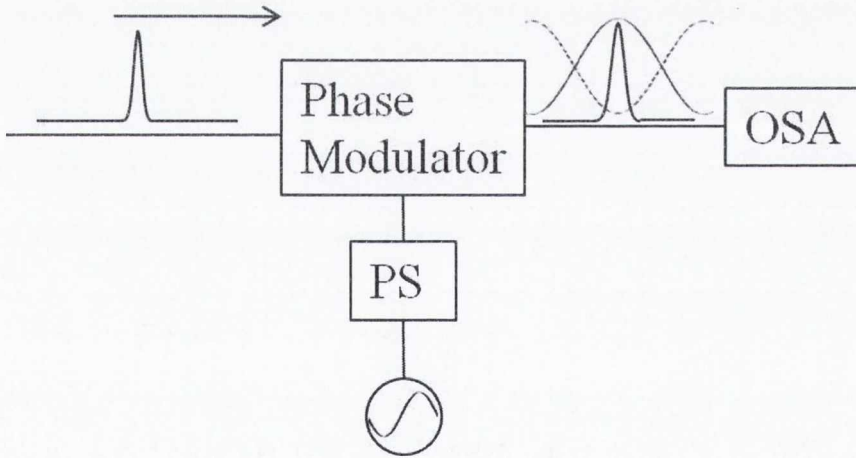


Figure 3.13 Simplified chronocyclic tomography measurement scheme. When the phase modulation is aligned with a sufficiently short pulse it experiences a quadratic phase modulation.

Pulses of ~2ps in duration have been measured in this way. Measurement of pulses of 15ps and above cannot be achieved with the same accuracy as the shorter pulses. This is because the assumption of quadratic phase modulation no longer holds due as the pulse no longer overlaps with the parabolic portion of the phase modulation curve.

3.5.6 Sinusoidal phase modulation

The simplified chronocyclic tomography method is limited in its capacity to measure longer pulses due to the requirement for quadratic phase modulation. Another issue is the fact that the pulse must be aligned with either the peak or trough of the phase modulation for the quadratic phase modulation assumption to hold. It is possible to solve both these issues a similar scheme [61] but using a different alignment of the pulse peaks with sinusoidal phase modulation. The phase modulation frequency is again synchronised with the repetition frequency of the pulse. Four temporal phase modulation values, $\pm\cos(\Omega t)$ and $\pm\sin(\Omega t)$ are investigated and their respective spectra, $I^{(c)}(\pm\psi, \omega)$ and $I^{(s)}(\pm\psi, \omega)$ are taken. $I^{(c)}$ and $I^{(s)}$ refer to the sine and cosine phase modulations. Ω is the angular frequency of the phase modulation. The advantage in this case is that the phase modulations are offset by $\pi/2\Omega$ and there is no requirement for alignment of pulse's peak with a particular point on the phase modulation curve. These spectra are then used to calculate the derivative spectra, which are used to find the amplitude and phase of the electric field. The electric field may be retrieved by using the relation between the two-frequency correlation function of the electric field $E(\omega+\Omega)E^*(\omega)$ and the derivatives of the phase-modulated spectra taken with respect to ψ as shown in equation (3.28)

$$\begin{aligned}\Delta\omega &= \left. \frac{dI^{(c)}(\psi, \omega)}{d\psi} \right|_{\psi=0} + i \left. \frac{dI^{(s)}(\psi, \omega)}{d\psi} \right|_{\psi=0} \\ &= i \tilde{E}(\omega + \Omega) \tilde{E}(\omega) \sin c\left(\beta^{(2)} L \left[(\omega + \Omega)^2 - \omega^2 \right] / 4\right) \\ &\quad - i \tilde{E}(\omega) \tilde{E}^*(\omega) \tilde{E}^*(\omega - \Omega) \sin c\left(\beta^{(2)} L \left[\omega^2 - (\omega - \Omega)^2 \right] / 4\right)\end{aligned}\quad (3.28)$$

$\beta^{(2)}$ is the second order dispersion of the phase modulator and L is the length of the phase modulator. $\tilde{E}(\omega)$ is the electric field propagating at the midpoint of the modulator and is related to the input field $E(\omega)$ through

$$\tilde{E}(\omega) = E(\omega) \exp \left[i \beta^{(2)} \omega^2 L / 4 \right] \quad (3.29)$$

In the case where $\beta^{(2)}$ is negligible (3.28) can be simplified to

$$\Delta(\omega) = i \left[E(\omega + \Omega) E^*(\omega) - E(\omega) E^*(\omega - \Omega) \right] \quad (3.30)$$

The frequency correlation function is calculated from the linear difference relations in equations (3.28) and (3.30). From this, the spectral phases may be recursively calculated. The spectral electric field is then calculated using the calculated spectral phase and measured spectral intensity.

3.6 Interferometric pulse measurements

Interferometry involves converting the phase information of an optical signal into intensity information, which allows measurement of the phase difference between different frequency components of the pulse. Measurement of the phase differences over the full pulse spectrum allows the phase change across the pulse to be determined.

3.6.1 Spectral shearing interferometry

Spectral shearing interferometry (SSI) [62] involves interfering pulse with a replica of itself that has been subject to a frequency shift or “shear”. A field $\tilde{E}(\omega)$ is interfered with a replica field $\tilde{E}(\omega - \Omega)$, which has been sheared by Ω along the frequency axis. Measurement of the interferometric component yields the field

$$E_{\text{INT}}(\omega) = \sqrt{I(\omega)}\sqrt{I(\omega-\Omega)} \exp\{i[\varphi(\omega) - \varphi(\omega-\Omega)]\} \quad (3.31)$$

the argument of which is the phase difference $\Delta\varphi = \varphi(\omega) - \varphi(\omega-\Omega)$. $\Delta\varphi$ is then integrated to find the spectral phase. $I(\omega)$ is then recovered by an independent measurement of the spectrum.

Experimental realisation may be achieved through “Spectral Phase Interferometry for Direct Electric Field reconstruction” (SPIDER) [63]. A pulse train is split into two signals as shown in Figure 3.14.

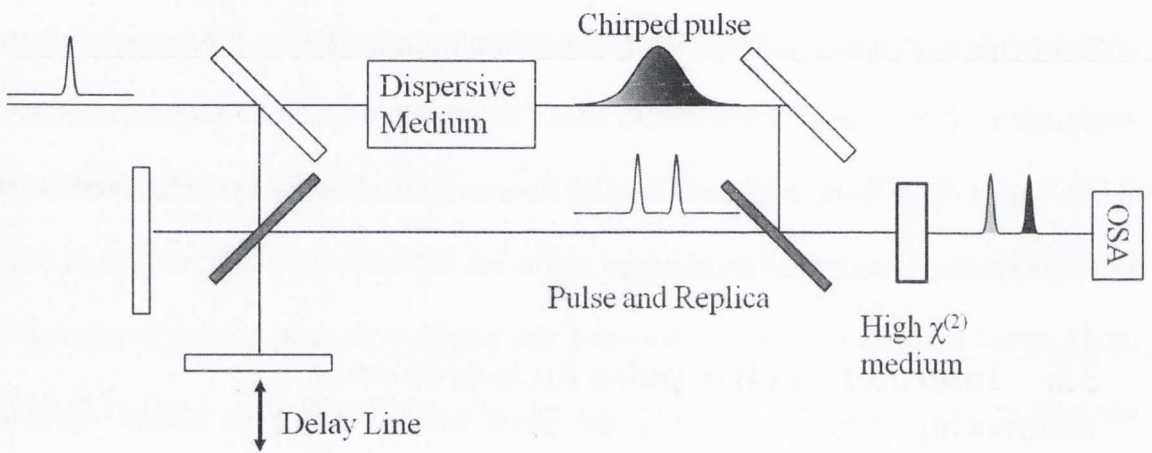


Figure 3.14 setup for SPIDER scheme, OSA: optical spectrum analyser.

One is passed through a dispersive medium which causes a large chirp. The other signal is split by means of a Mach-Zehnder interferometer such that there is a delay τ between the pulse and its replica. These are then recombined with the dispersed pulse in a material with a high second order susceptibility $\chi^{(2)}$, which causes a sum frequency generation. Each of the delayed pulses interacts with a different frequency component of the chirped pulse and experiences a different up conversion in the nonlinear medium. The signal is then passed into an optical spectrum analyser where the measured spectrum is the interference of the two spectrally sheared pulses. The use of nonlinear optical effects means that the input pulses' peak intensity must be high, especially as the signal is split to such an extent.

3.6.2 Electro-optic spectral shearing interferometry.

A more recent version of SSI focuses on the use of an electro optic modulator [64]. A signal is split and the two pulse replicas are given a delay τ . The pulses are recombined before passing through a phase modulator, which is driven by an RF frequency of $1/2\tau$. With appropriate alignment with the phase modulation as shown in Figure 3.15

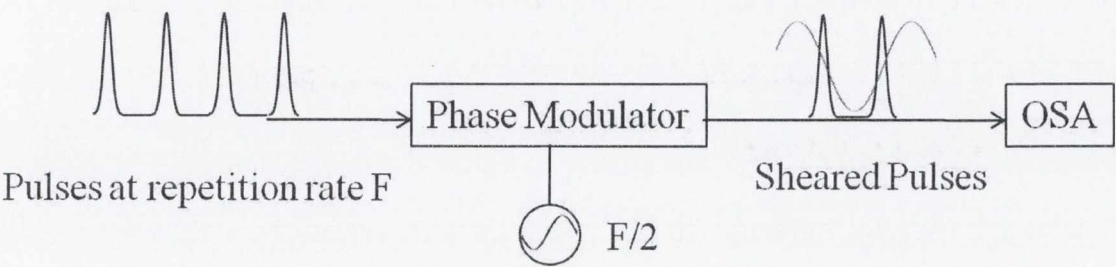


Figure 3.15 Setup for electro-optic spectral shear interferometry

The recombined pulses are subject to an equal but opposite linear temporal phase shift. The effect of this (as was shown in chapter 2) is to impose a frequency shift of $\pm\Omega$ on the pulse. An optical spectrum analyser is used to measure the spectral intensity of the sheared pulses and the full field retrieval may be achieved using the same treatment as the SPIDER technique.

3.6.3 Spectral phase interferometry using an MZI

Another interferometric method involves the use of a Mach-Zehnder modulator [65] and forms the basis of a commercially available high-resolution phase sensitive spectrum analyser. The pulses are passed through the modulator as shown in Figure 3.16.

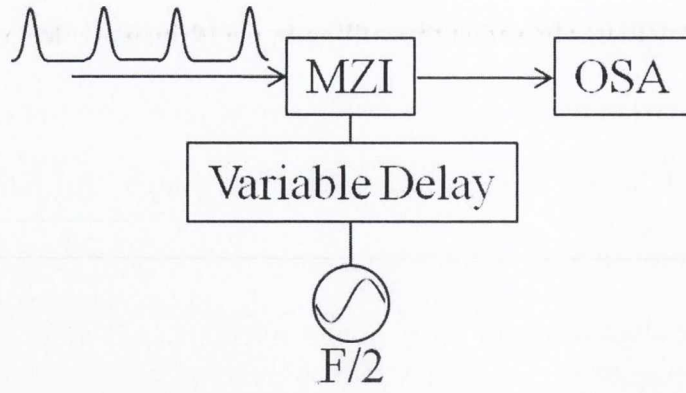


Figure 3.16 SPI by Mach-Zehnder modulator (MZI).

The modulator is biased at the null point and driven at half the repetition of the pulse train with a low driving voltage such that $v \ll v_\pi$. The driving signal may be shifted in time using a variable delay line. Prior to the modulator the pulse spectrum consists of a series of lines separated by the repetition frequency of the pulse as shown in Figure 3.17

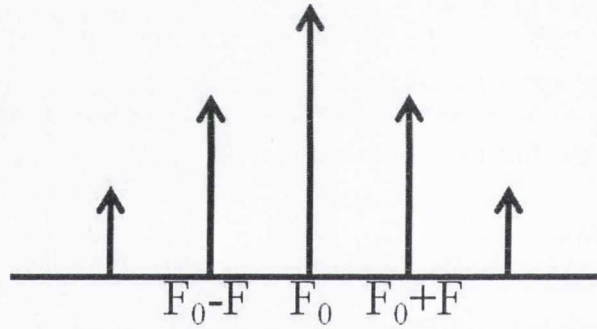


Figure 3.17 Schematic of the spectrum of a pulse with repetition rate F and carrier frequency F_0 .

The spectral field of the pulse prior to the modulator is given as the Fourier transform of $E(t)$

$$\tilde{E}_{in}(\omega) = \sum_k E_k \delta[\omega - (\omega_0 - k\omega_m)] \quad (3.32)$$

where $E_k = A_k \exp(i\Phi_k)$ and δ refers to a Dirac function. K is an integer and refers to the order of each spectral line, ω_m is the repetition frequency of the pulse multiplied by 2π .

The output field is obtained by multiplying the input field by the transfer function of the

Mach Zehnder modulator biased at the null point and driven at a frequency of $F/2$ which is given as

$$T(t) = m \cos[\pi \omega_m (t - \tau)] \quad (3.33)$$

τ is the time lag produced by the delay line and m is the depth of the phase modulation which is the extent of the driving voltage compared to V_π . This must be kept to approximately 30% of V_π in order to avoid third harmonics becoming non negligible. The Fourier transform of the transfer function is convolved with $\tilde{E}_{in}(\omega)$ to obtain the field of the pulse following the modulator.

$$\tilde{E}_{out}(\omega) = \tilde{E}_{in}(\omega) \otimes \tilde{T}(\omega) \quad (3.34)$$

The effect of the modulation is such that each spectral line in the pulse forms two side bands at a spacing of $F/2$ and the centre line is suppressed due to biasing at the null point. Each subsequent line also forms a pair of side bands as shown in Figure 3.18.

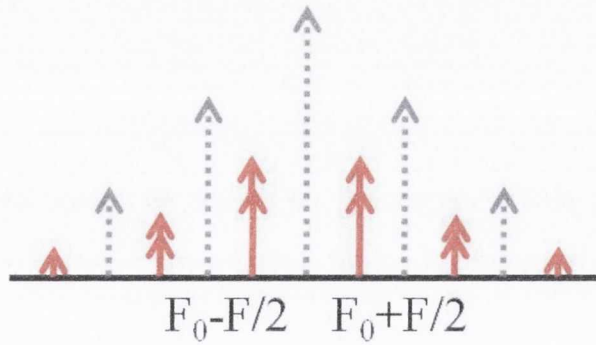


Figure 3.18 Schematic of the output spectrum of a Mach-Zehnder interferometer driven at a frequency $F/2$

Neglecting the now suppressed original spectrum, the output field can then be written as equation (3.35).

$$\begin{aligned}
\tilde{E}_{\text{out}}(\omega) = & \\
& \frac{m}{2} \sum_k E_k \exp\left(-i \frac{\omega_m}{2} \tau\right) \delta\left[\omega - \omega_0 - (k - 0.5) \omega_m\right] \\
& + \frac{m}{2} \sum_k E_k \exp\left(+i \frac{\omega_m}{2} \tau\right) \delta\left[\omega - \omega_0 - (k + 0.5) \omega_m\right]
\end{aligned} \tag{3.35}$$

This describes the newly formed spectral lines. The newly formed lines interfere with each other and the resultant intensity is related to the phase difference between these lines and the delay between the pulse and phase modulation. Equation (3.35) may be simplified as follows

$$\tilde{E}_{\text{out}}(\omega, \tau) = \frac{m}{2} \sum_k B(k, \tau) \delta\left[\omega - \omega_0 - (k - 0.5) \omega_m\right] \tag{3.36}$$

where

$$B(k, \tau) = A_k \exp i\left(\Phi_k + \frac{\omega_m}{2} \tau\right) + A_{k+1} \exp i\left(\Phi_{k+1} - \frac{\omega_m}{2} \tau\right) \tag{3.37}$$

The output spectrum from the modulator is given as

$$\tilde{I}_{\text{out}}(\omega, \tau) = \left|\tilde{E}_{\text{out}}(\omega)\right|^2 = \frac{m^2}{4} \sum_k c(k, \tau) \delta\left[\omega - \omega_0 - (k - 0.5) \omega_m\right] \tag{3.38}$$

Where c may be written as

$$C(k, t) = A_{k+1}^2 + A_k^2 + 2A_k A_{k+1} \cos \Phi(k, \tau) \tag{3.39}$$

And Φ is given as:

$$\Phi(k, \tau) = 2\pi\omega_m (\tau - \tau_0) + (\Phi_{k+1} - \Phi_k) \quad (3.40)$$

The output intensity of each line $C(k, t)$ is described by a sinusoidal function of τ with a phase $\phi_{k,k+1}$. By varying τ and measuring the spectrum, $\phi_{k,k+1}$ is found. So equation (3.40) yields

$$\phi_{k,k+1} = \omega_m \tau_0 (\Phi_{k+1} - \Phi_k) \quad (3.41)$$

The phase of the most intense line of the pulse located at $k=0$, Φ , is set to zero so that the phase differences between each line may be measured with this as a reference. τ_0 is the instant where $\Phi_1 = \Phi_{-1}$, then

$$\tau_0 = \frac{\phi_{0,1} + \phi_{-1,0}}{2\omega_m} \quad (3.42)$$

So the phase of a line k relative to $k=0$ may be determined using recurrent relations as shown in (3.43) and (3.44)

$$\Phi_k = \Phi_{k-1} + \omega_m \tau_0 - \phi_{k+1,k} \quad k > 0 \quad (3.43)$$

$$\Phi_k = \Phi_{k+1} - \omega_m \tau_0 + \phi_{k,k+1} \quad k < 0 \quad (3.44)$$

So the phase differences between adjacent lines k are computed from (3.43) in the direction of positive k values if the centre and from (3.44) in the negative k values. Usually spectra are measured at four different values of τ for this scheme. The intensity variations of the spectral lines as a function of τ as measured by the spectrum analyser and the phase difference across all k values may be calculated.

3.7 Summary

The manner in which ultrashort optical pulses are measured was introduced in this chapter. Due to the use of RZ modulation formats and shorter pulses in communications the standard method of direct detection using photodiodes was deemed to be insufficient

for measuring ultrashort pulses. Various techniques for the temporal and spectral characterisation of optical pulses were examined. While it is possible to resolve the spectrum with relative ease, the temporal FWHM and shape can only be approximated by autocorrelation measurements but some only give an approximation as to the pulse shape and temporal width and oftentimes this may be accompanied by prior knowledge of how the pulse was generated. Various means for the complete characterisation of the pulse field were investigated, which could be categorised into spectrographic measurements such as FROG and linear spectrograms, tomographic measurements such as the simplified chronocyclic tomography method and interferometric methods such as SPIDER, electro-optic spectral shear interferometry and a spectral phase interferometry method using a Mach-Zehnder modulator which has seen applications in the APEX communications spectral analyser. Limitations in these measurements were explored and these dictate the usage of the techniques for certain pulse durations, duty cycles and in some cases measurement times may be important.

Chapter 4 OSNR monitoring

4.1 Optical signal to noise ratio and ASE

The transmission of data in long haul optical networks requires that the attenuation in optical fibres be compensated. In the past, the optical signal was converted to an electrical signal and was reamplified, reshaped, retimed (“the three Rs”) and then reconverted using an optical regenerator and transmitted over an optical link. Such a conversion performed at regular intervals was costly and detrimental to the maximum transmissible bitrates due to the “electronic bottleneck”. In order to increase transmission speeds and network transparency, the fibre losses have to be compensated in the optical domain by the periodic placement of Erbium doped fibre amplifiers (EDFA) [66]. This removed the requirement for conversion of the signal to the electrical domain. The increase in transparency provided by this development is somewhat offset by two particular issues;

- Conversion of the signal to the electrical domain allowed the measurement of signal quality parameters at these stations. As result of this is that a faulty component could be traced back to one particular repeater unit or a particular span of fibre. This does not occur automatically in a network with periodic optical amplifiers.

- Transparent transmission does not allow retiming or reshaping of the signal, meaning that impairments such as timing jitter and dispersion propagate through the fibre link [67].

4.2 Optical signal to noise ratio (OSNR)

The cascading of optical amplifiers in networks causes the accumulation of amplified spontaneous emission (ASE) noise in signal channels. As the signal traverses several amplifier stages, this ASE becomes amplified [67]. As the noise level increases, it becomes difficult to distinguish between the on and off states of an intensity modulated

signal. This manifests itself as an increase in the bit error rate of the signal [68]. It is important to monitor the noise level as its magnitude has a profound effect on the signal quality. Knowledge of this noise level is also important in the diagnosis of faulty equipment as awareness that all components are operating correctly is insufficient to ensure the quality of the received signal [13]. This is because the components have a range of operating parameters and one can compensate another further down the line. For example, amplifier performance degradation due to excess tilt may be compensated by use of dynamic gain equalisation so the signal power level remains constant even though the signal quality may be severely degraded.

4.3 OSNR measurement

The optical signal to noise ratio (OSNR) is a quantity that provides a measure of the ASE present in a received signal and as such is one of the main indicators of general signal health in optical communications networks [13]. Knowledge of the OSNR evolution throughout a fibre link may be used as a fault location mechanism as a sudden decrease in the OSNR at a given point signifies a fault and allows network engineers to trace the fault back to the previous amplifier unit. The OSNR may also be used as a feedback tool to dynamically balance amplifier gain [13]. The OSNR is the ratio of the signal power to the ASE noise power in a 0.1nm optical bandwidth. Its magnitude at the channel wavelength is usually found using an optical spectrum analyser. The noise is usually measured at a certain frequency offset from the centre frequency of the channel. A linear interpolation is performed to find the ASE level at the carrier frequency [69,70] as depicted in Figure 4.1

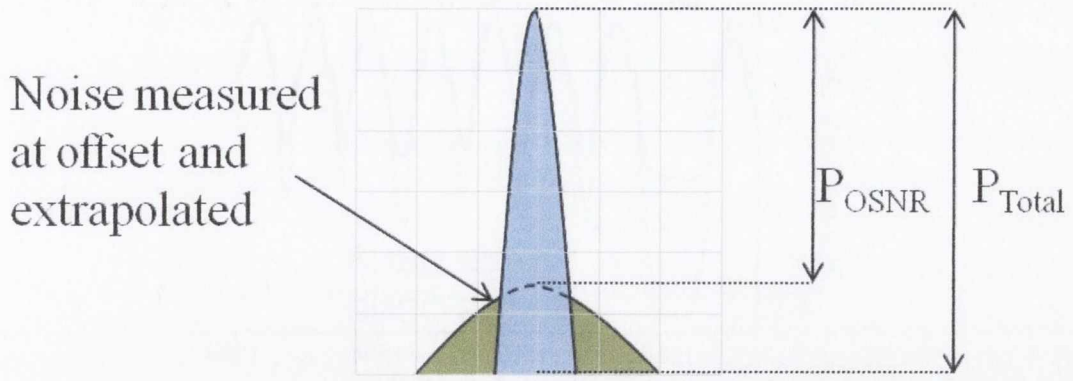


Figure 4.1 Interpolation method for determination of the in-band OSNR

Given the range and magnitude of the OSNR, it is convenient to express it on a logarithmic scale.

$$OSNR = 10 \log_{10} \left(\frac{P_s}{P_n} \right) \quad (4.1)$$

One issue that arises is when the spectral efficiency increases, the inter-channel spacing decreases and difficulties arise in finding the noise floor between adjacent channels as the noise level between them may be much lower than the minimum optical power between the channels due to spectral overlap [13]. This may be appreciated by observing the effect of switching off one of the channels as shown in Figure 4.2. It can be seen that, were the minimum power between channels taken as the out of band noise, the in-band OSNR would have been significantly underestimated. In this case, spectrum based estimation of the in-band OSNR monitoring would not be suitable for providing a reliable estimate of the in-band OSNR.

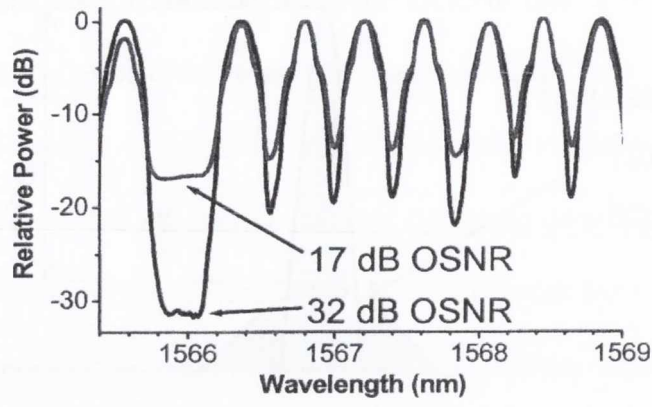


Figure 4.2 Effect of switching off a channel on the OSNR measured by the interpolation method between densely-packed channels. Figure from [13].

4.4 OSNR in reconfigurable optical networks

Reconfigurable optical networks take advantage of the ability to add or drop channels along a fibre link using reconfigurable optical add-drop multiplexers (ROADMS) [71]. A consequence of this is that neighbouring channels may have had vastly differing transmission histories. As a result, the ASE level of a given channel may not be inferred from the linear interpolation method as the neighbouring channels may have traversed different amplifier spans before reaching the monitoring point. This renders the result of a linear interpolation measurement meaningless. The issue now is to directly determine the noise at the carrier frequency.

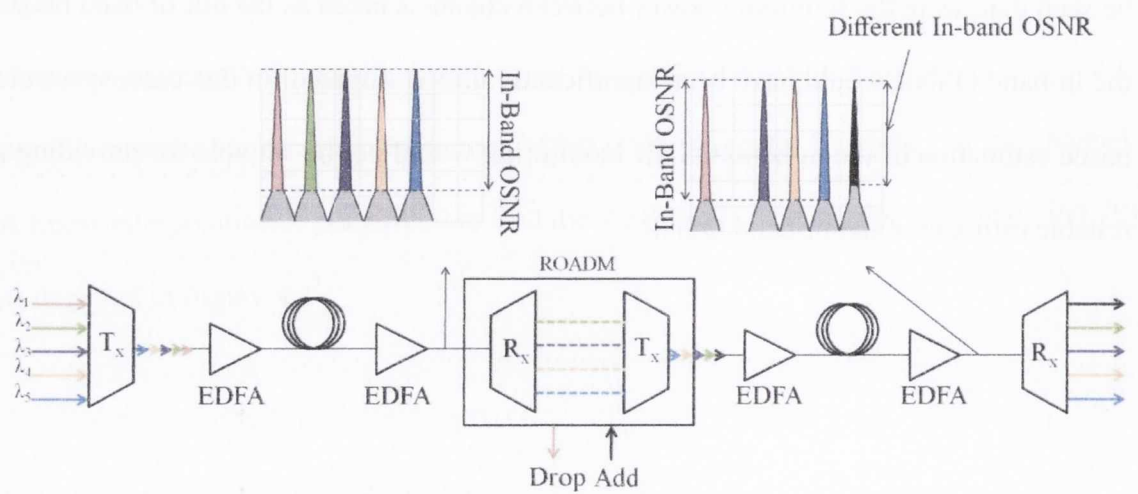


Figure 4.3 Schematic of the effect of channel history on the in-band OSNR in a hypothetical transparent reconfigurable network.

Figure 4.3 illustrates the difficulty mentioned above. Five WDM channels are multiplexed and launched from the transmitter T_x . One channel is dropped and another added at the ROADM. At the receiver (R_x), the in-band OSNR of the added channel differs from the others meaning that an interpolation measurement would not be appropriate to determine the in-band OSNR.

4.5 In-band OSNR measurement

As the linear interpolation method is unsuitable for reconfigurable optical networks, it becomes necessary to measure the noise directly in band. Several techniques exist to perform this measurement relying on various physical differences between signal and ASE noise. The more prominent examples are examined below.

4.5.1 Polarisation nulling

Perhaps the most common method currently used for monitoring of in-band OSNR is the polarisation-nulling method [72, 73, 74]. This is based on the assumption that the data signal is fully polarised and the ASE is unpolarized.

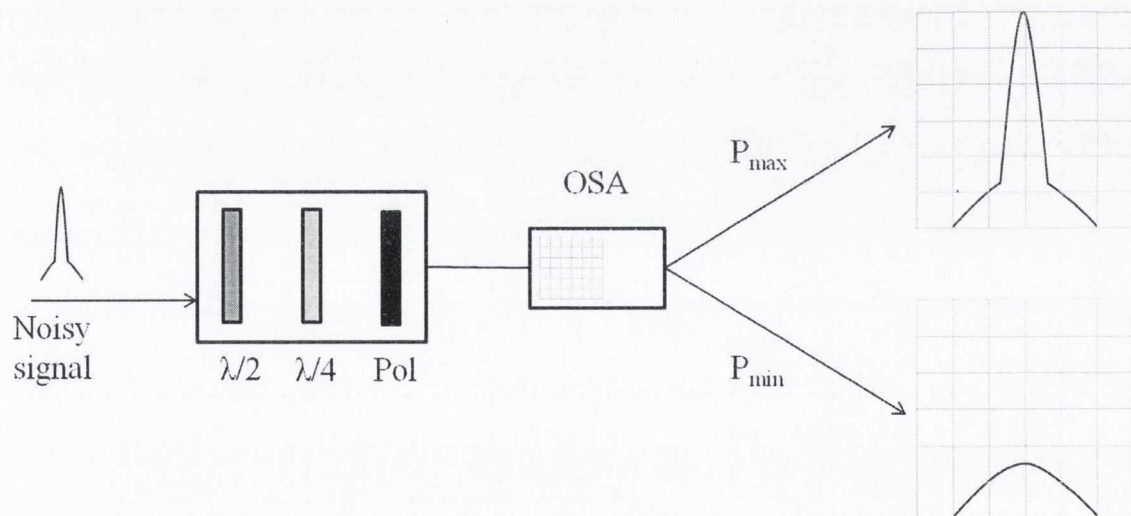


Figure 4.4 Monitoring of in-band OSNR by polarisation nulling

The noisy signal is passed through a polarisation controller followed by a polariser and spectrum analyser as shown in Figure 4.4. As the Poincare sphere is scanned by the polarisation controller a maximum and minimum transmission will be observed on the

OSA. The maximum corresponds to the case where the signal polarisation is aligned with the polariser, in which case

$$P_{\max} = P_s + 0.5P_n \quad (4.2)$$

The factor of 0.5 is due to the equal partition of noise power between the two orthogonal polarisation modes in the fibre. A minimum in transmitted power signifies the case where the signal polarisation is orthogonal to the polariser and only noise is transmitted through the polariser, In this case the power is

$$P_{\min} = 0.5P_n \quad (4.3)$$

P_s and P_n are the original signal and noise powers respectively. The measured OSNR is easily recovered as (4.4).

$$\text{OSNR} = \frac{P_s}{P_n} \left(\frac{B_n}{B_r} \right) = \frac{P_{\max} - P_{\min}}{2P_{\max}} \quad (4.4)$$

B_n is the noise equivalent bandwidth and B_r is the resolution bandwidth. B_n is determined by the spectral width channel filter. Measurement of the OSNR is performed simply by measuring $P_{\max}$ and $P_{\min}$ on the spectrum analyser. A number of other variations of this scheme may be used; these involve splitting the signal into two orthogonally polarised components and independently measuring the power in each of the components. However, commercially available schemes use the OSA form.

Unfortunately, the assumption of unpolarised noise does not always hold for the noisy signal due to polarisation dependent loss in the fibre. Also, depolarisation of the signal due to polarisation mode dispersion means that the assumption of a fully polarised signal is not always reasonable either. Also, as the signal polarisation is likely to fluctuate, it is necessary for the polarisation controller to scan the full Poincare sphere if a real-time measurement is to be performed.

4.5.2 Orthogonal polarisation heterodyne mixing

Orthogonal polarisation heterodyne mixing [75] involves splitting the signal under test into equal power components after the channel filter. Each signal arm passes through narrow bandwidth optical bandpass filters with slightly different centre frequencies f_1 and f_2 . The bandwidths of the filters, B_1 and B_2 are such that there is no spectral overlap between the two signals. Both signal arms have a power tap and one has a polarisation scrambler. The split signals are then recombined and passed into a photodiode. The electrical signal is passed through an electrical bandpass filter which is set at the frequency $\Delta F=f_1-f_2$. An RF power detector is used to measure the power of the beating between the combined signals. A schematic of such a measurement is shown in Figure 4.5

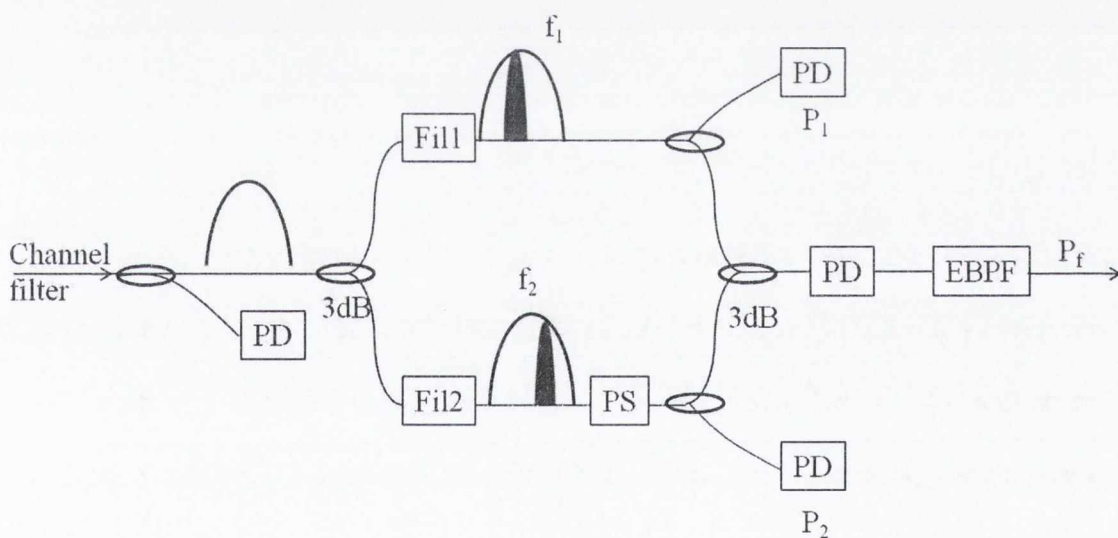


Figure 4.5. In-band OSNR monitoring using orthogonal polarisation heterodyne mixing. PD: photodiode, Fil1 and Fil2: optical bandpass filters. PS: polarisation scrambler. EBPF: electrical bandpass filter.

As the polarisation scrambler adjusts to the point where the two signals are orthogonal the signal-signal beating disappears leaving only the signal-ASE and ASE-ASE beating terms, which corresponds to the lowest RF power value. Measurement of the RF power following the bandpass filter and the optical power in both split signals allows the noise power orthogonal to the signal after filter “FIL1” to be calculated as in equation (4.5)

$$N_{\perp} = \frac{(\alpha P_1 + P_2) - \sqrt{(\alpha P_1 + P_2)^2 - 16\alpha P_{\Delta F} B_2 / (R^2 R_L B_e)}}{4\alpha} \quad (4.5)$$

P_1 and P_2 are the powers of the signals after the filters f_1 and f_2 respectively, $P_{\Delta F}$ is the minimum electrical power detected after the electrical filter, R is the photodiode responsivity, R_L is the resistance of the load resistor used in conjunction with the photodiode, α is the ratio of the noise power after fil2 and fil1 Assuming $B_2 \geq B_1$ and that the noise is unpolarised the power after the channel filter is given as in equation (4.6)

$$P_0 = P_s + 4 \frac{N_{\perp}}{B_1} B_{ch} \quad (4.6)$$

B_{ch} is the bandwidth of the channel filter. P_s is the signal power and the second term in equation is the noise power, where the factor of 4 is to compensate for the reduction of the optical power fourfold due to passing through a pair of 3dB couplers (assuming the power tap was negligible). The OSNR can then be calculated as (4.7)

$$OSNR = \frac{P_0}{4N_{\perp}} \frac{B_1}{B_{OSNR}} - \frac{B_{ch}}{B_{OSNR}} \quad (4.7)$$

B_{OSNR} is usually 0.1nm as mentioned in section 4.2.

This scheme is immune to the effects of polarisation mode dispersion but the assumption of unpolarised noise was made in the treatment of the data so the measurement is not immune to polarised noise.

4.5.3 Stimulated Brillouin scattering

Stimulated Brillouin scattering is a nonlinear effect that involves the efficient transfer of power from a pump wave to a frequency shifted backscattered Stokes wave propagating in the opposite direction. The process arises from spontaneous inelastic scattering off

acoustic waves which are thermally excited. The moving interference beat pattern resulting from the pump and frequency shifted wave causes a pressure wave in the material by electrostriction. This in turn modulates the refractive index of the fibre, stimulating further scattering of the pump and backscattered waves. Stimulated Brillouin scattering occurs when an optical pump beam of sufficient power (this is known as the Brillouin threshold) is incident on a nonlinear medium[76]. A highly nonlinear fibre (HNLF), typically of the order of a kilometre in length is used in conjunction with a circulator and optical spectrum analyser as shown in Figure 4.6. The noisy signal is fed into the HNLF at sufficient power to pump SBS, which may be used to measure the OSNR due to the fact that when noise is introduced along with the signal, the pump wave experiences a loss of coherence. This causes a reduction in the efficiency of power being transferred to the Stokes wave. As the noise power increases, the backreflected power due to the Stokes wave decreases. An OSA is then used to measure the frequency-shifted signals' intensities which can then be used to obtain a measure of the in-band OSNR [77,78].

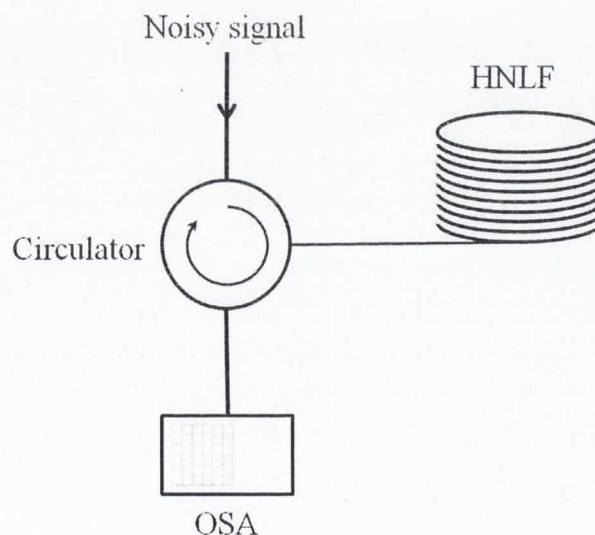


Figure 4.6. In-band OSNR monitoring using stimulated Brillouin scattering HNLF: highly nonlinear fibre.

From the literature it appears that optical powers of approximately 45mW are needed to excite SBS to a meaningful degree [79], meaning that the signal may need to be amplified before the monitoring system. One advantage is that the process is only weakly influenced by polarisation changes

4.5.4 Two-Photon absorption

Two-photon absorption is a nonlinear process where the energy and momentum of a pair of photons combine to form an electron-hole pair in a medium [80]. This occurs when the photon energy $h\nu$ falls within the range $1/2E_g \leq h\nu \leq E_g$, where E_g is the bandgap of the material in question (usually a semiconductor). The resultant photocurrent is proportional to the incident electric field to the fourth power. A result of this is that the TPA photocurrent is a function of the peak power of the signal and of its duty cycle, a property that allows its use in autocorrelation of short pulses as discussed earlier. A TPA based monitor in its simplest form consists of a semiconductor device which is constructed to optimise the generated TPA photocurrent using either a TPA semiconductor microcavity structure such as that developed by our group. The microcavity structure incorporates Bragg reflectors which act to enhance the two-photon absorption process at the cavity resonance frequency. TPA enhancement of 10^4 have been demonstrated over bulk semiconductor devices [81]. An avalanche photodetector may also be used [82] in conjunction with a data acquisition device. The dependence of the photocurrent on the signal and noise powers can be written as:

$$(I_{\text{TPA}})_S \propto \frac{3}{2} P_S^2 \quad (4.8)$$

$$(I_{\text{TPA}})_N \propto 3P_N^2 \quad (4.9)$$

For a given intensity, the TPA generated by ASE is twice that generated by signal.

The total time averaged photocurrent generated by the noisy signal is given as

$$I_{ph} = C_2 \bar{P}^2 \left[1 + \left(\frac{r}{1+r} \right)^2 \left(\frac{1}{2D} - 1 \right) \right] + C_1 \bar{P} \quad (4.10)$$

$\bar{P}$ is the average optical power, r is related to the OSNR as

$$r = 10^{\frac{OSNR}{10}} \left(\frac{0.1}{NEB} \right) \quad (4.11)$$

D is the effective duty cycle of the signal, C_1 and C_2 are constants relating to the single and two-photon absorption efficiencies respectively and are a function of the monitoring unit and hence may be easily established. If the duty cycle is known it is possible to find the r and hence the OSNR. Due to the inherent polarisation dependence, implementation of this scheme may be challenging [83].

4.5.5 Mach-Zehnder interferometer

It has been established that measurements based on the differing polarisation properties of signal and ASE can be susceptible to partly polarised noise and PMD changing the optical signal. Nonlinear effects used to measure the in-band OSNR will almost always suffer from the requirement for high input powers. These issues may be addressed through exploitation of the coherence difference between signal and ASE. To this end it is possible to use an asymmetric Mach-Zehnder fibre interferometer [84, 85, 86] and the interferogram of the noisy signal is used to determine the OSNR. ASE, by virtue of its larger spectral width, will have different interference properties to signal. Typically the interferometric OSNR measurement takes the form of Figure 4.7

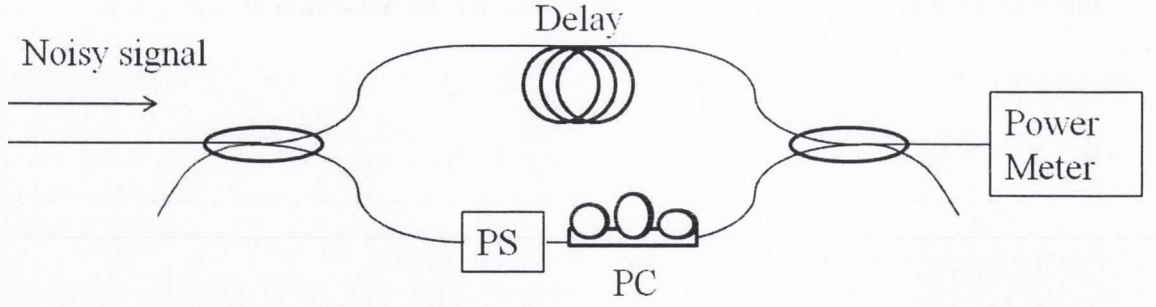


Figure 4.7. Optical delay Mach-Zehnder interferometer PS: phase shifter, PC polarisation controller.

The signal under test is split with a 3dB coupler and one arm is subject to an optical delay of the order of or greater than the bit period of the signal while the other arm has a phase delay of the order of the optical cycle time ($\sim 5\text{fs}$). A polarisation controller is used to ensure maximum extinction due to the interference. The signals are then recombined and a power meter is used to measure the transmitted optical intensity as a function of the phase delay. The optical delay used here is such that the ASE will interfere minimally. The optical power at the output of the interferometer P_n can be calculated as in (4.12).

$$P_{\text{out}} = cP_{\text{in}} \left[1 + r - \gamma_s \cos(\Delta L.k_s) - r\gamma_n \cos(\Delta L.k_n) \right] \quad (4.12)$$

where P_s and P_n are the signal and noise powers respectively, c is a constant relating to the loss in the interferometer L is the optical path difference between the two interferometer arms γ_s and γ_n is the amplitude autocorrelation function of the signal and noise at the delay value L . L is sufficiently long so as to ensure that $\gamma_n = 0$. r is the ratio of the ASE power to the signal power. Ignoring γ_n , P_{out} becomes

$$P_{\text{out}} = cP_{\text{in}} \left[1 + r - \gamma_s \cos(\Delta L.k_s) \right] \quad (4.13)$$

With a phase shift of 2π in one of the arms the OSNR may be written

$$\frac{1}{\text{OSNR}} = r = \frac{\epsilon(1 - \gamma_s) - (1 + \gamma_s)}{1 - \epsilon} \quad (4.14)$$

where “ ϵ ” is the ratio of the maximum to minimum optical or the “extinction ratio” power output. Such a system has displayed immunity to impairments such as chromatic dispersion, polarisation mode dispersion and polarised noise. One major issue with this system is that the value of γ_s must be calibrated regularly. This involves switching off the noise and measuring the interference of the signal alone. The same γ_s value is then used when the signal is loaded with noise. The problems with this are twofold, namely that it is practically very difficult to switch off the noise in a communications channel and that γ_s is dependent on the properties of the modulator used to generate the signal. If the modulator is biased such that its extinction ratio is degraded then γ_s will change [87] and the OSNR measurement will accrue further error. It was found that a calibration factor for each modulator could be used to provide a more accurate γ_s but this complicates matters further and may be incompatible with reconfigurable networks. For the interferometric method to be useful, this issue must be addressed.

4.6 Summary

In this chapter, the importance of the OSNR was established and the conventional method for estimating the in-band OSNR was explained. It was seen that this is unsuitable in the context of transparent reconfigurable optical networks due to transmission history differences between channels. With this in mind a series of direct in-band OSNR monitoring schemes were examined. The polarisation nulling method was seen to be susceptible to the influence of polarised noise and polarisation mode dispersion. It also requires that the Poincaré sphere be scanned in order to complete a measurement. Orthogonal polarisation heterodyne mixing addresses the issue of PMD but the measurement is limited by the influence of polarised noise. Two-photon

absorption and stimulated Brillouin scattering are nonlinear effects which may be used for the in-band OSNR measurement but require high input powers in order to exploit the nonlinearities. A Mach-Zehnder interferometer may also be used, which addresses the issues of chromatic and polarisation mode dispersion but must be calibrated to the modulator used to generate the signal under test, which is unrealistic in a reconfigurable network.

Chapter 5 Harmonic Decomposition Method for Full characterisation of Ultrashort Optical pulses

In chapter 3, the use of schemes employing sinusoidal phase modulation as a means of achieving phase sensitive pulse characterisation as described in [88,89] was introduced. Such measurements were seen to be hampered by the requirement of a low index phase modulation in the case of [89] and a requirement for the specific alignment of the phase modulation to the pulse peaks [88]. It is possible to use a similar measurement scheme, also employing sinusoidal optical phase modulation to address these issues. As with the other sinusoidal phase modulation characterisation techniques, measurement of optical spectra at different phase modulation values forms the basis of this scheme, only in this case a harmonic decomposition is used in order to extract the pulse phase and intensity from the spectral measurements. The results obtained are compared to the sinusoidal phase modulation scheme of Kang and Dorrer and also with a commercially available SHG-FROG unit.

5.1 Measurement unit

The arrangement is shown in Figure 5.1

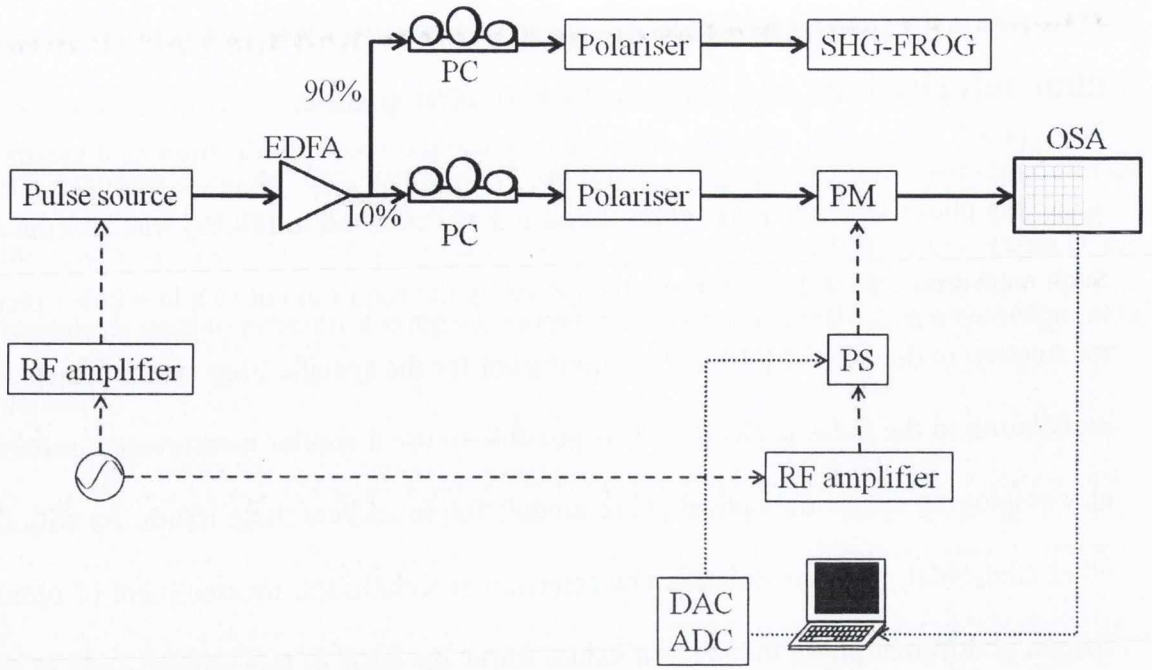


Figure 5.1 Setup for sinusoidal optical phase modulation scheme using harmonic decomposition. Solid arrows signify the path of the optical signal, dotted lines show the electrical connections, dashed lines show the path of the RF signal. PM: phase modulator, DAC/ADC: Data acquisition card, PS: RF phase shifter,

The pulse source under test passes through an optical phase modulator which is driven by a sinusoidal signal at the repetition rate of the pulse source. The delay between this modulation and the pulse peaks may be adjusted using an RF phase shifter which is controlled by the function generator capability of the DAC/ADC card. The modulated pulse train then passes into an optical spectrum analyser, from which the spectrum is recorded onto a computer.

5.2 Theory

The electric field of the pulse train may be expressed as (5.1)

$$E = \sum_k \sqrt{P_k} \exp(j\phi_k) \exp(j2\pi(kv + v_0)t) \quad (5.1)$$

where P_k is the spectral intensity of the pulse and ϕ_k is the corresponding spectral phase. v_0 is the carrier frequency of the signal and v is the repetition frequency of the pulse train. k refers to the order of the modes of the pulse, which exist at successive

separations of ν from ν_0 so the field of the most intense spectral line ($k=1$) is given as
(5.2)

$$E = \sqrt{P_0} \exp(j\phi_0) \exp(j2\pi(k\nu_0)t) \quad (5.2)$$

and when $k=1$

$$E = \sqrt{P_1} \exp(j\phi_1) \exp(j2\pi(\nu + \nu_0)t) \quad (5.3)$$

and so on until k reaches a number such that E becomes negligible. ϕ_0 and ϕ_1 are the phase when $k=0$ and $k=1$ respectively. Figure 5.2 illustrates this.

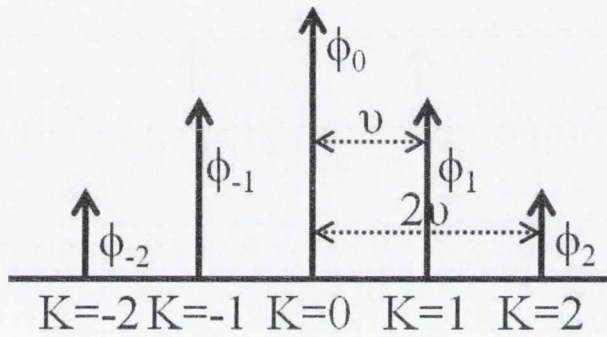


Figure 5.2 Schematic of the spectrum of a pulse with centre frequency ν_0 with lines at a spacing $k\nu$, where k is the order of the spectral line and ν is the pulse repetition frequency.

Measurement of the spectral intensity is quite straightforward although recovering the phase differentials between spectral lines k is the major issue. To this end, the pulse train passes through an optical phase modulator driven with a sinusoidal signal with frequency ν , the modulation function may be written as in equation (5.4)

$$\exp(jA \sin(2\pi\nu t + \psi)) \quad (5.4)$$

A is the amplitude of the phase modulation and ψ is the phase delay between the modulation signal and the pulse train. After the phase modulator, the electric field of the pulse train becomes

$$E_{\text{mod}} = E \exp(jA \sin(2\pi vt + \psi)) = E \sum_m J_m(A) \exp(jm\psi) \exp(j2\pi mvt) \quad (5.5)$$

J_m is an m^{th} order Bessel function. The effect of the phase modulation is to form a series of side bands from each spectral line with spacing equal to the frequency of the clock driving the phase modulator. This is illustrated in Figure 5.3 for the $k=0$ line

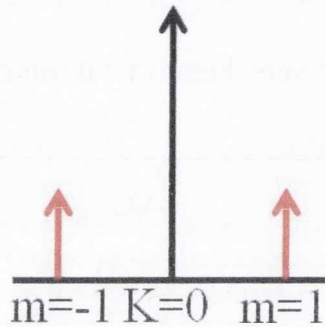


Figure 5.3, formation of sidebands as a result of the phase modulation.

Given that the frequency of the clock driving the phase modulator is the same as the pulse repetition rate, there is spectral overlap between the sidebands and the original modes of the pulse and as a result, interference occurs and the resultant lines' intensities are a function of the phase difference between the spectral lines of the original pulses. This interference of adjacent spectral components leads to a modified electric field as indicated in (5.6)

$$E = \sum_k \sqrt{Q_k} \exp(j\Phi_k) \exp(j2\pi(kv + v_0)t) \quad (5.6)$$

Φ_k is the resultant spectral phase of each line k . The modified spectral intensity, Q_k may then be written as (5.7)

$$Q_k = \left| \sum_m J_m(A) \sqrt{P_{k-m}} \exp(j(\phi_{k-m} + m\psi)) \right|^2 \quad (5.7)$$

Rearranging equation , the pulse spectral intensity becomes

$$Q_k = \sum_m J_m^2(A) P_{k-m} + 2 \sum_{m,n(m>n)} J_n(A) J_m(A) \sqrt{P_{k-n} P_{k-m}} \cos((\phi_{k-n} - \phi_{k-m}) - (m-n)\psi) \quad (5.8)$$

The first term on the right is the convolution of the pulse's spectral intensity with that of the phase modulation and contains the pulse spectral intensity. The second term contains the spectral phase differences namely $(\phi_{k-n} - \phi_{k-m})$, whose successive measurement over the pulse spectrum is used to reconstruct the spectral phase. The quantity ψ is varied by adjusting the delay between the phase modulation and the pulse train using the RF phase shifter. This causes the relative phase of the adjacent lines to change and as a result their intensity will be dependent on ψ as indicated in Figure 5.4

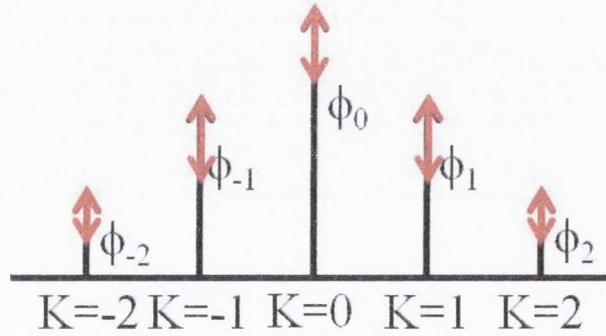


Figure 5.4 Schematic representing the spectral intensity variation as a result of modulating ψ .

At each selected value of ψ , the spectral intensity Q_k is measured, and by means of a harmonic decomposition as described by (5.9), the zeroth and first harmonics may be determined as in (5.10) and (5.11) respectively.

$$q_{k,s} = \frac{1}{2\pi} \int_0^{2\pi} Q_k(\psi) \exp(js\psi) d\psi \quad (5.9)$$

Where s refers to the order of the harmonic.

$$q_{k,0} = \sum_{m=-M}^M J_m^2(A) P_{k-m} \quad (5.10)$$

$$q_{k,1} = \sum_{m=-M}^M J_{m-1}(A) J_m(A) \sqrt{P_{k-m+1} P_{k-m}} \exp(j(\phi_{k-m+1} - \phi_{k-m})) \quad (5.11)$$

The order of the Bessel function was terminated at M so $J_M^2(A)$ would be negligible (this is judged as the point where $J_M^2(A)$ is 40dB lower than the peak of the sequence of $J_m^2(A)$). (5.10) and (5.11) may be considered as linear equations of the form $Y=CX$. In (5.10) C is a diagonal sparse matrix with diagonal elements consisting of $J_m^2(A)$ and X is the spectral intensity of the pulse, P_k . It can be seen that no phase information is provided by $q_{k,0}$. The dimensions of C are $K \times K$, where $k=1,2,3,\dots,K$, which is the number of spectral lines accounted for in the pulse spectrum after the phase modulation. In the case of (5.11), C is formed by $J_{m-1}(A)J_m(A)$ and X is formed by $\sqrt{(P_{k-m+1}P_{k-m})} \exp(j(\phi_{k-m+1} - \phi_{k-m}))$. The phase of X directly gives the spectral phase difference $\phi_{k+1} - \phi_k$ between each k and k+1, these can be concatenated to find the spectral phase ϕ_k . A consideration worth noting is the choice of k values. In one case, k can range from 1 to K-1, in which case m is from -M to M. This will be called case I. In the other case, k is from 2 to K, in which case m is from -M+1 to M+1, which will be known as case II. Both choices will yield identical results at the centre of the spectrum, where the intensity is high. However, in the presence of noise, there will be a small difference in results obtained at both ends of the spectrum. The effect of noise was minimised by taking results using case I for frequencies below the centre frequency of the pulse and case II above the centre frequency. When C is known for case I, it is possible to

compute C for case 2 by flipping the vector $q_{k,1}$, changing K to 2 and C to $-C$ and then X is found through matrix division and flipped once again in order to find C .

5.3 Simulation

The scheme was tested by way of a numerical simulation for three different types of pulses, all with a repetition frequency of 10GHz. These were (i) 2ps Gaussian pulses broadened by an accumulated dispersion of 5ps^2 with a duty cycle of 7% (ii) 2ps pulses with 35ps^2 of dispersion with a resultant duty cycle of 49% (iii) Pulses that were assumed to have been generated by driving one arm of an ideal MZ modulator biased at the null point, and driven by a 10GHz sine wave with modulation amplitude equal to π . The resultant pulses have a repetition frequency of 20 GHz with a duty cycle of 67%. Pulses (i) and (ii) have very broad spectra, with 81 spectral lines within 40 dB of the spectral peak. Pulse (iii) has a relatively narrow spectrum and just has 13 spectral lines within 40 dB of the peak spectral intensity. Non-correlated Gaussian noise with zero average and a variance of 0.01% of the peak of the unmodulated pulse spectral intensity was added to the phase modulated pulse spectra. 16 phase modulated spectra with delays spaced equally over the 100ps covered by the 10GHz sine wave were recorded. These were used to calculate the first and second harmonics using the Romberg integration method. The reconstruction error was calculated in the spectral domain from the root mean squared difference between the complex pulse field and that set in the simulation [90]. This was averaged over 40 simulation repetitions. The spectral intensity used in the pulse recovery is directly taken as the set values while the spectral phase was calculated from the retrieved spectra. By way of comparison, Kang and Dorrer's scheme in [89] was also tested. The aforementioned scheme required that only four phase modulated spectra be taken at four different phase modulation values separated by $\pi/2$. This scheme was seen to perform well for the 2ps pulse with low chirp. The

other pulses only returned satisfactory results when the modulation amplitude was limited to a small range. The scheme under investigation was seen to perform consistently over the three pulse types. The simulated RMS field error vs phase modulation amplitude is shown in Figure 5.5.

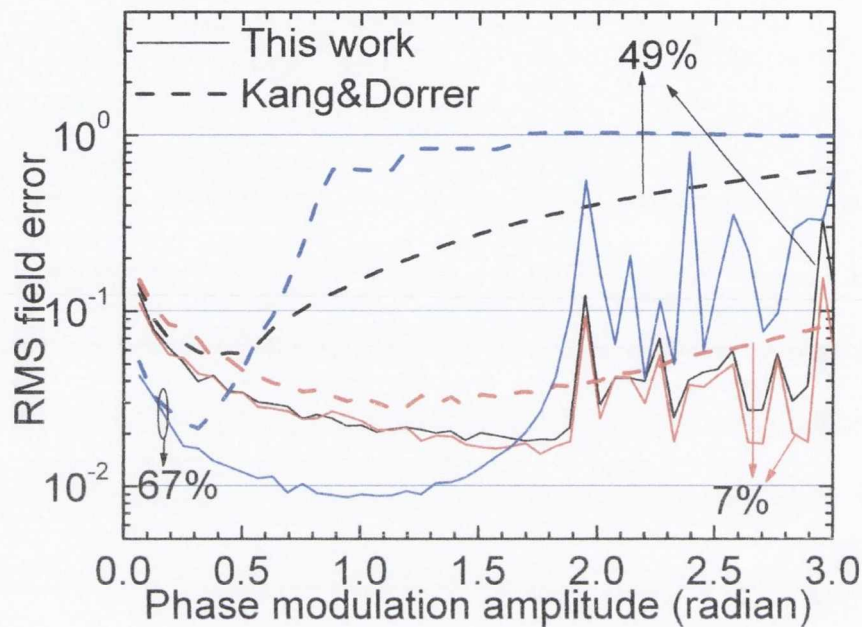


Figure 5.5. RMS field error comparison for the non-iterative method for the pulses with 7% duty cycle (red solid curve), 49% duty cycle (black solid curve) and 67% (blue solid curve) the corresponding dashed lines are for the same pulses measured using the sinusoidal phase modulation scheme

The RMS field error was obtained by calculating the RMS error between the set and retrieved fields. It can be seen that an RMS field error of <0.1 is attainable for phase modulation values of less than 1.9 radians for the simulated pulses. Spikes in the error curves occur when the phase modulation amplitude increases past 1.9 radians. This is because the matrix $J_{m-1}(A)J_m(A)$ may approach singularities when the modulation amplitude exceeds this value. Maintaining a phase modulation amplitude of below 1.9 radians assures that the measurement will be stable and only limited by noise.

5.4 Measurements

A tunable mode locked laser (u^{zt} TML1550) was used to generate 2ps pulses with a repetition frequency of 10GHz. The pulse train passed through a section of dispersion compensation fibre and was amplified using an EDFA. A variable optical amplifier (VOA) was used to control the optical power level. The pulse train was then passed through a polarisation controller and a polariser before being fed into Lithium Niobate phase modulator which was modulated with voltage such that the maximum phase shift was 1.32 radians. The phase modulator was driven by the same 10GHz clock as that which was applied to the TML1550. A voltage controlled RF phase shifter was used to change the relative delay between the phase modulation and the pulse train. Following the phase modulator, an optical spectrum analyser (OSA) with a resolution bandwidth of 0.02nm was used to acquire the pulse spectra which were recorded on the PC. The relative phase between the pulse train and phase modulation was shifted through 16 uniform steps over a span of 2π . For each of these steps the spectrum was recorded. The pulse spectrum was also recorded in the absence of the phase modulation to obtain the spectral intensity. The spectral phase was then determined using equations (5.10) and (5.11) and Fourier transformed into the time domain. A commercially available SHG-FROG [91] was also used to characterise the pulses. The results obtained from the scheme and the comparison to the SHG-FROG method are shown in Figure 5.6.

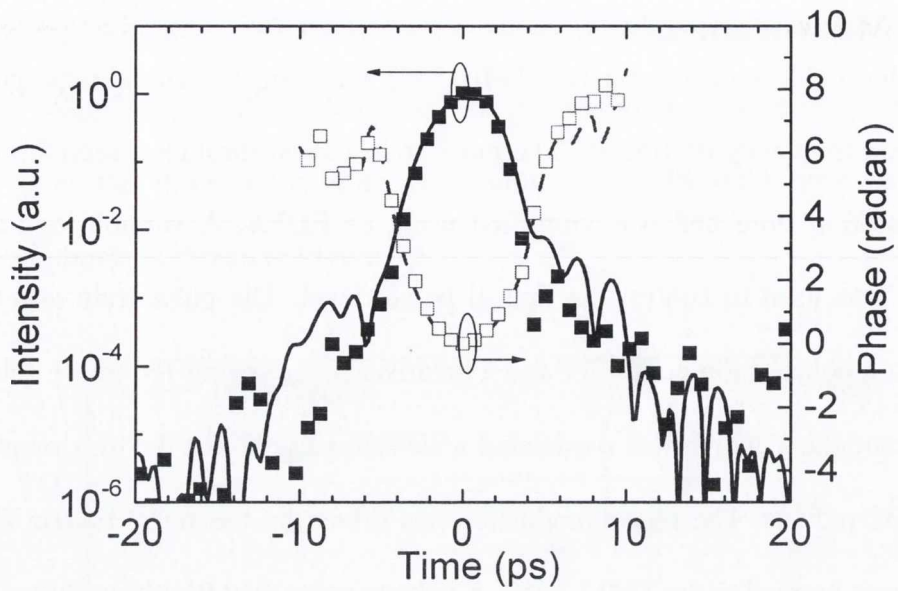


Figure 5.6 Recovered temporal intensity and phase using this method. The red curve is the recovered phase using this method compared to that recovered from a SHG-FROG measurement (red squares). The black line is the temporal intensity recovered using this scheme compared to FROG (black squares).

It can be seen that a good agreement with the FROG technique was achieved for the characterisation of 2ps pulses. This is encouraging as it is possible to obtain very similar results with much lower characterisation times and input power levels.

5.5 Summary

The noniterative method for complete pulse characterisation was introduced in this chapter based on a harmonic decomposition. A theoretical comparison was made between this scheme and Kang and Dorrer's sinusoidal phase modulation scheme for pulse recovery. The harmonic decomposition method was seen to be applicable to a greater range of pulse durations and phase modulation amplitude. Also, the use of a large phase modulation amplitude caused a larger variation in the spectral intensity as the delay between the phase modulation and the pulse train was adjusted compared to the aforementioned methods. This would lead to a more favourable signal to noise ratio in the acquired spectra. Experiments confirmed that a good agreement with the SHG-FROG method could be obtained for 2ps pulses at a repetition rate of 10G. A similar

scheme as mentioned in chapter 3 and [92] using a mach Zehnder modulator may also be used but requires the resolution of spectra at half the repetition rate of the pulse.

Chapter 6 Full Pulse Field Recovery using Dynamic Phase Modulation

It was seen in chapter 5 that a harmonic decomposition could be used to recover the intensity and phase information of an optical pulse when a sinusoidal phase modulation was used in conjunction with several measurements of the pulse spectrum. In this case, the spectra were measured as the phase modulation delay relative to the pulses was stepped through 16 values over total phase delay of 2π . This was seen to compare favourably with a standard SHG-FROG measurement. It is possible to implement another scheme based on the same harmonic decomposition method using a dynamic modulation of the RF phase shifter at a frequency f which modulates the delay between the sinusoidal phase modulation and the pulse train. In this case it is only necessary to record one spectrum to extract the harmonics and hence the spectral phase and intensity.

6.1 Measurement scheme

The dynamic phase modulation scheme is shown in figure 6.1.

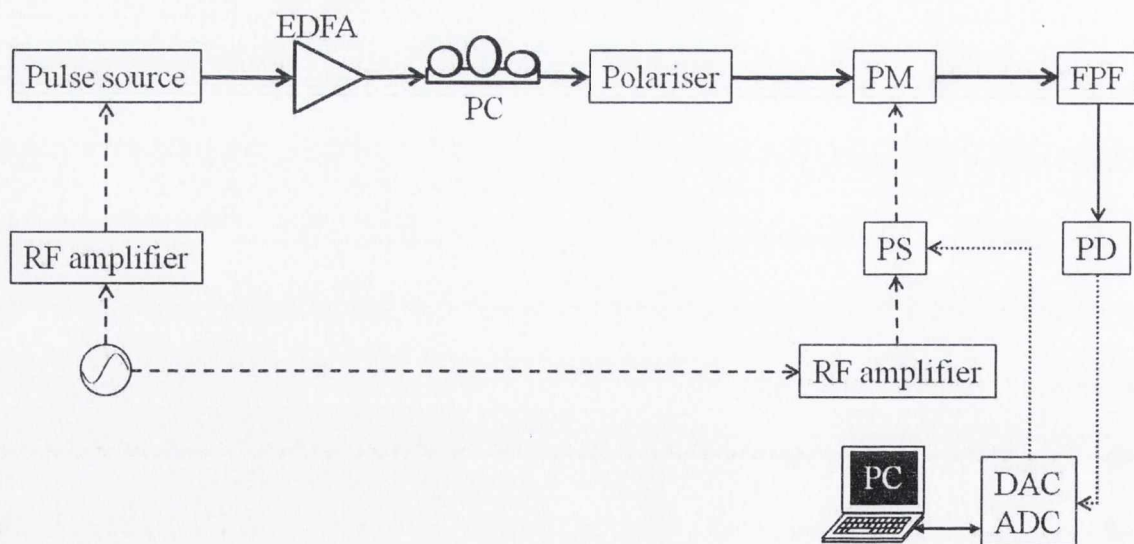


Figure 6.1 Measurement setup for dynamic phase modulation scheme. PC: polarisation controller, PM: phase modulator, FPF: Fabry-Perot filter, PS: phase shifter, PD: photodiode, DAC/ADC: digital to analog/analog to digital converter. Solid arrows signify the optical path, dashed arrows signify RF cables, dotted arrows show the electrical connections.

The key difference between this and the setup outlined in chapter 5 is the use of a voltage controlled Fabry-Perot filter in conjunction with a photodiode instead of an

optical spectrum analyser and continuously modulating the phase delay, ψ instead of stepping it through 16 discrete values. The bandwidth of the filter is sufficiently narrow as to resolve the spectral lines of the pulse, whose spacing will be equal to the pulse repetition frequency. The pulse source under test is passed through an optical phase modulator. As was discussed in chapter 5, the electric field of the pulse train prior to the phase modulator is given in (6.1)

$$E = \sum_k \sqrt{P_k} \exp(j\phi_k) \exp(j2\pi(kv + v_0)t) \quad (6.1)$$

P_k is the spectral intensity of the pulse and ϕ_k is the corresponding spectral phase. v_0 is the carrier frequency of the signal and v is the repetition frequency of the pulse train. k refers to the order of the modes of the pulse, which exist at successive separations of v from v_0 . Again, a sinusoidal phase modulation given by (6.2).

$$\exp(jA \sin(2\pi vt + \psi)) \quad (6.2)$$

is enforced on the pulse train, where A is the amplitude of the phase modulation and ψ is the delay between the pulse train and the phase modulation. Following the phase modulator the field is

$$E_{\text{mod}} = E \exp(jA \sin(2\pi vt + \psi)) = E \sum_m J_m(A) \exp(jm\psi) \exp(j2\pi mvt) \quad (6.3)$$

where J_m is the m th order Bessel function. Again, like the scheme in chapter 5, a series of sidebands are generated as a result of the phase modulation with their spacing equal to the spacing of the spectral lines k . Again, each of the lines interferes with the existing spectral lines. Q_k the modified spectral intensity is given as (6.4).

$$Q_k = \left| \sum_m J_m(A) \sqrt{P_{k-m}} \exp(j(\phi_{k-m} + m\psi)) \right|^2 \quad (6.4)$$

which may be rearranged to

$$Q_k = \sum_m J_m^2(A) P_{k-m} + 2 \sum_{m,n(m>n)} J_n(A) J_m(A) \sqrt{P_{k-n} P_{k-m}} \cos((\phi_{k-n} - \phi_{k-m}) - (m-n)\psi) \quad (6.5)$$

The first term in is the convolution of the pulse spectral intensity with the phase modulation which contains information on the spectrum only. The second term contains information on the phase differences between each of the spectral lines.

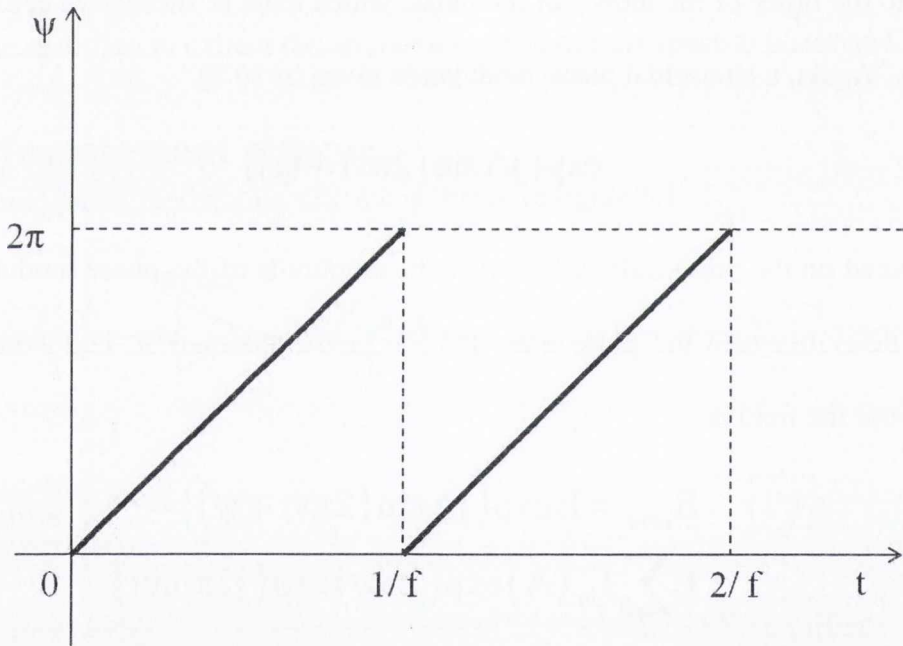


Figure 6.2 Modulation applied to the RF phase shifter by the function generator.

It is noted that in this case, ψ is modulated as represented by Figure 6.2. In contrast with the scheme introduced in chapter 5, it is shifted linearly with time from zero to 2π at a frequency f . The the spectral intensity is then modified to

$$Q_k = \sum_m J_m^2(A) P_{k-m} + 2 \sum_{m,n(m>n)} J_n(A) J_m(A) \sqrt{P_{k-n} P_{k-m}} \cos((\phi_{k-n} - \phi_{k-m}) - 2\pi(m-n)ft) \quad (6.6)$$

where f is the modulation frequency of the phase delay ψ .

Using the following harmonic decomposition, which differs slightly from that in chapter 5 the zeroth and first harmonics may be extracted, i.e. $s=0, s=1$.

$$q_{k,s} = f \int_0^{f-1} Q_k(t) \exp(j2\pi sft) dt \quad (6.7)$$

f is equal to the frequency at which the phase modulation is shifted. Because the exponential is equal to 1 in the case of $q_{k,0}$, the zeroth harmonic has no phase term and is given in equation (6.8)

$$q_{k,0} = \sum_{m=-M}^M J_m^2(A) P_{k-m} \quad (6.8)$$

The first harmonic is shown in equation (6.9)

$$q_{k,1} = \sum_{m=-M}^M J_{m-1}(A) J_m(A) \sqrt{P_{k-m+1} P_{k-m}} \exp(j(\phi_{k-m+1} - \phi_{k-m})) \quad (6.9)$$

Once the zeroth and first harmonics are retrieved, the recovery of the pulse spectrum and phase proceeds in the manner indicated in chapter 5.

6.1.1 Recovery of $q_{k,0}$ and $q_{k,1}$

The frequency response of the Fabry-Perot filter, $F(v)$ has a Lorentzian shape as described by (6.10).

$$f(v) = \frac{v_{FWHM}^2}{4v^2 + v_{FWHM}^2} \quad (6.10)$$

Where ν_{FWHM} is the 3dB bandwidth of the filter. The pulse spectrum after the filter is a convolution of the filter function $F(\nu)$ with the spectrum before the filter. This is described in (6.11).

$$\left(\sum_k Q_k(t) \delta(\nu - \nu_k) \right) \otimes f(\nu) \quad (6.11)$$

where each of the quantities mentioned are the same as in chapter 5. The filter may be tuned using an applied triangle wave voltage of frequency f' as shown in Figure 6.3.

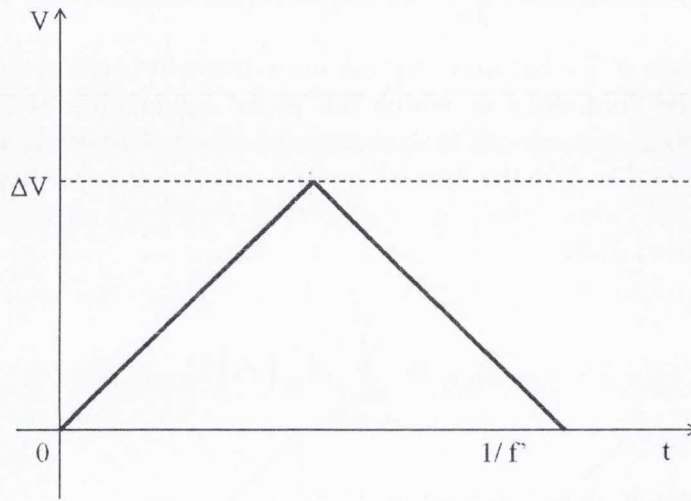


Figure 6.3 Tuning of the Fabry-Perot filter at a frequency f'

The maximum applied voltage ΔV is such that the filter may be tuned over the entire pulse spectrum. The output frequency of the filter is given as $\nu = \beta t$, where β is the tuning speed of the triangle wave, given by $\beta = \Delta V f'$. Following the filter, the photodiode detects a time dependent voltage as shown in equation (6.12)

$$V(t) \propto \left(\sum_k Q_k(t) \delta(\beta t - \beta t_k) \right) \otimes F(\beta t) \quad (6.12)$$

This signal is then filtered to remove frequency components of $V(t)$ greater than $2f$ so only the zeroth and first harmonics should remain. After the filter the voltage is described as follows

$$\begin{aligned}
V(t) \propto & \left(\sum_k q_{k,0} \delta(\beta t - \beta t_k) \right) \otimes F(\beta t) \\
& + \left[\left(\sum_k q_{k,1} \delta(\beta t - \beta t_k) \right) \otimes F(\beta t) \right] \exp(-j2\pi f t) + c.c
\end{aligned}
\tag{6.13}$$

Where $q_{k,0}$ and $q_{k,1}$ are the zeroth and first harmonics respectively.

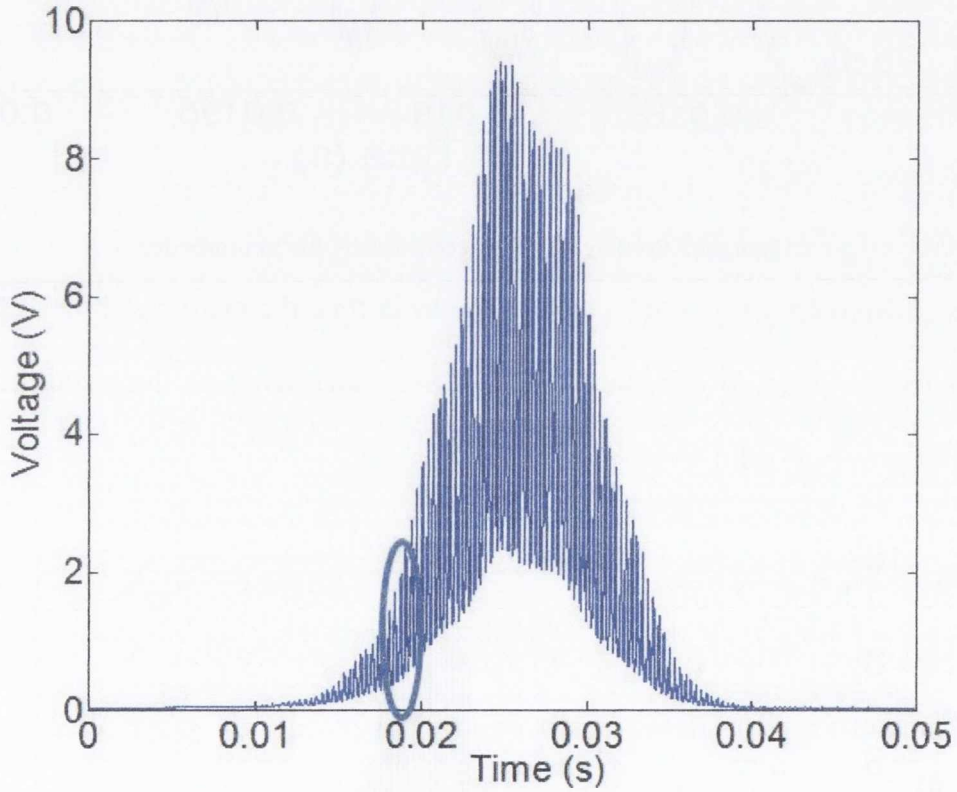


Figure 6.4. Recorded voltage from the photodiode as a function of time.

Figure 6.4 shows an example of $V(t)$ recorded from the photodiode for a 2ps pulse from a modelocked laser. Taking the section encircled by the blue oval in Figure 6.5 allows observation of the structure of $V(t)$ for such a pulse.

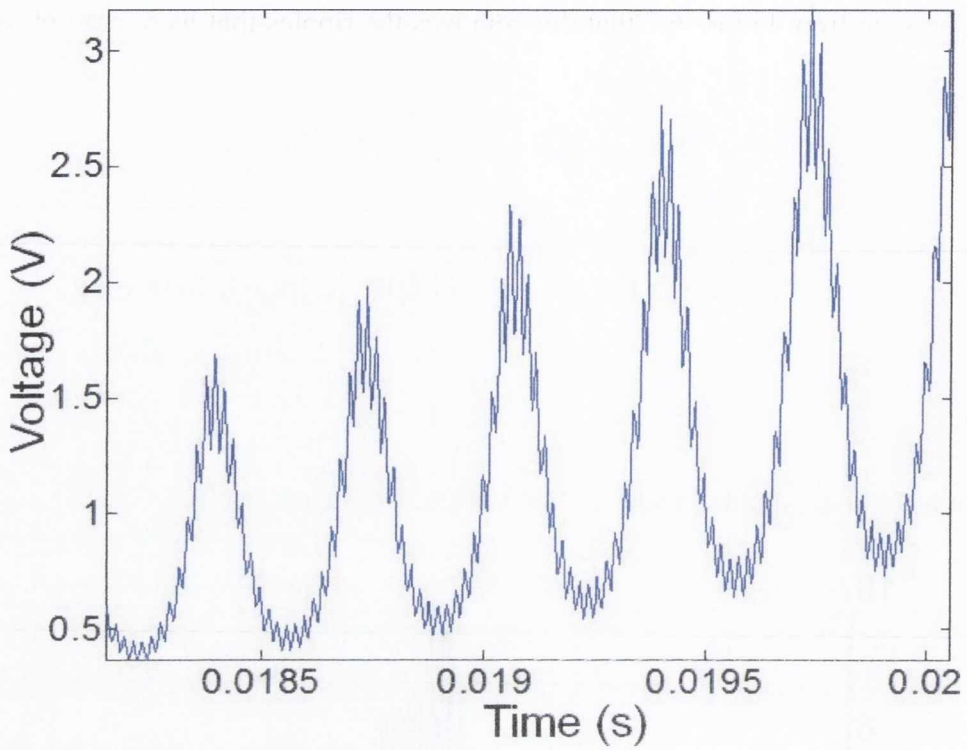


Figure 6.5 Section of $V(t)$ as recorded by the photodiode.

The modulation apparent on the peaks is due to shifting the phase, ψ of the 10G clock. This signal is subject to a digital low pass filter which retrieves the zeroth harmonic. Figure 6.6 shows $V(t)$ following the digital filter

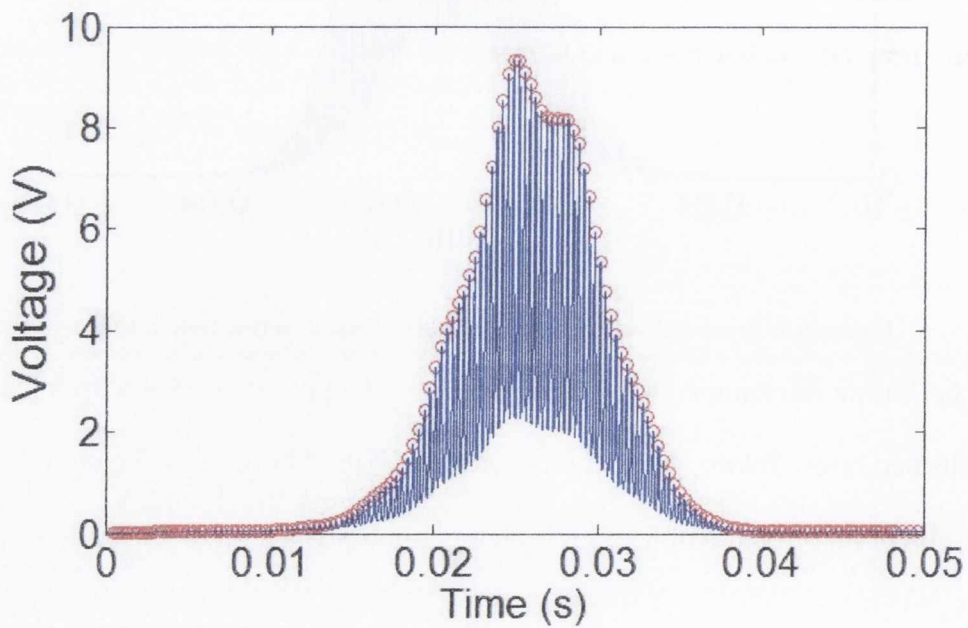


Figure 6.6 $V(t)$ following lowpass digital filtering in order to elucidate the zeroth harmonic.

It can be seen from Figure 6.7 that this removes the ripples that were present on Figure 6.5.

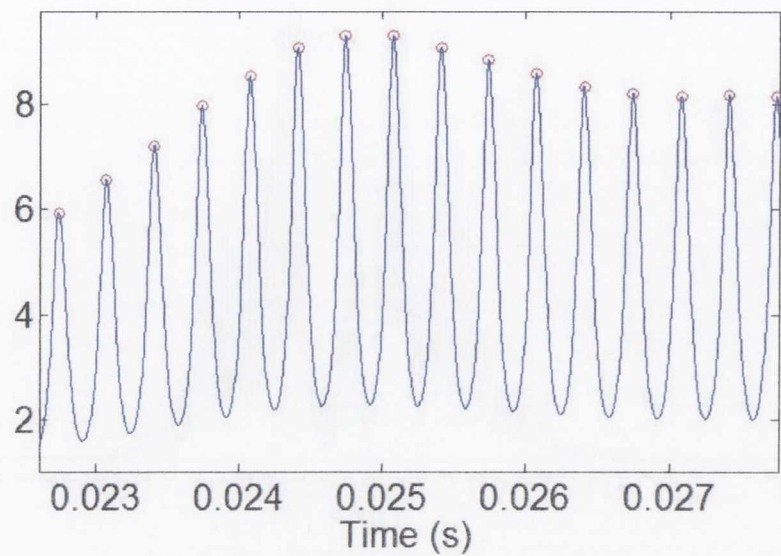


Figure 6.7 Filtered zeroth harmonic of $V(t)$

The pulse spectral intensity may be found from this as shown in equation (6.8) but practically it may be measured directly using a spectrum analyser. Retrieval of the first harmonic may be achieved by multiplying $V(t)$ by a phase function $\exp(j\pi f t)$ where f is the frequency at which the phase shifter is being driven. The recovered first harmonic from $V(t)$ is shown in Figure 6.8

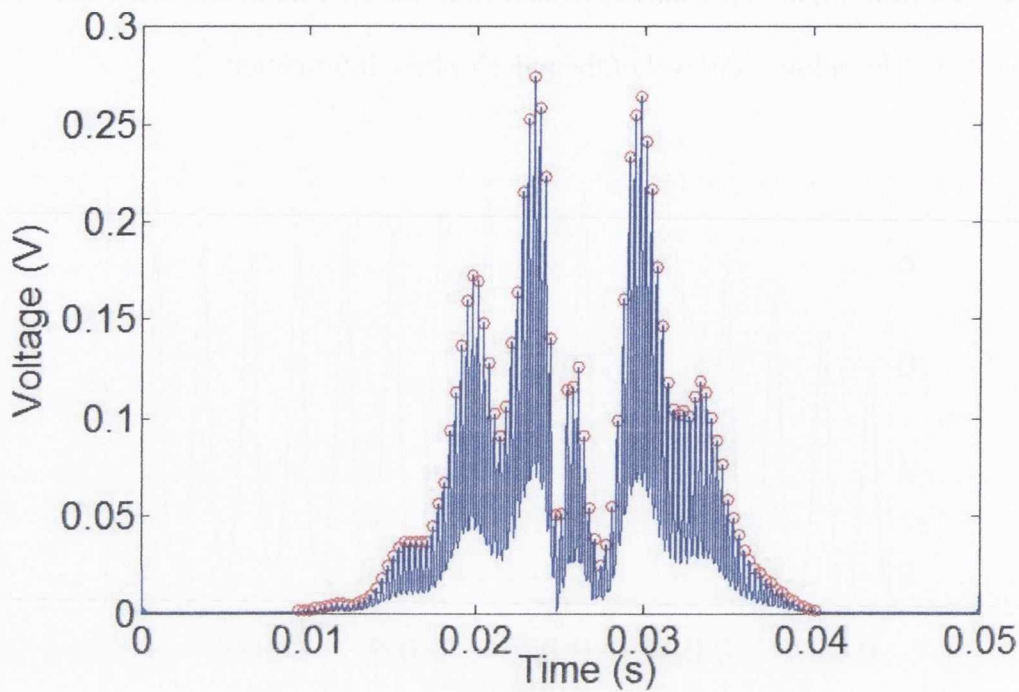


Figure 6.8. Retrieved first harmonic of $V(t)$ following multiplication by a phase function.

The magnitude of the first harmonic is much lower than that of the zeroth. The structure of a section of the first harmonic is shown in Figure 6.9

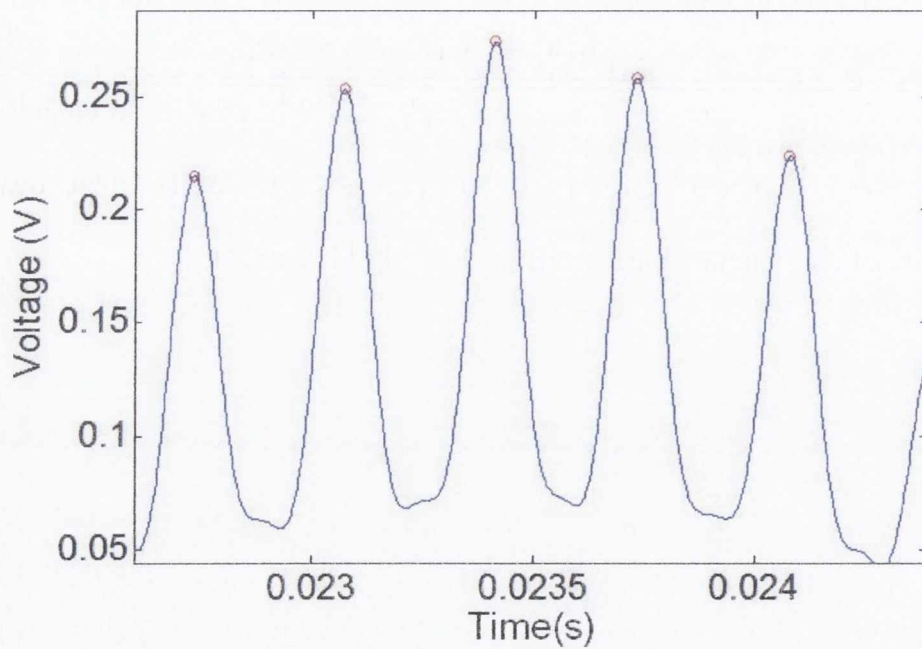


Figure 6.9 Structure of the first harmonic of $V(t)$

The pulse spectral group delay may be found from the first harmonic using the linear equation (6.9). This allows retrieval of the pulse's phase information.

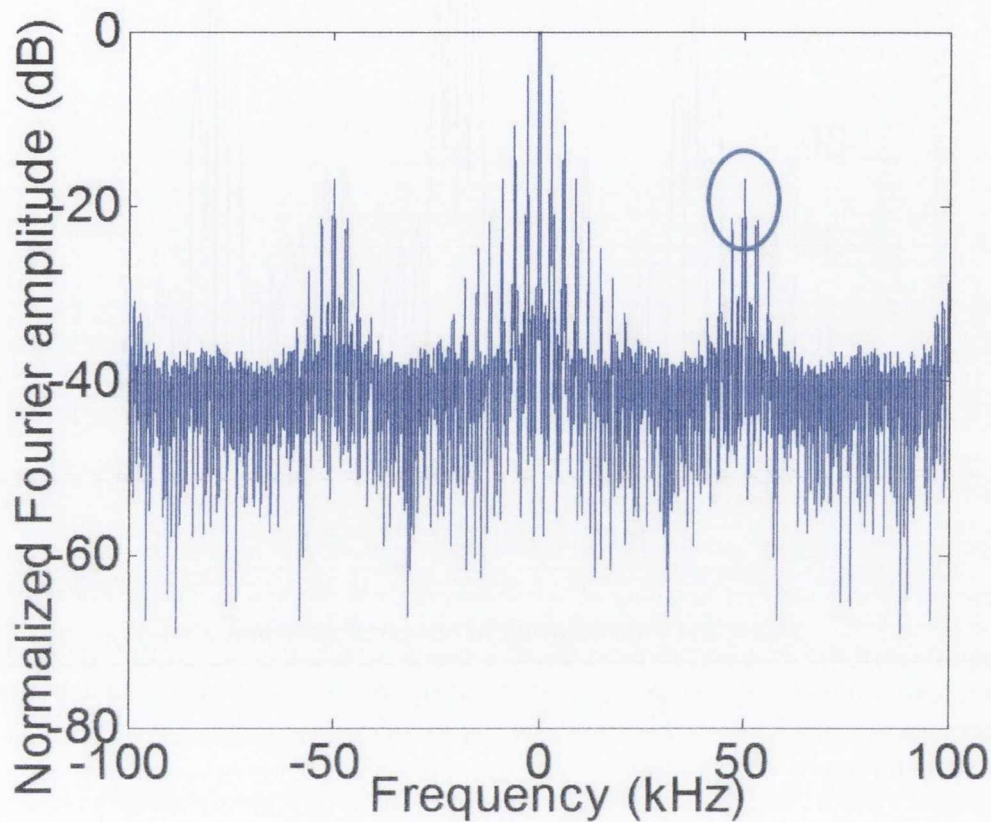


Figure 6.10 Spectrum of $V(t)$ when ψ is modulated at 50kHz

Figure 6.10 shows the result of a Fourier transform on $V(t)$, which is the modified pulse spectrum, where the zeroth and first harmonics are clearly visible. The blue oval encircles the peaks of the first harmonic, which is shown in Figure 6.11.

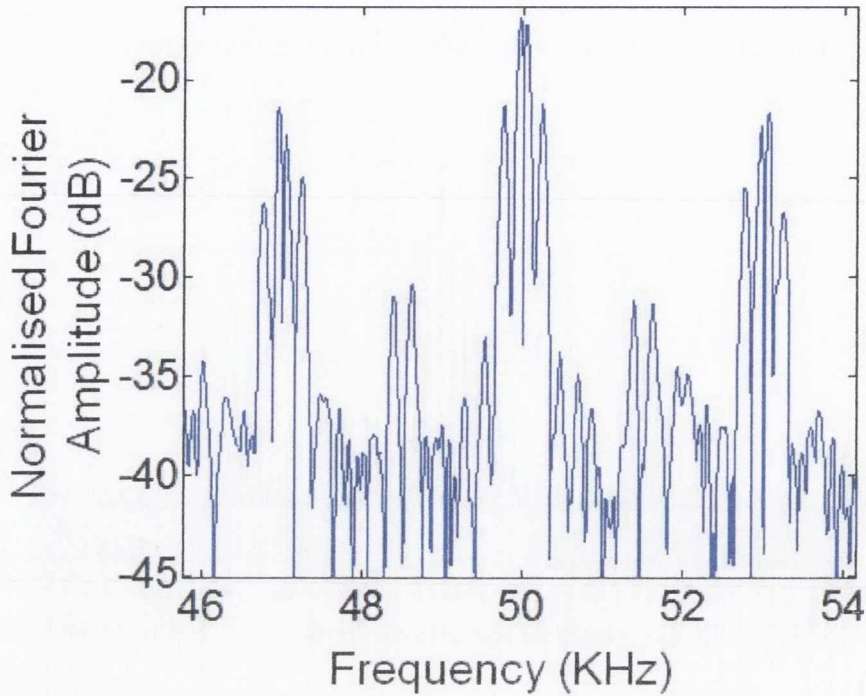


Figure 6.11 First harmonic of recovered spectrum.

Given the proximity to the noise floor of the detector, retrieval of the first harmonic may be challenging.

6.2 Simulation of scheme

A simulation of the scheme was performed using 2ps Gaussian pulses at a repetition frequency of 10GHz with 5ps^2 of dispersion added. The phase of the 10GHz clock phase was modulated at 100kHz using a sawtooth function. The Fabry Perot filter was assumed to have a 3dB bandwidth of 2GHz and was scanned at 12.5Hz. The analog to digital conversion of the photodiode voltage was sampled at 400kHz. As a result of this, four points were recorded for each period of the phase modulation. A low-pass filter was implemented after the cut-off point with a frequency of $1.6f$. A digital 8th order Chebyshev low-pass filter was implemented on the computer with a cut off point of $0.5f$. The phase modulation amplitude, A was 1 radian. A 40dB signal to noise ratio is assumed for the photodiode detection. $V(t)$ retrieved from the simulation is shown in Figure 6.12

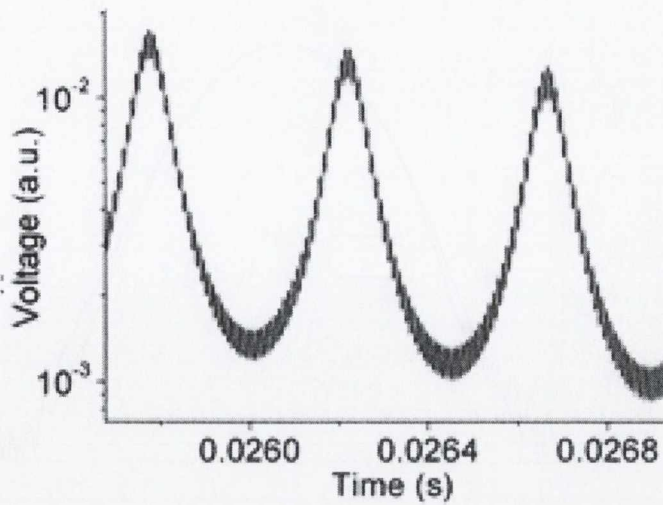


Figure 6.12 Spectrum retrieved from simulation.

The ripples on top of the curves are as a result of modulating ψ with a frequency F . The recovered spectrum is the Fourier transform of Figure 6.12 and is shown in Figure 6.13.

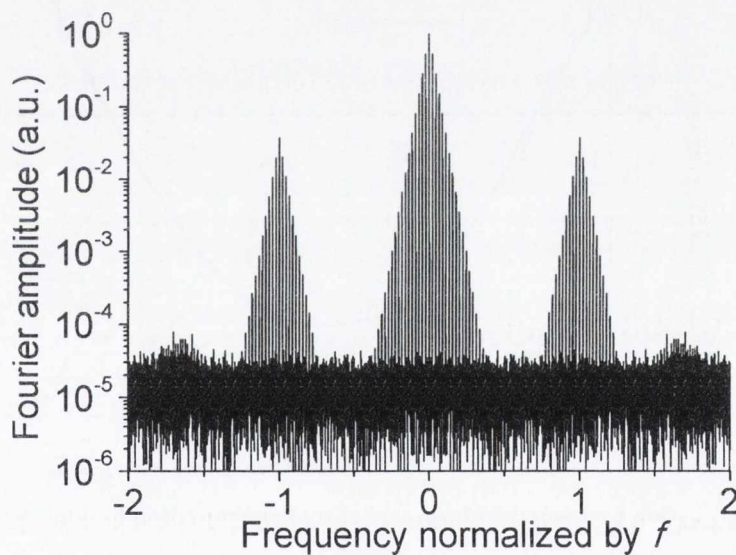


Figure 6.13 The recorded zeroth and first harmonics. This is the Fourier transform of $V(t)$

The zeroth and first harmonics are clearly visible with the first harmonic approximately 20dB lower in intensity than the zeroth. The retrieved temporal intensity is shown in Figure 6.14 while the retrieved temporal phase is shown in Figure 6.15

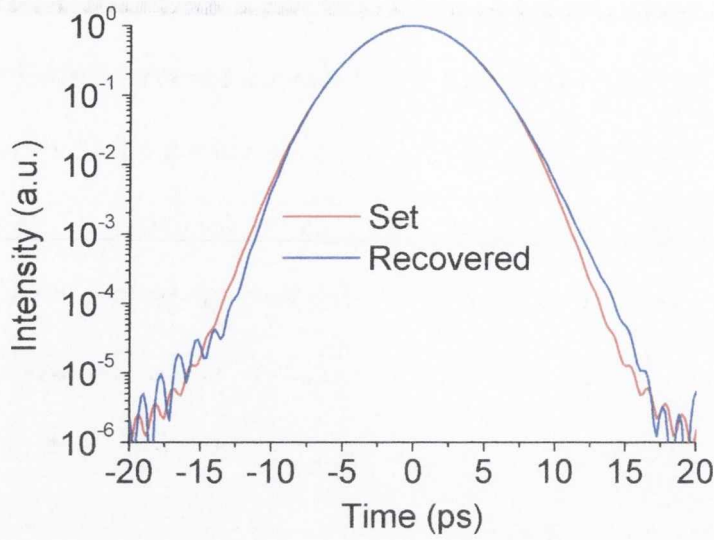


Figure 6.14 Comparison between the set temporal intensity and that recovered using the simulation.

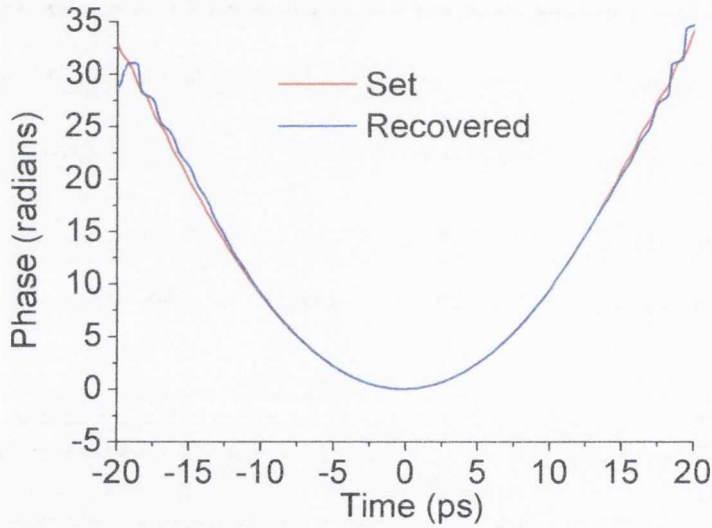


Figure 6.15 Comparison between the phase set before the simulation to that recovered using the scheme.

A good agreement can be seen between the set intensity and phase and that retrieved from the simulation of the scheme.

6.3 Measurement

The pulse source under test was a tunable modelocked laser (u²t TMLL) driven by an amplified RF clock which produced 2ps pulses at with a repetition rate of 10GHz. Two different lengths of dispersion compensation fibre were also used to impose a chirp on the pulse train. The clock signal was also passed through an RF phase shifter and

amplified before being applied to the optical phase modulator. A sawtooth function generated by the data acquisition card (DAC) was used to drive the RF phase shifter at a frequency, f , of 20kHz. The amplitude of this signal was such that a phase change between 0 and 2π was imposed on the 10G signal. The optical pulse train then passed through the phase modulator before being filtered by the tunable Fabry-Perot filter. The filter bandwidth was 3GHz and was tuned at 2Hz by the triangle wave shown in Figure 6.3. Data was only recorded on the rising edge of the triangle wave. The photodiode voltage is then sampled by the data acquisition card (ADC) at a rate of 200k samples per second. For retrieving the zeroth harmonic, a digital low pass filter was used, which had a cut-off of 8kHz. For the first harmonic the low pass filter cut-off was 2.5kHz. A comparison was made with the harmonic decomposition method shown in chapter 5 in order to test the validity of measurements (as this scheme was shown to compare favourably with a SHG-FROG measurement). Also, the scheme was compared to a version Kang and Dorrer's sinusoidal phase modulation method.

6.4 Results

Results obtained using this scheme are compared with the method outlined in chapter 5 and Kang and Dorrer's sinusoidal phase modulation scheme as shown in chapter 3. The recovered spectral intensity and phase are plotted as are the temporal intensity and phase.

6.4.1 2ps pulses with very low chirp

Figure 6.16-Figure 6.19 show the recovered spectral intensity and phase, temporal intensity and phase respectively. The green points are data retrieved using the sinusoidal phase modulation scheme in [89], the blue points are from the static phase modulation scheme shown in chapter 5 and the red points are from the method outlined in this chapter.

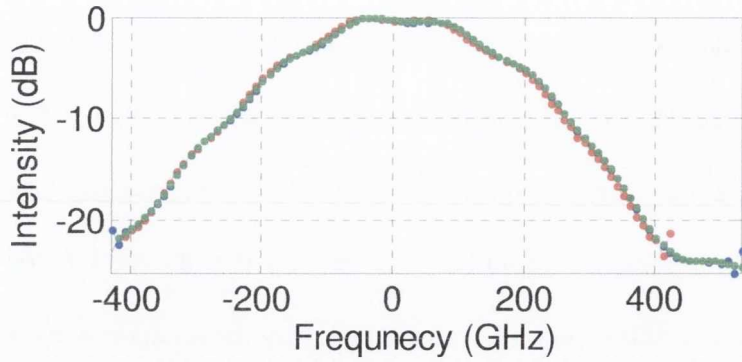


Figure 6.16 Spectral intensity of 2ps pulses retrieved using the three methods. Blue: static Harmonic deconvolution. Red: dynamic scheme. Green, Kang and Dorrer.

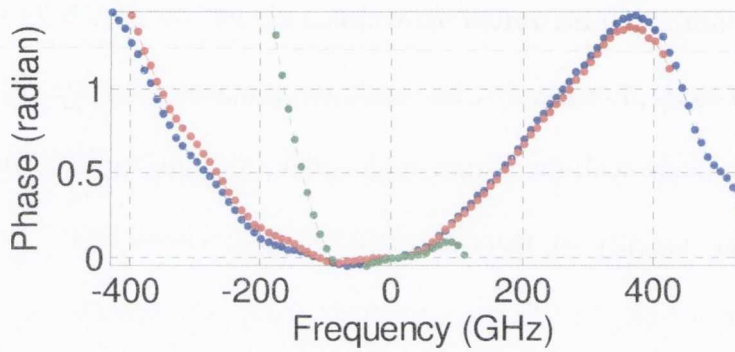


Figure 6.17 Spectral phase of 2ps pulses retrieved using the three methods. Blue: static Harmonic deconvolution. Red: dynamic scheme. Green, Kang and Dorrer.

Spectral phase retrieval using Kang and Dorrer's scheme appears to be problematic for 2ps pulses due to signal to noise issues.

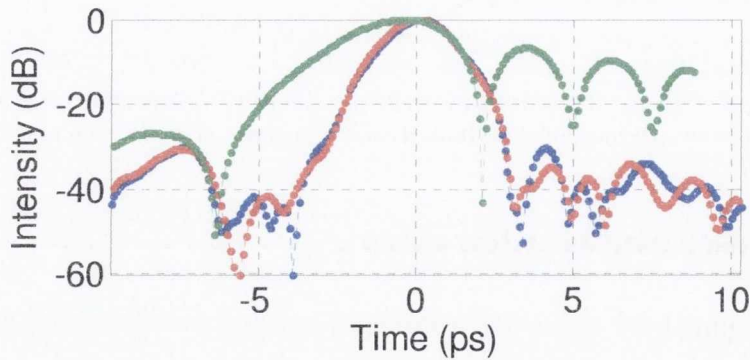


Figure 6.18. Temporal intensity of 2ps pulses retrieved using the three methods. Blue: static Harmonic decomposition. Red: dynamic scheme. Green, Kang and Dorrer.

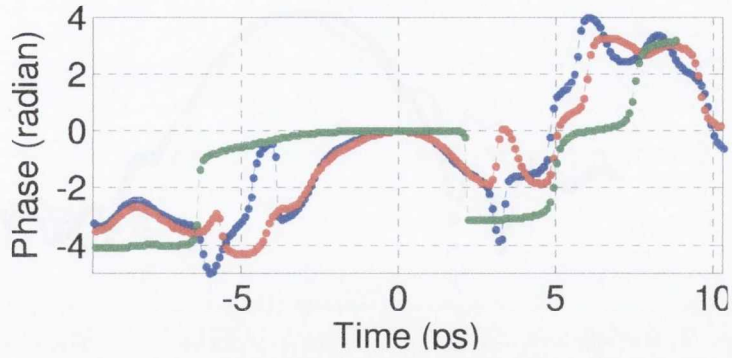


Figure 6.19 Temporal phase of 2ps pulses retrieved using the three methods. Blue: static Harmonic decomposition. Red: dynamic scheme. Green, Kang and Dorrer.

6.4.2 2ps pulses with added dispersion

The pulses from the u^2t TMML were passed through a small section of dispersion compensation fibre in order to cause temporal broadening. Figure 6.20-Figure 6.23 show the retrieved intensity and phase from this situation

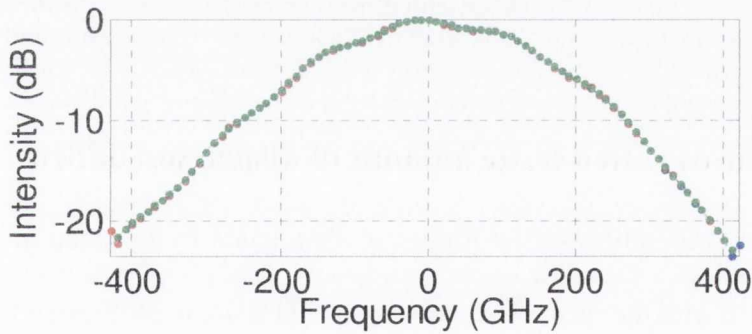


Figure 6.20 Spectral intensity of 2ps pulses with dispersion retrieved using the three methods. Blue: static Harmonic decomposition. Red: dynamic scheme. Green, Kang and Dorrer.

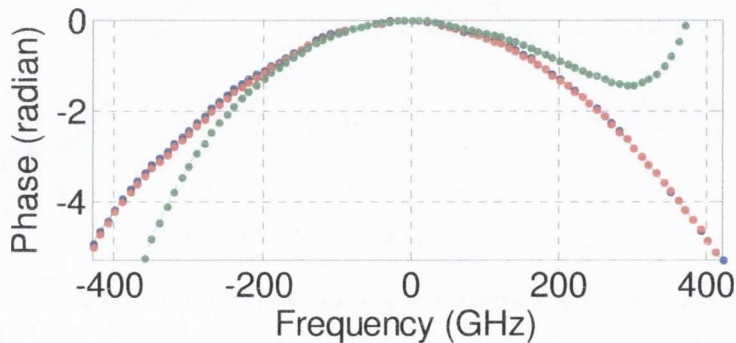


Figure 6.21 Spectral phase of 2ps pulses with dispersion retrieved using the three methods. Blue: static Harmonic decomposition. Red: dynamic scheme. Green, Kang and Dorrer.

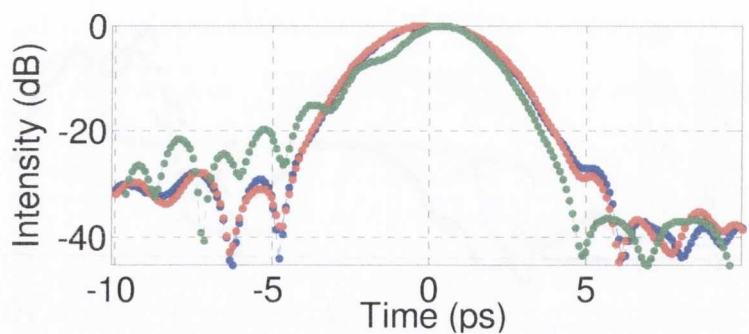


Figure 6.22 temporal intensity of 2ps pulses with dispersion retrieved using the three methods. Blue: static Harmonic decomposition. Red: dynamic scheme. Green, Kang and Dorrer.

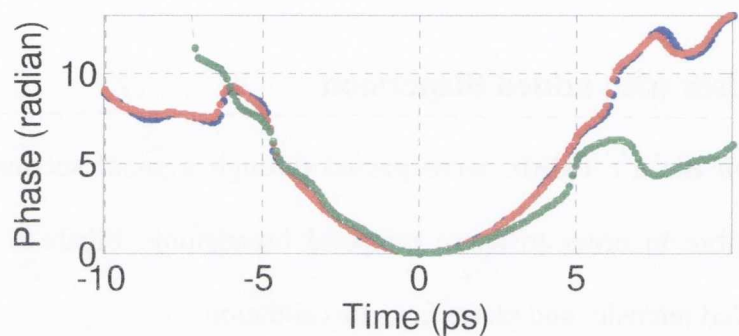


Figure 6.23 temporal phase of 2ps pulses with dispersion retrieved using the three methods. Blue: static Harmonic decomposition. Red: dynamic scheme. Green, Kang and Dorrer.

6.4.3 2ps pulses with a large amount of dispersion added

A longer dispersion compensation fibre was then added to the setup to further broaden the pulse spectra with the recovered data shown in Figure 6.24 through Figure 6.27

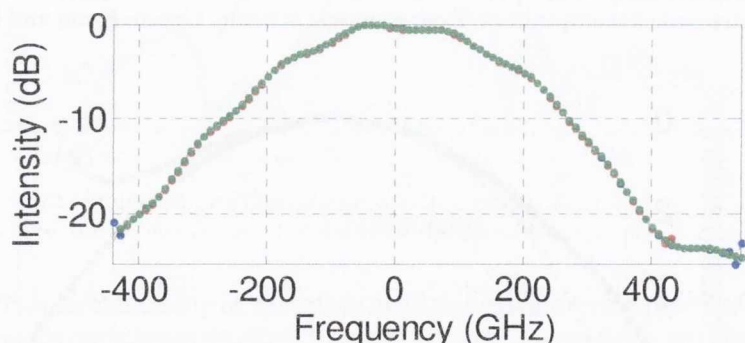


Figure 6.24 Spectral intensity of 2ps pulses with dispersion retrieved using the three methods. Blue: static Harmonic decomposition. Red: dynamic scheme. Green, Kang and Dorrer.

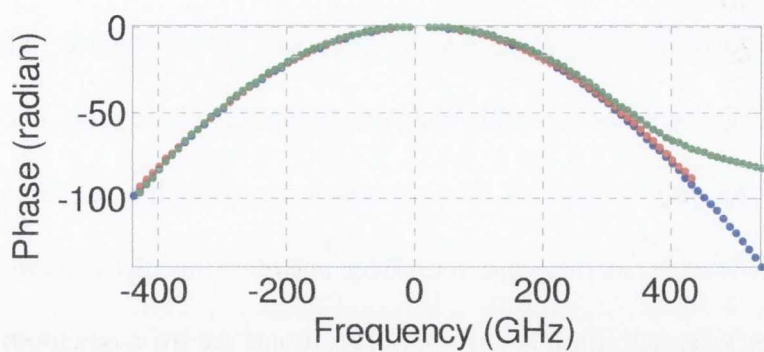


Figure 6.25 Spectral phase of 2ps pulses with dispersion retrieved using the three methods. Blue: static Harmonic decomposition. Red: dynamic scheme. Green, Kang and Dorrer.

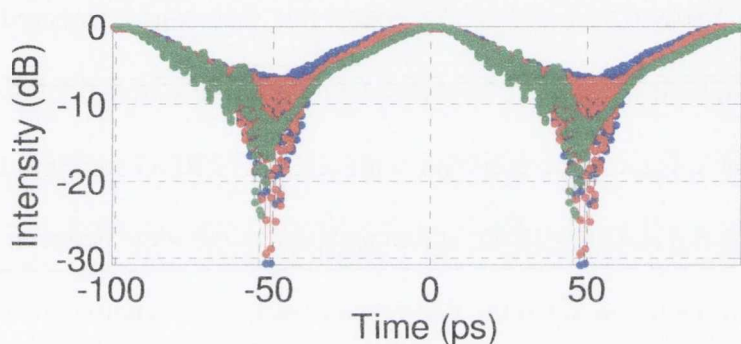


Figure 6.26 temporal intensity of 2ps pulses with dispersion retrieved using the three methods. Blue: static Harmonic decomposition. Red: dynamic scheme. Green, Kang and Dorrer.

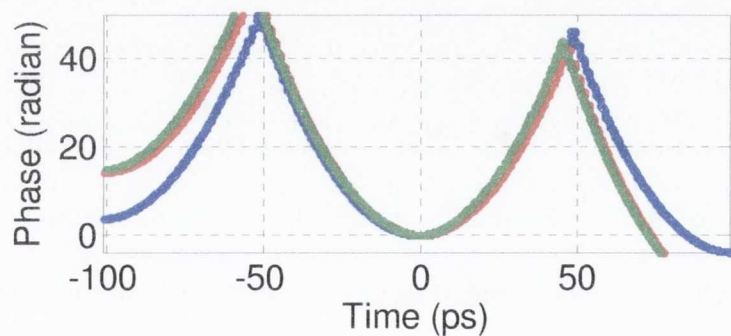


Figure 6.27 temporal phase of 2ps pulses with dispersion retrieved using the three methods. Blue: static Harmonic decomposition. Red: dynamic scheme. Green, Kang and Dorrer.

Increased agreement with Kang and Dorrer's scheme can be seen when the pulse is significantly broadened.

6.5 Summary

The dynamic phase modulation method for recovery of the electric field of ultrashort optical pulses was investigated in this chapter. An optical phase modulator driven at the same speed as the pulse repetition frequency was used, much like the method outlined in chapter 5. However, in this case the delay between the phase modulation and the pulse train was adjusted continuously between 0 and 2π by a sawtooth type wave at 20kHz, which caused a modified pulse spectrum from which the zeroth and first harmonics were extracted. The spectral and temporal phase and intensity could then be recovered using these harmonics. The recovered data was compared to Kang and Dorrer's sinusoidal phase modulation scheme and with the scheme outlined in chapter 5 (which itself was subject to comparison with an SHG-FROG measurement of similar ultrashort optical pulses). Excellent agreement could be seen between the static and dynamic phase modulation scheme. The advantage of the dynamic scheme lies in the fact that it only requires the measurement of one pulse spectrum, while the static scheme requires measurement of 16. The four spectrum technique of Kang and Dorrer seemed to be more susceptible to signal to noise issues (with the setup used in these experiments anyway) to the extent that 16 spectra had to be taken to allow the acquisition of meaningful results. The recovered spectral intensities unsurprisingly compared well to those extracted from the two harmonic decomposition schemes, however, the recovered phase and temporal intensity was seen to be quite poor in comparison and only appeared to yield reasonable results with broadened pulses. A possible reason for this is that the low phase modulation amplitude, which was needed to ensure the modulation remained sinusoidal, made it difficult to discern changes in the phase modulated spectra due to signal to noise issues. The resolution of the detection may also have been a factor in this regard.

Chapter 7 OSNR Monitoring using an Interferometer with Orthogonally Polarised Inputs

In [93] it was shown that the in-band OSNR could be measured using a pair of Michelson fibre interferometers with no requirement for prior knowledge of the signal coherence properties. It is also possible to fulfil these requirements using a single interferometer with a linear polariser before the OSNR module. The resultant measurement is a combination of the polarisation nulling scheme in [94] and a standard interferometric OSNR measurement as in [84,85]. For the double interferometer scheme discussed in chapter 8, it is required that two measurements of the signal visibility at different optical delays be taken. A similar principle may be applied if two measurements of the interferometer visibility are taken at orthogonal polarisations in order to find a measure of the signal amplitude autocorrelation function, which can then be used to retrieve the OSNR.

7.1 Monitoring unit

The measurement setup is shown in Figure 7.1 and consists of a fibre polariser followed by a single Michelson fibre interferometer. Faraday mirrors [95] are utilised to remove the influence of the input polarisation state on the interferometer extinction. The delay between fibre arms was measured as 14.4ps. A variable phase delay element (of the order of femtoseconds) was employed. For this purpose, a piezoelectric tube was used and one arm of the interferometer was wrapped around it. Upon applying a voltage to the tube it would expand and stretch the fibre thereby causing a refractive index change which produces a phase delay in one arm. This was driven by a low frequency (~5Hz) triangular wave. The transmitted intensity was measured using a low-speed photodiode, the voltage of which was sampled at 400kHz using the analog input of a data acquisition (DAQ) card.

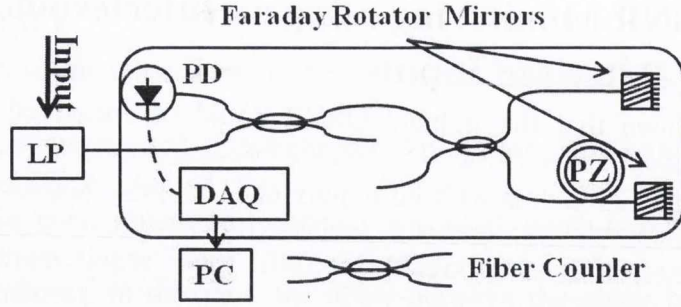


Figure 7.1 Single interferometer OSNR monitor using a linear polariser PZ: piezo fibre stretcher, LP: linear polariser, PC: computer, PD photodiode, DAQ: data acquisition card.

7.2 Theory

In this scheme, we make measurements at two orthogonal polarizations. For an arbitrary polarisation the total power after the polariser LP can be expressed as $P_1 = P_n/2 + P_{s1}$. P_n and P_s refer to the signal and noise power respectively. For the orthogonal polarisation state, the total power after the polariser is given as $P_2 = P_n/2 + (P_s - P_{s1})$.

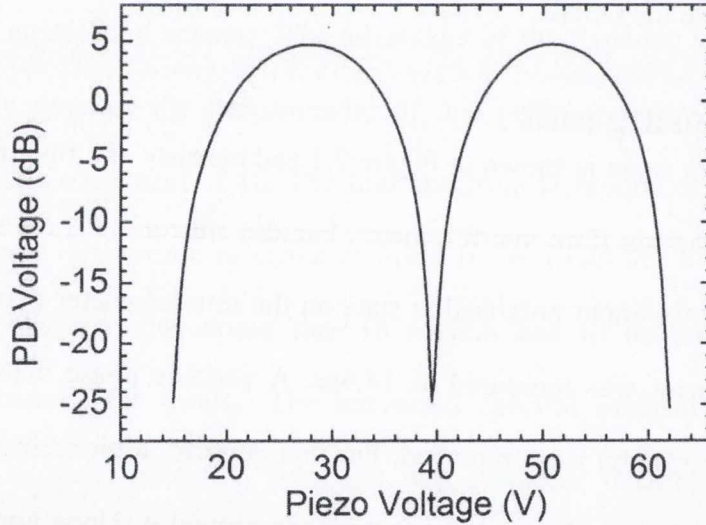


Figure 7.2, interferogram of the noisy signal.

Figure 7.2 shows the interferogram of the noise loaded signal. The maximum and minimum transmitted powers, $S_{\max}$ and $S_{\min}$ for one polarisation state can be expressed in equations (7.1) and (7.2).

$$S_{\max 1} = P_{s1} + \frac{P_n}{2} + \gamma_s P_{s1} + \gamma_n \frac{P_n}{2} \quad (7.1)$$

$$S_{\min 1} = P_{s1} + \frac{P_n}{2} - \gamma_s P_{s1} - \gamma_n \frac{P_n}{2} \quad (7.2)$$

and the maximum and minimum optical powers of the orthogonal polarisation states are expressed in equations (7.3) and (7.4).

$$S_{\max 2} = (P_s - P_{s1}) + \frac{P_n}{2} + \gamma_s (P_s - P_{s1}) + \gamma_n \frac{P_n}{2} \quad (7.3)$$

$$S_{\min 2} = (P_s - P_{s1}) + \frac{P_n}{2} - \gamma_s (P_s - P_{s1}) - \gamma_n \frac{P_n}{2} \quad (7.4)$$

γ_s and γ_n are the amplitude autocorrelation functions of the signal and noise respectively as a function of delay, Figure 7.3 shows the distinction between γ_s for an NRZ-OOK and DPSK signals and γ_n for ASE filtered to 0.25nm. The greater spectral width of the DPSK signal means that its coherence is lower than the OOK signal.

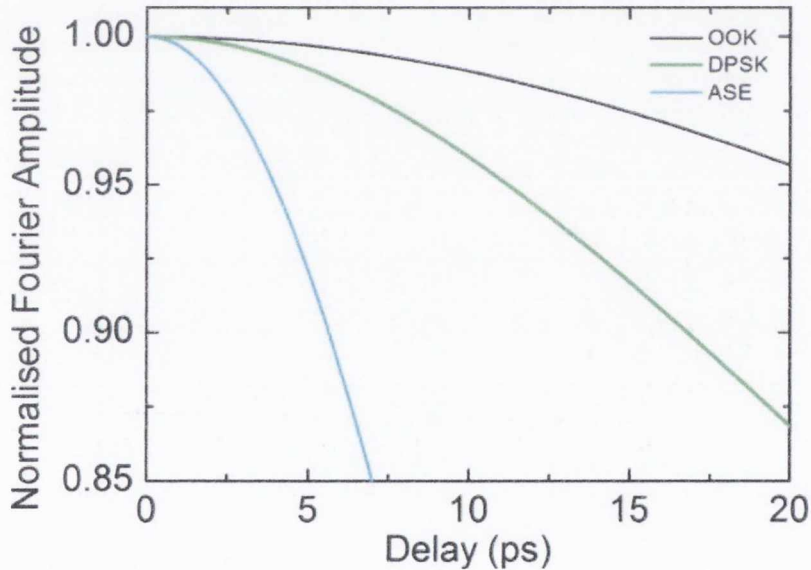


Figure 7.3 Amplitude autocorrelation function of an OOK signal (black), DPSK signal (green) and ASE noise (turquoise)

The difference and average transmitted powers for the one polarisation state can be written as equations (7.5) and (7.6) respectively.

$$\Delta S_1 = \frac{(S_{\max 1} - S_{\min 1})}{2} = \gamma_s P_{s1} + \gamma_n \frac{P_n}{2} \quad (7.5)$$

$$\bar{S}_1 = \frac{S_{\max 1} + S_{\min 1}}{2} = P_{s1} + \frac{P_n}{2} \quad (7.6)$$

Equations (7.7) and (7.8) show the values for the orthogonal case

$$\Delta S_2 = \frac{(S_{\max 2} - S_{\min 2})}{2} = \gamma_s (P_s - P_{s1}) + \gamma_n \frac{P_n}{2} \quad (7.7)$$

$$\bar{S}_2 = \frac{S_{\max 2} + S_{\min 2}}{2} = (P_s - P_{s1}) + \frac{P_n}{2} \quad (7.8)$$

So the difference in ΔS values for the two polarization states may then be written as (7.9).

$$\Delta S_1 - \Delta S_2 = \gamma_s (2P_{s1} - P_s) \quad (7.9)$$

and the difference in the average powers for the two polarisation states may be written as

$$\bar{S}_1 - \bar{S}_2 = 2P_{s1} - P_s \quad (7.10)$$

The subscripts 1 and 2 refer to the arbitrary polarisation state and its orthogonal state respectively. With this being the case the value of the amplitude autocorrelation function of the signal at the optical delay, γ_s can then be determined as (7.11)

$$\gamma_s = \frac{(\Delta S_1 - \Delta S_2)}{(\bar{S}_1 - \bar{S}_2)} \quad (7.11)$$

Also, it is noted that the sum of the average transmitted power for the two polarisation states is equal to the total optical power of the input signal

$$\bar{S}_1 + \bar{S}_2 = P_s + P_n \quad (7.12)$$

and the sum of the difference values may be written as equation (7.13)

$$\Delta S_1 + \Delta S_2 = \gamma_s P_s + \gamma_n P_n \quad (7.13)$$

This allows the signal visibility to be calculated for a given OSNR value in (7.14)

$$\mu = \frac{(\Delta S_1 + \Delta S_2)}{\bar{S}_1 + \bar{S}_2} \quad (7.14)$$

The optical noise to signal ratio, $r = P_n/P_s$ can then be written as (7.15).

$$r = \frac{\gamma_s - \mu}{\mu - \gamma_n} \quad (7.15)$$

γ_n is defined by the power transmission function of the channel filter. The Fourier transform of the filter function yields the value of the amplitude autocorrelation function at the delay time. γ_s , which is typically determined by measuring the signal interferogram in the absence of noise and using the same value for the noisy signal. This can be problematic if a modulator is poorly biased [96]. However, this is not an issue with the scheme presented here as γ_s is measured directly as seen in equation (7.11). The OSNR may then be calculated as shown in (7.16) (normalized to 0.1nm bandwidth)

$$\text{OSNR} = 10\log_{10}\left(\frac{\text{NEB}(\text{nm})}{r \times 0.1(\text{nm})}\right) = -10\log_{10}(r) + 10\log_{10}\left(\frac{\text{NEB}(\text{nm})}{0.1(\text{nm})}\right) \quad (7.16)$$

NEB is the noise equivalent bandwidth of the channel filter.

7.3 OSNR measurements

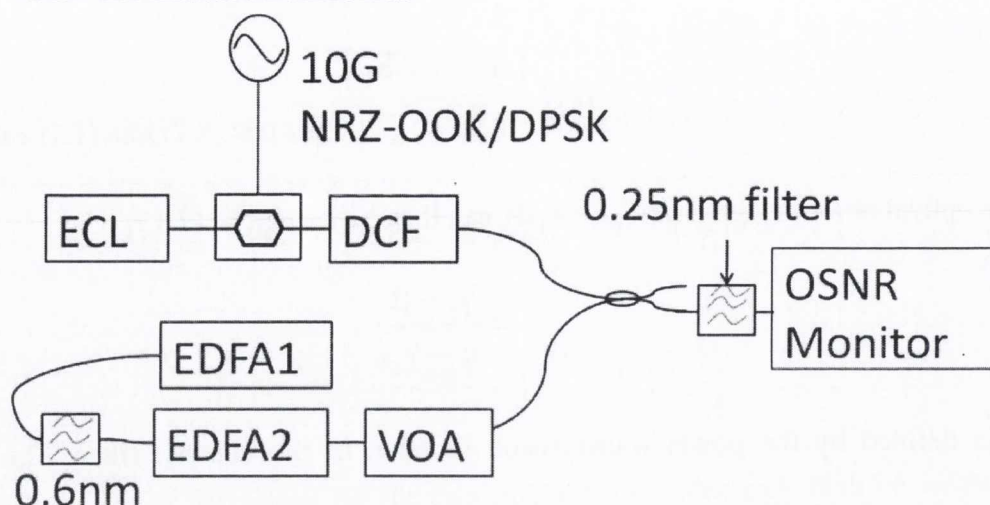


Figure 7.4 OSNR monitoring setup. ECL: External cavity laser, DCF: dispersion compensation fibre, VOA: variable optical amplifier.

Measurements were performed using the scheme shown in Figure 7.4. Signal was provided by an external cavity laser (ECL) tuned to 1549.4nm and modulated at 10Gb/s using a Mach Zehnder modulator and a dispersion compensation fibre (DCF) could be switched in to add 680ps/nm of dispersion to the signal. Noise came from two cascaded Erbium doper fibre amplifiers (EDFA1 and EDFA2) with a 0.6nm filter placed between them. A variable optical attenuator was used to control the noise power and hence the OSNR. Signal and noise were then combined using a 3dB coupler and passed through a tunable filter with bandwidth of 0.25nm before entering the module. The reference OSNR was measured using a power meter which provided a reading of signal and noise prior to commencing measurements. By keeping the signal power constant and attenuating the noise power by a known amount it was possible to have a measure of the set OSNR before the polariser at each point.

7.3.1 Intensity modulated signal results

A non return to zero (NRZ) on-off keyed (OOK) signal, which was generated by biasing the modulator at $0.5V_{\pi}$ and driving it at V_{π} .

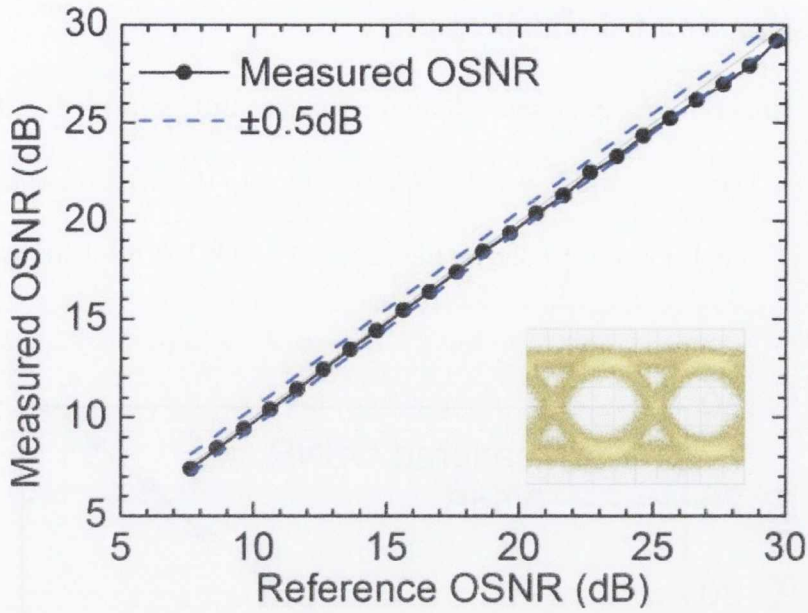


Figure 7.5 Results obtained from measurement on 10G NRZ-OOK signal with eye diagram inset.

Figure 7.5 shows the result of the OSNR measurement performed on the NRZ-OOK signal. Following this, the signal was passed through a section of dispersion compensation fibre whose effect was to impose 680ps/nm of dispersion on the signal. The result of the OSNR measurement on this signal is shown in Figure 7.6.

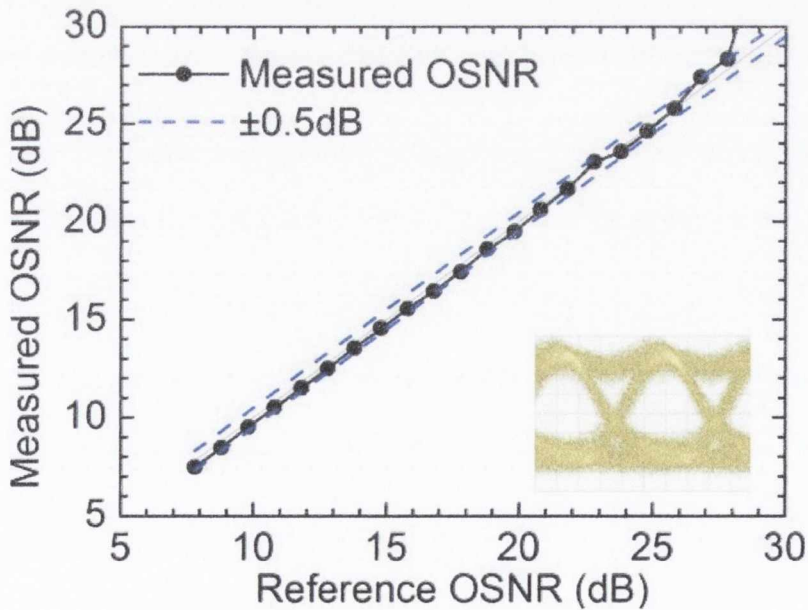


Figure 7.6 Results obtained from NRZ-OOK signal after 680ps/nm dispersion. Eye diagram inset.

7.3.2 Phase modulated signal results

The Mach-Zehnder modulator was DC biased at the null point and driven with a voltage of $2V_{\pi}$. The result of this was a non return to zero (NRZ) differential phase shift keyed (DPSK) signal. Figure 7.7 shows the result of an OSNR measurement performed on this signal.

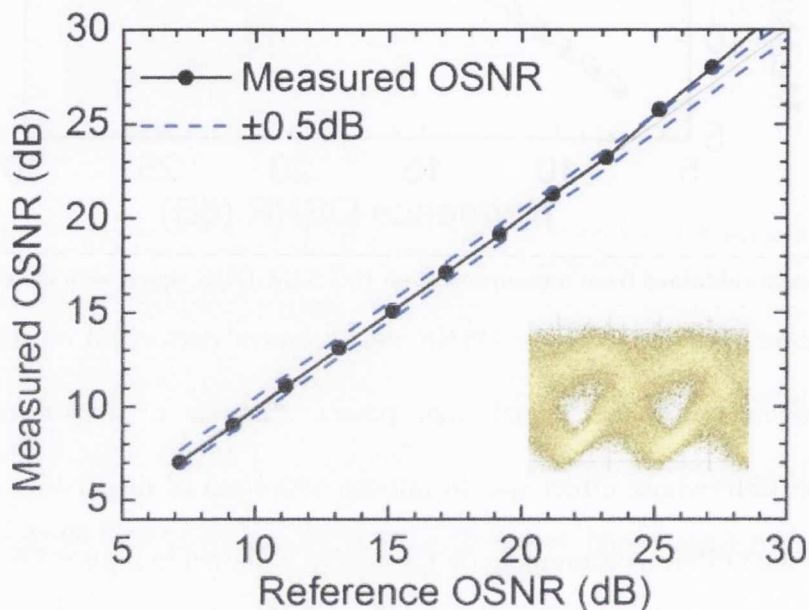


Figure 7.7 results obtained from NRZ-DPSK signal with eye diagram inset.

The effect of adding 680ps/nm of dispersion to the NRZ-DPSK signal is shown in Figure 7.8.

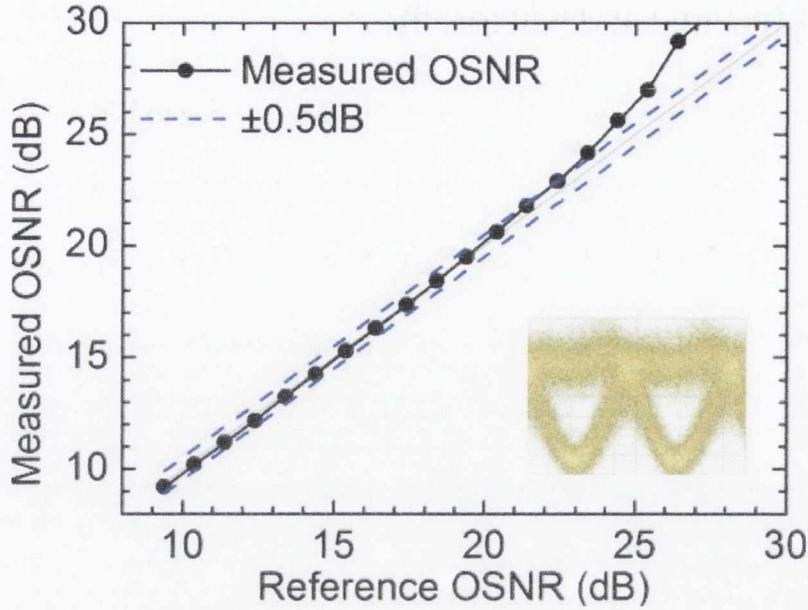


Figure 7.8 results obtained from NRZ-DPSK signal with 680ps/nm dispersion with eye diagram inset.

7.4 Summary

A measurement range of approximately 7 to 25dB OSNR was attainable using this scheme for different modulation schemes without any prior knowledge of the signal coherence properties. A single Michelson interferometer was used in conjunction with a fibreised linear polariser and measurements were performed at two orthogonal polarisations states. In each case, the signal amplitude autocorrelation function was the same, however, the OSNR for each polarisation state was different. This allowed computation of both the amplitude autocorrelation function of the signal and the signal visibility. Along with the amplitude autocorrelation function of the noise, these quantities were sufficient to calculate the in-band OSNR of the signal within the appropriate measurement range and accuracy constraints. It has also been shown that the signal is immune to the effects of a high level of chromatic dispersion. This is similar to what was reported by us in [93] previously using a pair of interferometers with different optical delays.

Chapter 8 In band OSNR monitoring using a pair of Michelson Fibre interferometers

As discussed in chapter 4, knowledge of the in-band OSNR of a communications channel is an important consideration for a reconfigurable optical network due to differing channel transmission histories. As has been stated, current measurement techniques [74] exhibit several drawbacks that would impair our ability to perform an accurate measurement. Another monitoring technique utilises an interferometer with an optical delay and, instead of relying on the differing polarisation properties of signal and ASE noise, differences in their respective spectral (and hence, coherence) properties are exploited. These interferometers generally utilise the Mach-Zehnder configuration in fibreised form [85,84]. In some cases a deterioration in signal coherence leads to the accuracy of results becoming compromised [87] as the interferometers are calibrated assuming that the signal coherence remains constant. This issue may be circumvented either by applying a calibration factor to each modulator as in [4]. To avoid the necessity for such a calibration, a scheme using a pair of interferometers with different optical delays may be used. In this chapter, such a system is tested in the presence of amplitude and phase modulated signals, polarisation multiplexed signals and the effect of various transmission impairments on the scheme is also examined. The importance of such measurements pertains to satisfying the needs of future networks in which various advanced modulation formats may be used, so prior knowledge of the coherence properties of the signal may be unavailable. It is important to investigate the effect of these modulation schemes and impairments on the OSNR monitor in order to investigate its adaptability to such situations.

8.1 Michelson fibre interferometers

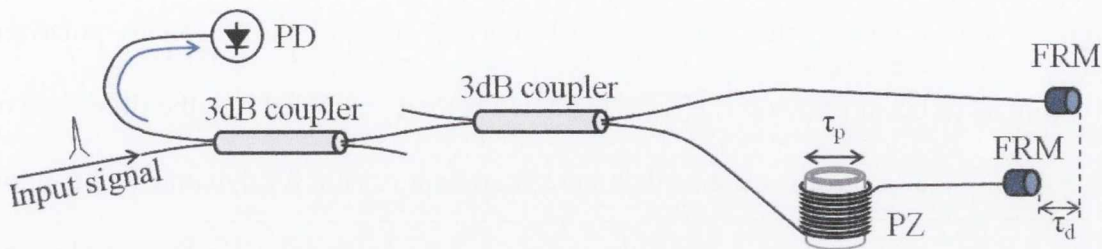


Figure 8.1 A single Michelson fibre interferometer. FRM: Faraday rotator Mirror, PD: photodiode, PZ: piezoelectric tube, τ_p and τ_d are the time and phase delays.

Figure 8.1 shows a single Michelson fibre interferometer. A noisy optical signal enters to the left as shown in the diagram and is split into two arms by a 3dB coupler. One of the fibre arms is slightly longer than the other to facilitate a time delay, τ_d between the two arms, which is of the order of 10ps. The other fibre arm is wrapped around a piezoelectric tube. Both fibre arms are then coupled to Faraday rotator mirrors.

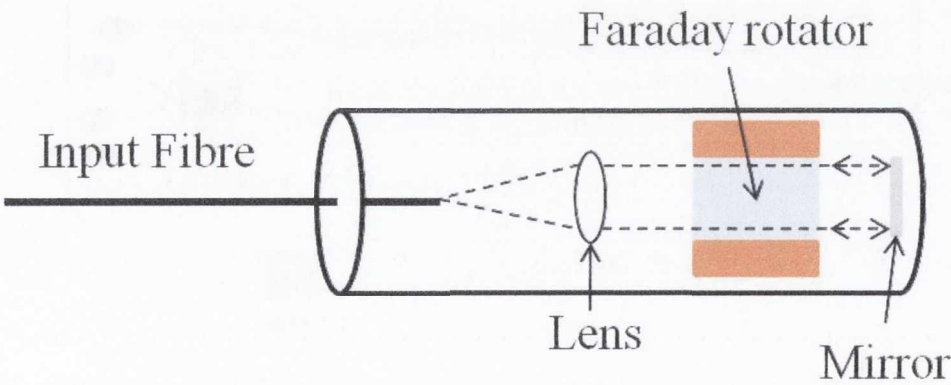


Figure 8.2 Schematic of a Faraday rotator mirror

Faraday rotator mirrors, as shown in Figure 8.2, utilise a lens to collimate the output of the fibre, the collimated beam is then passed through a “Faraday rotator”, which is basically an optical medium in a magnetic field, which by means of the Faraday effect

[97], enforces a rotation of the polarisation of a propagating beam by 45° . Upon reflection, the beam is also rotated by 45° making the total rotation 90° . This, along with the reverse in the “handedness” of the polarisation brought about by the reflection from the mirror, ensures that the returning beam will always be orthogonally polarised to the incident beam [98,99]. The reflected beam is then recoupled into the fibre. Use of the Faraday rotator mirrors ensures that there is no requirement for dynamic polarisation control in the interferometer. The reflected beams then recombine in the coupler and their combined intensity is measured using a high sensitivity, low speed photodiode. As the piezoelectric tube in Figure 8.1 oscillates, it stretches the fibre in one arm of the interferometer, which causes a delay, τ_p , which is of the order of the optical cycle time (a few fs). This acts as a variable phase element. As this delay is scanned the interference pattern (interferogram) may be measured by recording the photodiode voltage as a function of time.

8.2 OSNR monitoring module

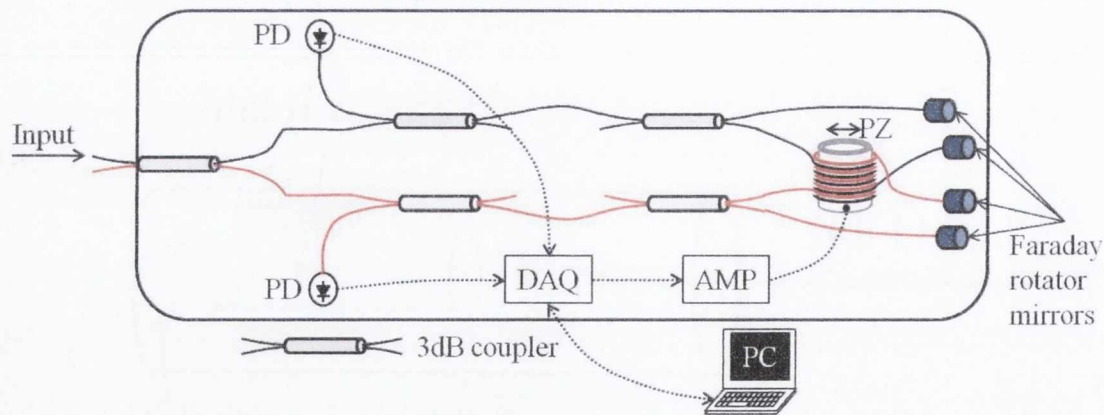


Figure 8.3 OSNR monitoring unit DAQ: data acquisition card, AMP: voltage amplifier, PD: photodiodes

Figure 8.3 shows a schematic of the monitoring unit, which consists of a pair of Michelson fibre interferometers as outlined in Figure 8.1. Light enters the unit following a channel filter and is split into two paths, one for each interferometer (red

and black paths as indicated in Figure 8.3. There is a length difference between the two optical paths in each interferometer which leads to a time delay τ_d . The delay in the black interferometer was 8ps and the delay in the red was 17ps. The piezoelectric tube is driven by a triangle wave produced by the DAQ card and amplified to approximately 100V using the voltage amplifier marked AMP in Figure 8.3. The measured intensity as a function of time is digitised using the analog to digital conversion in the DAQ card, recorded by the computer and is plotted. This plot is known as the interferogram of the signal. The ratio of the maximum recorded voltage to the minimum recorded voltage is called the extinction ratio ϵ , which in turn may be used to measure the OSNR from the treatment shown in section 8.2 if certain other parameters are known. The fixed optical delay happens to fall within the of the order of the manufacturing tolerance of the fibres (1mm corresponds to a delay of 10ps). This delay could be characterised by scanning the interferometers' transmission as a function of wavelength and calculating the delay from the free-spectral range using equation (8.1).

$$\tau_d (\text{ps}) = \frac{\lambda^2 (\mu\text{m})^2}{\lambda_{\text{FSR}} (\mu\text{m}) c (\mu\text{ms}^{-1})} \quad (8.1)$$

where λ is the centre wavelength of the signal, λ_{FSR} is the free spectral range and c is the speed of light.

8.3 Theory behind scheme:

The expression for the electric field of the signal prior to the interferometer is given in equation (8.2).

$$e_{in}(t) = E(t) \exp(j2\pi f_0 t) \quad (8.2)$$

f_0 is the carrier frequency of the signal. The signal at interferometer output is then written as equation (8.3).

$$e_{out}(t) = e_{out}(t)_d + e_{out}(t)_p \quad (8.3)$$

where d and p refer to the signal in the interferometer arms with the variable delay and phase respectively. Written in terms of (8.2), $e_{out}(t)$ becomes

$$e_{out}(t) = \sqrt{K_d} E(t + \tau_d) \exp(j2\pi f_0(t + \tau_d)) + \sqrt{K_p} E(t + \tau_p) \exp(j2\pi f_0(t + \tau_p)) \quad (8.4)$$

K_d is a parameter that represents the power of the signal passing through the interferometer arm with the optical delay. K_p is a parameter that represents the power of the signal passing through the interferometer with the variable optical phase. τ_d is the optical delay as a result of the length difference between the two interferometer arms and τ_p is the phase delay, which is the order of the optical cycle (~ 5 fs) and is the quantity that is modulated as a result of the piezoelectric crystal.

Noise can be described in a similar fashion and prior to the interferometer the expression for its electric field can be written as in equation (8.5)

$$n_{in}(t) = N(t) \exp(j2\pi f_0 t) \quad (8.5)$$

After passing through the interferometer, the noise field may be written as equation (8.6)

$$n_{out}(t) = \sqrt{K_d} N(t + \tau_d) \exp(j2\pi f_0(t + \tau_d)) + \sqrt{K_p} N(t + \tau_p) \exp(j2\pi f_0(t + \tau_p)) \quad (8.6)$$

Assuming the interferometer is polarisation insensitive (due to the Faraday mirrors), a scalar description of signal and noise is adopted. The voltage detected by the low speed photodiode, which is related to the square of the electric field of the output field (intensity), can then be written as equation (8.7).

$$V = C \overline{|e_{out}(t) + n_{out}(t)|^2} \quad (8.7)$$

Where the line above the expression refers to the fact that the low speed photodiode performs a time-averaged measurement of V . C is a constant related to detector efficiency. The average noise power detected by the photodiode can be expressed as the square of the time averaged electric field amplitude of the noise as shown in equation (8.8).

$$\overline{N^*(t)N(t)} = P_n \quad (8.8)$$

P_n is the average noise power. The noise autocorrelation is equal to the Fourier transform of its power spectrum $F(f)$ as shown in equations (8.9) and (8.10). The amplitude autocorrelation of the noise field γ_n as a function of the delay τ_d (τ_p is too short to have an influence on γ_n) is related to the Fourier transform of the noise power spectrum as $F(f)$ as shown in equation (8.11).

$$\overline{N^*(t)N(t+\tau)} = \int F(f) \exp(j2\pi f \tau) df \quad (8.9)$$

$$\overline{N(t)N^*(t+\tau)} = \int F(-f) \exp(j2\pi f \tau) df \quad (8.10)$$

$$\int \frac{F(f) + F(-f)}{2} \exp(j2\pi f \tau) df \equiv \gamma_n(\tau) P_n \quad (8.11)$$

It is assumed that $F(f)$ is defined by the power transmission function of the channel filter (a quantity that is easily measurable using an optical spectrum analyser). γ_n , the autocorrelation of the noise amplitude may be determined through equation (8.11). γ_n is real if the channel filter is symmetrical i.e. when $F(f) = F(-f)$. Were the signal to pass through several ROADMs before reaching the channel filter then the noise power spectrum may not be simply determined by the power transmission function of the

channel filter due to the possibility of the spectrum being narrower than the channel filter. In this case it may be necessary to switch off the signal and measure the interferogram visibility of the noise (which will be present due to the cascaded fibre amplifiers in the network). For the purposes of this analysis, it is assumed that γ_n is known. From equation , it can be seen that γ_n is a real and even function of the delay τ and is normalised to a peak value of 1 at delay time $\tau=0$.

A similar treatment may be performed for the signal, the average signal power is equal to the square of the electric field as shown in equation (8.12)

$$\overline{E^*(t)E(t)} = P_s \quad (8.12)$$

The amplitude autocorrelation of the signal is equal to the Fourier transform of the signal power spectrum as indicated in (8.13) and (8.14)

$$\overline{E^*(t)E(t+\tau)} = \int S(f) \exp(j2\pi f \tau) df \quad (8.13)$$

$$\overline{E(t)E^*(t+\tau)} = \int S(-f) \exp(j2\pi f \tau) df \quad (8.14)$$

The arithmetic average of the signal power spectrum as a function of frequency either direction from the centre frequency is Fourier transformed in equation (8.15)

$$\int \frac{S(f) + S(-f)}{2} \exp(j2\pi f \tau) df \equiv \gamma_s(\tau) P_s \quad (8.15)$$

$S(f)$ is the signal power spectrum, P_s is the average signal power. $\gamma_s(\tau)$, the autocorrelation of the signal amplitude, will also be real, an even function of τ_d and will have a peak value of 1 at $\tau_d=0$. With this in mind, the voltage detected by the photodiode can be written as in equation (8.16)

$$V = C \left[(K_d + K_p)(P_s + P_n) + 2\sqrt{K_d K_p} \times \cos(2\pi f_0(\tau_d - \tau_p)) (P_s \gamma_s(\tau_d - \tau_p) + P_n \gamma_n(\tau_d - \tau_p)) \right] \quad (8.16)$$

Due to the fact that $\tau_p \ll \tau_d$, equation (8.16) may be simplified to (8.17).

$$V = C[(K_d + K_p)(P_s + P_n) + 2\sqrt{K_d K_p} \times \cos(2\pi f_0(\tau_d - \tau_p)) (P_s \gamma_s(\tau_d) + P_n \gamma_n(\tau_d))] \quad (8.17)$$

For a fixed τ_d and a varying τ_p , the voltage measured by the photodiode will follow a sine curve, which is the interferogram of the noisy signal. The maximum and minimum values of this variation can be written as in equations (8.18) and (8.19)

$$V_{\max} = C \left[(K_d + K_p)(P_s + P_n) + 2\sqrt{K_d K_p} (P_s \gamma_s(\tau_d) + P_n \gamma_n(\tau_d)) \right] \quad (8.18)$$

$$V_{\min} = C \left[(K_d + K_p)(P_s + P_n) - 2\sqrt{K_d K_p} (P_s \gamma_s(\tau_d) + P_n \gamma_n(\tau_d)) \right] \quad (8.19)$$

The extinction of the interferogram is given as $\rho = V_{\max}/V_{\min}$ and is expressed on a log scale due to the large of values it can assume. An example of the interferogram of a noisy signal is shown in Figure 8.4 which shows the sinusoidal variation of the photodiode voltage as a function of the driving voltage of the piezo (albeit on a log scale). Using the maximum and minimum voltage values it is possible to define the quantities V_{ave} and V_{diff} as shown in equations (8.20) and (8.21)

$$V_{\text{ave}} \equiv \frac{V_{\max} + V_{\min}}{2} = C(K_d + K_p)(P_s + P_n) \quad (8.20)$$

$$V_{\text{diff}} \equiv \frac{V_{\max} - V_{\min}}{2} = 2C\sqrt{K_d K_p} (P_s \gamma_s(\tau_d) + P_n \gamma_n(\tau_d)) \quad (8.21)$$

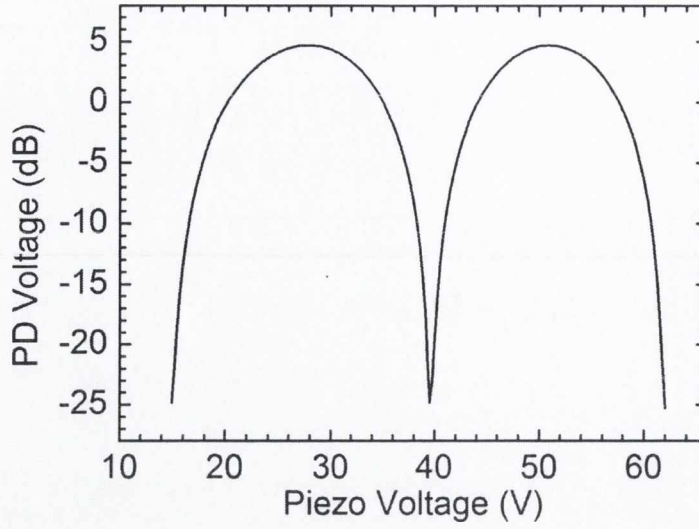


Figure 8.4 interferogram of a noisy signal when the piezoelectric crystal is driven with a triangle wave

The visibility of the interferogram, μ , is defined as $\mu = V_{\text{diff}}/V_{\text{ave}}$. μ is related to the extinction ratio of the interferogram simply through $\mu = (\rho - 1)/(\rho + 1)$ and $\rho = (1 + \mu)/(1 - \mu)$. It can be seen that if μ is close to 1, for example 0.998, the extinction ratio would be quite high ~ 1000 (30dB). r , the optical noise to signal ratio can then be defined as $r \equiv P_n/P_s$ so μ can found from equations (8.20) and (8.21).

$$\mu = \frac{2C\sqrt{K_d K_p}}{C(K_d + K_p)} \frac{(P_s \gamma_s(\tau_d) + P_n \gamma_n(\tau_d))}{(P_s + P_n)} \quad (8.22)$$

We can substitute the first term in (8.22) as follows

$$Z = \frac{2\sqrt{K_d K_p}}{(K_d + K_p)} \quad (8.23)$$

When this has been performed, some substitutions can be made in (8.22) based on the definition of r . μ may then be written as

$$\mu = Z \frac{\frac{1}{r} \gamma_s(\tau_d) + \gamma_n(\tau_d)}{\frac{1}{r} + 1} \quad (8.24)$$

And finally, μ can be written as equation (8.25)

$$\mu = Z \frac{\gamma_s(\tau_d) + r\gamma_n(\tau_d)}{1 + r} \quad (8.25)$$

So calibrating out the power imbalances in the interferometer arms may be performed by dividing by the quantity Z , this defines the quantity M .

$$\frac{\mu}{Z} = M \quad (8.26)$$

M is the interferogram visibility when the power difference in the two interferometer arms has been calibrated out by dividing by Z . This being the case, r can be written as equation (8.27).

$$r = \frac{\gamma_s(\tau_d) - M}{M - \gamma_n(\tau_d)} \quad (8.27)$$

As discussed earlier, $\gamma_n(\tau_d)$ is already known through the channel filter spectrum or measured directly by switching off the signal. Knowing $\gamma_s(\tau_d)$ would allow r to be determined and the OSNR may be written as equation (8.28)

$$\text{OSNR} = 10\log_{10}\left(\frac{\text{NEB}(\text{nm})}{r \times 0.1(\text{nm})}\right) = -10\log_{10}(r) + 10\log_{10}\left(\frac{\text{NEB}(\text{nm})}{0.1(\text{nm})}\right) \quad (8.28)$$

NEB is the noise equivalent bandwidth of the channel filter, which for our purposes is taken as the full width at half maximum of the filter. The second term on the right of equation (8.28) normalises the OSNR to within a 0.1nm optical bandwidth as mentioned in chapter 4. In the case of methods currently available which utilise a single interferometer with a fixed optical delay[2,3], the delay between the arms is made sufficiently large such that $\gamma_n(\tau_d)=0$. $\gamma_s(\tau_d)$ is obtained by switching off the noise in which case only signal is present and $r=0$ so $\gamma_s(\tau_d)=M$. Following addition of noise, the same $\gamma_s(\tau_d)$ is used to measure the OSNR. It has been shown that this method is immune to various network impairments such as depolarisation of the signal due to polarisation mode dispersion (PMD), partial polarisation of the noise due to polarisation dependent loss (PDL) from optical elements installed in the optical link. One important issue with this format is that, in practice, it is difficult to switch off the noise as a long optical link necessitates the use of fibre amplifiers. As a result, the signal will almost always accumulate additional noise. For this reason, an accurate determination of $\gamma_s(\tau_d)$ is almost impossible to attain. Additionally, certain circumstances could lead to $\gamma_s(\tau_d)$ changing, one such condition involves the DC bias or drive voltage of the modulator used to generate the signal changing. Additional filtering and self (cross) phase modulation, which change the signal spectrum, could also change $\gamma_s(\tau_d)$. With this in mind, it is important to either periodically monitor the signal amplitude autocorrelation or perform a self-referencing measurement. This is not possible in the case of a single interferometer with fixed optical delay.

Examination of the function which describes $\gamma_s(\tau_d)$ shows that it nearly always has a parabolic shape around zero delay. Employing optical delays much shorter than the bit period $1/\nu$ allows $\gamma_s(\tau_d)$ to be expanded around τ_d as follows

$$\gamma_s(\tau_d) = 1 - \sum_{k=1}^K c_k \tau_d^{2k} \quad (8.29)$$

Only terms with powers in $2k$ are present due to the fact that $\gamma_s(\tau_d)$ is an even function of τ_d . This can then be substituted into equation (8.27) to give

$$\sum_{k=1}^K c_k \tau_{d,q}^{2k} + r(M_q - \gamma_n(\tau_{d,q})) = 1 - M_q \quad (8.30)$$

The subscript q refers to the fact that a series of measurements that are performed with different optical delays represented by $\tau_{d,q}$ where q is a number from 1 to Q . This means that equation (8.30) is a series of linear equations with unknown values of r and c_k which can be solved if $Q \geq K+1$. Generally selecting $k=1$ and performing two measurements at two different delay values i.e. $Q=2$, can achieve results within the appropriate range and accuracy (30dB ± 0.5 dB OSNR). It is assumed in what follows that it is necessary to measure two interferograms of the noisy signal at two different delays which are much smaller than the bit period of the signal (100ps for a 10G signal).

8.4 Measurements:

8.4.1 CW signal

The basic experimental setup is shown in Figure 8.5. The box marked “OSNR monitor” refers to Figure 8.3 Signal was provided by an external cavity laser tuned to 1549.4nm. Noise was provided by a pair of erbium doped fibre amplifiers (EDFA) cascaded together with the output of EDFA1 filtered to 4nm and used as the input for EDFA2. The output of EDFA2 was filtered using a 100G DWDM filter (NEB=0.6nm). This was combined with the signal using a 3dB coupler and fed into the input of the OSNR

monitor. The OSNR was changed by maintaining a constant signal power and attenuating the noise power using a variable optical attenuator (VOA). The reference OSNR was determined by initially measuring the signal and noise power using a power meter. By attenuating the noise by a known amount for each point it was possible to know the actual OSNR throughout the measurement. The phase delay in the interferometer was changed by applying a triangular wave of 1Hz to the piezoelectric tube and the photodiodes' voltages were sampled by the DAQ card at 200kHz per channel. Ten interferograms (see Figure 8.4) were taken for each OSNR value and the average visibilities for each of the two interferometers was used to calculate the OSNR using equation (8.30)

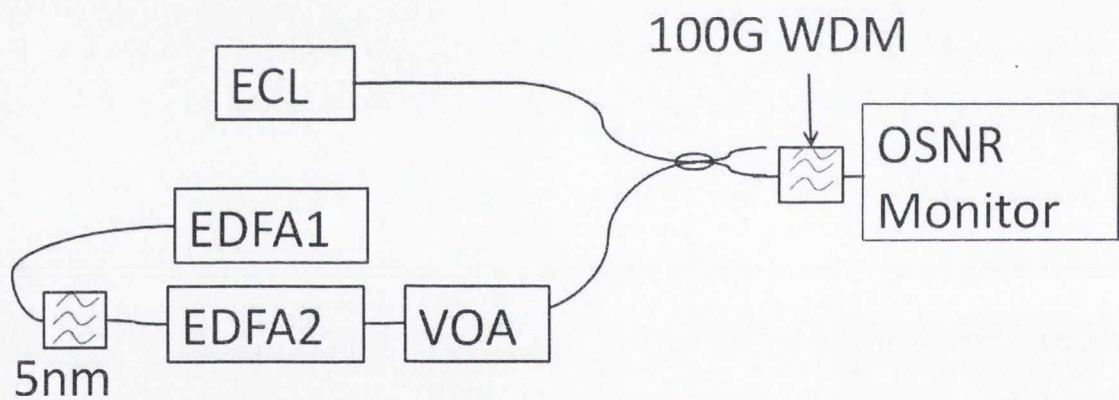


Figure 8.5. Setup for monitoring OSNR of CW signal. ECL: external cavity laser

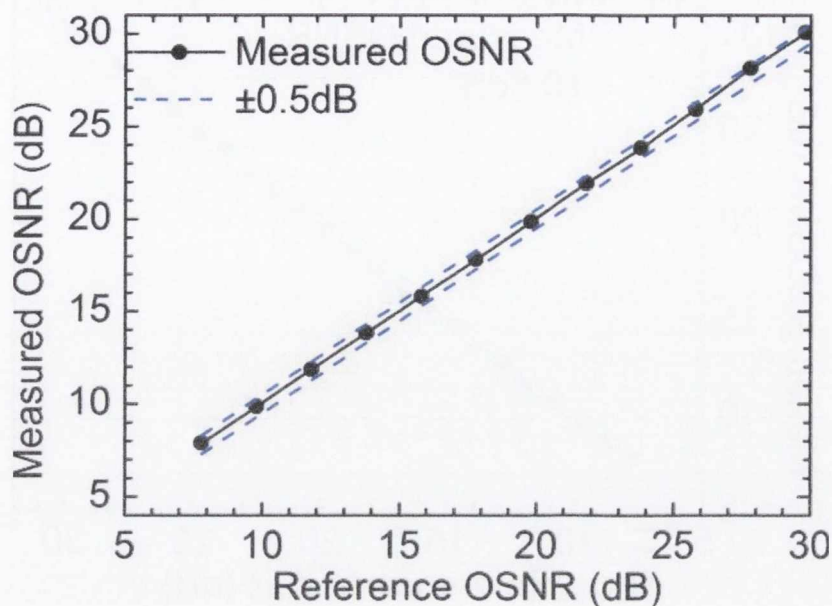


Figure 8.6. OSNR measurement using CW signal. Outer dashed lines represent an error of $\pm 0.5\text{dB}$

Figure 8.6 shows the result of the measurements with using a CW signal for a range of OSNR values from 7 to 30dB

8.4.2 Modulated signals

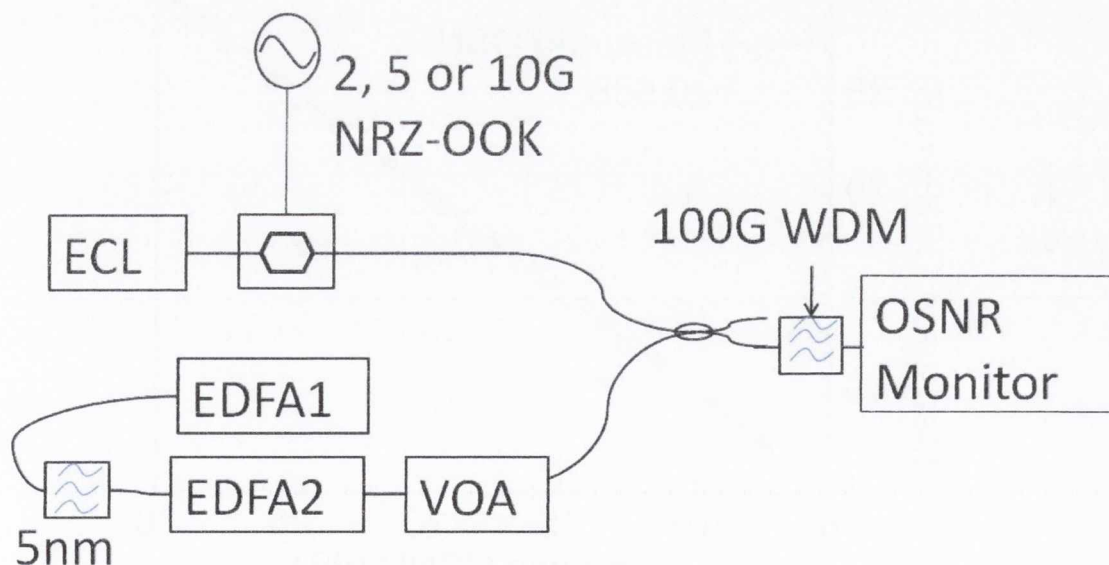


Figure 8.7 setup for OSNR measurement on modulated signal. VOA: variable optical attenuator, ECL: External cavity laser

Figure 8.7 shows the setup for OSNR monitoring of the modulated signals. A Mach-Zehnder modulator was used to generate NRZ-OOK PRBS (pseudorandom binary sequence) signals at repetition rates of 2, 5 and 10G.

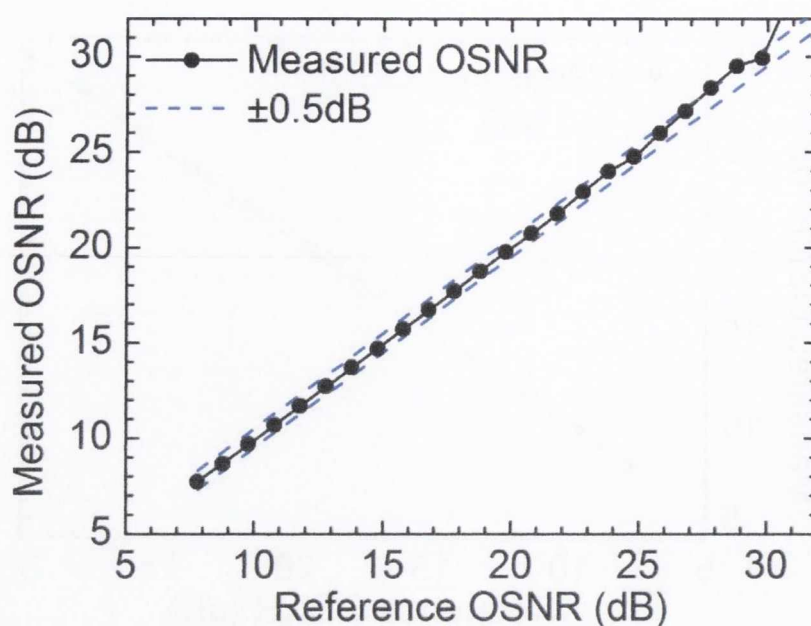


Figure 8.8. OSNR measurement on 2G NRZ-OOK PRBS signal

Figure 8.8 shows the result of the OSNR measurement on the 2G signal between 7 and 30dB OSNR. Figure. 8.9 shows the result of the measurement on a 5G NRZ-OOK signal and Figure 8.10 shows the result of the measurement on a 10G NRZ-OOK signal.

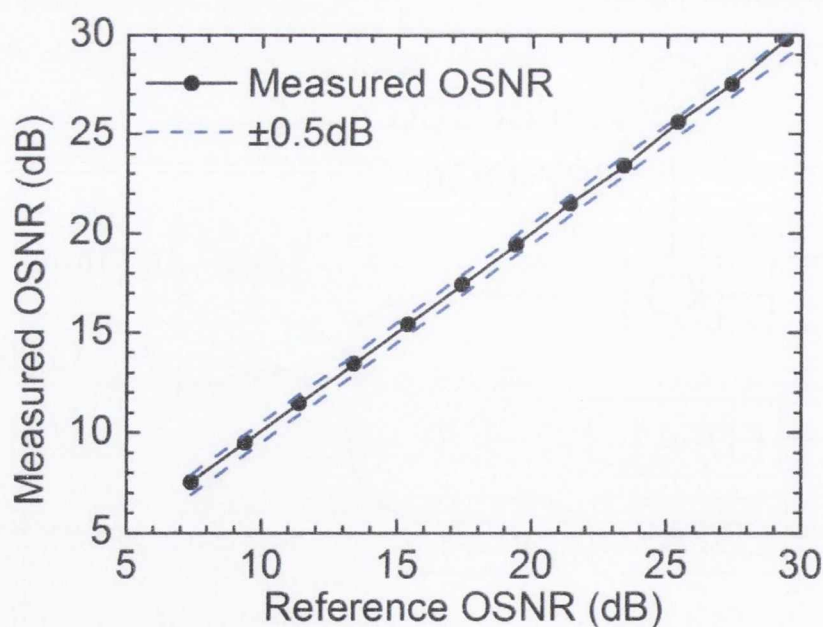


Figure. 8.9. OSNR measurement on 5G NRZ-OOK PRBS signal

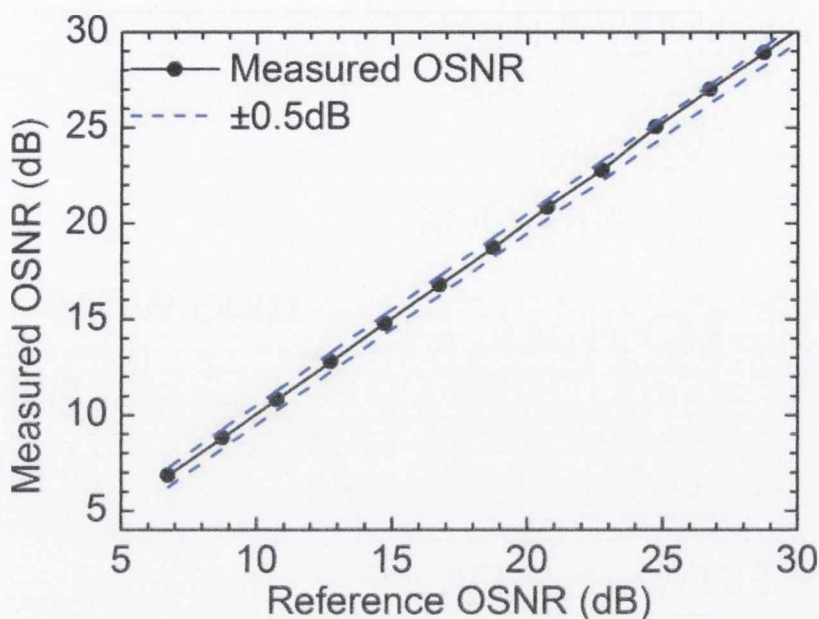


Figure 8.10. OSNR measurement on 10G NRZ-OOK PRBS signal

8.4.3 Signals with degraded extinction ratios

For this measurement, the setup remained the same as in 8.3.2, the only difference was that the Mach-Zehnder interferometer used to generate the 10G signal was overbiased by 2V. Figure. 8.11 shows the results obtained from the measurement on this signal

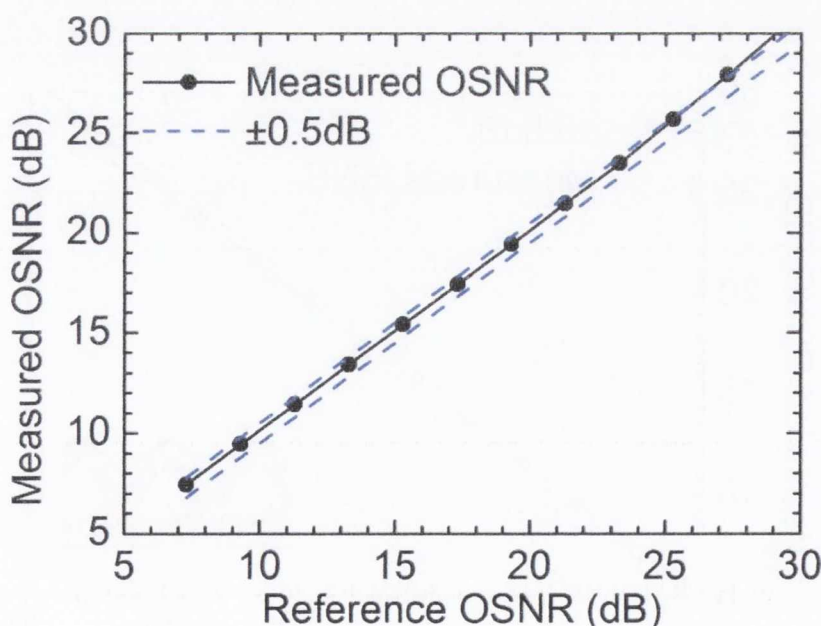


Figure. 8.11 OSNR measurement on 10G NRZ-OOK signal where the modulator was overbiased leading to a degraded extinction ratio.

8.4.4 Signal with added dispersion

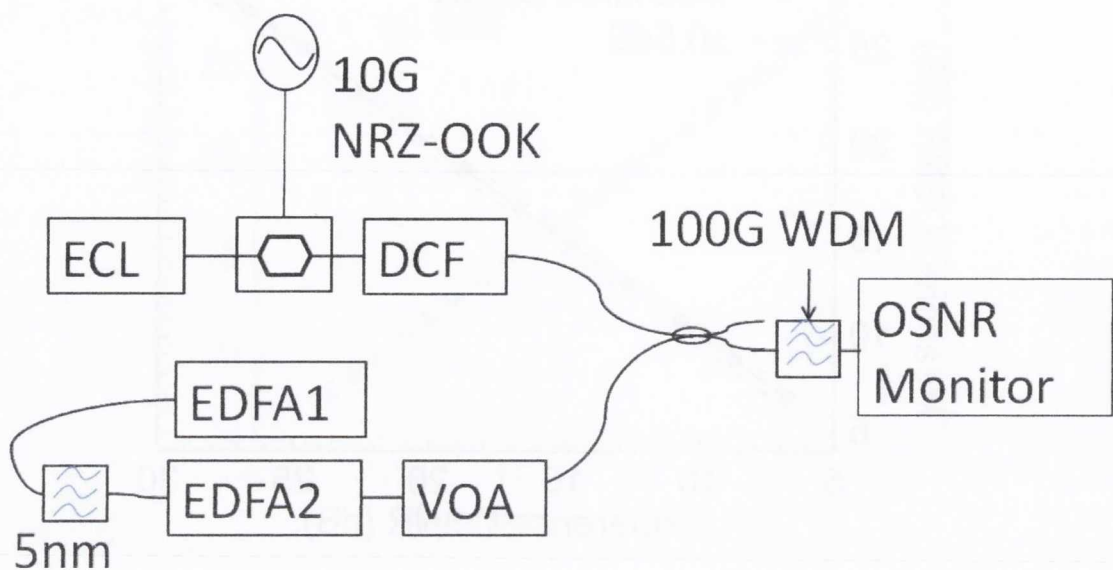


Figure 8.12 Setup for OSNR measurement on modulated signal with added dispersion. DCF: dispersion compensation fibre.

A dispersion compensation fibre was used to add 600ps/nm of dispersion to the modulated signal as shown in Figure 8.12. The result of the OSNR measurement on this signal is shown in Figure 8.13 with the corresponding signal eye-diagram inset showing the broadening of the pulses due to the dispersion.

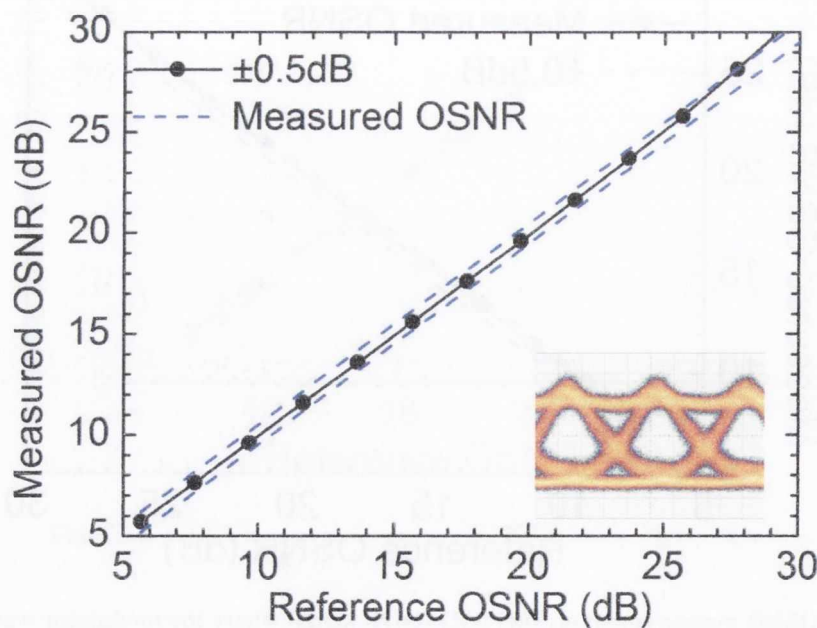


Figure 8.13. OSNR measurement on 10G NRZ-OOK signal with 600ps/nm of chromatic dispersion with the signal eye-diagram inset

8.4.5 CW signal with Polarised noise

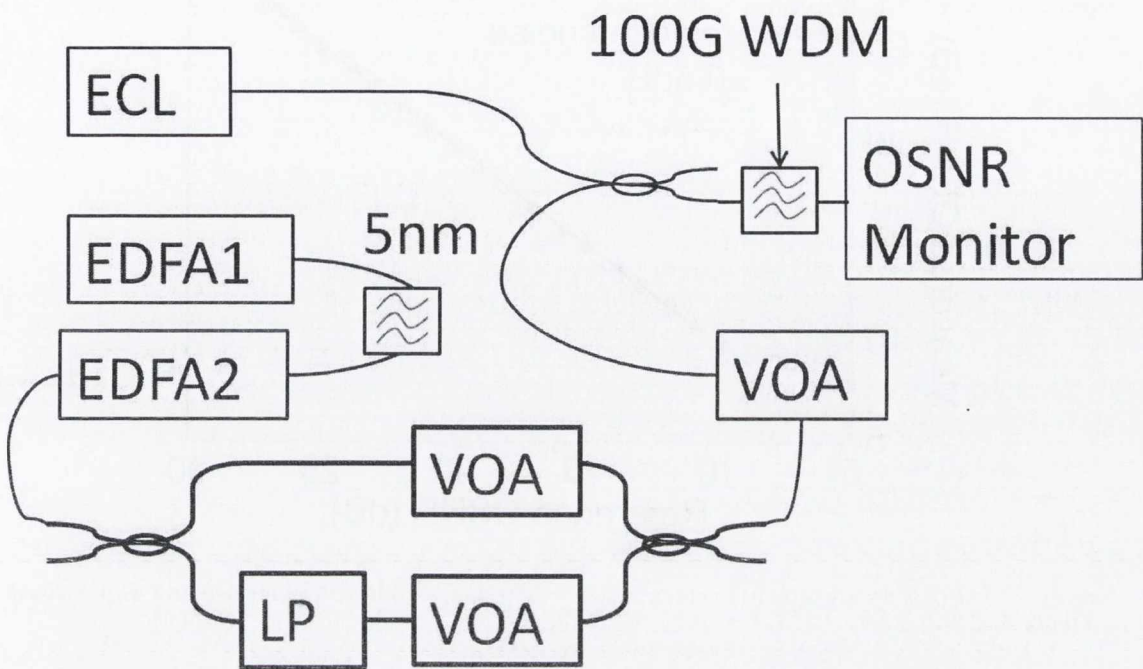


Figure 8.14 OSNR monitoring setup for signal with polarised noise LP: linear polariser, VOA: variable optical attenuator

Figure 8.14 shows the setup for measuring the effect of polarised noise on the OSNR monitor. The unpolarised noise from the amplifiers is split using a 50:50 coupler and one arm was fed through a linear polariser followed by a variable optical attenuator while the other was left unpolarised and also passed through a VOA and combined with the polarised ASE. In this way, it was possible to adjust the degree of polarisation of the noise to the desired value. The effect of a range of degrees of polarisation (DOP) values from 0 to 1 was investigated in increments of 0.25. shown in Figure. 8.15 is the OSNR measurement performed when the noise is fully polarised and completely unpolarised.

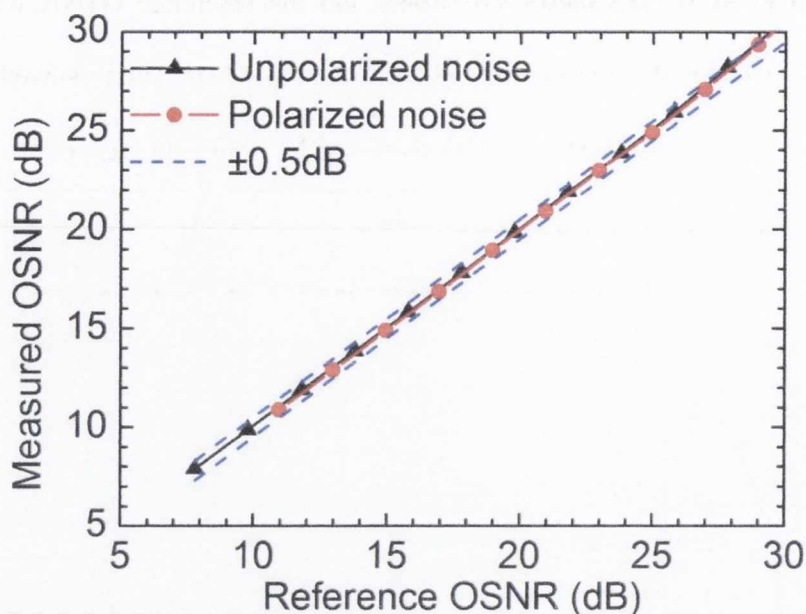


Figure. 8.15 OSNR measurement performed on NRZ-OOK signal with polarised and unpolarised noise

8.4.6 Phase modulation results

The monitor was identical to that used in the previous section, however, the experimental setup is slightly modified and is as shown in Figure 8.16.

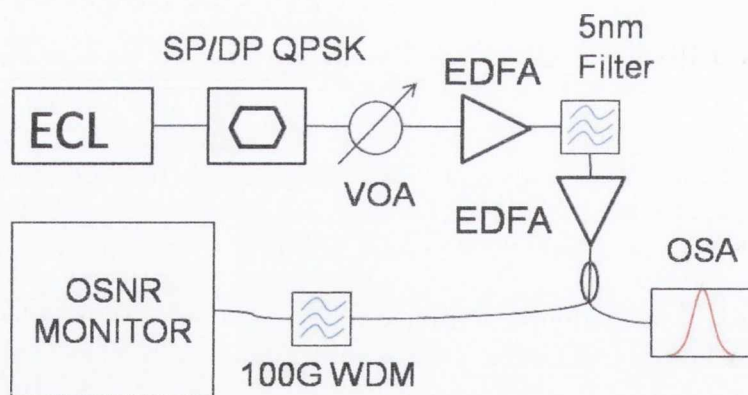


Figure 8.16 setup for OSNR monitoring of QPSK signal

The output from an external cavity laser was modulated with a 12.5G QPSK signal and attenuated using a variable optical attenuator. This was then amplified and filtered to 5nm using an EDFA. This is used as the input for a second EDFA. The signal was split at this point and one arm went to an optical spectrum analyser (OSA) and the other was filtered using a 100G WDM filter and fed into the same module as in Figure 8.3. The

OSNR was changed by attenuating the signal and the reference OSNR was measured using the linear interpolation method with the OSA. Measurements were performed using single and dual-polarisation QPSK signals. The result of this is shown in Figure 8.17

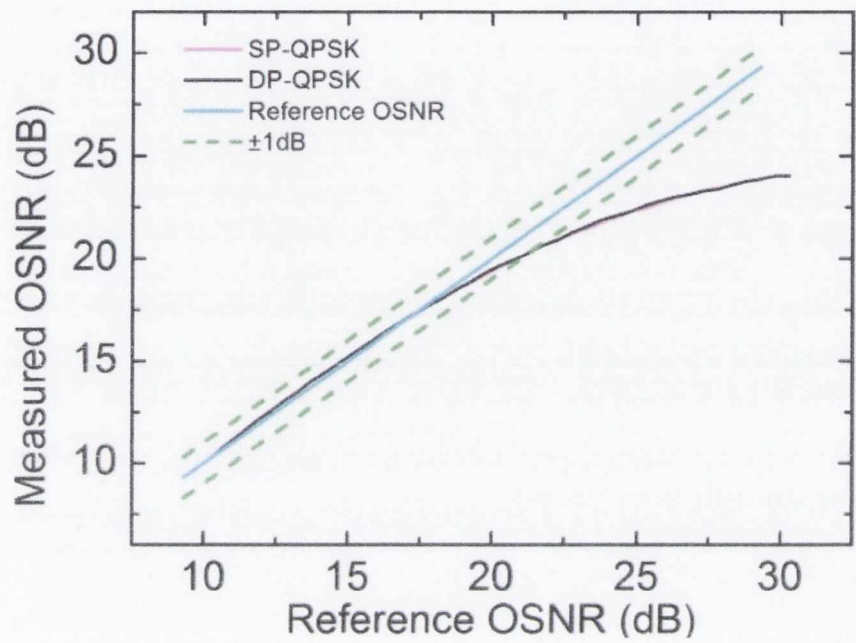


Figure 8.17 Single (pink) and dual (black) polarisation QPSK results

8.5 Discussion

It has been shown in Figure 8.6 and Figure 8.7-Figure 8.10 that the monitor was capable of measuring in-band OSNR values of up to $30\text{dB} \pm 0.5\text{dB}$ from a noisy CW signal and 10G NRZ-OOK signal respectively. When 600ps/nm of dispersion was added the effects were seen to be negligible with a similar measurement range being attained as shown in Figure 8.13. The effects of fully and partially polarised noise were investigated in Figure 8.14. This can affect measurement accuracy in both the polarisation-nulling method and the orthogonal polarisation heterodyne mixing method. In our case, there was little to no effect on measurement accuracy over the measurement range. A poorly biased Mach-Zehnder modulator was employed in the case of Figure. 8.11, the effect of this would be to change $\gamma_s(\tau_d)$ and cause measurement errors if a single interferometer were to be used. Owing to the fact that our system has no reliance

on the assumption that $\gamma_s(\tau_d)$ remains constant, a changing $\gamma_s(\tau_d)$ had a minimal effect on results obtained. It was seen that increasing the bit period of the signal to a value much larger than the optical delay by utilising a 2G and 5G NRZ-OOK signal had little influence on measurements as shown in Figure 8.8 and Figure. 8.9 respectively. The optical delay values used in these measurements would not be suitable for use in a 40G system. However, the use of shorter delays would accommodate this with reasonable ease with no further modifications to the setup. The accuracy of the system could be further improved by using photodetectors with lower noise and a larger dynamic range as this can help to measure the extinction ratios of the interferograms more precisely. Another issue is with the parabolic function used to approximate $\gamma_s(\tau_d)$. This is simple and results in a pair of linear equations. A more sophisticated system may involve the use of a Pade approximation where γ_s evolves according to equation (8.31).

$$\gamma_s(\tau_d) = \frac{1 - c_1 \tau_d^2}{1 + c_2 \tau_d^2} \quad (8.31)$$

The use of such an approximation means that a third unknown, c_2 is introduced to the system. This requires knowledge of how the signal was generated and as such defeats the purpose of this scheme.

8.6 Phase modulated signal discussion

It is evident from Figure 8.17 that the measurement range attained using the QPSK signal does not compare well to that which was observed using the OOK modulated signals. The reasons for this become apparent when the spectra and amplitude autocorrelation functions of these signals are investigated. Figure 8.18 shows the spectra of the OOK and QPSK signals along with that of the noise. It can be seen that the QPSK signal has a larger spectral width than the OOK signal.

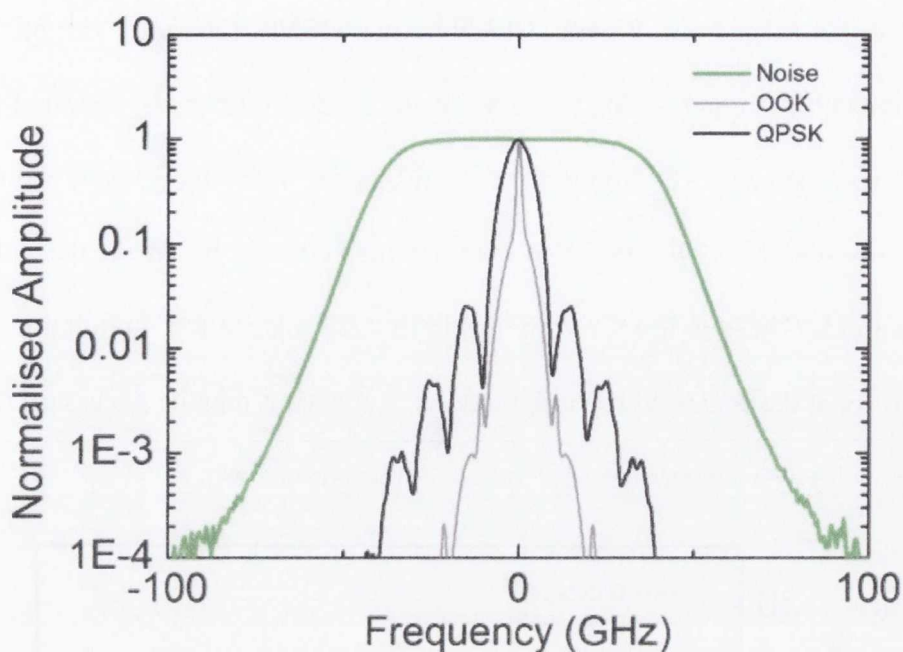


Figure 8.18 spectrum of the filtered ASE noise (green), QPSK signal (black) and OOK signal (grey)

Taking the Fourier transforms of these spectra results in the plot shown in Figure 8.19

The wider spectrum of the QPSK signal means that its $\gamma_s(\tau_d)$ falls off at a greater rate than the OOK signal. The result of this is that the measured OSNR will roll off as in

Figure 8.17.

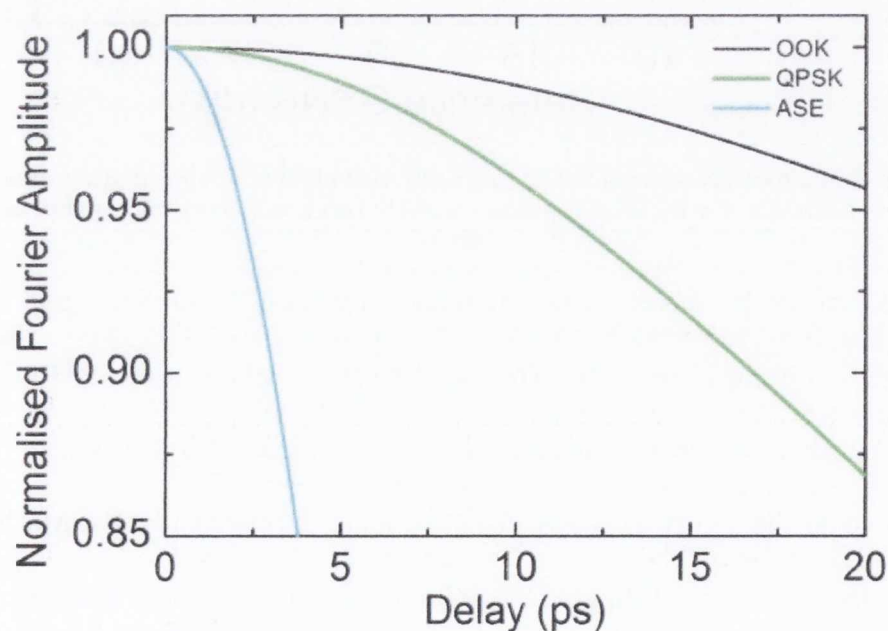


Figure 8.19 Amplitude autocorrelation functions for an OOK signal (black), QPSK signal (green) and ASE noise (turquoise)

It can also be seen from Figure 8.17 that the OSNR measurement on the dual polarisation QPSK signal yielded nearly identical results to those obtained from the single polarisation case. The importance of this can be understood when one considers that the currently available OSNR monitoring methods are based on the polarisation nulling method. These will not be applicable to dual polarisation modulation schemes. In order to solve the issues with phase modulated signals it may be necessary to employ several delay interferometers in order to better approximate $\gamma_s(\tau_d)$

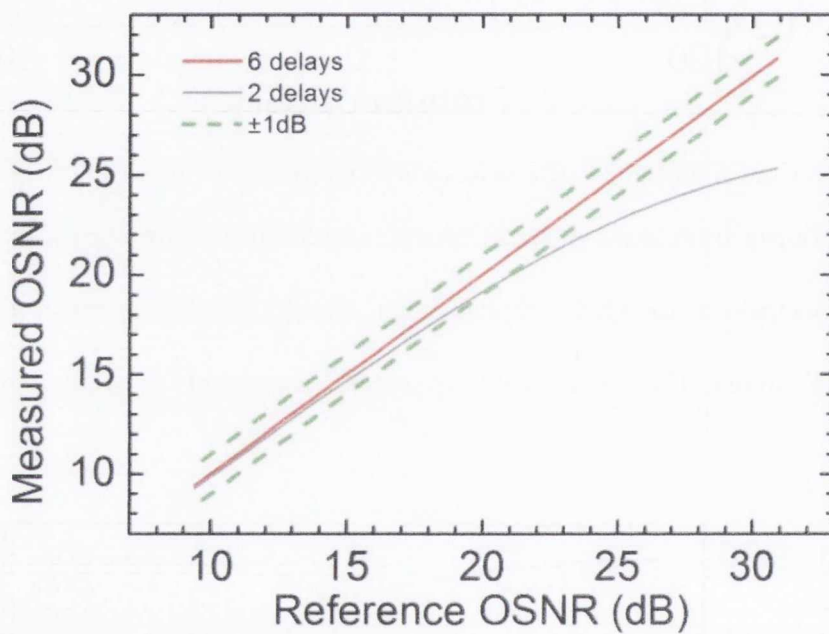


Figure 8.20 Simulation of results obtained using a different number of optical delay values. The red line shows the result of using six delay values while the grey line shows the result of using two delays

Figure 8.20 shows the simulated OSNR measurement on the QPSK signal using two optical delays of 8 and 17 ps (grey). The use of 6 optical delays from 5 to 15ps inclusive in 2ps increments was also simulated. The noise autocorrelation was taken from the filter data while the signal data was taken as being generated using a Mach-Zehnder modulator and the results of this are shown in the red plot. It is seen that such a system would allow an OSNR measurement up to 30dB and possibly beyond within 1dB accuracy. This would be simple to implement using a variable fibre delay line with a single interferometer.

8.6.1 Compatibility with 40G signals

The two interferometer delay values used here would not be suitable for a 40G signal due to the higher spectral width causing the parabolic approximation to fail at lower delay values. Simulations were performed with shorted delays on 10 and 40G signals that were assumed to have been generated with a mach Zehnder modulator with an extinction ratio of 26dB. The delay values used were 6.5 and 7ps.

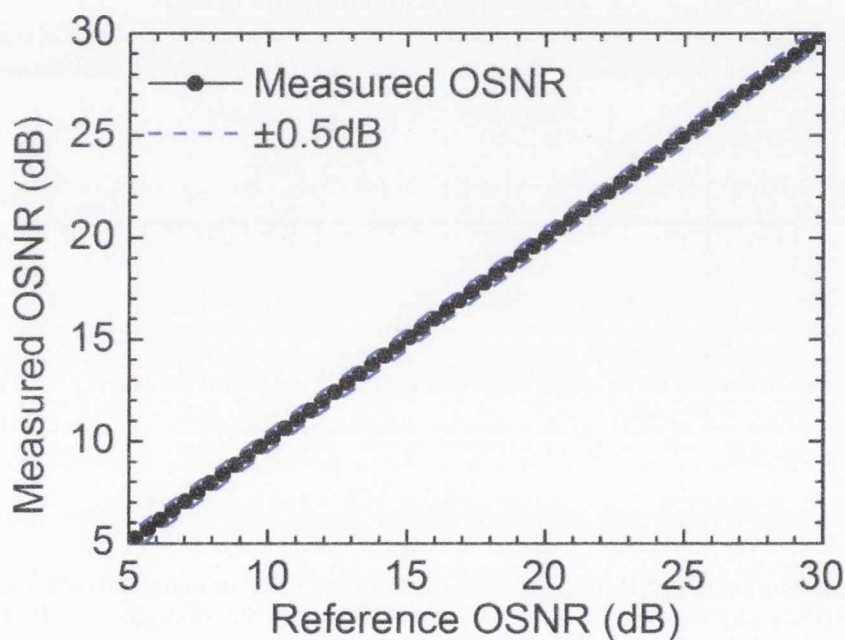


Figure 8.21 Results of a simulated OSNR measurement using two delays of 6.5 and 7ps on a 10G signal

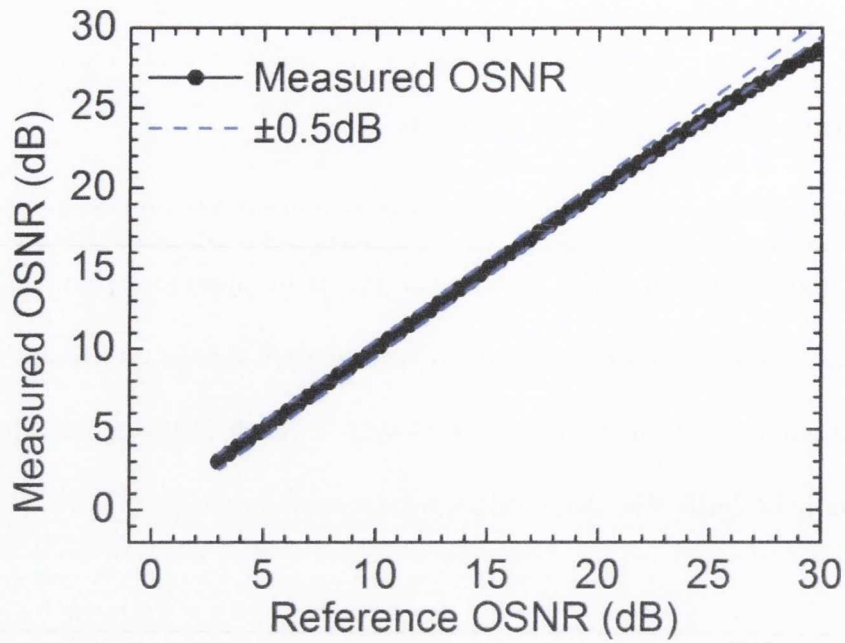


Figure 8.22 Result of simulated OSNR measurement on 40G signal using delay values of 6.5 and 7ps

It can be seen from Figure 8.21 and Figure 8.22 that there exists a set of delays at which OSNR values of up to about 27dB may be measured within 0.5dB, which is an important consideration if the data rate of the signal is unknown for a given channel in a transparent network.

Chapter 9 Conclusions

In this work, a series of novel techniques for the characterisation of optical pulses were developed and examined. Full characterisation of the electric field of optical pulses was demonstrated using two simple techniques based on measurement of the spectrum of a train of pulses aligned with different components of a sinusoidal phase modulation. The first method involved measuring the spectrum of the output of a phase modulator driven with a sinusoidal RF voltage synchronised with the repetition frequency of the pulses. The delay between the phase modulation was stepped over 2π radians in sixteen uniform steps and the resultant variation in the intensity of the spectral lines of the pulses as a function of delay was used to determine the phase differences between each line and hence the total phase change across the pulse. A harmonic decomposition was used to elucidate the spectrum and phase of the pulse train. Results were compared to a commercially available FROG unit. It was seen that the field of optical pulses down to 2ps in duration could be measured in this manner. Longer pulses were also measured by broadening the 2ps pulses by passing them through varying sections of dispersion compensated fibre.

A second phase sensitive pulse monitoring scheme was also examined, this also exploited the harmonic decomposition method for recovery of the pulses' electric fields. A similar arrangement was used involving a phase modulator but the manner in which the data was acquired was slightly different. In this case the delay between the phase modulation and the pulse train was continuously modulated at a frequency of the order of 10kHz which lead to a modified spectrum, which in this case was measured using a photodiode in conjunction with a voltage controlled, narrow bandwidth Fabry-Perot tunable filter. Only one measurement of this modified spectrum was necessary to allow the harmonic decomposition recover the pulse phase and spectrum, which allowed even shorter characterisation times than the stepped delay scheme. The recovered data was

compared to two other schemes for pulses with three different dispersion levels. The first is Kang and Dorrer's sinusoidal phase modulation method from [89] and the second was the harmonic decomposition method. Results were seen to compare favourably with those obtained from the stepped delay harmonic decomposition scheme but Kang and Dorrer's scheme was unable to measure the shorter optical pulses with reasonable accuracy and only recovered useful data when the pulses were subject to a large chirp.

Two methods for measurement of the in-band OSNR of a communications signal were also developed. The first was based on a combination of a fibre polariser and a Michelson fibre interferometer. Measurement of the signal interferogram of an arbitrary polarisation state and its orthogonal state allowed the OSNR to be recovered with no need for prior knowledge of the signal's coherence properties. This allowed measurement of the in-band OSNR of 10G NRZ-OOK signals and NRZ-DPSK signals up to 25dB with and without added chromatic dispersion. However, the nature of this measurement was such that it would not be suitable for measurement of the OSNR of polarisation multiplexed signals and would be susceptible to polarised noise and signal depolarisation.

Another method involving the use of a pair of Michelson interferometers with two short optical delays was developed. This was capable of measuring the in-band OSNR of a 10G NRZ-OOK signal with a PRBS data pattern up to 30dB ± 0.5 dB and was seen to be immune to the influence of polarised noise, chromatic dispersion, modulator bias. Simulations were performed to check the ability of the unit to measure the OSNR of 40G NRZ-OOK signals. It was seen that the optical delay values currently used would not allow measurements on such a system with reasonable accuracy. The effect of modification of the optical delay of the interferometers was examined and it was seen

that delays of 6.5 and 7ps would allow measurement of the OSNR of 40G signals to greater accuracy. The effect of phase modulated signals on the system was also examined. It was seen that the larger spectral width of the phase modulated signals severely degraded the OSNR measurement. Due to the reliance of current OSNR monitoring techniques on the polarisation properties of signal and ASE, there is currently no available means of measuring the OSNR of polarisation multiplexed signals. To this end, OSNR measurements were performed on polarisation multiplexed QPSK signals with the results matching almost exactly those obtained for the single polarisation QPSK signal. Simulations show that two approaches may help alleviate the issue of the degradation of the OSNR measurement due to the wider spectrum of phase modulation signals. The first approach involves modelling of the signal autocorrelation using a Pade approximation, this was seen by way of a simulation to increase the measurement range of the unit. However, knowledge of the signal modulation format is required to allow recovery of the in-band OSNR. Simulations have also shown that choice of different delay values would also allow for a better approximation of the signal autocorrelation.

9.1 Future work

The major issue with the dynamic delay modulation harmonic decomposition method is the recovery of the first harmonic from the modified pulse spectrum. Amplification of the photodiode voltage is the obvious solution but this would prevent measurement of the zeroth harmonic due to its magnitude being outside the dynamic range of the photodiode. Selective amplification of the first harmonic is necessary in this regard. Another issue pertaining to this method is the influence of the second and higher harmonics, even though a digital low pass filter is implemented on the data following acquisition, some of the second harmonic signal remains but this may be mitigated prior

to the data acquisition using an analogue filter. Also, as the delay between the phase modulation and the pulses is altered, the RF power at the output of the phase modulator varies also, causing the depth of modulation to change as the delay is shifted. The use of a limiting amplifier after the phase modulator should alleviate this issue somewhat. In terms of measurements, it will be necessary to characterise communications pulses repeating at 10G and 40G with different modulation formats. The phase modulator is driven sinusoidally at the same signal as that used to generate the pulses, which is not a realistic scenario if the scheme formed part of an OPM system. For this reason it will be important to use some form of clock recovery so there is no reliance of prior knowledge of the signal.

It can be seen from chapter 8 that the in-band OSNR monitoring module suffers impaired performance in the face of advanced modulation formats such as QPSK. Also, further work needs to be done to allow measurement of 40G signals. Simulations have shown that the delay values used on the 10G NRZ and the QPSK signals would not be adequate for measurement of 40G signals. Were the monitor to be deployed in an OPM system, it should be possible to measure the OSNR of signals with any modulation format or bit rate without changing the delays. With this in mind it will be necessary to test various delay values. A more convenient way to allow this would be to use a single interferometer with a variable time delay. Other issues which may affect the measurement pertain to signal filtering prior to the unit. While it is possible to characterise the noise autocorrelation through measurement of the visibility in the absence of signal, this is best avoided as it involves switching off the signal in the channel under test. It is assumed that the noise autocorrelation amplitude is determined by the channel filter but even with the simulations performed in chapter 8 it will be necessary to test the effect of several cascaded ROADMS on the measurement. If the

effect is sufficiently deleterious, it will be necessary to design a filter such that the effect of several ROADMS may be mitigated.

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