# Fast Wavelength Switching Transceiver for a Virtualized Coherent Optical Network

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*(Invited Paper)*

***Abstract*—Fast reconfigurable digital coherent transceivers will become a key subsystem in future virtualized optical networks as they provide the ability to tailor the channel frequency, symbol rate, and modulation format. This transceiver agility is of paramount importance for network resilience, bandwidth on demand applications, and congestion aware routing. We demonstrate a fast wavelength switching coherent transceiver based on semiconductor tunable lasers and a parallelized digital signal processing implementation. The inherent frequency and phase noise associated with semiconductor tunable lasers is analyzed and a dc pilot tone assisted phase noise mitigation technique is proposed for a fast wavelength switching dual-polarization orthogonal frequency-division multiplexed transceiver. This enables the use of commercially available digital supermode distributed Bragg reflector lasers in both the transmitter and receiver in an 800-km wavelength routed coherent network that exhibits a variable path history.**

***Index Terms*—Coherent communications, tunable lasers, wavelength switching.**

## I. Introduction

THE conventional architecture of coherent optical networks is entering a new age, where the fundamental control and distribution of information is tailored in software and based on a global network optimum. This is being achieved by decoupling the network control program from the physical layer, thereby allowing for the abstraction of intelligence from network routers and switches to a centralised global controller. This simple paradigm enables the realisation of a virtual network topology and seeks to unify how services are delivered and managed across an entire domain using a single protocol.

Routing or dynamic reconfiguration is typically performed on a packet-by-packet basis in large electrical layer 2 (data link) or layer 3 (network) switches and the underlying optical physical layer (layer 1) is viewed as static pipes interconnecting edge nodes. However, global traffic growth is increasing at ∼25% per year and the capacity of current optical networks will eventually run up against hard physical limits. Traditionally, capacity increases for optical networks have been achieved using wavelength division multiplexing (WDM) techniques, however gains in information capacity from WDM experiments are beginning to plateau.

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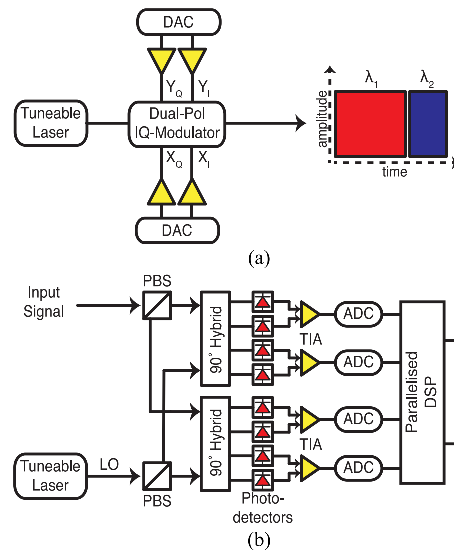


Fig. 1. Fast wavelength switching coherent transceiver.

Therefore, the requirement for more intelligent networking and optical routing at the physical layer is becoming more apparent, as this provides greater bandwidth utilisation and enables the network to be operated within tighter performance margins. Optical routing can be achieved by performing sub-wavelength switching in the optical domain and facilitates the dynamic allocation of point-to-point links. A key subsystem required to realise a truly reconfigurable physical layer is the fast wavelength switching coherent transceiver, which is capable of dynamically adjusting the channel frequency, modulation format and symbol rate in nanosecond time frames.

This paper will focus on recent developments in fast wavelength switching coherent transceivers and outline the DSP techniques required to mitigate the large frequency and phase noise associated with the semiconductor tunable lasers (TLs) that provide key functionality within the transceiver. The DP-OFDM modulation format is proposed for a fast wavelength switching network and the performance of the transceiver is demonstrated in an 800 km optical burst switched (OBS) ring network.

## II. Wavelength Tunable Transceiver

The basic configurations of the transmitter and receiver subsystems within a fast wavelength switching coherent transceiver are illustrated in Fig. 1. The transmitter (see Fig. 1(a)) consists of a widely tunable semiconductor laser, which should typically

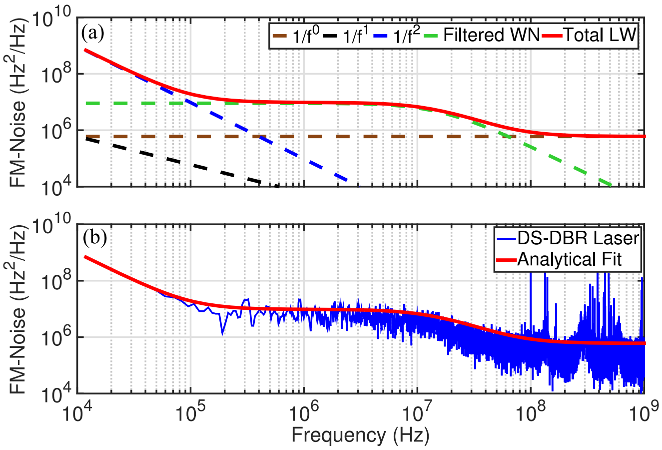


Fig. 2. (a) Analytical FM noise spectrum of a multi-section DBR laser and (b) experimentally measured FM noise spectrum of a DS-DBR laser (operated in static mode) with an analytical fit.

exhibit a switching time of <100 ns, high frequency stability and low phase noise. The output of the semiconductor TL is passed directly into a dual-polarisation nested Mach–Zehnder modulator and multi-level electrical signals are supplied by high speed digital-to-analog convertors. In such an arrangement, the modulation format, symbol rate, channel frequency and output power can be independently adjusted to meet the requirements of a specific link within the network.

At the receiver (see Fig. 1(b)), a second semiconductor TL is used as a local oscillator (LO) and the frequency is tuned to coincide with the incoming channel of interest. The coherently received signal is digitised using analogue-to-digital convertors, after which a parallelised digital signal processing (DSP) implementation [1] is employed for chromatic dispersion compensation, equalisation and frequency noise mitigation.

### A. *Frequency Modulation (FM) Noise*

One of the key advantages of semiconductor TLs is the exploitation of the free carrier plasma effect to vary the refractive index of passive tuning sections [2], which enables the alignment of multiple reflection peaks. This creates a wavelength selective loss within the cavity [3], [4] and through the fine adjustment of a short passive phase section, ensures single mode operation with a high side mode suppression ratio and very fast switching times. However, this tuning mechanism also leads to one of the main drawbacks associated with these devices. Noise present on the switching signals applied to the TL passive sections causes carrier density fluctuations, which results in a corresponding refractive index and loss fluctuation. This variation is directly translated into frequency noise, which is proportional to both the number of laser passive sections and the tuning efficiency of each section.

An analytical approximation [5] of the FM noise spectrum of a tunable multi-section DBR laser is shown in Fig. 2(a). The white noise (WN) component ($1/f^0$) corresponds to the Lorentzian phase noise of the TL and is determined by the fundamental Schlawlow–Townes–Henry linewidth. Low frequency noise ($1/f^1$ and $1/f^2$) is the dominant noise term below 100 kHz; while significant filtered WN arising from injection recombination shot noise [2] in the passive tuning sections is present at frequencies up to 100 MHz. The magnitude of the filtered WN is also dependent on the number of TL passive sections, their tuning efficiency and bandwidth. The total TL FM noise is the summation of each individual noise component, as seen in Fig. 2(a).

In this work, a commercially available DS-DBR TL from Oclaro [4] was employed as both the transmitter laser and as the LO laser in the fast wavelength switching transceiver. The laser consists of a uniform phase grating with intermittent $\pi$-phase shifts at the rear of the device, a continuous phase tuning section, a gain section and a multi-contact (eight contacts) chirped grating at the front of the device. The output power (∼15 dBm) is maintained by an integrated semiconductor optical amplifier (SOA) section. In operation, the front grating is used as the coarse wavelength (supermode) selection mechanism by creating a broad reflection peak which overlaps with one of the peaks from the rear phase grating comb. Fine tuning of the wavelengths of the longitudinal modes is achieved by injecting current into the phase section.

The FM noise spectrum of a single channel at 1550 nm was experimentally measured using a coherent heterodyne technique [6] and is illustrated in Fig. 2(b). The bandwidth of the filtered WN originating from the laser passive sections was reduced by placing series resistors and parallel capacitors on the laser evaluation board to provide low-pass filtering. The filtering removes electronic noise from the tuning-section, at the expense of a slightly increased switching time. The visible peaks at frequencies between 100 MHz and 1 GHz correspond to multiple suppressed sub-peaks that are present ∼40 dB below the peak of the received signal power spectral density (PSD) [6]. The Lorentzian phase noise was ∼1.2 MHz and varied between 700 kHz and 1.5 MHz across the entire conventional wavelength band. An analytical approximation, based on the measured bandwidth response of each passive section, is also displayed in Fig. 2(b).

In a fast wavelength switching coherent transceiver, each component contributing to the overall FM noise spectrum of the TL must be tracked and mitigated in order to avoid significant performance penalties due to phase uncertainty and cycle slips in the carrier phase estimation algorithm in the receiver DSP.

## III. FM Noise Mitigation Techniques

A number of FM noise mitigation techniques have been proposed in the literature for fast wavelength switching burst mode coherent receivers. Pre-emphasis of the laser drive signals was proposed by Simsarian *et al.* to reduce both the switching time of the semiconductor TL [7] and also to reduce the instantaneous frequency drift of a DS-DBR laser from 1 GHz to <100 MHz [8]. Although signal pre-emphasis relaxes the requirements of the digital frequency offset (FO) estimator in the receiver DSP, it is a challenging technique to optimise for every TL switching combination and also has be performed for both transmitter

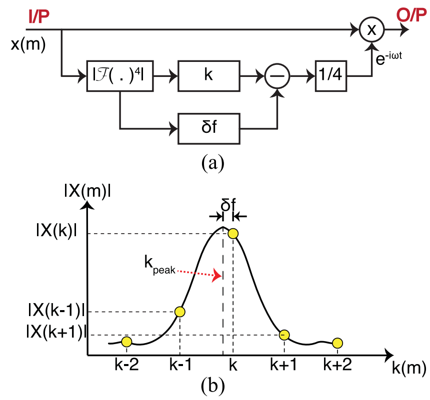


Fig. 3. (a) Block based FO estimation technique and (b) schematic of approximate-fit algorithm to estimate FO correction factor [12], where $k$ is the index of the FFT sample with the largest magnitude, $X(k)$.

and receiver lasers independently. Although this may prove prohibitive as a frequency noise mitigation technique, signal pre-emphasis is required for the transmitter TL, to avoid a longitudinal mode hop when using SOA blanking during a switching event [9].

An alternative instantaneous FO mitigation technique for a fast wavelength switching transceiver is illustrated in Fig. 3(a). The estimator is a block based algorithm, which is implemented in parallel after the equaliser in the receiver DSP [10]. Each equalised block is passed into the FO estimator and the data is removed using a 4th-order non-linearity algorithm [11]. A Fourier transform is subsequently performed and the frequency corresponding to the peak of the resulting PSD is taken as the estimate of the FO.

However, the resolution of the measurement is inversely proportional to the block length, therefore a quadratic interpolation technique proposed by Jacobsen and Kootsookos [12] is used to increase the accuracy of the estimate. This is achieved by calculating a correction factor ($\delta f$), using three FFT samples ($X(k-1)$, $X(k)$ and $X(k+1)$) centred around the index of the largest FFT magnitude sample, $k$, as seen in Fig. 3(b). Once corrected, the FO is assumed to be constant across the block length and the process is repeated for the remaining parallelised blocks. This provides a time resolved instantaneous FO estimation that is subsequently removed from the equalised signal. A key advantage of this technique is that it is fully receiver based, can be implemented in parallel and estimates the combined instantaneous frequency of both the transmitter and receiver lasers simultaneously.

An alternative approach to FM noise mitigation for fast switching coherent transceivers is to exploit the use of training symbols (TSs). An OFDM-based data independent TS was used by Dischler [13] to estimate the instantaneous FO in a 32- and 64QAM system. The training sequence consisted of a synchronisation symbol proposed by Schmidl and Cox [14], which enabled the estimation of the FO, which was assumed to remain constant for the duration of the burst. This allowed the reception of differently shaped M-ary QAM signals, however external cavity lasers were used at both the transmitter and within the coherent receiver. Although ECLs provide a very stable optical carrier with low phase noise, the large form factor and slow switching times (>10 ms) associated with these devices render them impractical for reconfigurable dynamic optical networks.

As previously demonstrated, when considering semiconductor TLs, the instantaneous FO must be tracked as a function of time across the entire burst. Gripp *et al.* demonstrated the use of TSs for correlation based polarisation separation and FO estimation [15]. The TS consisted of a 70 ns header, where only one quadrature phase shift keyed (QPSK) symbol was transmitted. This negated the requirement to remove the data at the receiver using the 4th-order non-linearity algorithm prior to FO estimation. This technique tracked the FO through the burst header and enabled the coherent reception of 5 GBd DP-QPSK packets.

It is important to note that instantaneous FO estimators do not mitigate the white FM noise component (Lorentzian phase noise) of a semiconductor TL. Lorentzian phase noise can cause cycle slips in the receiver DSP, which occur when there are errors in the $\pi/2$ phase unwrapping in the carrier and phase recovery algorithm. After a cycle slip occurrence the receiver will begin to decode the received symbols in the incorrect quadrants, causing catastrophic errors until the receiver is reset. Differential decoding can be employed to avoid catastrophic errors due to cycle slips [10], however this technique will not reduce the significant performance penalties in systems that employ higher order modulation formats and also becomes non-trivial for square QAM formats higher than 16QAM. Conversely, the Lorentzian phase noise of a semiconductor TL can be fully compensated at the transmitter using an electro-optical feed-forward technique [16], while the digital coherent enhancement technique can be used for receiver LO laser phase noise compensation [17]. Both mitigation techniques have demonstrated excellent phase noise compensation, however this is accompanied by a significant increase in transceiver complexity.

We propose the use of OFDM for a fast wavelength switching coherent transceiver as it offers numerous advantages applicable to reconfigurable optical networks. The framing and TSs that are inherent to the OFDM format can be exploited for burst detection, timing synchronisation and channel estimation. In addition to this, OFDM is compatible with the significant FM noise associated with fast tunable semiconductor lasers, as it easily allows for the insertion of an unmodulated dc pilot tone to aid with phase and frequency noise compensation. The use of the cyclic prefix (CP) also solves the problem of fast channel dispersion estimation that is required in order to accommodate the variable path history that may occur in a system with dynamic bandwidth allocation.

## IV. DP-OFDM Transceiver

The experimental setup for the fast wavelength switching DP-OFDM transceiver and 800 km OBS network is illustrated in Fig. 4. The DS-DBR laser in the transmitter was dynamically switched between two 50 GHz spaced channels on the ITU grid ($\lambda_1$: 1553.22 nm and $\lambda_2$: 1552.83 nm), with a burst length equal to the OFDM frame period ($\sim$7 $\mu$s). The total frequency drift of both transmitter and receiver TLs, experienced after a switching

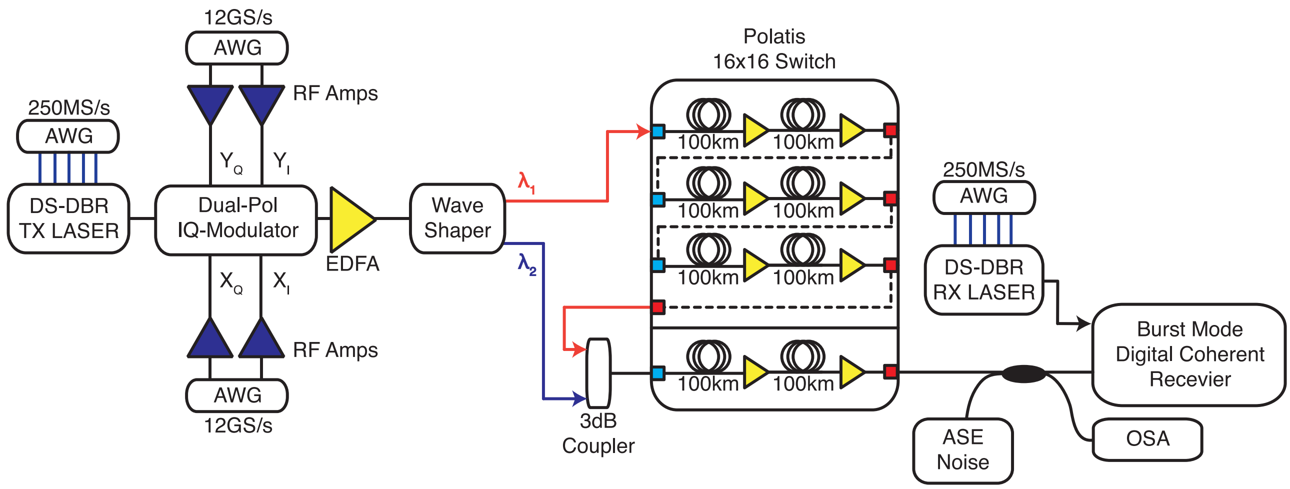


Fig. 4. Experimental setup of fast wavelength switching DP-OFDM transceiver and 800 km OBS network with variable path history.

event, was ∼300 MHz for each burst channel; however this can increase up to ∼2 GHz when larger switching bandwidths are considered [10]. The OFDM frames for each burst channel were generated offline in MATLAB and consisted of a frame header, TSs (both at the start of the frame and periodically throughout the frame) and data symbols. The frame header was based on the Schmidl and Cox scheme [14] and consisted of two even symbols transmitted on a single polarisation. This was used for both burst detection and to estimate the angle of the received polarisation.

The frame header was followed by 20 OFDM TSs, transmitted simultaneously on both polarisations, which are used for channel estimation. An FFT size of 512 was used, of which 256 carriers were encoded with QPSK modulation and a single pair of TSs were inserted periodically every 34 symbols for polarisation tracking. The training and data symbols were mapped onto the 256 carriers with the exception of ±15 carriers around dc, which provided a guard band to accommodate for the RF pilot tone that was used for phase noise compensation (a technique similar to that proposed by Jansen *et al.* [18]).

The OFDM signal was subsequently two times up-sampled by padding with zeros, before an inverse fast Fourier transform was used to generate the time domain OFDM symbols. This provided an ODFM symbol length ($L_{sym}$) of 512 samples, to which a CP ($L_{cp}$) of 30 samples was appended in order to accommodate chromatic dispersion. The OFDM waveforms were quantised and uploaded to two 12 GS/s arbitrary waveform generators (AWGs) that provided the in-phase and quadrature components for each polarisation, which were subsequently applied to an integrated dual polarisation IQ modulator. The RF pilot was placed at the centre of the OFDM spectrum by adding a dc bias to the X-polarisation IQ modulator drive voltage. The actual bit rate per WDM carrier after the OFDM and forward error correction (FEC) overhead had been removed was 18 Gb/s. It is important to note that the upper limitation of this bit rate was due to the electrical bandwidth response of the AWGs available to us at the time of this work.

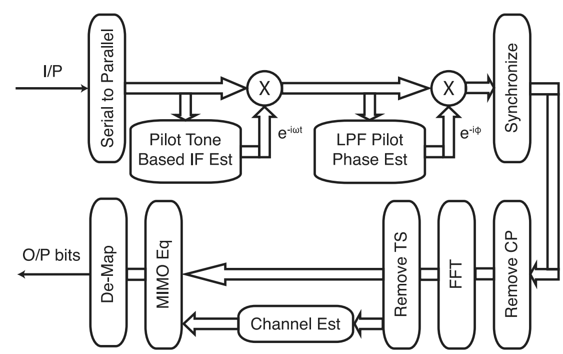


Fig. 5. DP-OFDM receiver DSP.

The transmission line consisted of eight 100 km spans of standard single mode fibre (SSMF) (arranged in 200 km links), which were each followed by a gain-flattened EDFA. A Polatis 16 × 16 low loss optical switch was used to reconfigure the transmission spans and to re-position the digital coherent burst mode receiver. The variable path history of the network was adjusted by first setting the transmission distance (0–600 km) for CH1, before both channels were combined using a 3 dB coupler to traverse the final 200 km simultaneously.

The DP-OFDM signals were captured using a polarisation diverse coherent receiver that consisted of a 90° optical hybrid, four balanced photo-receivers (TeleOptix) with a 3 dB electrical bandwidth of 25 GHz and a 50 GS/s real-time sampling oscilloscope with 20 GHz analogue electrical bandwidth. The coherent burst mode receiver utilised a second DS-DBR laser as the LO and the frequency was tuned to coincide with the transmitted burst channels. The DSP was carried out offline and used block based processing that is suitable for implementation on an ASIC and a schematic of the important DSP blocks are illustrated in Fig. 5.

### *A. Receiver DSP*

The received signals were initially sampled and converted from serial to parallel into blocks of length, $L_b = L_{sym} + L_{cp}$.

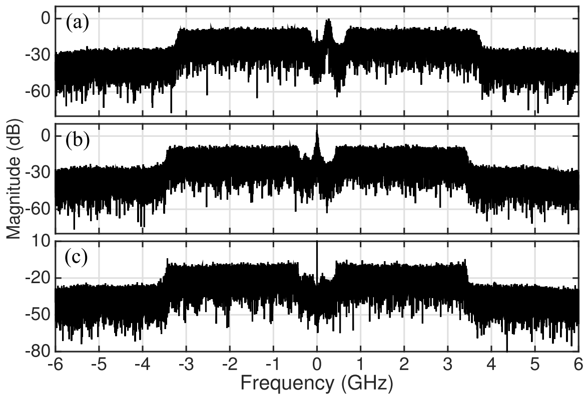


Fig. 6. Received OFDM spectrum, (a) at the receiver, (b) after block based FO compensation and (c) after pilot tone based carrier phase compensation.

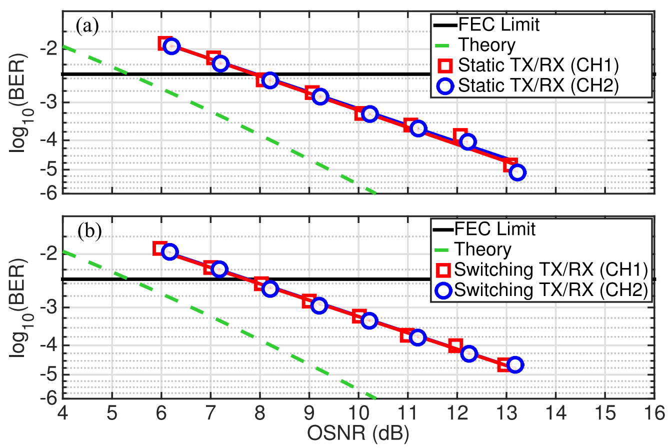


Fig. 7. BER performance for, (a) two static wavelength channels and (b) two dynamic burst channels.

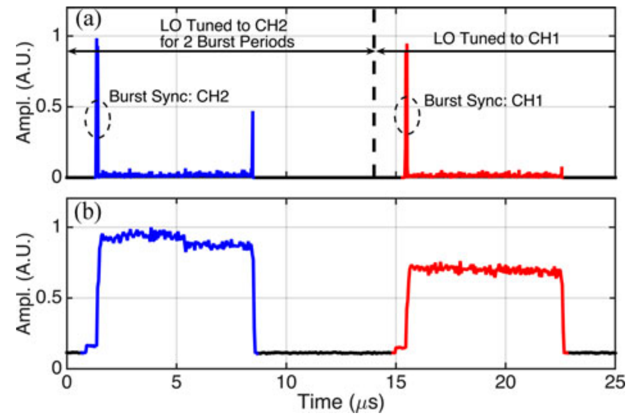


Fig. 8. (a) Synchronisation of two burst channels (dotted line illustrates when the DS-DBR LO laser switched from burst CH2 to CH1) and (b) corresponding burst amplitude.

The intermediate frequency variation of the beat note between the transmitted signal and the receiver LO was tracked per block using the digital FO estimator illustrated in Fig. 3, with the exception of the 4$^{\text{th}}$-order nonlinearity algorithm, as the dc pilot tone was unmodulated. The estimated FO was subsequently removed from each block, which down converted the OFDM signal to baseband, as seen in Fig. 6(b). The RF pilot tone was then filtered from the signal using a 140 MHz bandwidth low-pass filter, conjugated and multiplied with the signal block to compensate the phase variation, as seen in Fig. 6(c). The frame timing (synchronisation) was subsequently recovered using the Schmidl and Cox approach [14] and the CP was removed. The TS based channel estimation scheme proposed by Liu and Buchali [19] was used to estimate the channel, where the response was averaged over 20 symbols. A MIMO equaliser compensated the channel response and 16 data pilots were used to compensate any remaining phase error within each symbol.

## V. Results and Discussion

The back-to-back performance of the fast wavelength switching DP-OFDM transceiver is shown in Fig. 7. When the DS-DBR lasers in both the transmitter and the receiver were operated in static mode (not switching), an implementation penalty of 2.6 dB was experienced for both WDM channels relative to theory at the FEC threshold of $3.8 \times 10^{-3}$. No additional OSNR penalty was incurred when both the transmitter and LO DS-DBR lasers were dynamically switched between the two burst channels, as seen in Fig. 7(b). This back-to-back characterisation provided a base line level of performance for the OFDM transceiver and it is important to note that the dc pilot tone based phase noise compensation scheme was a pre-requisite in order to achieve a BER below the FEC threshold.

In order to investigate the performance of the transceiver in a network that exhibits a variable path history, the transmitter DS-DBR laser was sequentially switched between the two burst channels. The differential transmission distance between the two burst channels was set using the reconfigurable transmission line shown in Fig. 4. Burst CH2 ($\lambda_2$) was transmitted over a fixed distance of 200 km, while burst CH1 ($\lambda_1$) was transmitted over variable distances from 200 to 800 km. Therefore, the differential transmission distance between the two channels was varied from 0 to 600 km. The CP length of 30 symbols was sufficient for a transmission distance greater than 1000 km. At the receiver, the LO DS-DBR laser resided at each wavelength channel for two burst periods before switching. This was required for asynchronous detection, as the burst channels experienced different propagation delays through the network, therefore the LO waited at a desired wavelength channel until the burst was received. In a practical implementation of this system, a header containing the wavelength switching information would be transmitted in advance on a control channel to the receiver.

Burst detection was achieved by searching for the OFDM frame synchronisation symbols at the start of each burst. Fig. 8(a) illustrates the correlation of the Schmidl and Cox synchronisation symbols for the two channels and the corresponding burst amplitude is shown in Fig. 8(b). Once each optical burst was detected the remaining DSP blocks were applied, as described in the Receiver DSP section. The OSNR penalty at

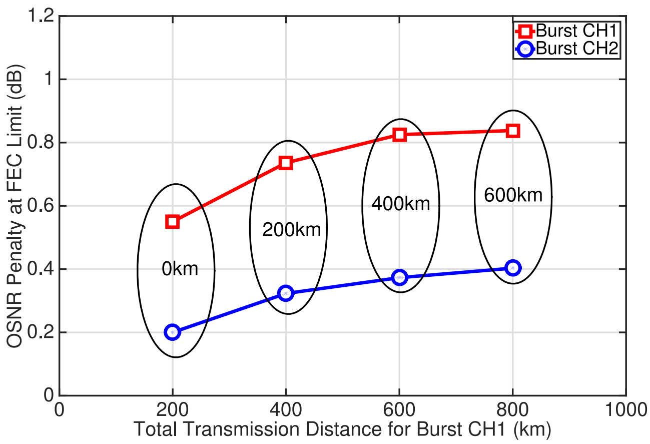


Fig. 9. OSNR penalty at the FEC threshold as a function of the total and differential transmission distance for both burst channels.

the BER threshold of $3.8 \times 10^{-3}$, relative to the B2B case (see Fig. 7(b)), for the 800 km OBS network is shown in Fig. 9. After both channels propagated over 200 km of SSMF (differential transmission distance of 0 km) a slightly larger OSNR penalty of ~0.6 dB was experienced for CH1 relative to the B2B performance, while CH2 incurred an additional penalty of only 0.2 dB. The 0.4 dB difference in required OSNR for both channels was attributed to the Polatis free-space optical switch, which was only in the transmission path for burst CH1.

However, as the differential transmission distance between both burst channels increased, the difference in required OSNR of 0.4 dB remained approximately constant. The slope of the OSNR penalty as a function of transmission distance was also similar for both burst channels. A constant penalty was expected for burst CH2 as it propagated over a constant transmission distance of 200 km, however accumulated ASE noise from burst CH1 was added to CH2 in the 3 dB coupler, which resulted in an increase in OSNR penalty as a function of the differential transmission distance. The maximum excess penalty for burst CH1 (relative to the B2B switching case) remained below 1 dB after transmission over 800 km of SMF. This penalty was caused due to a trade-off between low launch power to minimise fibre non-linearity and OSNR degradation due to the 100 km span loss. The OSNR penalty could be improved by either reducing the span length or by utilising ultra low loss fibre.

The excellent transmission performance of the fast wavelength switching DP-OFDM transceiver demonstrates that the CP inherent to the OFDM modulation format completely compensates the accumulated chromatic dispersion of each optical burst channel and demonstrates that fast dispersion estimation is not required for an OFDM based OBS network. In addition to this, the frame synchronisation symbols inherent to OFDM were employed for optical burst detection and dc pilot tone aided phase noise compensation was a key enabling compensation technique for the OFDM transceiver.

## VI. Conclusion

Optical packet or burst switching was originally proposed as a technology that could streamline both the control protocol and hardware for the next generation optical internet. It was viewed as a key technology to accommodate the bursty nature of IP traffic, while simultaneously providing an agile optical network that could effortlessly re-distribute system bandwidth [20]. However, the merits of OBS, such as dynamic reconfiguration and bandwidth on demand have not proved sufficient for this technology to become widely adopted. The main reason for this is that the time variability of traffic patterns within traditional networks typically fall into predictable or quasi synchronous categories [21].

The emergence of coherent optical networks also saw a revival in coherent optical burst switching, where optical bursts are routed based on wavelength. There have been several real-time demonstrations of coherent OBS receivers [22], [23], while commercially available TLs have been demonstrated in a fast wavelength switching coherent transceiver [24]. However, the inherent difficulties associated with implementing an OBS network that is capable of reconfiguring bandwidth on nanosecond timescales will prevent this technology from becoming a commercial reality within the foreseeable future and will therefore be confined to a research theme.

However, rapid reconfiguration using wavelength agile transceivers will become a corner stone of future software defined networks, especially for network resilience where reduced down time after a failure is extremely important. This will also extend virtualisation into the optical physical layer, which will provide the functionality for dynamic bandwidth provisioning, capable of accommodating the next generation of bandwidth intensive applications.

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