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Special Issue on Transparent Optical Networking

Guest Editorial

The development of the transparent optical-fiber based transmission and networking have revolutionized modern telecommunications and consequently have influenced dramatically the working methods and everyday life of the societies and individuals. The excellent progress with that is reviewed and reported in this Special Issue on Transparent Optical Networking which contains fifteen carefully selected papers based on the contributions to the 10th Anniversary International Conference on Transparent Optical Networks ICTON 2008, Athens, Greece, June 22-26, 2008, www.irtl.waw.pl/icton. ICTON is receiving technical co-sponsorship by the IEEE Photonics Society (formerly LEOS).

All papers in this Special Issue have been submitted by Editors’ invitations.

The first paper entitled Control and Management Issues in All-Optical Networks by R. Rejeb, M.S. Leeson, C. M. Machuca, and I. Tomkos, introduces a management system for ensuring the secure and continued functioning of the network.

The second paper on Optical Transmission of OFDM Ultra-wideband Signals beyond 40 Gb/s by Y. Ben-Ezra, M. Ran, B.I. Lembrikov, U. Mahlab, M. Haridim, A. Leibovich, proposes for the first time the highly efficient method of RF and optical signal mixing based on two different architectures: the parallel-RF/serial-optics architecture characterized by all-optical mixing for sub-carrier multiplexing, and the parallel-RF/parallel-optics architecture based on the array of 12x10 GHz components with directly modulated VCSELs and 12 multimode optical fibers.

The third paper on Characterization of Wavelength Tunable Lasers for Future Optical Communication Systems by P. M. Anandarajah, A. Kaszubowska-Anandarajah, R. Maher, K. Shi and L. P. Barry, investigates the possibility of using tunable lasers (TLs) in DWDM wavelength packet switched networks by focusing on the characterisation of the instantaneous frequency drift of a TL due to wavelength tuning and direct modulation. Characterization of the line-width of the TLs is presented to verify the feasibility of using TLs in systems employing advanced modulation formats.

The fourth paper on Signal Processing Algorithms in 100Gbit/s Optical Coherent and Non-coherent Receivers with PSK Modulation by Ch. Hebebrand and W. Rosenkranz, addresses multi-level PSK modulation formats with coherent and non-coherent detection, and reviews signal processing algorithms like mitigation of phase noise, the clock and carrier recovery algorithms as well as the equalizer structures and their performance.

The fifth paper entitled A New All-Optical Switching Node Including Virtual Memory and Synchronizer by S. Batti, M. Zghal, and N. Boudriga, presents an architecture for an all optical switching node suitable for optical packet and optical burst switching which provides appropriate contention resolution schemes and QoS guarantees. A new concept of a ‘virtual memory’ is developed to allow controllable and reasonable periods for delaying optical traffic. Implementation issues are discussed, and the implementation and performance of an all-optical synchronizer able to synchronize arriving data units to be aligned on the clock signal associated with the beginning time of slots in the node with an acceptable error is reported.

The sixth paper entitled Electronic Predistortion for Compensation of Fiber Transmission Impairments - Theory and Complexity Considerations by S. Hellerbrand and N. Hanik, revises principles of electronic pre-distortion technique to combat transmission impairments on fiber-optic communication links, and different methods to implement this method are explained. The Authors focus on the implementation complexity of this scheme and they highlight how the complexity scales for increasing channel data rates.

The seventh paper on Photonic Routing Systems Using All-optical, Hybrid Integrated Wavelength Converter Arrays by L. Stamoulidis, E. Kehayas, P. Bakopoulos, D. Apostolopoulos, P.Zakynthinos, D. Petrantanakis, H. Avramopoulos, A. Poustie, and G. Maxwell, presents the realization of demanding functionalities required in high-capacity photonic routers using compact and multi-functional photonic processing components including: clock recovery, data/label recovery, wavelength routing and contention resolution; all implemented with multi-signal processing using a single photonic chip – a quadruple array of SOA - MZI wavelength converters with a chip area of only 15 x 58 mm². The capability of this technology to build WDM signal processing systems with the simultaneous operation in a four wavelength burst-mode regenerator is presented, and the potential to provide photonic on-chip systems is demonstrated with the first hybrid integrated all-optical burst-mode receiver prototype.

The eighth paper on Efficient Signal Processing in OFDM-based Indoor Optical Wireless Links by J. Grubor and K.-D. Langer, proposes a rate-adaptive optical wireless transmission system based on orthogonal frequency division multiplexing for indoor communications. The Authors show that a dynamically adaptive system can greatly enhance performance when compared to static operation, and they demonstrate how a loading algorithm, which optimally performs in power-limited systems, needs to be adjusted if the specific terms of the optical wireless channel are to be rigorously obeyed. Finally it is demonstrated that the transmission rate can be significantly improved even further by accepting a minor increase in the error rate as a result of controlled clipping, and the results are compared with the upper system capacity limit.

The ninth paper entitled Characterisation of the PMD Distribution along Optical Fibres and Improvement of the Backbone Fibre Infrastructure by a POTDR by A. Ehrhardt, D. Fritzche, M. Paul, L. Schuerer, D. Breuer, W. © 2010 ACADEMY PUBLISHER doi:10.4304/jnw.5.2.129-131
Weiershausen, V. Fuerst, N. Cyr, H. Chen, and G.W. Schinn, proposes the use of a new random-scrambling polarization optical time domain reflectometry (POTDR) measurement technique to investigate the spatial distribution of the cumulative PMD in deployed fibres. Results help to identify high-PMD fibre pieces or sections which need to be replaced to enable 40 Gbit/s transmission and beyond, rather than substitution of a whole fibre link. Techno-economical investigations show the high economic potential of this method leading to significant reduction of expenses for infrastructure improvements, thus enabling high data rates beyond limitation given by PMD.

The tenth paper on System Impact of Cascaded All-optical Wavelength Conversion of D(Q)PSK Signals in Transparent Optical Networks by R. Elschner, C.-A. Bunge, and K. Petermann, compares techniques for all-optical wavelength conversion of differentially phase-modulated signals using four-wave mixing and super-continuum generation. For the super-continuum generation, a relation between the conversion efficiency and the nonlinear phase distortion is derived and it is shown that this technique is not suitable for the conversion of phase-modulated signals. The suppression of Brillouin scattering and its impact on phase-distortions is discussed, and a detailed discussion of its cascadability in transparent optical networks concludes the paper.

The eleventh paper, On local CAC Schemes for Scalability of High-speed Networks by J. Aracil, J.A. Hernández, A.J. Elizondo, R. Duque, and O. Gonzalez de Dios, investigates local CAC (Connection Admission Control) schemes where the admission decisions are performed at the network edges, based on pre-calculated admission quotas, as opposite to centralized CAC approaches that could suffer from scalability problems if the number of requests for connections is excessive.

The twelfth paper on Storage and Mirroring in Single and Dual Section Metro WDM Rings under Different Traffic Scenarios by T. E.H. El-Gorashi and J. M. H. Elmighani, introduces a novel data mirroring technique for Storage Area Networks (SANs) in a metropolitan wavelength division multiplexing (WDM) ring scenario. Performance is evaluated under two different slot schemes accommodating variable size packet traffic: variable-size (VS) and super-size (SS) slot schemes.

The thirteenth paper on Optical Code Processing System, Device, and its Application by N. Wada, reports on the recent progress of optical code processing technology. Ultra-high speed time domain, spectral domain, hybrid domain, and multiple optical code processing devices and systems are shown. OCDMA-PON, OPS network, and ultra high-speed optical clock generation are demonstrated.

The fourteenth paper on SLA-Aware Survivability by H. Waldman and D.A.A. Mello, discusses the provision of differentiated guarantees to a population of users who share a network with different requirements for their connections. The basic concept underlying the proposed solutions is the required availability of the connections, both in the long term and during the period covered by the Service Level Agreement. An adequate metric for the latter is provided by the interval availability. The Authors discuss how Markov chains may be used to model interval availability during the SLA period.

Guest Editors:

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Marian Marciniak has graduated in solid state physics from Marie Curie-Sklodowska University in Lublin, Poland, in 1977. He received a Ph.D. degree (with distinction) in optoelectronics from Military University of Technology in Warsaw in 1989, and a Doctor of Sciences degree in physics/optics from Warsaw University of Technology, Institute of Physics, in 1997.

In 1996 he joined the National Institute ofTelecommunications in Warsaw where he actually leads the Department of Transmission and Optical Technologies. Since 2008 he is a professor at Kielce University of Technology, Poland. Since 2004 he serves as an Honorary International Advisor to the George Green Institute for Electromagnetics Research at the University of Nottingham, UK.

His early activities included extended studies of optical waveguiding linear and nonlinear phenomena with focus on developing original analytic and numerical methods including beam-propagation methods. He has formulated a theory for mode cut-off conditions in lossy waveguides. He has performed advanced studies of radiation field propagation in optical waveguides and he has formulated conditions for efficient radiation-to-guided mode conversion. He has studied extensively the propagation of light in optical waveguides containing simultaneously amplifying and lossy layers and he was
the first to point out the oscillatory or exponential field behaviour in such waveguides depending on the absolute value of gain and loss. Actual research interests include transparent and all-optical packet- and burst-switched networks, ultrafast and nanoscale/subwavelength photonics, ubiquitous networking & computing, and the future global optical & wireless networking and services. Recently he has introduced and developed a concept of a hybrid photonic real-time service & packet network. He is an author of over 300 technical publications including a number of invited conference presentations and over 20 books authored, co-authored and/or edited by himself.

He is widely involved in the European research on optical communication networks. He serves as the Chair of the Management Committee of COST Action MP0702 Towards functional sub-wavelength photonic structures (2008-2012).

He is a Senior Member of the IEEE - Lasers & Electro-Optics, Communications, and Computer Societies, a member of The American Association for the Advancement of Science, The Optical Society of America, SPIE - The International Society for Optical Engineering and The American Chemical Society. In early 2001 he originated the IEEE/LEOS Poland Chapter and he served as the Chairman of the Chapter until July 2003. As a LEOS volunteer he is the originator and Co-Chair, together with Prof. Trevor M Benson from University of Nottingham, UK, of a Nonlinear Dynamics in Photonic Systems topic for IEEE/LEOS Winter Topicals 2009, January 2009 in Innsbruck, Austria.

He is a Delegate to the International Telecommunication Union, Study Group 15 Optical and Other Transport Network Infrastructures, and to the International Electrotechnical Commission, Technical Committee 86 Fibre Optics.

He is a Deputy Editor for Optical and Quantum Electronics from 2008. He is a Member of the Editorial Board of Microwave & Optoelectronics Technology Letters journal, Wiley, USA, and the Journal of Telecommunications and Information Technology, Poland. He has been a Guest Editor of Special Issues of journals: Journal of Telecommunications and Information Technology, Applications of Nonlinear Optical Phenomena (2000), Optical Switching and Networking (2009), Journal of Networks, Transparent Optical Networking (2009), IEEE Journal of Quantum Electronics, a Feature Section on the Nonlinear Dynamics in Photonic Systems (2009).

In 1999 Prof. Marciniak originated the International Conference on Transparent Optical Networks ICTON http://www.itl.waw.pl/icton series.

Dr. Ioannis Tomkos, has the rank of Full Professor at Athens Information Technology Center, where he serves as its Associate Dean (since 2004) and is an Adjunct Faculty at the Information Networking Institute of Carnegie-Mellon University, USA. In the past (1999 - 2002) he held a senior scientist position at Corning Inc. USA. He joined AIT in 2002 where he founded and serves as the Head of the “High Speed Networks and Optical Communication (NOC)” Research Group that was/is involved in many EU funded research projects (including 5 running FP7 projects) within which Dr. Tomkos is representing AIT as Principal Investigator and has a consortium-wide leading role (e.g. Project Leader of the EU ICT STREP project DICONET, Technical Manager of the EU IST STREP project TRIUMPH, Chairman of the EU COST 291 project, WP leader of many other projects).

Dr. Tomkos has received the prestigious title of “Distinguished Lecturer” of IEEE Communications Society for the topic of transparent optical networking. Together with his colleagues and students he has authored about 270 peer-reviewed archival articles (over 140 IEEE sponsored), including about 75 Journal publications. Dr. Tomkos has served as the Chair of the International Optical Networking Technical Committee of IEEE Communications Society and the Chairman of the IFIP working group on “Photonic Networking”. He is currently the Chairman of the OSA Technical Group on Optical Communications. He has been General Chair, Technical Program Chair, Subcommittee Chair, Symposium Chair or/and member of the steering/organizing committees for the major conferences (e.g. OFC, ECOC, IEEE GlobeCom, IEEE ICC, ONDM, etc.) in the area of telecommunications/networking (more than 50 conferences/workshops). In addition he is a member of the Editorial Boards of the IEEE/OSA Journal of Lightwave Technology, the OSA Journal of Optical Networking, the IET Journal on Optoelectronics, and the International Journal on Telecommunications Management.
Control and Management Issues in All-Optical Networks

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Abstract—As more intelligence and control mechanisms are added into optical networks, the need for the deployment of a reliable and secure management system using efficient control techniques has become increasingly relevant. While some of available control and management methods are applicable to different types of network architectures, many of them are not adequate for all-optical networks. These emerging transparent optical networks have particularly unique features and requirements in terms of security and quality of service thus requiring a much more targeted approach in terms of network management. In particular, the peculiar behavior of all-optical components and architectures bring forth a new set of challenges for network security. In this article, we briefly overview security and management issues that arise in all-optical networks. We then discuss the key management functions that are responsible for ensuring the secure and continued functioning of the network. Consequently, we present a framework for the realization of an appropriate management system that can meet the challenges posed by all-optical networks.

Index Terms—all-optical networks, optical network security, fault and attack management.

I. INTRODUCTION

All-Optical Networks (AONs) are emerging as a promising technology for very high data rates, flexible switching and broadband application support. Specifically, they provide transparency capabilities and new features allowing routing and switching of traffic without any regression or modification of signals within the network. Although AONs offer many advantages for high data rate communications, they have unique features and requirements in terms of security and management that distinguish them from traditional communication networks. In particular, the unique characteristics of AON components and network architectures bring forth a set of new challenges for network security. By their nature, AON components are particularly vulnerable to various forms of denial of service, Quality of Services (QoS) degradation, and eavesdropping attacks. Since even short (in terms of duration) faults and attacks can cause large amounts of data to be lost, the need for securing and protecting optical networks has become increasingly significant [1-3].

In emerging AONs, efficient monitoring and estimation of signal quality along a lightpath¹ are of highest interest because of their importance in diagnosing and assessing the overall health of the network. One of the main reasons for continuous monitoring and controlling of the proper functioning of AON components is related to the fact that transmission in AONs is limited by a number of effects such as optical crosstalk, amplified spontaneous emission noise, laser saturation, fiber nonlinearities, reflections, jitter accumulation, and signal bandwidth narrowing caused by filter concatenation. Despite a careful design of AONs, enabling pre-estimation of the worst possible impact on QoS degradations and the provision of sufficient margins to tolerate it, a resourceful attacker or nefarious user can exploit these transmission limiting effects to perform disruptive attacks upon the whole network. Another important characteristic of AONs is that the impact of subtle forms of attacks and miscellaneous transmission impairments accumulate while traveling through the network and can thus degrade the signal quality enough to reduce the QoS without precluding all network services. Such disruptive faults or attacks are totally

¹ A lightpath is defined as an end-to-end optical connection between a source and a destination all-optical node.
beyond control and is quite difficult to be detected by monitoring the network equipment.

In this article, Section II and III give a brief overview on the security and management issues that may arise in AONs from their employed components, respectively. Section IV presents the key management functions that are responsible for ensuring the secure and continued functioning of the network. In Section V we propose a framework for the realization of an appropriate Network Management System (NMS) that can meet the challenges posed by AONs. Section VI introduces several control plane architectures taking into consideration still open and unsolved development issues. Finally, we present the conclusions of this article.

II. SECURITY ISSUES

In the context of this work, an attack is defined as an intentional action against the ideal and secure functioning of the network, whereas a fault is defined as an unintentional action against the ideal and secure functioning of the network. Failures are referred to as the faults and attacks that can interrupt the ideal functioning of the network [2].

Security attacks upon AONs may range from a simple physical access to more complex attacks exploiting: a) the peculiar behaviors of optical fibers, b) the unique characteristics of AON components, and c) the shortcomings of available supervisory techniques and monitoring methods. Attacks can be classified as eavesdropping or service disruption. In this scenario, they are different in nature ranging from malicious users (i.e., users inserting higher signal power) to eavesdroppers. Thus, attacks differ from conventional faults and should be therefore treated differently. This is because they appear and disappear sporadically and can be launched elsewhere in the network. In particular, the attacker may thwart simple detection methods, which are in general not sensitive enough to detect small and sporadic performance degradations. Furthermore a disruptive attack, which is erroneously identified as a component failure, can spread rapidly through the network causing additional failures and triggering multiple erroneous alarms. Security attacks therefore must be detected and identified at any node in the network where they may occur. Moreover, the speed of attack detection and localization must be commensurate with the data transmission rate [3].

Furthermore, transparency in AONs may introduce significant miscellaneous transmission impairments such as optical crosstalk, amplified spontaneous emission noise, and power divergence [4]. In AONs, those impairments accumulate as they propagate and can impact the signal quality so that the received bit error rate at the destination node might become unacceptable high.

III. MANAGEMENT ISSUES

Network management is an indispensable constituent of communication systems since it is responsible for ensuring the secure and continued functioning of any network. Specifically, a network management system should be capable of handling the configuration, fault, performance, security, and safety in the network.

Network management for AONs faces additional security and management challenges. One of the main premises of AONs is the establishment of a robust and flexible control plane for managing network resources, provisioning lightpaths, and maintaining them across multiple control domains. Such a control plane must have the ability to select lightpaths for requested end-to-end connections, assign wavelengths to these lightpaths, and configure the appropriate resources in the network. Furthermore, it should be able to provide updates for link state information to reflect which wavelengths are currently being used on which fiber links so that routers and switches may make updated routing decisions. An important issue that arises in this regard is how to address the trade-off between service quality and resource utilization. Addressing this issue requires different scheduling and sharing mechanisms to maximize resource utilization while ensuring adequate QoS guarantees. One possible solution is the aggregation of traffic flows to maximize the optical throughput and to reduce operational and capital costs, taking into account qualities of optical transmission in addition to protection and restoration schemes to ensure adequate service differentiation and QoS assurance. A control plane should therefore offer dynamic provisioning and accurate performance monitoring, plus efficient restoration in the network and most of these functions need to move to the optical domain. Connection provisioning, for example, should enable a fast automatic setup and teardown of lightpaths across the network thereby allowing dynamic reconfiguration of traffic patterns without conversion to the electrical domain.

Another related issue arises from the fact that the implementation of a control plane requires information exchange between the control and management entities involved in the control process. To achieve this, fast signaling channels need to be in place between switching nodes. These channels might be used to exchange up-to-date control information that is needed for managing all supported connections and performing other control functions. In general, control channels can be realized in different ways; one might be implemented in-band while another may be implemented out-of-band. There are, however, compelling reasons for decoupling control channels from their associated data links. An important reason for this is that data traffic carried in the optical domain is transparently switched to increase the efficiency of the network and there is thus no need for switching nodes to have any understanding of the protocol stacks used for handling the control information. Another reason is that there may not be any active channels available while the data links are still in use, for example when bringing one or more control channels down gracefully for maintenance purposes. From a management point of view, it is unacceptable to teardown a data traffic link, simply because the control channel is no longer available. Moreover, between a pair of
switching nodes there may be multiple data links and it is therefore more efficient to manage these as a bundle using a single separated out-of-band control channel.

In recent years, the notion of an optical control plane has received extensive attention and has rapidly developed to a detailed set of protocol standards, currently being standardized by the International Telecommunication Union—Telecommunication Standardization Sector (ITU-T) and others [5]. Nevertheless, several additional issues in terms of security and network management are still unsolved [3]. One of the main management issues revolves around the fact that optical performance monitoring techniques for AONs are not a well developed and mature technology. As a result, performance and impairment information cannot be used in order to ensure better QoS. For example, when discussing routing in AONs, it is usually assumed that all routes have adequate signal quality (ensured by limiting AONs to sub-networks of limited size). This approach is very practical and has been applied to date when determining the maximum length of optical links and spans, for example. Specifically, operational considerations such as failure isolation also make limiting the size of domains of transparency very attractive. Another example concerns the Routing and Wavelength assignment (RWA) problem, where network blocking has traditionally been estimated using analytical and simulation approaches without taking transmission impairments into consideration [4]. Efficient connection provisioning, however, requires more than simply advertising the availability of wavelengths and routes to switching nodes. Hence, RWA approaches require additional control information to be taken into consideration including the characteristics and performance measurements of established lightpaths in the network. Thus, the combination of both RWA and performance information may offer the prospect of improvements in the provisioning of lightpaths in AONs [6]. However, dealing with this challenge necessitates improving available RWA mechanisms to update and advertise the performance measurements needed for assigning efficient routes and pre-computing their restoration paths.

IV. NETWORK SURVIVABILITY

A crucial feature of any communication network is its survivability which refers to the ability to withstand component failures and to continue providing services in disruption conditions. Providing resilience against failures is therefore an important requirement for many high-speed networks. As these networks carry more and more data, the amount of disruption caused by a network fault or attack becomes more and more significant.

Protection switching is the key technique used to ensure network survivability. Protected connections require a secondary path so that when a fault degenerates enough the primary path, the information can be transmitted through the secondary path (which can be reserved or used by preemptive traffic depending on the protection scheme). Another used technique requiring less network resources and redundant capacity is restoration. In this approach, the secondary path is computed when the fault occurs and therefore, although it may be slower than protection, it adapts better to the available capacity that the network has in that moment. Another important advantage of restoration is that it can handle multiple simultaneous failures; whereas protection techniques are designed for handling a preset number of failures; typically single fiber failures. Both techniques are usually implemented in a distributed manner to ensure faster restoration of services after the occurrence of single (or multiple) fault(s). Since many data-centric services may not require hard guarantees on the recovery time, restoration techniques are more suitable than protection techniques.

From a management viewpoint, network survivability involves primary the protection of secure data. As shown in Fig. 1, network survivability can be broadly divided in two categories, namely, fault survivability and attack survivability.
survivability. The main goals of fault survivability are to provision lightpaths in anticipation of faults, locate the faults, and to restore the affected connections. Attack survivability can be, in turn, subdivided into two separate but still interrelated types: Physical security and semantic security. The former promises to ensure integrity and privacy of information, as well as the QoS by protecting the network against service disruption and service degradation. The latter focuses on the protection of information even when an attacker has access to the transmission data channel.

Fault and attack management is one of the crucial functions and a prerequisite for the above mentioned protection and restoration schemes. In this regard, there are many complications that prevent the continuous control and monitoring of all supported lightpaths simultaneously. An important implication of using AON components in communication systems is that available methods used to manage and monitor the health of the network may no longer be appropriate. Therefore, without additional control mechanisms, an attack upon the core network might not be detectable. In an AON, a lightpath will be capable of carrying analogue and digital signals with arbitrary bit rates and protocol formats. Such a lightpath typically traverses multiple data links and there are many components along its route that can fail. Since a transmission fiber carries a number of high-bandwidth channels, its failure will cause simultaneous multiple channel failures, resulting in large amounts of data being corrupted or compromised. Since AON components will not by design be able to comprehend signal modulation and coding, intermediate switching nodes are unable to regenerate data, making segment-by-segment testing of communication links more demanding. As a direct consequence, failure detection and localization using existing integrity test methods is made particularly difficult.

In recent years, several studies related with the management of faults and attacks in AONs have been reported [7-12]. Worthy of mentioning is a series of management methods proposed in [3, 7]. In particular, the authors have presented an extensive analysis of AON vulnerabilities to security attacks. As countermeasures, they have proposed new methods for detecting, localizing, and identifying of attacks that may be practiced upon AONs.

Another approach [2, 8] proposes an algorithm that solves the multiple failure location problem in transparent optical networks where the failures are more deleterious and affect longer distances. The proposed solution also covers the non-ideal scenario, where lost and/or false alarms may exist. Although the problem of locating multiple faults has been shown to be NP-complete, even in the ideal scenario where no lost or false alarms exist, the proposed algorithm keeps most of its complexity in a pre-computational phase. Hence, the algorithm only deals with traversing a binary tree when alarms are issued. This algorithm locates the failures based on received alarms and the failure propagation properties, which differ with the type of failure and the kind of network equipment used in the network.

Another algorithm has been proposed in [1, 9] to correlate multiple security failures locally at any node and to discover their tracks through the network. The algorithm is distributed and relies on a reliable management system since its overall success depends upon correct message passing and processing at the local nodes. To identify the source and nature of detected performance degradation, the algorithm requires up-to-date connection and monitoring information of any established lightpath, on the input and output side of each node in the network. This algorithm mainly runs a generic localization procedure, which will be initiated at the downstream node that first detects serious performance degradation at an arbitrary lightpath on its output side. Once the origins of the detected failures have been localized, the NMS can then make accurate decisions (for example, which offender lightpaths should be disconnected or rerouted) to achieve finer grained recovery switching actions. However, most of these works are still in progress and need to be embedded and tested within standardized control planes.

V. MANAGEMENT FRAMEWORK

Given the complexity of network management and diversity requirements, a valid question to ask is whether a centralized NMS would be better than decentralized management methods. Although most of available management systems are implemented in a centralized manner, this manner of implementation is rather slow and inadequate for optical networks because of the large
communication overhead involved in the process. Furthermore, it becomes too difficult for a single centralized NMS to manage the entire network when this becomes very large. In contrast, decentralized methods are usually much faster and more efficient than a centralized NMS, even in small networks with only a few nodes. This is not only because distributed methods involve low communication overheads, but also because certain management functions require rapid and accurate actions to be taken in response to failed conditions. For example, if rapid rerouting is required, it is not possible to rely on a centralized management system to compute a recovery route for each failed lightpath on demand after a failure has been detected. Distributed systems would clearly not have this difficulty.

Network management is performed in a hierarchical manner involving multiple management systems, which are usually implemented in a master-slave relationship between managers and associated agents. The key concept applied in this manner of implementation is that of a layering architecture, illustrated in Fig. 2a. Layering refers to the ability of a network to nest finer-granularity over coarser-granularity by hiding details and complexities in different management levels. Thus, these levels would advertise only a standardized abstraction of their state information. In such a layered model, the management functions with associated information can be decomposed into several layers of abstractions. At the border between two adjacent layers, for example Layer n-1 and Layer n, the management view is presented to Layer n-1. It is presented in the form of management information that is contained within the agent at Layer n. For security and other reasons, the management view that is presented to a higher layer need not unveil all details of a lower layer. An agent at an arbitrary layer is then responsible for providing the management information that is needed at the immediately higher layer. Since the principle of layering can be applied in a recursive fashion, the management view of Layer n+1 can be presented to Layer n, and so on.

In such a hierarchical model, the individual components to be managed are referred to as Network Elements (NEs). For AONs, these include Optical Line Terminals (OLTs), Optical Add/Drop Multiplexers (OADM), optical amplifiers and Optical-Cross-Connects (OXC). Each NE is managed by an Element Management System (EMS). The NE is represented by an information model that contains a variety of attributes associated with it and a set of control and measurement parameters for monitoring purposes. Typically, a NE consists of built-in agents that are responsible for performing control functions such as monitoring and reporting any performance degradation. As shown in Fig. 2b, an EMS is usually connected to one or more agents and communicates with neighboring EMSs using separated control channels. Hence, the EMS has a view only of the NEs that it can manage and may not have a comprehensive view of the entire network.

In this model, a lightpath is considered as a sequence of detection points. An agent therefore, may be able to manage one or more detection points, simultaneously. For example, an OXC node, which is composed of several active and passive optical components, can be considered as a transmission segment or a subset of several detection points that may be managed by one or more agents at the same time. Accordingly, the corresponding EMS may communicate with certain agents to examine or alter some information, without the difficulty of handling each detection point individually. It will in fact need to collect a view of performance information, which is only part of the information illuminated at each detection point. In short, an agent may act as a proxy for certain EMSs hosted elsewhere which only need to contact the agent to get the information they need. In conclusion, additional management functions need to be implemented at the agent and EMS components. Special attention needs to be paid by analyzing and specifying interfaces, protocols and the control information that should be exchanged between EMS-EMS and EMS-Agent pairs. Consequently, the vulnerability analysis of specified interfaces and protocol stacks from a security perspective is an important prerequisite. To do so, it is necessary to generate a detailed picture of the state information of transmission links in a manner that can be communicated to EMSs and other control entities involved in the management process. To satisfy this requirement, control information should be expanded to include additional performance measurements, which will be carried and shared among EMSs. In particular, this needs a comprehensive synthesis of optical attributes to distinguish the parameters that change frequently, requiring continuous exchanges between EMSs, from those that change infrequently and often depend on the configuration properties of the individual NEs.

VI. CONTROL PLANE ARCHITECTURES

The design of an optical network is an important and very practical issue. As stated above, a desirable architecture should feature, inter alia, flexible management, automatic lightpath protection and restoration, and the ability to compile an inventory. Moreover, network architectures should support the gradual introduction of new technologies into the network without time consuming and costly changes to embedded technologies. However, network architectures currently used may be categorized in two main models, namely the overlay model and the peer model. Although both models consist essentially of an optical core that provides wavelength services to client interfaces, which reside at the edges of the network, they are intrinsically different and offer up two key concepts for managing traffic flows in the network.

The overlay model hides the internals of the optical network and thus requires two separate, yet interoperable, control mechanisms for provisioning and managing optical services in the network. One mechanism operates within the core optical network and the other acts as interface between the core and edge components which support lightpaths that are either dynamically signaled across the core optical network or statically provisioned.
without seeing inside the core’s topology. The overlay model therefore imposes additionally control boundaries between the core and edge by effectively hiding the contents of the core network.

The peer model considers the network as a single domain, opening the internals of the core optical network to the edge components making the internal topology visible able to participate in provisioning and routing decisions. Whilst this has the advantage of providing a unified control plane, there are some significant considerations:

- The availability of topological information to all components involved makes this model less secure.
- New standard control mechanisms are required since available proprietary ones cannot be employed.
- Additionally approaches for traffic protection and restoration are required.

Another model, known as the hybrid model, combines both the overlay and peer approaches, taking advantages from both models and providing more flexibility. In this model, some edge components serve as peers to the core network and share the same instance of a common control mechanism with the core network through the Network-Network Interface (NNI). Other edge components could have their own control plane (or a separate instance of the control plane used by the core network), and interface with the core network through the User-Network Interface (UNI).

From a control plane point of view, the notion of control domain is very useful. A large optical network, as shown in Fig. 3, may be portioned into moderate control domains mainly for the following reasons [5]:

- To enforce administrative, management and protocol boundaries making them reliable enough.
- To ensure rapid and accurate actions to be taken in response to failed conditions (e.g. performing failure localizing processes in commensurate time).
- To increase the scalability of management functions and control planes.

The management information in a typical domain, as shown in Fig. 4, is distributed among domain’s EMSs where each one has only a partial knowledge of the whole domain control and management information. However, there are still open and unsolved problems in the development of secure AONs that should be carefully addressed. One particular security issue is related to the
interfaces UNI and NNI within the control plane employed. Consequently, the analysis of protocol stacks from a security perspective is an important prerequisite. Another issue related to network protection is a comparative study of the trade-off between network complexity and traffic restoration time.

VII. CONCLUSION

In this article, we have briefly reviewed management issues in AONs with particular emphasis on complications that arise from the unique characteristics of AON components employed in communication systems. We have also discussed the key management functions that are responsible for ensuring the secure and continued operation of the network. Accordingly, we have presented a management framework for the realization of an appropriate management system that could meet the challenges posed by AONs.

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Optical Transmission of OFDM Ultra-wideband Signals beyond 40 Gb/s

(Invited Paper)

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Abstract—We for the first time propose the highly efficient method of RF and optical signal mixing based on two different architectures: the parallel-RF/serial-optics architecture characterized by all-optical mixing for sub-carrier multiplexing, and the parallel-RF/parallel-optics architecture based on the array of 12x10 GHz components with directly modulated VCSELs and 12 multimode optical fibers. The main advantages of both architectures are simplicity and low-cost implementation. We have carried out numerical simulations of ultra-wideband signals propagation in the proposed systems and proved the high efficiency and feasibility of the proposed method.

Index Terms—100 GbE, OFDM, ultra-wideband signals.

I. INTRODUCTION

In optical subcarrier multiplexed (SCM) systems, several channels are multiplexed around an optical carrier using frequency division multiplying (FDM) technique [1], [2]. Each channel can be digitally intensity modulated into a bit-stream or coded using a multilevel modulation scheme at the radio frequency (RF) domain, enabling an increase of the channel information capacity. The frequency multiplexed signal is then converted to an analog optical signal using a very wide optical modulator. In order to correctly transpose the frequency multiplexed signals, a high degree of linearity and a very wide RF bandwidth are required. One of the main drawbacks of SCM technique is the RF fading effect coming from interaction between the RF subcarrier and the chromatic dispersion of the fiber. During the propagation of the double side band signal through a dispersive optical fiber, the upper and the lower side bands will undergo different phase shifts due to the different group velocities. The squared photodetection applied to this double side band signal will exhibit the RF fading effect at determined fiber distances. Dispersion induced RF fading effect is significant in optical links without dispersion compensation. In order to improve the spectral efficiency and to reduce the chromatic dispersion penalties, two subcarrier modulation schemes can be used: single-side band modulation (SSB) and tandem single side band modulation (TSSB) [3]. These two techniques can be realized by using a dual drive Mach-Zehnder modulator (MZM) [3]. However, MZM induces nonlinear signal distortions.

In this paper, we investigated theoretically and experimentally the three aspects of Orthogonal Frequency Division Multiplexing (OFDM) technology:

1) the parallel and serial architecture;
2) the optical link;
3) a novel component for the UWB MB OFDM signal detection.

OFDM technology is a key building block for mitigation of the non-linear signal distortion. Parallel and serial architectures are explored in order to construct multi-band (MB) OFDM signals capable of delivering a multi-gigabit analog signal. The basic element of the both proposed architectures is the optical link consisted of a directly modulated vertical cavity surface emitting laser (VCSEL), multimode optical fiber (MMF) and a p-i-n photodiode (PD). Transmission performances of such a system are analyzed theoretically. We have carried out the numerical simulations of UWB MB OFDM signal transmitted over this optical link. The measurements and simulations have been carried out for the time frequency code 5 (TFC5), TFC6 and TFC7 bands of MB OFDM UWB signals determined by the frequency intervals (3168 ÷ 3696) MHz, z, (3696 ÷ 4224) MHz, and (4224 ÷ 4752) MHz, respectively [2]. We address beyond 40Gb/s data rates by parallel transmission over more than 128 conventional WiMedia/ECMA [2] baseband channels, each having 528 MHz. One of the key advantages of the proposed approach is the ability to provide hybrid fiber-wireless solution, where the wireless segment at the available ultra-wideband (UWB) transmission is fully compliant with UWB regulations. We proposed a novel low-cost silicon chip for the UWB MB OFDM signal generation using a novel SiGe technology as the key building block for the proposed system. We have investigated the Ultra-wideband (UWB) MB OFDM signal generation using a novel SiGe technology as the key building block for the proposed system.
based on SiGe/Si structure is analyzed theoretically. The numerical simulation results for MB OFDM UWB signal transmission are presented in Section 5. The experimental results are discussed in Section 6. Conclusions are presented in Section 7. The details of mathematical derivations are presented in Appendix.

II. TYPES OF ARCHITECTURE

A. Parallel RF/Serial Optic Architecture

A novel concept for a scalable radio-over-fiber (ROF) system enable to bring up to 61.44Gb/s is shown in Fig. 1. The system is scalable in such a way that it enables the various channels and bands. The development of novel O/E and E/O components and subsystems for the extended band UWB signal transmission over the fiber is necessary. For instance, photo detection up to 64GHz may be achieved through the lateral illumination and resonant-cavity-enhancement of SiGe heterojunction phototransistor (HPT). Additionally, an ultra-wideband and highly linear E/O modulator is needed for the implementation of the proposed architecture. A single-mode fiber (SMF) can be used in long-haul applications.

B. Beyond 40 Gb/s Parallel RF/Parallel Optic Architecture

In the alternative scheme of the parallel RF/parallel optics architecture shown in Fig. 2 each directly modulated low-cost multimode VCSEL with a 10Gb/s bandwidth transmits its signal over a separate MMF. This architecture based on 12x10 GHz transceiver for digital 100GbE was proposed in [4]. We enhanced this architecture for ROF applications. In contrast to the parallel RF/serial optics architecture with SMF suitable for long-haul applications, this version based on MMF is appropriate for short-range applications. The parallel RF/parallel optics architecture is expected to operate at the wavelength of 850nm over MMF of the length of about several hundred of meters. The input lanes are directly connected to 12 Laser Drivers (LDs), which are in turn connected to a 12-element VCSEL array. The output lanes are directly connected to a 12-element p-i-n diode array. After the detection the RF signals are amplified by 12 transimpedance amplifiers (TIAs) as it seen from Fig. 2.

III. THEORETICAL MODEL OF AN OPTICAL LINK

The both proposed architectures are based on the optical link as it was mentioned in Section 1. The proposed link shown in Fig. 3 consists of an optical link and wireless channels. In this section we consider existing theoretical models of the optical link containing the directly modulated VCSEL as a transmitter, MMF, p-i-n PD, and a wireless channel. MMF link models have been discussed in a number of works [5]-[8].

Consider the standardized 850nm laser-optimized 50µm MMF model [5]. Transmitter, MMF and connections are the most important factors determining the 3dB optical bandwidth of the link [5]. The transmitter, VCSEL is a key device in local area networks using MMFs [9]. The VCSEL’s well known advantages are following: low power consumption; high-speed modulation with low driving current; narrow circular beam for direct fiber coupling; low cost and small packaging capability; single longitudinal mode operation with vertical microcavity.
The operational characteristics of the directly modulated VCSEL are described by the rate equations for the photon density \( P(t) \), electron density \( N(t) \) and the phase \( \phi(t) \) since the amplitude modulation in semiconductor lasers is accompanied by the phase modulation determined by the linewidth enhancement factor (LEF) \( \alpha_c \) [5, 10]-[13].

\[
\frac{dP}{dt} = \left[ \Gamma a (N - N_0) - \alpha_{tot} \right] v_g P - \frac{P}{\tau_p} + \beta \Gamma N \frac{\tau_c}{\tau_e} + F_P(t)
\]

\[
\frac{dN}{dt} = \frac{I(t)}{qV} - \frac{v_g a (N - N_0)}{(1 + \epsilon^2)} N - \frac{N}{\tau_c} - B \times N^2 - CN^3 + F_N(t)
\]

\[
\frac{d\phi}{dt} = \frac{1}{2} \alpha_c \left[ \Gamma v_g a (N - N_0) - \frac{1}{\tau_p} \right] + F_\phi(t)
\]

where \( a \) is the differential gain, \( N_0 \) is a transparency electron concentration, \( \Gamma \) is the confinement factor, \( v_g \) is the group velocity of light, \( V \) is the active region volume, \( \tau_{p,e} \) are the photon and electron lifetimes, \( \epsilon \) is the gain compression factor, \( \beta \) is the spontaneous emission fraction combined into a laser mode, \( q \) is the electron charge, \( I(t) \) is the VCSEL bias current, \( B \) is the bimolecular recombination factor, \( C \) is the Auger recombination factor, \( \alpha_{tot} \) is the total loss coefficient given by

\[
\alpha_{tot} = \alpha_{loss} + \frac{1}{L} \ln R
\]

\( \alpha_{loss} \) is the VCSEL absorption coefficient, \( L \) is the VCSEL active region length, and \( R \) is the reflectivity of the mirrors. The terms \( F_P(t), F_N(t), F_\phi(t) \) are the Langevins forces assumed to be the Gaussian random processes with zero expectancy and the following correlation function

\[
\langle F_i(t) F_j(t') \rangle = 2D_{ij} \delta (t - t')
\]

where \( i, j = P, N, \phi \), the angle brackets denote the ensemble average, and \( D_{ij} \) is the diffusion coefficient in the Markovian approximation. The main contribution to the laser noise is caused by the diffusion coefficients \( D_{PP} \) and \( D_{\phi\phi} \) given by

\[
D_{PP} = \frac{\Gamma v_g a (N - N_0)}{V} P; D_{\phi\phi} = \frac{\Gamma v_g a (N - N_0)}{V P}
\]

Single mode rate equations (1)-(3) have been found to be a very good approximation to the large signal behavior for MMF [5]. For the analogous applications, the relation between the SNR and the relative intensity noise (RIN) is given by

\[
SNR = \frac{m^2}{2RIN}
\]

where \( m = \Delta I / (I - I_{th}) \) is the electrical modulation depth, \( I, I_{th} \) are the input and the threshold currents, respectively. RIN is given by

\[
RIN = \frac{\langle \delta P_{opt}^2(t) \rangle}{\langle P_{opt}(t) \rangle^2}
\]

where \( \langle \delta P_{opt}^2(t) \rangle \) is the mean square optical power fluctuation and \( \langle P_{opt}(t) \rangle \) is the average optical power.

MMF links performance is affected by degradation due to finite rise and fall times at the transmitter and the receiver, the intermodal and intramodal dispersion, and noises specific to MMF links or multimode lasers [5]. The proposed model mainly concentrates on the signal degradation due to the intermodal dispersion because the largest part of the link power budget consumption is caused by pulse spreading caused by the intermodal dispersion [5]. At the operating wavelength \( \lambda = 850\,\text{nm} \), the 50\,\mu\text{m}, 19 mode groups, each of which can have its own group velocity \( v_g \) [5]. In actual MMFs there exists the coupling between the modes due to the fiber imperfections. However, only the coupling of modes within a mode group is significant over the short length scales of hundreds of meters, while the modal dispersion between mode groups is neglected, and the coupling between them is absent [5]. The attenuation of the coupling modes within each group \( \mu \) is described by the attenuation rate \( \gamma_\mu \), and the amplitude of a pulse launched into group \( \mu \) is proportional to the factor \( \exp (-\gamma_\mu z) \) as it propagates through MMF [5]. As a result, the bandwidth and the MMF transfer function strongly depend on the excitation conditions determining how much power will be coupled into each mode group, and the signal at the receiver output is determined by the launch conditions, MMF properties, and the link configuration [5].

The transverse modes of a VCSEL are assumed to be the Gaussian beam modes \( u_{pl}(r, \varphi, z, w_0, k) \) centered at the origin \( r = 0 \) of MMF and parallel to the \( z \) axis. They are given by [5]

\[
u_{pl}(r, \varphi, z, w_0, k) = \frac{w_0}{w} \left( \frac{\sqrt{2}}{w} \right)^l I_p \left( \frac{2}{k w^2} r^2 \right) \times \exp \left[ -i (k z - \Phi_{pl} + l \varphi) - r^2 \left( \frac{1}{w_0^2} + \frac{1}{2k R} \right) \right]
\]

where \( p \geq 0, l \geq 0 \) are the radial and angular mode numbers, \( w_0 \) is the spot size at the waist, \( k = 2\pi/\lambda \) is the free space wavenumber, \( I_p \) are the generalized Laguerre polynomials, and

\[
\Phi_{pl}(z, w_0, k) = (2p + l + 1) \arctan \left( \frac{2z}{k w_0^2} \right)
\]

\[
w(z, w_0, k) = w_0 \left[ 1 + \left( \frac{2z}{k w_0^2} \right)^2 \right]^{\frac{1}{2}}
\]

\[
R(z, w_0, k) = z \left[ 1 + \left( \frac{k w_0^2}{2z} \right)^2 \right]^{\frac{1}{2}}
\]

For the few-moded VCSEL the Gaussian beam model is a reasonable approximation [5]. A VCSEL \( u_{pl} \) mode at the air-fiber interface is transformed into a different Gaussian beam mode which then excites the various modes \( \psi_{lm}(r, \theta) \) of MMF corresponding to the transverse components \( E_{x,y} \) of the electric field in the fiber. The modes \( \psi_{lm}(r, \theta) \) are given by [5]

\[
\psi_{lm}(r, \theta) = f_{lm}(r) \nu (l \theta) \rho
\]

where \( l \geq 0 \) and \( m > 0 \) are the eigen-values of the radial and angular parts of (13), the index \( \nu \) denotes angular dependence.
\[ \sin \theta \text{ or } \cos \theta, \text{ and polarization } \mathbf{p} = x, y. \] The modes satisfy the normalization condition
\[ \int_0^\infty \int_0^{2\pi} r dr d\theta f_{in}^2(r) \rho^2(l\theta) = 1 \] (14)
The coupling amplitudes \( a_{pl}^{lmn} \) of the incident Gaussian beam mode with the fiber mode \( \psi_{lmnp} (r, \theta) \) are given by [5]
\[ a_{pl}^{lmn} = \int_A d^2 x \psi_{lmnp}(x) u_{pl'}(x', x'_0, k') \] (15)
where the integration is carried out over the area \( A \) of the fiber end face. Assuming that impulses from the transmitter induce electric fields at the input end face of MMF have the form \( \sum_{pl} c_{pl} u_{pl'}(x', x'_0, k') \delta(t) \) we write the impulse response of MMF \( h(z, t) [5] \)
\[ h(z, t) = \sum_{\mu} w_{\mu} \exp(-\gamma_{\mu} z) \delta(t - \tau_{\mu} z) \] (16)
where \( c_{pl} \) are the complex amplitudes, \( w_{\mu} = \sum_{lmn} w_{lmn'} \) is the mode power distribution (MPD), and
\[ w_{lmn'} = \sum_{pl'} |c_{pl'} a_{pl}^{lmn'}|^2 \] (17)
Consider now a typical p-i-n PD. Its quantum efficiency \( \eta \) is given by [13]
\[ \eta = \frac{P_{oabs}}{P_{opt}} = \zeta (1 - r) (1 - \exp(-\alpha_{PD}d)) \] (18)
where \( \zeta \) is p-i-n PD internal quantum efficiency close to unity, \( P_{oabs} \) is the incident optical power at the input of PD, \( P_{opt} \) is the optical power absorbed in PD, \( r \) is the reflection coefficient of the PD surface, \( \alpha_{PD} \) is the PD material absorption coefficient, and \( d \) is the thickness of the PD absorption intrinsic layer. The p-i-n PD bandwidth \( \Delta f \) is determined by the carrier transit time and time constant of the p-i-n PD equivalent circuit. It is given by [14]
\[ \Delta f = \left[ \left( \frac{2\pi d}{3.5\pi_d} \right)^2 + \left( 2\pi \epsilon_0 \epsilon_r S \left( R_a + R_l \right) \right)^2 \right]^{-1/2} \] (19)
where \( \pi_d \) is the average charge carrier drift velocity in the PD absorption intrinsic layer, \( \epsilon_0 \) is the free space permittivity, \( \epsilon_r \) is the PD permittivity, \( S \) is the PD photosensitive area, \( R_a, R_l \) are the series and load resistances in the PD equivalent circuit, respectively.

UWB wireless channel description is based on the modified Saleh - Valenzuela (SV) model [15]. In this model the analytical representation of a discrete multipath impulse \( h_i(t) \) can be presented as follows [15]
\[ h_i(t) = X_i \sum_{l=0}^{C_i} \sum_{k=0}^{K_{i,l}} \alpha_{kl}^i \delta(t - T_{i,l} - \tau_{i,k}) \] (20)
where \( \alpha_{kl}^i \) are the multipath gain coefficients, \( T_{i,l} \) are the delays of the \( l \)th cluster, \( \tau_{i,k}^l \) is the delay for the \( k \)th multipath component relative to the \( l \)th cluster arrival time. Shadowing effect obeys the log-normal distribution and is represented by \( X_i \) where \( i \) refers to the \( i \)th realization. The channel coefficients \( \alpha_{kl}^i \) are given by [15]
\[ \alpha_{kl} = p_{kl} \xi_{l} \beta_{kl} \] (21)
where \( \xi_{l} \) is the fading associated with the \( l \)th cluster, and \( \beta_{kl} \) is the fading associated with the \( k \)th ray of the \( l \)th cluster. The cluster arrival time distribution \( p(T_l|T_{l-1}) \) and the ray arrival time distribution \( p(\tau_{kl}|\tau_{k(l-1)}) \) are given by, respectively [15]
\[ p(T_l|T_{l-1}) = \Lambda \exp(-\Lambda (T_l - T_{l-1})), l > 0 \] (22)
and
\[ p(\tau_{kl}|\tau_{k(l-1)}) = \lambda \exp(-\lambda (\tau_{kl} - \tau_{k(l-1)})), k > 0 \] (23)
where \( \Lambda \) is the cluster arrival rate, and \( \lambda \) is rate arrival rate. The modified SV model derives as an output mean and root mean square (RMS) excess delays, number of multipath components, and power decay profile.

The numerical simulations of the optical link and wireless channel using the models discussed above had been carried out in the framework of the advanced design system (ADS) package, version 2006, update 2, product of Agilent. The package contains the UWB toolbox based on "Multiband OFDM Physical layer Specification", (WiMedia Alliance document, Release 1.1, July 14, 2005).

IV. NOVEL OCMC COMPONENT

In the case of the parallel RF/serial optic architecture the detection of multiplexed MB OFDM UWB modulated optical signal is required. The output UWB RF signal should be the minimum distorted replica of the original multiplexed MB OFDM UWB envelope of the optical signal at the input to the optical fiber. The optical handling of microwave (MW) devices such as directional couplers, phase shifters, attenuators, ultra-fast MW switches, etc. has been thoroughly investigated both theoretically and experimentally and used successfully in many applications in the framework of Microwave Photonics approach [16]-[26]. The advantages of this approach are following: low cost solution, low power consumption, high responsivity, flat spectral response over the desired band, low noise characteristics, possibility of creation of compact components which can be easily integrated with other electronic and photonics systems [17].

However, the conditions of the detection process are essentially different from the steady state optical control of the MS load. The input signal of the system in our case is the multiplexed MB OFDM UWB modulated optical radiation fed from an optical fiber. Typically, the optical carrier power \( P_{opt} \) is comparatively low: \( P_{opt} \sim 1\mu W \). For a multimode optical fiber with an optical beam radius \( r_b \sim 10\mu m \) it yields a comparatively low intensity \( I = P_{opt} / (\pi r_b^2) \sim 3W/mm^2 \). Instead of it, the detected UWB RF signal voltage over the optically controlled load should serve itself as a source of the MW radiation. We propose a novel OCMC device for this purpose. In the proposed method we used OCMC consisting of an open ended microstrip (MS) line with a semiconducting substrate, as sketched in Fig. 4. The optical beam, modulated
by UWB RF signal, is illuminated on the substrate near the open end of the MS line.

The down conversion from the optical domain to the MW domain can be modeled by an optically controlled load connected at the open end of the MS line. The variations of the photocurrent at the optically controlled load of the MS produce an electromagnetic (EM) waves that propagate along the MS line towards the output port of OCMC from which they are probed by a coaxial line of the same characteristic impedance, $Z_0$. The efficiency of the optical-microwave frequency down conversion depends on the ability to collect the photocarriers at the bottom contact. In the case of silicon technology, the thickness conventional substrates is in the range of $350-500\mu m$ which is quite large compared to the diffusion length of the photocarriers $L_{n,p} = \sqrt{D_{n,p}\tau} \sim (10 \div 30)\mu m$. In the case of surface absorption characterized by large values of absorption coefficient $\alpha$ and consequently a very small absorption length $\sim \alpha^{-1}$ the effective depth the photocarriers can reach is determined by the diffusion and drift properties of the photocarriers. The feasibility of the proposed OCMC device was experimentally verified by an open-ended MS line with $Z_0 = 50\Omega$ implemented on a high resistivity $\rho > 3000\Omega cm$ slightly p-type Si substrate are shown in Fig. 4. The optical source was a tunable laser diode with wavelengths from $\lambda = 680$ up to $\lambda = 980nm$. The results for the OCMC response function at the different levels of the optical power are shown in Fig. 5. These results do not satisfy the requirements of the UWB RF signal detection.

An alternative approach has been proposed recently in a number of works [27]-[30]. It has been demonstrated experimentally that thin Ge-on-Si, SiGe/Si, or Si layers of a thickness about one up to several micrometers can operate successfully as UWB RF signal detectors providing a bandwidth of about $(10 \div 20)\; GHz$ [27]-[30]. A resonant cavity-enhanced Si photodetector permits to overcome the comparatively low absorption in Si by using the substrate with a distributed Bragg reflector (DBR) that provides 90\% reflection of an optical power back into the detector layer [27]. Silicon photodetectors monolithically integrated with preamplifier circuits have achieved error-free detection at up to $SGb/s$ at an optical wavelength $\lambda = 850nm$ [30]. For operation at longer wavelengths Ge-on-Si photodiodes with the bandwidth up to 21 GHz at $\lambda = 1.31\mu m$ are attractive for monolithic optical receivers [30]. A theoretical model of such thin film devices has not yet been developed to our best knowledge.

Consider an infinite in the $x, y$ directions layer of a thickness $d$ in the $z$ direction placed on a semi-infinite in the $z$ direction substrate. The geometry of the problem is presented in Fig. 6. The electric and magnetic fields of the incident and reflected optical waves $E_{1x}, H_{1y}$ in the free space $z < 0$, $E_{2x}, H_{2y}$ in the layer $0 \leq z \leq d$, and $E_{3x}, H_{3y}$ in the substrate $z > d$ are given by [31]

$$
\begin{align*}
  z < 0 \rightarrow E_{1z} & = [E_1^+ \exp(-ik_1z) + E_1^- \exp(ik_1z)] \\
  & \times \exp(i\omega_{opt}t) \\
  H_{1y} & = \frac{1}{Z_1} [E_1^+ \exp(-ik_1z) - E_1^- \exp(ik_1z)] \exp(i\omega_{opt}t)
\end{align*}
$$

Fig. 4. Schematic view of an optically controlled microstrip convertor (OCMC)

Fig. 5. The dependence of the normalized response function of Si based OCMC on a bandwidth for different values of an optical power. The thickness of OCMC substrate $d = 520\mu m$

Fig. 6. Illuminated SiGe layer on a Si substrate
\[0 \leq z \leq d \rightarrow E_{zx} = \left[ E_0^+ \exp(-\gamma z) + E_0^- \exp(\gamma z) \right] \times \exp(i\omega_{opt} t)\]

\[H_{xy} = \frac{1}{Z_3} \left[ E_0^+ \exp(-\gamma z) - E_0^- \exp(\gamma z) \right] \exp(i\omega_{opt} t)\]

\[z > d \rightarrow E_{zx} = E_0^+ \exp(-ikz) \exp(i\omega_{opt} t)\]

\[H_{xy} = \frac{1}{Z_3} E_0^+ \exp(-ikz)\]

Here the wave impedances of the media have the form

\[Z_1 = \sqrt{\frac{\mu_0}{\varepsilon_0}} = 377\Omega; Z_2 = |Z_2| \exp i\theta; Z_3 = \sqrt{\frac{\mu_0}{\varepsilon_0}}\varepsilon_{r_3}\]

\[\mu_0, \varepsilon_0\] are the free space permeability and permittivity, respectively, the absorption layer wave impedance \(Z_3\) is assumed to be complex, \(\varepsilon_{r_3}\) is the permittivity of the substrate, \(\omega_{opt}\) is the optical frequency.

\[k_1 = \frac{\omega}{c}, \gamma_2 = \frac{\alpha}{2} + i\beta, k_3 = \frac{\omega}{c}\sqrt{\varepsilon_{r_3}}\]

\(\beta\) is the propagation constant, and \(c\) is the speed of light in vacuum. The solution of the boundary problem yields the expression for the optical intensity \(I_{opt}^{tot}(z)\) in the thin film layer with absorption.

\[I_{opt}^{tot}(z) = I_0 \frac{2Z_3 \cos \theta}{|Z_2|} \cosh(\alpha(z - d)) - \sinh(\alpha(z - d)) \left( 1 + \frac{Z_3^2}{|Z_2|^2} \right)\]

where

\[I_0 = \frac{2Z_1 P_{opt}}{A_{eff}} \frac{\cos \theta}{|D|^2 |Z_2|} \]

\[P_{opt} = \frac{|E_0^+|^2 A_{eff}}{2Z_1} A_{eff} = \pi r_b^2\]

\[|D|^2 = \left| \sinh(\gamma_2 d) \left( 1 + \frac{Z_1 Z_3}{Z_2^2} \right) + \frac{(Z_1 + Z_3)}{Z_2} \cosh(\gamma_2 d) \right|^2\]

\(P_{opt}\) is the optical power of the incident wave in the free space \(z < 0\), and \(r_b\) is the light beam radius. The explicit expression of \(|D|^2\) in general case is hardly observable, and we do not present it here. It can be substantially simplified under the realistic quasi-resonance assumption for \(\lambda_{opt} \sim 1\mu m\) and \(d \sim (0.5 \div 2) \mu m\)

\[\sin \beta d = 0, \beta d = \pi m, m = 1, 2, \ldots\]

Then, a simplified expression of \(|D|^2\) takes the form

\[|D|^2 = \frac{(Z_1 + Z_3)^2}{|Z_2|^2} \cos^2 \left( \frac{\alpha}{2} d \right) + \sinh^2 \left( \frac{\alpha}{2} d \right) + \sinh^2 \left( \frac{\alpha}{2} d \right) \left[ 1 + 2 \frac{Z_1 Z_3 \cos 2\theta}{|Z_2|^2} + \frac{(Z_1 Z_3)^2}{|Z_2|^2} \right]\]

\[+ \sinh^2 \left( \frac{\alpha}{2} d \right) \left[ 1 + 2 \frac{Z_1 Z_3 \cos 2\theta}{|Z_2|^2} + \frac{(Z_1 Z_3)^2}{|Z_2|^2} \right] \frac{Z_1 Z_3 \cos \theta}{|Z_2|^2} \left[ 1 + \frac{Z_1 Z_3}{|Z_2|^2} \right]^2\]

The detailed derivation of expression (32) is presented in Appendix.

Evaluate now the concentration the photocarriers in the framework of the drift-diffusion model [32]-[34]. The continuity equations for the photoinduced electron and hole concentration \(n(z, t)\) and \(p(z, t)\) have the form, respectively.

\[\frac{\partial n}{\partial t} = -\mu_n E \frac{\partial n}{\partial z} + \frac{D_n}{\tau_n} \frac{\partial^2 n}{\partial z^2} + \frac{g_n(z, t) - n - n_0}{\tau_n}\]

\[\frac{\partial p}{\partial t} = -\mu_p E \frac{\partial p}{\partial z} + \frac{D_p}{\tau_p} \frac{\partial^2 p}{\partial z^2} + \frac{g_p(z, t) - p - p_0}{\tau_p}\]

\[g_n(z, t) = g_p(z, t) = g(z, t) = \eta \frac{\partial I(z, t)}{\partial z}\]

Then the generation rates of electrons and holes \(g_{n,p}(z, t)\) can be written as follows

\[g(z, t) = g_0(z) + g_1(z, t)\]

\[g_0(z) = \eta \frac{\partial f(t)}{\partial z}\]

\[g_1(z, t) = \eta \frac{\partial f(t)}{\partial z}\]

where \(\tau_{n,p}, D_{n,p}, \mu_{n,p}\) are the lifetime, diffusion coefficients, and mobilities of electrons and holes, respectively, \(n_0, p_0\) are the equilibrium electron and hole concentrations, \(\eta\) is a quantum efficiency, \(E\) is the electrostatic field, and \(f(t)\) is the UWB RF envelope of the optical carrier (32). The coordinate averaged Fourier transform \(|\mathcal{N}_1(\omega)|\) of the photocarrier concentration time-dependent part \(n_1(z, t)\) is actually the OCMC response function. It has the form.

\[\mathcal{N}_1(\omega) = \frac{1}{d} \int_0^d N_1(z, \omega) dz = \frac{F(\omega) \tau (1 - i\omega \tau)}{d (\omega^2 L_{aeq}^2 - 1) \left[ 1 + (\omega \tau)^2 \right]} \times \left\{ \frac{1}{\alpha} \left[ I_{01} \cos(\alpha d) - 1 \right] + I_{02} \sinh(\alpha d) \right\} + \frac{1}{L_{aeq}} \cos(d/L_{aeq}) + \frac{\alpha}{L_{aeq}} \sinh(d/L_{aeq}) \]

\[\times \left[ \frac{\partial g_0}{\partial z}(0) - \frac{s_0}{D_a} g_0(0) \right] L_{aeq} \left[ 1 + \cosh \left( \frac{d}{L_{aeq}} \right) \right] + \frac{g_0(d)}{\sinh \left( \frac{d}{L_{aeq}} \right)} + \frac{s_0 L_{aeq}}{D_a} \left( \cosh \left( \frac{d}{L_{aeq}} \right) - 1 \right) \right\}\]

The detailed derivation of \(|\mathcal{N}_1(\omega)|\) is presented in Appendix II. The results of the numerical evaluations of the response function \(|\mathcal{N}_1(\omega)|\) for the typical values of material parameters of SiGe/Si and Si are presented in Figs. 7, 8. The power absorption coefficient of Si\(_x\)Ge\(_{1-x}\) compounds \(\alpha \sim 10^3 cm^{-1}\) in the interval of \(\lambda_{opt} \sim 850nm\). The electron mobility reaches...
its maximum value of $\mu_n = 7700 \text{cm}^2/(\text{V}\cdot\text{s})$ for $\text{Si}_{0.5}\text{Ge}_{0.5}$. For smaller concentrations of Ge the charge carrier mobilities are closer to the ones of a pure Ge, while in the opposite case they tend to the values of a pure Si charge carrier mobilities.

The numerical estimations based on the proposed analytical model of the thin layer SiGe/Si OCMC structure with an detecting layer thickness of about $d = (0.5 \div 2) \mu m$ clearly show that a bandwidth of at least $60 \text{GHz}$ can be achieved as it is seen from Fig. 7. The resonant conditions (36) are essential for the layer thickness because the reflection from the SiGe/Si interface in such a case reaches its maximum value. The proposed structure is simpler as compared to resonant-cavity-enhanced (RCE) photodetectors with DBR layers in the substrate. Generally, the SiGe/Si structures are promising candidates for the high-speed optoelectronics receivers due to the high operation rate, comparatively optical high absorption coefficient, the possibility of operation in the near IR spectrum from 850nm to 1550nm, low noise and compatibility with Si based electronic components.

**V. SIMULATION RESULTS**

The numerical simulations have been carried out for the parallel RF/parallel optics architecture. We investigated the mixing of 10 RF channels each one with the $0.5 \text{GHz}$ bandwidth. The resulting signal was applied to the multimode $10 \text{GHz}$ VCSEL, the modulated optical signal was transmitted through the $50 \text{m}$ MMF and at the output detected by the p-i-n PD. The simulation results are shown in Fig. 9. The mixed RF spectrum at the VCSEL input, the modulated optical signal at the VCSEL output, and the detected RF spectrum are shown in the upper box, the middle box, and the lower box of Fig. 9, respectively. The internal structure of one of the RF channels located at $3.5 \text{GHz}$ central frequency at the corresponding transmission stages is shown in Fig. 10. This channel includes 128 subcarriers and is transmitting $496 \text{Mb/s}$ over $0.5 \text{GHz}$ bandwidth. In order to study the dispersion influence on the quality of the transmitted MB OFDM signals we have carried out the simulation for the different MMF lengths. The short MMF with a length of $50 \text{m}$ has an almost flat frequency response up to the frequency of $10 \text{GHz}$. The
strongly inhomogeneous behavior in such a case in the vicinity of 10GHz is caused by the VCSEL bandwidth limitations. The p-i-n PD used in these measurements has the bandwidth of about 25GHz. The bandpass filter behavior of the MMF caused by the multimode dispersion is strongly manifested for longer MMFs. The magnitude and the phase of the 650m MMF are shown in Fig. 11. In the TFC7 frequency range the transfer function magnitude curve is flat. On the contrary, in the TFC6 frequency range the transfer function magnitude has a notch. The transmission of the multiplexed MB OFDM signals is limited by the MMF length of about 100m.

VI. THE EXPERIMENTAL RESULTS

A. MB OFDM Signal Transmission

MB OFDM UWB signal was directly applied to the VCSEL and after the propagation through the MMF was detected by the p-i-n PD. The objective of the measurements was to study the performance of the proposed link by means of the packet error rate (PER). The measurements have been carried out for the TFC5, TFC6 and TFC7 band of MB OFDM UWB signals. Fig. 12 presents the PER versus MMF length for the optical link only. The PER dependence versus the MMF length for the different MB OFDM UWB signals shows a peculiar behavior. The PER of the TFC7 band located at higher carrier frequency (4.488GHz) stays flat and has values of an order of magnitude of $10^{-6}$ for MMF length up to 1km. However, the PER in the case of TFC5 band located at the 3.5GHz carrier frequency and the TFC6 band located at the 4.0GHz carrier frequency increases dramatically for the MMF lengths longer than 300m.

In order to understand this behavior of the PER versus MMF length we have measured the MMF transfer function for different MMF lengths. The short MMF with a length of 10m has an almost flat frequency response up to the frequency of 10GHz as it is shown in Fig. 13. The strongly inhomogeneous behavior in such a case in the vicinity of 10GHz is caused by the VCSEL bandwidth limitations. The p-i-n PD used in these measurements has the bandwidth of about 25GHz. The bandpass filter behavior of the MMF caused by the multimode dispersion is strongly pronounced for longer MMFs.

According to Fig.13, the MMF transfer function has strong notches in the frequency range of TFC5 band at the fiber lengths longer than 500m. These strong notches affect the signal spectrum and lead to the significant increase of the PER. PER versus MMF length for the combined MMF and wireless link is shown in Fig. 14. In these measurements carried out for the different MMF lengths we kept constant the distance between two antennas of the wireless channel. These experimental results are in good accord with the simulations results mentioned above.

B. All-optical Up-conversion of MB OFDM Signals

For the parallel RF/series optics architecture shown in Fig. 1 all-optical up-conversion can be applied instead of the conventional RF up-conversion. The low cost all-optical up-conversion is realized using VCSEL’s nonlinearity. In such a case, VCSEL is biased simultaneously by a local oscillator (LO) and UWB signal. The second and third-order intermodulations are situated in frequency ranges $f_{LO} \pm f_{UWB}$, $2f_{LO} \pm f_{UWB}$, $f_{LO} \pm 2f_{UWB}$ where $f_{LO}$, $f_{UWB}$ are the LO and the UWB central frequencies, respectively. The second-order intermodulation products are dominant in the vicinity of the VCSEL threshold bias current. However, due to the large bandwidth of UWB OFDM signal the frequency $f_{UWB} + f_{LO}$
may fall into the up-converted UWB signal spectrum and for this reason it cannot be singled out. This bandwidth overlapping limits the possibility of the up-conversion for a wide range of the UWB OFDM signal frequencies.

On the contrary, in the case of the third order nonlinearity the distortion terms of the type $2f_{UWB} + f_{LO}$ and $2f_{LO} + f_{UWB}$ fall outside the UWB OFDM signal bandwidth. For this reason, it is possible to provide a much better performance of the up-conversion of a wide range of UWB signal frequencies $f_{UWB}$ by means of the third-order intermodulation. We used the UWB signal power of $P_{UWB} = -14dBm$, the LO power of $-5dBm$, and the bias current of $3mA$.

The measured error-free constellation diagrams and undistorted spectra of the both up-converted UWB signals for TFC5, TFC6, and TFC7 bands with the central frequencies $f_{UWB} = 6.6GHz$ and $f_{UWB} = 7.128GHz$, respectively, are shown in Figs. 15, 16. These results prove the feasibility of the proposed low cost all-optical up-conversion.

VII. CONCLUSIONS

In conclusion, we proposed two possible architectures for the high spectral efficiency optical transmission of OFDM UWB signals beyond 40Gb/s: the parallel RF/serial optics architecture and parallel RF/parallel optics architecture. We have carried out the numerical simulations for the parallel RF/parallel optics architecture and predicted its highly quality performance. We investigated theoretically and experimentally the optical link consisted of the directly modulated VCSEL, MMF, p-i-n PD and a wireless channel. We presented the detailed theoretical analysis and numerical results for a novel OCMC detecting device based on the SiGe/Si structure. We demonstrated experimentally the highly efficient and low cost all-optical up-conversion of UWB signals.

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APPENDIX

EVALUATION OF THE OPTICAL INTENSITY IN A THIN LAYER WITH ABSORPTION

The boundary conditions for the electric and magnetic fields at the layer surfaces $z = 0$ and $z = d$ yield [31]:

$$z = 0 \rightarrow E_1^+ + E_1^- = E_2^+ + E_2^-$$

$$\frac{1}{Z_1} [E_1^+ - E_1^-] = \frac{1}{Z_2} [E_2^+ - E_2^-]$$

$$z = d \rightarrow E_2^+ \exp(-\gamma_d d) + E_2^- \exp(\gamma_d d) = E_1^+ \exp(-i k_d d)$$

$$\frac{1}{Z_2} [E_2^+ \exp(-\gamma_d d) - E_2^- \exp(\gamma_d d)]$$

$$= \frac{1}{Z_3} E_3^+ \exp(-i k_d d)$$

We assume that the incident wave amplitude $E_1^+$ is known. Then, eliminating $E_1^-$ and $E_1^+$ we obtain

$$E_2^+ = \frac{E_2^+ \exp(\gamma_d d) [1 + \frac{Z_3}{Z_2}]}{[\sinh(\gamma_d d) [1 + \frac{Z_3}{Z_2}] + \frac{(Z_1 + Z_3)}{Z_2} \cosh(\gamma_d d)]}$$

$$E_2^- = \frac{E_2^+ \exp(-\gamma_d d) [1 - \frac{Z_3}{Z_2}]}{[\sinh(\gamma_d d) [1 + \frac{Z_3}{Z_2}] + \frac{(Z_1 + Z_3)}{Z_2} \cosh(\gamma_d d)]}$$

Substituting (50) and (51) into (26) and (27) we obtain the expressions for the electric and magnetic field in the layer.

$$E_{2x}^+ (z, t)$$

$$= \frac{E_2^+ \exp(-\gamma_d (z - d)) [1 + \frac{Z_3}{Z_2}] \exp i wt}{[\sinh(\gamma_d d) [1 + \frac{Z_3}{Z_2}] + \frac{(Z_1 + Z_3)}{Z_2} \cosh(\gamma_d d)]}$$

$$E_{2x}^- (z, t)$$

$$= \frac{E_2^+ \exp(\gamma_d (z - d)) [1 - \frac{Z_3}{Z_2}] \exp i wt}{[\sinh(\gamma_d d) [1 + \frac{Z_3}{Z_2}] + \frac{(Z_1 + Z_3)}{Z_2} \cosh(\gamma_d d)]}$$

$$H_{2y}^+ (z, t) = \frac{E_{2x}^+(z, t)}{Z_2}; H_{2y}^- (z, t) = -\frac{E_{2x}^-(z, t)}{Z_2}$$

The time averaged total optical intensity $I_{opt}^{tot}$ in the layer consists of the intensities $\langle P^+ \rangle$ and $\langle P^- \rangle$ of the incident wave and reflected wave, respectively. It has the form [31]

$$I_{opt}^{tot} (z) = \langle P^+ \rangle + \langle P^- \rangle$$

where

$$\langle P^+ \rangle = \text{Re} \left( \frac{1}{2} E_{2x}^+ (z, t) \left( H_{2y}^+ (z, t) \right)^* \right)$$

$$= \text{Re} \left\{ \frac{E_2^+ \exp(-2 \text{Re} (\gamma_d) (z - d)) [1 + \frac{Z_3}{Z_2}]}{2Z_2^2 \sinh(\gamma_d d) [1 + \frac{Z_1 + Z_3}{Z_2}] \cosh(\gamma_d d)} \right\}$$

$$\langle P^- \rangle = \text{Re} \left( \frac{1}{2} E_{2x}^- (z, t) \left( H_{2y}^- (z, t) \right)^* \right)$$

$$= \text{Re} \left\{ \frac{E_2^+ \exp(2 \text{Re} (\gamma_d) (z - d)) [1 - \frac{Z_3}{Z_2}]}{2Z_2^2 \sinh(\gamma_d d) [1 + \frac{Z_1 + Z_3}{Z_2}] \cosh(\gamma_d d)} \right\}$$

Taking into account that according to (31) $2 \text{Re} (\gamma_d) = \alpha$

II.

APPENDIX

EVALUATION OF THE PHOTOINDUCED CARRIER CONCENTRATION

Substituting (32) into equation (43) we obtain

$$g_0 (z) = I_{01} \sinh(\alpha (z - d)) - I_{02} \cosh(\alpha (z - d))$$

where

$$I_{01} = \frac{\eta \alpha}{\hbar \nu} \frac{2Z_3 \cos \theta}{|Z_2|}; I_{02} = \frac{\eta \alpha}{\hbar \nu} \frac{1 + \frac{Z_3^2}{|Z_2|^2}}{2}$$

Typically, in the photoinduced plasma, the electron and hole relaxation time $\sim 10 ns$ is much smaller than the carrier lifetime, and electroneutrality condition can be applied [22], [23], [34].

$$n = p$$

At the illuminated surface of the semiconductor the strong injection mode and ambipolar diffusion are realized when $n \gg n_0, p_0$ and the ambipolar mobility $\mu_a$ vanishes [20], [21]

$$\mu_a = \mu_h \mu_p (p - n) \mu_p p + \mu_n n = 0$$

In our case the thin layer is entirely occupied by the strong injection mode region. Under such conditions continuity equations (38), (39) reduce to the ambipolar diffusion equation [34]

$$\frac{\partial n}{\partial t} = D_a \frac{\partial^2 n}{\partial z^2} - \frac{n}{\tau} + g (z, t)$$

where it is assumed that $\tau_n = \tau_p = \tau$. According to expressions (42)-(44) we separate the steady-state and time dependent parts $n_{ph0} (z)$, $n_1 (z, t)$ of the photocarrier concentration $n$.

$$n = n_{ph0} (z) + n_1 (z, t)$$

Substituting (63) into equation (62) we obtain two equations for $n_{ph0} (z)$ and $n_1 (z, t)$, respectively

$$D_a \frac{\partial^2 n_{ph0}}{\partial z^2} - \frac{n_{ph0}}{\tau} + g_0 (z) = 0$$

$$\frac{\partial n_1}{\partial t} = D_a \frac{\partial^2 n_1}{\partial z^2} - \frac{n_1}{\tau} + g_0 (z) f (t)$$
where the ambipolar diffusion coefficient $D_a$ is given by [20]

$$D_a = \frac{2D_nD_p}{(D_n + D_p)}$$  \hspace{1cm} (66)

We use the boundary conditions of the mixed type. We assume a finite surface recombination rate $s_0$ on the top surface $z = 0$ which yields [23]

$$\frac{\partial n}{\partial z}_{z=0} = \frac{s_0}{D_a}n(z = 0)$$  \hspace{1cm} (67)

On the other hand, a kind of an ohmic contact at the interface of the layer and the substrate[27] $z = d$ prescribes the condition

$$n(z = d) = 0$$  \hspace{1cm} (68)

We are interested in the time-dependent part of the photocarrier concentration $n_1(z, t)$ which is responsible for the UWB RF signal detection. Hence we should solve equation (65) with the boundary conditions (67) and (68). In general case of the UWB RF signal $f(t)$ we carry out the Fourier transform of equation (65) with respect to time. We obtain

$$D_a \frac{\partial^2 N_1(z, \omega)}{\partial z^2} = \left( i\omega + \frac{1}{\tau} \right) N_1(z, \omega) + g_0(z) F(\omega) = 0$$  \hspace{1cm} (69)

where

$$N_1(z, \omega) = \int_{-\infty}^{\infty} n_1(z, t) \exp(-i\omega t) \, dt$$  \hspace{1cm} (70)

and

$$F(\omega) = \int_{-\infty}^{\infty} f(t) \exp(-i\omega t) \, dt$$  \hspace{1cm} (71)

The boundary conditions (67) and (68) can be applied to the general solution of (69). The result has the form.

$$N_1(z, \omega) = \frac{F(\omega)\tau (1 - i\omega\tau)}{(\sigma^2 L_{aeq}^2 - 1) \left[ 1 + (\omega\tau)^2 \right]} \times \left\{ -g_0(z) + \left[ \frac{1}{L_{aeq}} \cosh \left( \frac{d}{L_{aeq}} \right) + \frac{s_0}{D_a} \sinh \left( \frac{d}{L_{aeq}} \right) \right] \right\} \times \left[ \frac{\partial g_0}{\partial z}(0) - \frac{s_0}{D_a} g_0(0) \right] \sinh \left( \frac{z - d}{L_{aeq}} \right) + g_0(d) \left[ \frac{1}{L_{aeq}} \cosh \left( \frac{z}{L_{aeq}} \right) + \frac{s_0}{D_a} \sinh \left( \frac{z}{L_{aeq}} \right) \right]$$  \hspace{1cm} (72)

where

$$L_{aeq}^2 = \frac{D_a \tau (1 - i\omega\tau)}{1 + (\omega\tau)^2}$$  \hspace{1cm} (73)

The expression (72) for $N_1(z, \omega)$ averaged over the layer thickness $d$ can be used as the frequency response of the illuminated layer when $f(t) = \delta(t)$ and consequently $F(\omega) = 1$. Using the explicit expression (58) for $g_0(z)$ we obtain expression (45).
Characterization of Wavelength Tunable Lasers for Future Optical Communication Systems

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Abstract—The use of tunable lasers (TL) in dense wavelength division multiplexed (DWDM) networks for optical switching, routing and networking has gained a lot of interest in recent years. Employment of such TLs as tunable transmitters in wavelength packet switched (WPS) networks is one of the possible applications of these devices. In such systems, the information to be transmitted could be encoded onto a destination dependent wavelength. The authors investigate the possibility of using TLs in DWDM WPS networks by focusing on the characterisation of the instantaneous frequency drift of a TL due to wavelength tuning and direct modulation. Characterization of the linewidth of the TLs is also presented to verify the feasibility of using TLs in systems employing advanced modulation formats.

Index Terms— tunable lasers, dense wavelength division multiplexing, frequency drift, direct modulation, optical communications

I. INTRODUCTION

The use of widely tunable lasers in Dense Wavelength Division Multiplexed (DWDM) systems, packet switched schemes and access networks, has gained increased interest in recent years. They are being introduced as alternatives to fixed wavelength sources to reduce inventory and to provide a more dynamic device for external or direct modulation [1]. In the above mentioned systems, the information to be transmitted could be encoded onto a destination dependant wavelength, generated by the tunable laser. This would allow for dynamic bandwidth provisioning and thus increasing the flexibility of the system. There are several different types of tunable lasers currently vying for supremacy in this market such as External Cavity Lasers (ECL), Vertical Cavity Lasers (VCSEL), Grating assisted co-directional Coupler with rear Sampled Reflector (GCSR) laser and the Sampled Grating Distributed Bragg Reflector (SG-DBR) laser [2]. A major drawback associated with the external cavity laser is the complex manufacturing process that inevitably leads to high production costs, while the VCSEL and GCSR lasers suffer from low output powers, thus reducing their effectiveness in a long haul communication system.

The SG-DBR laser is an ideal candidate due to its large tuning range (40 nm), high output power, large Side Mode Suppression Ratio (SMSR) and its ability to be monolithically integrated with other semiconductor devices such as a Semiconductor Optical Amplifier (SOA) or an Electro-Absorption Modulator (EAM) [3]. The SG-DBR is comprised of a front mirror, back mirror, gain and phase section. The front and back mirrors are sampled at different periods such that only one of their reflection peaks can coincide within the range of the gain spectrum at a given set of currents. These peaks are spaced apart in wavelength, at a period inversely proportional to the period of the sampling. Through this technique, the desired ITU (International Telecommunication Union) channel can be selected by tuning the two mirrors, with the closest reflection peak of each mirror aligned at the desired channel [4].

When the tunable laser is set to switch from one ITU channel to another, it generates several spurious components, which interfere with other channels when the tunable laser is employed in a DWDM system. This can lead to unacceptable penalties in system performance, as shown previously in [5, 6]. In order to eliminate the unwanted spurious components the tunable laser module employs a monolithically integrated SOA at the output. This device remains unbiased for 60 ns after the beginning of the switch, attenuating the output of the tunable laser during the initial phase of the wavelength tuning. This essentially blanks the spurious components for a significant time during the switch from one ITU frequency to another, thus circumventing any cross channel interference.

Another important aspect of the tunable lasers performance, when switching between wavelengths, is the wavelength stability of the device. As the laser tunes into its desired wavelength, there is a settling drift before the channel finally stabilizes. This drift can have a large impact on the performance of a dense wavelength division multiplexed system (causing adjacent channel interference) and therefore must be fully characterized, in terms of both the magnitude of the drift and the time it takes to settle to the destination wavelength.

Thus far, external modulation has been the most popular technique used with tunable lasers, allowing
high-speed intensity modulation of the transmitter. Although external modulators provide high-speed, stable data modulation, they introduce loss on the transmitted signal due to their large coupling and insertion losses. The extra optical component also adds to the cost and complexity of the transmitter [7]. On the other hand, direct modulation is one of the simplest and most efficient techniques that could be used to modulate a lightwave signal. Additional benefits associated with direct modulation also include the reduced cost and a smaller transmitter footprint. It is therefore viable to investigate the performance of a directly modulated tunable laser to verify its usefulness in a DWDM system. When a static channel generated from the tunable laser is directly modulated; the direct modulation itself will cause a time dependant frequency drift. This drift can cause poor system performance, due to cross channel interference, thus making it imperative to characterize the magnitude and settling time of the drift in order to optimize system performance.

This paper is divided into three main sections. Section II focuses on the characterization of the frequency drift due to switching events in the tunable laser. The drift is characterized for both magnitude and the time it takes the channel to settle into its targeted ITU frequency. Section III analyzes the direct modulation induced frequency drift on the tunable laser and its effect on the performance of a wavelength division multiplexed system. The latter is achieved by passing the modulated channel through an Optical Band-Pass Filter (OBPF) to mimic a wavelength division multiplexed system receiver. Finally, section IV presents a linewidth characterization, which is vital, if the TL is to be used as a source in systems employing advanced modulation formats. A tuning map as a function of the front and back gratings is presented, which allows the choice of an optimum operating point for such systems (high SMSR and narrow linewidth).

II. FREQUENCY DRIFT DUE TO WAVELENGTH SWITCHING EVENTS

As mentioned previously, a vital aspect of tunable laser performance is wavelength stability after the laser switching event. Here we measure the instantaneous frequency drift of the tunable laser by using an optical filtering technique. The peak of the filter passband is initially set to the target wavelength, after which it is detuned by ~ 0.1 nm to create a sloped frequency discriminator. This process enables us to relate the signal’s power variation to the corresponding frequency variation, through the filter reflection profile, thus providing an accurate frequency drift as a function of time.

A. Experimental Setup

The settling drift of the tunable laser after blanking is characterized using the experimental setup depicted in Fig. 1. The tunable laser used for this work is an AltoNet 1200 fast wavelength switched transmitter that allows switching from one channel to any other channel, over the entire C-band wavelength range, in 200 ns. The tunable laser was supplied by Intune Networks and featured a custom made SMA connector to enable direct modulation of the gain section of the device. The drift of the TL after blanking was measured using a tunable fiber Bragg grating (FBG) with a 3 dB bandwidth of 27.5 GHz [8]. Initially, a power reference measurement (shown be the dotted line in Fig. 1) is recorded, where the output of the TL is detected by a 50 GHz photodiode in conjunction with a 50 GHz sampling oscilloscope. Subsequently, the switching signal is passed through the OBPF to filter out the targeted ITU channel.

![Figure 1. Frequency drift measurement set-up.](image1)

![Figure 2. (a) Fiber Bragg grating reflection profile, (b) turn on power transient of laser and signal power with filter detuned by 0.1 nm and (c) frequency drift measurement](image2)

B. Results and Discussion

As illustrated by Fig. 2a, the optical band-pass filter is detuned by 0.1 nm (black dot) to create a sloped frequency discriminator. When the light from the tunable laser is passed though the discriminator, the frequency fluctuations manifest as intensity variations as seen in Fig. 2b. The turn on power transient of the laser (thin line in Fig. 2b) is subtracted from the filtered signal to ensure that any variation of the received power is only due to the settling drift of the tunable laser. From the filter profile, we know the wavelength or frequency at the steady state level of the measured trace. Thus, through the filter reflection profile, we can relate the power variation to the corresponding frequency variation, therefore providing us with an accurate measurement of the drift as a function of time. The drift measurement was performed for a number of channel transitions. We present the result for the transition between channel 13 (192.1 THz) and channel 89 (195.9 THz). This was found to have one of the larger drifts of approximately 10 GHz in magnitude as illustrated in Fig. 2c.

As the tunable laser emerges from blanking it is approximately 10 GHz from its target frequency. The wavelength locker can be seen to turn on 30 ns after blanking. The locker causes a fast fluctuation in frequency for approximately 15 ns, after which the settling drift is characterized by a damped oscillation. The channel settles into its ITU frequency in about 400 ns. In a DWDM system, where channel spacing is as low as 12.5 or 50 GHz, the magnitude of this frequency drift would lead to poor system performance due to cross
channel interference. Hence, the impact of such a drift on the performance of a DWDM system is examined in the following section.

C. Performance Degradation due to Switching Drift

The set-up shown in Fig. 3 is employed to investigate the effect of the measured wavelength drift in a dense wavelength division multiplexed network. A two-transmitter DWDM system with channel spacing of 12.5 GHz is investigated. The tunable is set to switch between channel 42 and 52. A fixed laser is used as a second transmitter operating 12.5 GHz away from the tunable laser target channel at 1544.824 nm. Each channel is externally modulated with a 2.5 Gbit/s Non-Return-to-Zero (NRZ) pseudo-random bit sequence with a pattern length of 2\(^{-1}\).

The same data is used for both lasers, therefore it is necessary to de-correlate the information carried in each channel. This is achieved by passing one of the channels through 3 m of fiber. The two channels are then combined together and the fixed channel is filtered out using a Fabry-Perot (FP) tunable filter with a 3 dB bandwidth of 6 GHz. The demultiplexed channel then passes through a Variable Optical Attenuator (VOA), an optical filter, a photodiode and an electrical amplifier. The signal spectrum, eye diagram and the BER of the detected channel are also examined.

The BER of the filtered channel is measured as a function of the received optical power for various tunable laser configurations – when the interfering tunable laser was (a) set to channel 42 (> 500 GHz away from filtered channel), (b) set to channel 52 (12.5 GHz from filtered channel) and (c) switching from channel 42 to channel 52. The tunable laser is set to switch at a rate of 5 kHz. There is a minimal power penalty (<0.1 dB) when the tunable laser operates in a static mode at channel 52 (i.e. 12.5 GHz from the fixed channel) compared to when the tunable laser is static at channel 42. When the tunable laser switches between channels 42 and 52 the wavelength drift, after blanking, degrades the system performance by introducing an error floor at 10\(^{-7}\). This bit error rate is the average number of errors on the fixed channel over (a) the time at the beginning of the data transmission when the tunable laser is settling into channel 52, (b) the time when the tunable laser is settled into channel 52, (c) the time when the tunable laser is at channel 42 and (d) the time when the tunable laser is blanked. The switching tunable laser causes a burst of errors only at the beginning of the data stream, as it emerges from blanking and begins to settle into channel 52. At this time the drift is large enough such that it enters the filter passband of the fixed channel. As the tunable laser approaches its target wavelength the errors reduce [9], eventually giving no errors when the tunable laser is within ~ 3 GHz of its target wavelength. Thus if the tunable laser can be locked with less drift then the interference can be reduced.

The laser blanking time after the wavelength transition is initiated is subsequently increased from the default value of 60 ns to 200 ns. This is done in anticipation that the wavelength change will be closer to the target wavelength once the blanking time ends, however this will also result in a corresponding increase in the effective tunable laser switching time, thereby reducing network throughput. The improvement in system performance is also illustrated in Fig. 4. With the extended blanking time the system performs with a residual power penalty of ~ 1.1 dB (relative to the case when the tunable laser is static), at a reference bit error rate of 10\(^{-9}\), due to interference from the adjacent tunable laser channel. This power penalty could not be reduced by further increasing the blanking time as we believe it is primarily due to the frequency drift induced by the locking mechanism.

III. FREQUENCY DRIFT DUE TO DIRECT MODULATION

As mentioned earlier, direct modulation is the simplest and most cost efficient method to modulate the lightwave signal generated from a tunable laser. An inherent problem of direct modulation is a shift in the operating frequency due to a large variation of the carrier density upon the application of the modulating current. This shift in frequency will cause adjacent channel interference, thus decreasing the system performance. Therefore, it is of vital importance to investigate the magnitude of this shift and the effect it may have if utilized in a dense wavelength division multiplexed system.
A. Experimental Setup

The set-up used to characterize the frequency drift is similar to that employed to characterize the settling drift due to a switching event, with the addition of a Pulse Pattern Generator (PPG), as seen in Fig. 3a. The tunable laser is operated in static mode and the emission frequency is set to channel 38 (193.35 THz). The gain section (via an SMA input) of the tunable laser is then directly modulated with a series of different pattern sequences. As with the switching drift measurement, the optical band pass filter is de-tuned by 0.1 nm to create a sloped frequency discriminator [10]. The frequency discriminator essentially assists in characterising the offset in frequency of the tunable laser, from the set ITU frequency. By detuning the filter, the set ITU frequency is made to lie on a portion with a higher rejection (lower power) of the filter transfer characteristic. On the application of the modulating signal the frequency of the tunable laser is offset towards a lower rejection side of the filter transmission resulting in a higher amount of optical power being detected. Subsequently, as the in-built wavelength locker drags the output signal back to its target frequency, a smaller output power from the filter is recorded.

The power measurement is subsequently subtracted from the filtered signal to ensure that any variation in received power is due to the frequency drift of the tunable laser (and not due to the intensity modulation). To accurately measure the magnitude and settling time of the drift, a programmable bit sequence of thirty ones and thirty zeros is applied to the tunable laser. An average shift of 7 GHz (towards lower frequencies) from the specified frequency (unmodulated emission frequency of tunable laser at channel 38) is noticed as the modulation is applied to the laser.

B. Results and Discussion

Fig. 4 illustrates the frequency drift as a function of time. The magnitude of the drift is approximately 12 GHz which could have a large impact on an adjacent channel if employed in an ultra dense wavelength division multiplexed system. The offset frequency is then pulled back to its targeted frequency by the wavelength locker incorporated in the tunable laser module. This frequency lies on a lower portion of the filter transfer characteristic, thus exhibiting higher rejection, producing a lower output power. As can be seen from Fig. 4, the output frequency settles at its target wavelength after approximately 18 ns.

To perform a qualitative analysis of the effect the frequency drift will have in a dense wavelength division multiplexed system, Bit Error Rate (BER) measurements are performed. The experimental setup is depicted in Fig. 3b and again consisted of a static tunable laser channel (193.35 THz) modulated with a 1 Gb/s pseudo random bit sequence of length $2^{31}-1$. The peak to peak voltage of the modulating signal is set at one volt. An optical attenuator and an inline power meter (P$_{in}$) are incorporated into the set-up to monitor the varied received power falling on the pre-amplified receiver. The bit error rate measurements are carried out with the aid of an error detector for three different scenarios, namely: no filter (dotted line), with the 27.5 GHz optical band pass filter centered at the targeted ITU channel (193.35 THz) and with the filter centered at the average shifted frequency upon direct modulation (193.343 THz).

![Figure 6. Single channel frequency drift due to direct modulation](image)

![Figure 7. BER versus received power for three scenarios](image)

Fig. 5 illustrates the effect the frequency drift has on the performance of a typical dense wavelength division multiplexed system, by plotting the received power as a function of the bit error rate. The back to back reference curve highlights the systems performance in the case where no filter is used (●). It exhibited a receiver sensitivity of -33 dBm at an error rate of $10^{-9}$. In contrast, when the modulated signal is filtered with the optical band pass filter centered at the specified frequency of channel 38, the induced frequency deviations caused by the direct modulation result in intensity fluctuations at the output of the filter. These intensity fluctuations are reflected in a worsening system performance shown by the incurred power penalty of 6.7 dB at a bit error rate of $10^{-9}$ (●). A small improvement in performance (1.64 dB at a BER of $10^{-9}$), relative to the latter case, is achieved when the center frequency of the optical band pass filter is tuned to match the average shifted frequency of the tunable laser under modulation (Δ).

Eye diagrams are also recorded with a 50 GHz sampling oscilloscope for each of the discussed permutations. The eye diagrams further illustrate the bit
error rate degradation depicted in Fig 5. Fig. 6a shows the received eye when no filter is employed prior to the detected signal. A clear, wide eye opening is displayed, thus supporting the excellent performance with received powers in the order of -33 dBm at a BER of $10^{-9}$. However, in the case of Fig. 6b, where the filter is centered at the targeted frequency, the large intensity fluctuations caused by the drift in the directly modulated signal results in a partially closed eye, which in turn leads to a power penalty of 6.7 dB. Fig. 6c represents the received eye when the OBPF is tuned to the average shifted frequency, demonstrating improved system performance in comparison to Fig. 6b (1.64 dB at a BER of $10^{-9}$).

![Eye diagrams](image)

Figure 8. Eye diagrams (a) without filter, (b) with filter centered at targeted ITU frequency and (c) with filter centered at average shifted frequency

![Linewidth tuning map](image)

Figure 9. SG DBR Linewidth tuning map as a function of front and back mirror currents

IV. LINEWIDTH CHARACTERIZATION

In addition to network reconfigurability, tunable lasers are also being investigated as dynamic sources for advanced modulation formats. The introduction of fast reconfigurable optical networks and advanced modulation formats will place additional constraints on the tunable laser transmitters that can be used in such systems. In order to employ an SG-DBR laser as a continuous wave source for advanced modulation formats, it will become imperative to select an optimum operating point to achieve the narrowest possible linewidth to support phase modulated coding schemes. We have employed the self-heterodyne method to measure the linewidth of a Sampled Grating Distributed Bragg Reflector tunable laser. This method was used due to its simple setup, insensitivity to slow wavelength drift and ability to measure narrow linewidths (5kHz). The linewidth characterization setup is the same as in [11]. The Sampled Grating Distributed Bragg Reflector tunable laser used for this characterization was a commercially available device from Agility Communications and allowed access to all four laser sections.

The linewidth tuning map as a function of the front and back grating currents is illustrated in Fig. 9. The current of the front SG-DBR is scanned from 0 to 40mA with a 0.5mA step, while the current of the back grating is scanned from 0 to 60mA with a 1mA step. From Fig. 9 we can see that each mode is surrounded by points with relatively high values of linewidth at the mode boundaries, where frequency jumps occur [9]. Similar side mode suppression ratio and output power tuning maps can also be utilized in conjunction with the linewidth characterization to ensure that acceptable levels of transmitter performance are achieved. From Fig. 9, it is evident that the linewidth alternates significantly from 20 MHz to as low as 4 MHz within each mode. If these lasers are to be employed in systems that use advanced modulation formats it will become necessary to include the linewidth as an important characteristic when choosing the optimum operating point.

V. CONCLUSIONS

We have characterized the drift of a Sampled Grating Distributed Bragg Reflect laser for both, switching events (settling drift), and under the influence of direct modulation (frequency drift). Switching was achieved between two wavelength channels on the ITU grid. The magnitude and the settling time of the drift were characterized, as both factors would have serious implications in a dense wavelength division multiplexed system. A settling drift magnitude of 10 GHz was experienced with a settling time of 400 ns. In a dense wavelength division multiplexed system, where channel spacing is as low as 12.5 GHz, the magnitude of this drift could lead to poor system performance due to cross channel interference.

The frequency drift was also characterized when the TL was subjected to direct modulation. The magnitude of the drift has been recorded as 12 GHz with a peak to peak modulation of one volt. In dense wavelength division multiplexed systems this drift may cause performance degradation, again due to cross channel interference and also when passed through an optical filter. Results achieved demonstrate that a power penalty of 6.7 dB is incurred when using such a filter in comparison to an unfiltered case. A slight improvement in performance (1.64 dB at $10^{-9}$) is achieved when the filter is tuned to the average shifted frequency of the tunable laser under direct modulation.

The final characterization of the tunable laser involved the measurement of the linewidth as a function of the front and back mirror currents. A linewidth map has been created showing that the purity of the signal generated by tunable laser strongly depends on the value of these currents and can vary (within a mode) from 20 to 4 MHz.
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REFERENCES


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Signal Processing Algorithms in 100Gbit/s Optical Coherent and Non-coherent Receivers with PSK Modulation

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Abstract— This paper addresses multi-level PSK modulation formats with coherent and non-coherent detection and reviews signal processing algorithms like mitigation of phase noise, the clock and carrier recovery algorithms as well as the equalizer structures and their performance.

Index Terms— optical communications, modulation, phase shift keying, phase noise, equalizer

I. INTRODUCTION

Electronic signal processing is currently introduced in various subsystems of modern optical high speed transceivers. Traditionally electronic processing was basically limited to tasks like MUX/DEMUX, clock and data recovery and various synchronization loops as well as analog circuitry like modulator drivers or limiting amplifiers. In order to deal with advanced systems and with the need for increased performance at long reach and higher speed, additional signal processing requirements have appeared. Examples are FEC and electronic dispersion equalizers (EDC), electronic pre-compensation, differential encoders, clock and carrier recovery, and polarization control. Recent challenges employ pure digital processing, where ADCs and DACs are required, like OFDM modulation (orthogonal frequency division multiplexing) or MLSE (maximum likelihood sequence estimation) equalizers. In this paper some of these signal processing algorithms, which may be used in PSK modulated systems are considered in detail and their impact on performance and cost in terms of implementation effort is reviewed.

II. MITIGATION OF PHASE NOISE IN NON-COHERENT PSK-RECEIVERS

In a first part, phase noise is investigated. Phase noise is an important limiting effect in high-speed optical communication systems that make use of phase-shift keying modulation formats [1]. One method to mitigate nonlinear phase noise distortions is the compensation of the mean nonlinear phase shift (MEAN) [2]. Alternatively, multi-symbol phase estimation (MSPE) [3, 4] can be used to reduce the loss from direct detection compared to coherent detection by estimating a reference phase over several consecutive received symbols. Both phase noise compensation methods, MEAN (in a post detection version for DQPSK) and MSPE require signal processing at the receiver side after optical-to-electrical conversion.

We investigate the performance of different phase noise compensation options for a high speed RZ-DQPSK transmission system with various dispersion maps and direct detection. The efficiency of MEAN, MSPE and the combination of both is examined by Monte-Carlo simulations for varying pre- and post-compensation. Furthermore the performance of these three strategies for different average fiber input powers is investigated. Not only the optimum phase noise compensation strategy is identified, also the sensitivity of all three compensation variants to the chosen dispersion map is compared to the conventional receiver. For the best noise tolerant receiver option a maximum Q-gain of approximately 2 dB can be achieved for a wide range of fiber input power values [13].

A. Simulation Setup

The simulation setup for the 43 Gbit/s RZ-DQPSK multi-span transmission system is shown in Fig. 1. (Here 43 Gbit/s is used, but the methods are also applicable for 100 Gbit/s). At the transmitter side an optical I/Q-modulator based on a parallel Mach-Zehnder-Modulator (MZM) structure is used to generate the DQPSK signal. After RZ-pulse carving the optical signal is filtered (Gaussian filter 1st order, f_{FWHM} = 50 GHz), taking into account an optical multiplexer, and then amplified to achieve the desired average input power. The optical channel is characterized by a variable chromatic dispersion (CD) pre-compensation, eight fiber spans and a variable CD post-compensation. The pre- and post-compensation values are varied in steps of 50 ps/nm. Each span consists of 80 km SMF (D = 17 ps/nm/km, γ = 1.37 W^{-1} km^{-1}), a DCF and an EDFA, which fully compensates for the attenuation of the whole span. The OSNR of the system is adjusted to 14 dB. In each span the DCF compensates for 90% of the CD of the SMF (10% inline under-compensation). In the simulation the transmission fibres are assumed to be nonlinear. At the receiver side the signal is filtered (Gaussian filter 1.5th order, f_{FWHM} = 50 GHz) and then received by two Mach-Zehnder delay interferometers (MZDI) followed by a balanced detector. The inphase- and quadrature
components are then filtered by an electrical lowpass filter (Bessel 5th order, \( f_{3dB} = 15.05\, \text{GHz} \)). Both phase noise compensation methods, MEAN and MSPE, are implemented. Either one of them or both can be used for these investigations.

The MEAN method [2], which exploits the fact, that the mean nonlinear phase shift is proportional to the received power, is implemented in the electrical domain using an additional intensity detection branch. The intensity difference of two consecutive symbols has to be determined to get a value that is proportional to the received nonlinear phase noise. This is the difference of the nonlinear phase noise of two consecutive symbols due to the reception with MZDIs. The intensity difference is multiplied by a scaling factor \( \alpha \) and then used to rotate the phase of the inphase and quadrature component of the received signal.

The MSPE method as described in [3, 4] calculates recursively a new decision variable \( x(n) \) using the received signal \( u(n) \) and the previous variable \( x(n-1) \). A forgetting factor \( w \) slowly fades out the contribution of the previous symbols. This method averages out the phase noise of the received symbols, leading to a better phase reference and thus eliminating the loss of direct detection compared to coherent reception.

It can be seen that for 4 dBm the Q gain is almost the same for the three pre-compensations as well for MEAN as for MSPE. For MEAN there is almost no improvement, whereas the MSPE achieves a Q gain of 1-1.5 dB. For 6 dBm and 8 dBm the efficiency of MEAN increases for those values of pre-compensation where the performance of the conventional receiver is worse. Moreover, it can be seen that the performance of MEAN increases with increasing average fibre input power. For 6 dBm a maximum Q gain of 0.9 dB and for 8 dBm of 2.3 dB can be achieved. This leads to the conclusion that MEAN is most efficient for systems with performance loss due to the influence of nonlinearities, as in the case of non-optimised dispersion maps.

The improvement of MSPE, however, is almost the same for all investigated pre-compensations as well as for MEAN. But there is almost no improvement, whereas the MSPE achieves a Q gain of 1-1.5 dB. For 6 dBm and 8 dBm the efficiency of MEAN increases for those values of pre-compensation where the performance of the conventional receiver is worse. Moreover, it can be seen that the performance of MEAN increases with increasing average fibre input power. For 6 dBm a maximum Q gain of 0.9 dB and for 8 dBm of 2.3 dB can be achieved. This leads to the conclusion that MEAN is most efficient for systems with performance loss due to the influence of nonlinearities, as in the case of non-optimised dispersion maps.

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The efficiency with respect to the conventional receiver of the combination of MEAN and MSPE is shown in Fig. 3 for different pre-compensations and the three fibre input powers. The Q gain of the combined method is approximately equal to the sum of the Q gains of each method alone. The combined method is hence most efficient if the improvement of MEAN and MSPE is relative high, as e.g. for 6 dBm input power and -900
c. Summary

The performance of three strategies for compensation of phase noise effects: the compensation of the mean nonlinear phase shift, the multi-symbol phase estimation and the combination of both methods for 43 Gbit/s RZ-DQPSK multi-span transmission was obtained by Monte-Carlo simulations for various dispersion maps and different fibre input powers. In general, all three methods exhibit the same optimum dispersion map with respect to the residual dispersion as is given for the conventional receiver. Thus, it is not required to change the dispersion map if receivers are upgraded with one of these phase noise compensation methods. Furthermore, it was shown that the MEAN is most efficient for systems with high performance loss due to the influence of nonlinearities and therefore also for non-optimised dispersion maps (0.9 dB Q gain for 6 dBm input power and 2.3 dB for 8 dBm with 38 ps/nm residual dispersion). The improvement of this method for systems with low influence of nonlinearities however is very poor (0.35 dB Q gain for 4 dBm with 38 ps/nm residual dispersion).

The Q gain of MSPE is almost independent of the dispersion map that is used. The maximum Q gain decreases with increasing fibre input power (1.4 dB Q gain for 4 dBm input power, 0.9 dB for 6 dBm and 0.7 dB for 8 dBm with 38 ps/nm residual dispersion).

The improvement of the combination of MEAN and MSPE is approximately equal to the sum of the improvement from each individual method (1.8 dB Q gain for 4 dBm input power, 1.9 dB for 6 dBm and 2.7 dB for 8 dBm with 38 ps/nm residual dispersion). Therefore, depending on the system setup, it has to be decided if the additional gain of implementing both methods, MEAN and MSPE, is reasonable.

III. ELECTRONIC EQUALIZERS IN COHERENT PSK-RECEIVERS

In a second part we investigate electronic equalization in coherent receivers in combination with multi-level modulation and nonlinear fiber effects. The next bitrate in the Ethernet hierarchy is expected to be 100Gb/s. For this bitrate, multi-level modulation formats should be used to reduce the bandwidth requirements. In conjunction with a coherent receiver the multi-level formats can be detected in the electrical domain with digital signal processing (DSP), due to the availability of high-speed digital signal processors. Similarly carrier and phase recovery as well as the equalization can be achieved with DSP [5].

A. Simulation Setup

At the transmitter side either 53.5Gbaud RZ-QPSK [6], 35.7Gbaud RZ-8PSK [7] or 26.75Gbaud Star-RZ-16QAM are generated to achieve 107Gb/s for all modulation formats according to Fig. 4 (top). Star-RZ-16QAM is generated from RZ-8PSK with an additional Mach-Zehnder modulator (MZM). The advantage of this setup is the necessity of only binary driving signals, contrary to multi-level driving signals in the case of pure QI-modulation. All data signals are differentially encoded due to the phase ambiguity of the carrier recovery.

The transmission channel is modelled as a single span with variable length to adjust the chromatic dispersion (CD). A linear channel (only CD, D=17ps/nm/km) as well as a nonlinear channel (self-phase-modulation (SPM) and CD, γ=1.6215 1/W/km, α=0.21dB/km, P_{fiber in, average=5dBm}) is assumed.

At the receiver side an EDFA with a Gaussian optical bandpass filter (BP, f_{BP}=4.1x Baudrate) is used for noise loading. The received signal is combined with the signal of a local oscillator (LO) in a 2x4 90° hybrid (for this contribution ideal homodyne detection is assumed) and detected with two balanced photo-detectors. Afterwards the resulting electrical inphase and quadrature signals are lowpass (LP) filtered (Butterworth 3rd order, f_{LP}=1x Baudrate), sampled twice per symbol and then processed in a digital signal processing unit (fig. 5).

For EDC, linear transversal filters for all investigated equalizers are used. The performance of equalizers with only a feed forward filter (FFE[x]), where x denotes the
number of filter coefficients) and with an additional decision feedback equalizer (DFE[x], where feedback of the decision to the EDC is necessary (Fig. 5) are examined. The equalizer has a complex baseband structure and the coefficients are determined with a zero-forcing (ZF) approach based on the MMSE criterion with the use of a training sequence [8].

The symbol decision in the DSP is made on phases with an additional decision on the amplitude for Star-RZ-16QAM. After decision the differential decoding and mapping into two, three or four data streams (depending on the modulation format) takes place.

However in general the overall improvement with an additional DFE is rather small.

B. Simulation Results

Fig. 6 shows the OSNR penalty at a BER of $10^{-4}$ versus CD without EDC for all modulation formats for the linear and nonlinear channel.

Due to the short transmission distances the effective nonlinear length is small and therefore the influence of the nonlinear effects is small as well. The dispersion tolerance for RZ-QPSK (53.5G Baud) and RZ-8PSK (35.7G Baud) is almost equal. Obviously the expected improvement due to the reduced bandwidth of RZ-8PSK is just compensated for by the narrower symbol spacing. For Star-RZ-16QAM (26.75G Baud) the bandwidth is reduced further, whereas the minimum distance of the double-ring constellation is reduced only slightly compared to RZ-8PSK.

Now, EDC with the ZF-equalizer for the linear channel is applied (Fig. 7). First of all an improvement of at least one order of magnitude can be observed (note the abscissa scales of Fig. 6 and 7). Comparing the various modulation formats the dispersion tolerance increases with decreasing bandwidth of the format. For each modulation format the dispersion tolerance can be enlarged further by increasing the equalizer filter length. For Star-RZ-16QAM e.g. the dispersion tolerance at 2 dB OSNR penalty can be increased from ±550ps/nm with FFE[9] to greater than ±800 ps/nm with FFE[15].

The determined coefficients are a compromise for the compensation of both phase shifts. Thus for such multi-ring constellations equalization should be supported by a
compensation of the nonlinear phase shift (as e.g. proposed in [9]).

C. Summary

The performance of EDC in conjunction with coherent reception for RZ-QPSK, RZ-8PSK and Star-RZ-16QAM for the linear and nonlinear channel at 107Gb/s has been investigated. For EDC a zero-forcing equalizer is applied, using the MMSE criterion for deriving the coefficients. The dispersion tolerance (at 2dB OSNR penalty) for the investigated modulation formats using FFE equalizer with various tap counts is shown in Table I.

<table>
<thead>
<tr>
<th>taps</th>
<th>0</th>
<th>9</th>
<th>15</th>
<th>9</th>
<th>15</th>
</tr>
</thead>
<tbody>
<tr>
<td>QPSK</td>
<td>8 ps/nm</td>
<td>180 ps/nm</td>
<td>300 ps/nm</td>
<td>180 ps/nm</td>
<td>240 ps/nm</td>
</tr>
<tr>
<td>8-PSK</td>
<td>8 ps/nm</td>
<td>350 ps/nm</td>
<td>620 ps/nm</td>
<td>350 ps/nm</td>
<td>600 ps/nm</td>
</tr>
<tr>
<td>Star-16-QAM</td>
<td>17 ps/nm</td>
<td>550 ps/nm</td>
<td>800 ps/nm</td>
<td>350 ps/nm</td>
<td>380 ps/nm</td>
</tr>
</tbody>
</table>

IV. DIGITAL CARRIER RECOVERY IN COHERENT PSK-RECEIVERS

In coherent receivers, a local laser (LO) at the receiver front end is required, which down-converts the optical signal into or close to the baseband. Here we assume an intradyne receiver, where the nominal frequencies of both, the received signal and the local laser are almost equal, however there is no phase- or frequency locking of the LO. Thus we have to compensate for (i) the carrier frequency offset, (ii) the carrier phase, and (iii) the phase noise of the lasers involved (transmitter and receiver). We review carrier recovery schemes based on M-th power approach in 100Gb/s applications with QPSK-modulation and polarization multiplexing, where we can transmit with a Baudrate of one quarter of the bitrate, making a hardware implementation feasible in the near future. The basic operation and fundamental limits of the recovery scheme, as well as some optimization with respect to the implementation of the filters involved, are shown in the paper.

A. Simulation Setup

The receiver for a polarization multiplexing system is shown in Fig. 9. The incoming signal is split with a polarization beam splitter (PBS) into x- and y-polarization. Each polarization is then mixed separately with the local oscillator (LO). Due to some distortions on the transmission channel the received x- and y-polarization must not correspond to the two transmitted signals. The separation of the two transmitted signals is finally done during equalization in the digital signal processing unit.

The structure of the used carrier recovery is shown in Fig. 10 which is a Viterbi-and-Viterbi approach [10] for two polarisations. For each polarisation the incoming complex signal is raised to the power of four in order to eliminate the modulation $\Phi_{\text{mod}}$. These two signals are averaged and low pass filtered. The derived phase is divided by four and subtracted from the phase of the delayed incoming signal for each polarisation.

$\phi(\Phi_{\text{mod}}(kT) + \Delta \text{sc}(kT) + \Phi_{\text{noise}}(kT))$

$\Phi_{\text{mod}}(kT) + \Delta \text{sc}(kT) + \Phi_{\text{noise}}(kT)$

Fig. 11 shows the impulse response of the three investigated filter types. The first one is a rectangle (Fig. 11a), the second one provides a hyperbolic shape (Fig. 11b)) and the third one is a Wiener filter where the coefficients are dependent on the combined laser linewidth and ASE noise (Fig. 11 c) and d)) [11].

B. Simulation Results

We investigate the influence of the filter type as well as the influence of the filter order $D$ on the performance of the coherent QPSK receiver for different distortions, i.e. in the presence of ASE noise, laser phase noise and
carrier frequency offset between the received signal and the LO. In Fig. 12 especially the carrier frequency offset (detuning) is addressed whereas the OSNR is fixed at 17dB. It turns out that all three filter types perform roughly similar. A larger filter order (i.e. lower bandwidth) of $D=8...12$ is required for a good ASE noise cancellation. However if the detuning becomes larger, the filter bandwidth must be widened (by decreasing $D$) in order to be able to track a large amount of frequency offset, which results in degraded BER performance.

This drawback can be overcome by using a separate frequency recovery unit before the carrier recovery which eliminates the carrier frequency offset to zero.

Fig. 13 depicts the influence of phase noise, which is associated with a non-zero laser linewidth, for the three filter types and various filter orders $D$ and zero detuning. Up to a laser linewidth of 25 MHz the hyperbolic filter and the Wiener filter exhibits similar performance. For laser linewidth larger than 25 MHz the Wiener filter is the best choice. However, for all investigated filter types and laser linewidths a filter order of $D=8...12$ is sufficient to get an optimum result.

C. Summary

We have investigated the performance of a M-th power carrier recovery for different filter types and filter orders. It can be seen that performance decreases with increasing carrier frequency offset. The performance is almost independent of the filter type, however the filter order must be optimized. For phase noise due to laser linewidth the Wiener filter performs best for large laser linewidth and filter orders. A filter order of approximately 10 seems to be sufficient.

V. CONCLUSIONS

In future communication networks the transceiver design will increasingly require (electronic and digital) signal processing solutions. Powerful and fast electronic processors are required, where the speed requirements have to be solved with massive parallel processing. Also fast (Giga-samples/s) and accurate (word length) ADC and DAC are key components for enabling further progress and for exploiting the cost advantage of electronics vs. optics. This development is another example how advanced algorithms and methods from “Digital Communications Theory” are applied in optics as well.

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A New All-Optical Switching Node Including Virtual Memory and Synchronizer

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Abstract—This paper presents an architecture for an all optical switching node. The architecture is suitable for optical packet and optical burst switching and provides appropriate contention resolution schemes and QoS guarantees. A concept, called virtual memory, is developed to allow controllable and reasonable periods for delaying optical traffics. Related to its implementation, several engineering issues are discussed, including the use of loop-based optical delay lines, fiber Bragg gratings, and limited number of signal amplifications. In particular, two implementations using optical flip-flop and laser neuron network based control units are analyzed. This paper also discusses the implementation and performance of an all-optical synchronizer that is able to synchronize arriving data units to be aligned on the clock signal associated with the beginning time of slots, in the node, with an acceptable error.

Index Terms—optical burst switching, optical packet switching, optical synchronizer, virtual optical memory, optical delay lines

I. INTRODUCTION

Optical packet and burst switching (OPS/OBS) technologies are very promising solutions for the next generation Internet backbone [1, 2, 3]. Recently, a large activity has been performed for these technologies to build models, techniques, and tools to provide better quality of service and higher traffic control for optical communication. Unfortunately, the current state of the development of these technologies cannot offer explicit transfer guarantees. In fact, major lacks concerns the constraining QoS parameters and the contention resolution. The main reasons behind these limits are related to two points. The first one is the lack of optical memory and the second one is the inappropriateness of processing techniques applied to the optical signal.

The major contribution of this paper is four-fold: First, a novel concept, called virtual optical memory, is presented. Its implementation is discussed through the use of loop-based optical delay lines, fiber Bragg gratings, and limited number of signal amplifications. Second, this paper proposes an all optical synchronizer that allows arriving data units to be aligned on the same clock signal associated with the beginning time of slots, with an acceptable error. Third, various engineering issues related to the control, loop length, and amplification of the virtual optical memory and the synchronizer are addressed and their impact on node performance is analyzed. Finally, an optical node architecture suitable for optical packet and burst switching and QoS provisioning that builds on the architecture presented by the authors in [4] by adding mechanisms that use output full range wavelength converters [5] and sharing virtual optical memories. The proposed architecture node is a hybrid node supporting both technologies, and it operates with both (short) optical packet and (long) optical bursts at the same time. The architecture operates with long bursts from some tens/hundreds of microseconds to some milliseconds. Bursts are assumed to be segmented using conventional techniques [6].

While the architecture provides an efficient contention resolution scheme [7] and helps the provision of differentiated and optimized services to IP-based applications through the use of a set of novel traffic engineering protocols providing QoS guarantees, the virtual optical memory contributes efficiently in allowing dynamic management of the traffic parameters, admission control, contention resolution, real-time congestion control, and performances monitoring. By allowing a controllable delay on each node of an all-optical network, the contention can be resolved (or at least highly reduced) by allowing all the contending flows, except one, to be delayed within shared virtual memories. In addition, an admission control and congestion resolution can be implemented using a technique that estimates the state of the nodes based on a prior description of the traffic flowing through the nodes.

On the other hand, two schemes have been considered to make available QoS provision over the considered architecture [4]. The first scheme is a relative service differentiation based on burst segmentation and a segment-priority based contention resolution. The second scheme aims to provide an absolute QoS guarantee based on the dynamic management of traffic parameters instead of static priorities [8]. Fortunately, both schemes allow a better QoS provision and the optimization of resource utilization in the architecture proposed in this paper.

Our architecture adopts a time slotted (or synchronous) optical packet and burst switching to make
easier building and operating the network. At the input port of the node, optical bursts are synchronized on the contrary to the conventional OBS operation. This mode also allows better performances than unslotted transmission [5, 9, 10]. In a synchronous optical network, the switch fabric at each individual node can only be reconfigured at the beginning of a time slot, in the sense that it can be aware of the state of the network by collecting appropriate information at the beginning of each slot. The design of the proposed architecture has required the development of an ATM-like signaling protocol for optical data unit-based traffic handling, as well as a just-enough-time (JET)-like protocol for an optimized management of bursty traffic for the optical burst handling.

The first approaches to implement optical buffers used as memory are based on architectures using fiber delay lines (FDL) which are based only on fiber coils. As a signal cannot leave fiber until its end, several lengths are expected. This approach is too bulky. Others varieties of architectures have been proposed in the literature for the design of virtual memory based on re-circulating loop. For instance, in [11], authors presented a variable delay line using all-optical signal processing without describing its control process. Moreover, the presented architecture offers a fixed number of delay values. Our proposed architecture is able to delay a data flow for a relatively long period of time. It also allows the data flow to leave the memory almost immediately after a decision is made. To the best of our knowledge, our virtual optical memory architecture is the only one allowing a dynamic time delay values compared to existing architectures that are designed to provide only static time delay values. Our virtual optical memory is built using optical components such as optical circulator, wavelength converter, and optical amplifier. This virtual memory is able to delay a packet flow for a relatively long period of time. It also allows the packet flow to leave the memory almost immediately after a decision is made. This means that an optical data unit can be kept circulating in the virtual memory for a long period of time. However, such memory cannot authorize an infinite stay and does not authorize an immediate departure. To automatically manage this delay, a control unit is proposed. This makes the formulas developed for the blocking and waiting time within a traditional switch inappropriate for our architecture. In this paper, we propose the design of a control unit to provide an autonomic and transparent control on the activity of the virtual memory. To this end, two implementations using different technologies are proposed for the control. The first solution is based on optical flip-flop memory. The second solution uses a laser neural network (LNN). To evaluate their performances, we assess the response time of each one and its effect on the optical memory system dimensions. Integrated to the abovementioned architecture, the unit will define a virtual memory in optical packet switching network.

The remaining part of this paper is organized as follows. Section 2 presents the basics for all optical switching and networking and discusses the main existing architectures provided for optical switches and their drawbacks. Section 3 describes the proposed node architecture for packet and burst switching, develops the basic functions of each component and discusses some implementation issues. Section 4 presents the architecture of the all-optical synchronizer and its control unit. Section 5 discusses the design of the virtual memory, its control unit implementation and its performances. Section 6 concludes the paper.

II. BASICS ON NETWORKING AND SWITCHING IN OPTICAL SWITCHES

A wide variety of architectures has been proposed in the literature for the design of optical packet and burst switches and optical networks. Some of these architectures have experienced moderate success at the industrial level. In the following, we first present the generic architectures provided for OPS and OBS switches. Then, we discuss the main drawbacks of these architectures.

A. Main architectures

Three major switching techniques have been proposed in the literature for optical networks, namely the wavelength routing (WR), optical packet switching (OPS) and optical burst switching (OBS).

WR networks carry data between access stations in the optical domain without any intermediate optical/electronic conversion. This is realized by assigning a path in the network between the two stations and allocating a wavelength on all links on the path. Such an all-optical path is commonly referred to as a lightpath. Even though WR networks have already been largely deployed, they may not be the most appropriate for various communication services. In particular, it takes at least a round-trip delay to establish a lightpath and leads to poor wavelength utilization.

In OPS networks, the transported flows are broken into packets of small size before being transmitted. Routing information is added to the overhead of each packet so that intermediate nodes between the source and destination are able to forward the arriving packets. An OPS is capable of dynamically allocating network resources at the packet level granularity, theoretically, while offering excellent scalability. However, they need appropriate techniques to perform their functions, such as bit-level synchronization and fast clock recovery, as they are required for packet header recognition and packet delineation.

Optical burst networks represent an intermediate solution between WR and OPS networks. In OBS networks, the control plane and the data plane are separated, and the signaling is performed out of band. Although the control process is performed electronically, the data is transmitted in all-optical domain. This function is achieved by assembling data into bursts and transmitting the information related to a burst before it is sent.

In an OBS network, data bursts consisting of multiple packets are switched. Each burst is preceded by a burst
header packet (BHP) to configure the switches along the path followed by the burst. The latter follows the header without waiting for an acknowledgement announcing that the needed resources have been allocated. The edge nodes, in an OBS network, carry out electrical/optical and electrical conversion and burst assembly for each output wavelength channel, [12].

A generic OBS core node architecture is composed of a switching unit (SU) and a switching control unit (SCU) [13]. The SCU maintains a forwarding table and is responsible for configuring the SU. When the SCU receives a burst header packet, which is a small packet that precedes the related burst to inform the nodes about the characteristics of the burst, it identifies the intended destination and consults the forwarding table to find the appropriate output port. If the output port is available when the data burst arrives, the SCU configures the SU to let the data burst pass through. If the port is not available, the SU attempts a contention resolution based on the contention resolution scheme it implements. Among the schemes proposed to address this problem, one can mention two approaches: in the first, the ongoing burst is prioritized and the arriving burst is dropped. In the second approach, bursts are decomposed into segments and the contending segment of the lower priority burst is dropped.

Typically, the SCU is responsible for signaling, BHP interpretation, contention detection, and resolution, forwarding table lookup, wavelength conversion control, and SU control. Nowadays, the lack of fast, scalable, and robust optical bit-level processing technologies has led to the decision to process the BHP at the electronic level. This means that the majority of the developed architectures implement the SCU electronically and that all optical support takes care only of the transmission of bursts after their BHPs are processed.

On the other hand, some works have addressed the design of architecture for an OPS node. The architecture presented in [14] is basically composed of an input interface, a switching unit, an output interface, and a switch control unit. A packet arriving at a node is first demultiplexed into individual wavelengths and is sent to the input interface. The input interface is responsible for extracting the optical packet header and forwarding it to the switch control unit for processing to determine the output wavelength channel and configure the switch fabric unit to route the packet accordingly. It also delays the payload of the arriving packet using an optical delay line (ODL). Finally, the packet is reassembled by attaching the header with related payload and delivered to the output interface. In case of output port contention, the switch may need to delay the packet or transfer it to a new wavelength on the same optical path.

Contention occurs when multiple input traffic units want to be switched on the same output channel (fiber link, wavelength) at the same time. The contention problem can be addressed using three different approaches: deflection routing, optical buffering using ODLs, and wavelength conversion using wavelength converters, [5, 12, 13]. A wavelength converter may be fixed or tunable with full or limited range conversion capability. Also, they may be placed at the input and/or output of the switch. Moreover, each port of the switch may be equipped with its own dedicated converter, or may share converters with other ports.

Depending on the position of ODLs, optical switch architectures can be classified into three categories: input buffering, output buffering, and shared buffering [7, 10]. In input buffering, a set of ODLs is dedicated to each input port. In output buffering, a set of buffers is dedicated for each output port. In shared buffering, a unique buffering unit is shared by all switching ports. ODLs can be also classified into feed-forward and feed-backward. With feed-forward ODLs, a contending data unit can be delayed once; meaning that when such a packet emerges from the ODL after the specified delay, it must be switched to an output port or dropped. With a feed-backward ODL, a packet emerging from the ODL may be buffered multiple times by sending it back to the ODL. This situation may arise if the packet experiences contention.

B. QoS support

Different optical networks have been addressed to provide QoS including OBS network and OPS network, where QoS discusses how to provide differentiated services in order to support various requirements for different applications. Several schemes have been proposed to provide QoS in OBS networks and allow loss and/or delay differentiation [3]. Three QoS models can be distinguished: the relative QoS, the proportional QoS, and the absolute QoS [15, 16, 17].

To provide a relative QoS differentiation over an OBS network, an extra-offset-based scheme that provides relative loss differentiation can be used [15]. Utilizing this scheme, higher priority class bursts are given a larger offset time (which is the time separating the transmission of a BHP and the related burst) than the lower priority class bursts. By allowing a larger offset time, the probability of reserving the resources for the higher priority class bursts is increased and, consequently, the loss probability experienced by higher priority class bursts is reduced. However, even though it may offer a relative service differentiation over a bufferless OBS network, the extra-offset-time based scheme suffers from a set of shortcomings. In particular, high-priority bursts will experience higher delays and not be capable of meeting delay requirements. In addition, it has been shown that the offset-time based scheme tends to select the small bursts for low priority service classes.

To provide proportional QoS differentiation over an OBS network [16], an intentional burst dropping scheme can be introduced to provide proportionally differentiated loss probability, in the sense that a low priority burst is intentionally dropped when the proportionality equation is violated. This will give more opportunity for a high priority burst to be admitted. However, it can result in unnecessary dropping of low-priority bursts.

In addition, to provide a proportional packet delay differentiation over an OBS network, one can define an
assembling scheme allowing to keep a queue for each class of packets. A burst will be assembled and transmitted into the OBS backbone when a token is generated at time $t$. The token's generation is assumed to be Poisson in order to avoid possible synchronization among the burst generations from different sources [18].

Two mechanisms can be introduced to provide an absolute loss guarantee over an OBS network [17]; early dropping and wavelength grouping. The early dropping mechanism probabilistically drops the bursts of lower priority class in order to guarantee the loss probability of higher priority class traffic. Using this mechanism, an early dropping probability is computed for each traffic class based on the online loss probability and the maximum acceptable loss probability of the immediately-higher priority traffic class. In wavelength grouping mechanism, the traffic is classified into different groups, and a label is assigned to each group. Each group is provisioned a minimum number of wavelengths. One approach to group the traffic is to assign all traffic of the same service class to the same group.

C. Main drawbacks of the current architectures

OPS technology offers high capacity and data transparency but is currently in an experimental stage and is not largely implemented commercially, due to the lack of random access memory equivalents in the optical domain. The network control and routing is performed electronically and the routing information located in the packet header must be converted to electrical form for ensuring the routing process. While synchronization is an important issue in OPS networks, most of the proposed optical switches are synchronous and use constant length packets. This makes controlling and managing the switch easier and facilitates routing and buffering. The synchronization process is usually managed using ODL while the issue in OPS networks is the regeneration of the optical signal. Today's OPS networks cannot forward and buffer data traffic entirely in the optical domain. The optical signals are converted to the electrical form before switching and processing at each switch/router over the OPS network core.

OPS technology is not largely implemented commercially, due to the lack of an efficient optical layer management and the use of complex architecture. Since OBS networks provide connectionless transport, there is a possibility that the bursts contend with one another at an intermediate node. OBS networks usually work over bufferless optical architecture. When contention occurs, one contended burst is switched to its original path, whereas the other can be converted or deflected to an alternative path on the outgoing link, if any. If there is no available path, the entire burst is dropped resulting in a high burst loss which produces a negative effect on the OBS network behavior and QoS requirements. In addition, the use of wavelength conversion and deflection routing methods require adjusting the control plane to adapt the alternative optical path with the original path.

Another important issue in OBS networks is the design of optical burst switches insuring optical buffering. By implementing optical buffering, a buffer may be used to hold the burst for a viable time. Note that, in any optical buffer architecture, the buffer is severely limited by physical space limitations. Due to the size limitation of optical buffers, a node may be unable to effectively handle high load or bursty traffic conditions.

Finally, let us notice that the architectures developed for optical networks can be distinguished based on the mechanisms they implement to provide basic services such as QoS provision, contention resolution, and data transmission and processing. However, the majority of the existing architectures suffer from common limits. The main shared drawbacks include the following:

The architectures have limited contention resolution capability: this is mainly due to the following facts:
- The use of poor contention resolution mechanisms.
- The use of either a fixed wavelength converter or a set of tunable wavelength converters with limited conversion capacity (limited-range converter).
- The use of feed-forward ODL buffers rather than feed-backward ODLs.

The architectures are bulky: this drawback is typically due to:
- The use of dedicated ODL buffers and/or wavelength converters instead of shared resources.
- Contention resolution is performed at the granularity of a burst. This needs the use of unacceptable long ODLs.

The architectures provide very poor QoS support: this limitation is induced by the following facts:
- The high level of dropping operations.
- The absence of acknowledgement messages and reasonable buffers.
- The traffic state is not completely observed in advance, in the sense that, the traffic arriving at the ingress node is not usually predictable.

In order to alleviate these limits, the proposed optical node architecture should satisfy the following requirements:
- Be able to reduce the number of data unit droppings
- Use sharable and tunable resources to allow uniform and efficient buffering.
- Provide tools and mechanisms for the measurement and estimation of traffic parameters suitable to support QoS guarantees and dropping reduction.
- Allow dynamically controlled waiting period chosen within a wide range of variable delays.

III. A NEW ARCHITECTURE FOR ALL-OPTICAL SWITCHES

To our knowledge, a few works have addressed the design of a core node architecture that builds on OPS and OBS technologies to provide an integrated transport networking infrastructure for improving resource allocation and enhancing the switching process [19]. These works, however, do not cope with issues such as the utilization of all sorts of buffering, wavelength conversion capabilities, and QoS constraints.
In this section, we describe the major functions and components of a hybrid switch combining the advantages of the OPS and OBS technologies. The design of this architecture is highlighted by the introduction of additional optical components and new features to enable the traffic transport over a heterogeneous infrastructure.

A. Main components

Figure 1 depicts the architecture that we propose to implement optical nodes. The node is composed of N input ports and N output ports. Each port can handle w wavelengths using a set of multiplexers and demultiplexers. A wavelengths can be used either for signaling or data transport. The main components of the node architecture are: a switch fabric unit (SFU), a waiting unit (WU), a switch control unit (SCU), an input processing unit (IPU), and an output processing unit (OPU). In the following, we describe the key functions of the proposed switch.

The waiting unit: The WU is used for contention resolution occurring in the output ports. It helps QoS provision, but adds extra delays that may affect real-time services. It is composed of a set of shared virtual optical memory implementations (serving as multi-wavelengths ODL buffers). Two sizes are used for the main loop composing the virtual memory: the packet time slot duration (equal to the time slot $TS$) for packets, and another size corresponding to a duration equal to $nxTS$, where $n$ can be the number of packets a data segment can contain. The two intervals of time correspond to the minimal delay experienced by a packet, respectively segment, entering the WU. The data unit can be kept in the WU for longer periods depending on several rules that will be discussed in the sequel. This situation may arise if the contention is persistent. The WU also contains a set of full range wavelength converters used for ODL buffers conflicts resolution [20]. The implementation of the virtual optical memory will be discussed in a following section.

In introduction, we have underlined the advantages of virtual optical memory over the FDL. Considering N input ports switching node, each port supporting W wavelengths, to obtain a WU able to manage the $NxW$ inputs, two methods can be used: In the first, $NxW$ virtual optical memories are needed (as depicted in figure 1). As the virtual optical memories are based on recirculation, the use of bulky FDL component is avoided. In the second, a multiplexer/demultiplexer system can be used to make able a single virtual memory to admit traffics arriving over the N ports using the same wavelength. In this case, W virtual memories are sufficient, and bulky FDL component are avoided. In both cases, the use of virtual optical memory allows keeping the traffic for longer period of time than in bulky FDL.

The switch fabric unit: The SFU carries synchronized input optical data units to the intended output channel or to an appropriately selected virtual memory, in case of output port contention. Various schemes can implement the SFU. An example of such schemes can be based on the space switch, which arranges $n^2$ switch elements $Sij$, for $1 \leq i, j \leq n$, having two inputs and tow outputs, in an $n \times n$ matrix. The switch elements are connected in such a manner that allows $n$ optical signals are switched to $n$ output signal with blocking.
The switch control unit: The SCU is used to supervise the SFU activity. It collects the information related to the availability of each wavelength on every output port and the availability of virtual optical memories. Such information is useful for the reservation of the needed resources (e.g., output wavelengths and optical memories) and contention detection and management. The SCU creates and maintains a forwarding table and is responsible for configuring the SFU. It manages signaling packets between core nodes to optimize data unit transmission, reserves the suitable output channel and optical memories, configures the appropriate input and output processing units, and updates the forwarding table to switch the arriving unit on pre-established virtual optical circuit. The SCU supervises input/output processing units and controls the state of optical gates based on signaling information. It opens the OBG or the OPG gate depending on the incoming unit type. Today, the lack of fast, scalable, and robust optical bit-level processing technologies has led to implementation of the major functions of the SCU at the electronic level, [21].

The synchronizing unit (SU): The SU is in charge of synchronizing the input traffic. To facilitate the slotted transmission, all-optical synchronizer units allow arriving data units to be aligned on the same clock signal associated with the beginning time of slots with an acceptable error equal to a predefined value $e$. The goal is achieved by delaying the signal in a small size optical loop capable of delaying $e$ seconds and allowing the data unit to circulate in the loop $T/\epsilon$ times, where $T$ is duration of the time slot. The details of SU are discussed in the next section.

The optical synchronizer use permits the discretization of the offset time value due to the alignment on the clock signal. The virtual optical memory permits to estimate its value using conventional method that takes in consideration the similarity between electronic network and the use of virtual optical memory in optical network, for instance, the mean value analyses (MVA), the average residence time can be estimated. This estimation is used in admittance control.

The I/O processing units: An input processing unit is associated with each input channel. It is composed of a synchronizing unit that is used to synchronize the arriving data bursts and align them with switching time slots boundaries and an Electrical/Optical converter that is used for burst header packet conversion for electronic processing. An output processing unit is associated with each output channel. It is composed of a full range wavelength converter that is used for output port and an Electrical/Optical converter that is used to convert a burst header packet in the optical domain after being treated by the SCU.

B. Switching protocols

The proposed core node architecture allows the switching of optical data units, the packets and bursts. To this end, two protocols have been defined.

Packet switching protocol: Arriving packets are demultiplexed into individual wavelengths, if needed. Each packet is then processed by the corresponding IPU. Various tasks are applied. First, it passes through the synchronizing unit to be aligned to the first time slot boundary and its header is optically extracted by the optical label swapping (OLS) component. While the header is converted to electrical form, and so forwarded to the SCU for processing, the corresponding packet payload is inserted in an available virtual optical memory. The SCU reads the header information and determines the output port and wavelength involved in the data routing. In the case of output channel availability, the SCU configures the SFU to carry the packet payload to the corresponding OPU. At the same time, the packet header and payload are transmitted to the next node. The output channel unavailability leads to a contention problem which is resolved by the SCU according to the contention resolution scheme.

Burst switching protocol: The burst transmission is preceded by a control packet used to establish a virtual optical circuit (VOC). A VOC has the following format: $(VP_1, VC_1; VP_2, VC_2; \ldots; VP_n, VC_n)$ where $VP_i$ is the arriving port on node $i$, $VC_i$ is the arriving wavelength on node $i$ and $n$ is the length of the path. After admission control, the VOC is set and the nodes occurring in the part allocated are informed about the relative value of port and wavelength. The traffic is switched based on these values. When leaving a node, the control packet is equipped with an updated information related to the output port, output wavelength, and next destination. The routing table writes the entry:

\[<\text{in-port}, \text{in-wavelength}, \text{preceding node}; \text{out-port}, \text{out-wavelength}, \text{next node}>\]

The management of the VOCs is dependent on the priority of the optical data unit. While higher traffic priorities see their connections guaranteed during the transmission, lower priority units may see their VOC identifiers modified, during their travel, for the need to resolve contention. Optical burst and optical packets are treated differently in a node. In fact, bursts are preceded by information data unit which indicate the data path. But packets must be split in header and data to be processed. Gates are used to induce data units on the appropriate path. The SCU configures the appropriate IPU to receive the arriving unit. It mainly closes the optical packet gate (OPG) and opens the optical burst gate (OBG). If the intended output channel is available when the burst arrives, SCU configures the associated OPU (closes the OBG and opens the OPG), and instructs SFU to let the data burst pass through.

Like the packet switching, the output channel unavailability leads to attempt to resolve the contention of segments by the SCU.

Optical network should be able to support, process, and transfer two categories of IP traffics: traffic with variable rate and traffic with bursty condition. The main...
characteristic of the proposed node architecture is its ability to handle simultaneously multiple granularities such as the packet and burst levels. To make this function more efficient, we have proposed a novel signaling protocol and enhanced the existing signaling scheme performed in the optical switching networks [22]. For this, two signaling protocols have been proposed: the first, called ATM-like protocol, is used to set up optical paths for packet-based traffic handling. This signal scheme assumes that an arriving packet is forward in the WU while its header is processed. The second is called Jet-like signaling protocol, in the sense that the node is informed sufficiently in advance, and is dedicated to build optical paths for burst-based traffics.

Based on the lightpath created during the signaling phase, the proposed architecture develops multiple granularity schemes of switching depending on received data unit to switch. QoS provision can be performed by the use of the core node architecture and efficient contention resolution methods. A prioritized QoS differentiation has been introduced for contention resolution to enable better QoS provision based on the IP packet priority and through the use of wavelength conversion and optical buffering. This approach is improved with the use of dynamic QoS differentiation to support QoS satisfaction for applications with various constraints. With this approach, an IP packet traffic is accepted and handled based on a set of QoS requirements that the network will be able to guarantee its transfer along a transmission optical path.

**Contention resolution scheme:** There are two types of optical data units (ODU) handled in the node architecture. The first type is suited for optical burst switching. In fact, an ODU of the first type is nothing but a segment in a burst. For this, we assume that when a burst is formed at an ingress node it is logically composed as a sequence of a number of optical segments (or a finite sequence of bits). The second type of an ODU is nothing but a sequence of data packets of equal size along with a short field in its beginning that carries information about the path followed by the ODU. The ODUs are assumed to flow in a connection-oriented mode. All ODUs are supposed to have fixed size. Optical bursts are managed in a Jet-like manner. Optical packets are managed in an ATM-like manner. For both, the treatment is usually preceded by the establishment of VOC, which is assumed to be managed in an ATM-like manner.

The BHP preceding an optical burst carries information about the ODUs the burst contains. This information is used by a node receiving the burst to separate an ODU in the burst or the remaining part, of a burst, starting with an ODU. This way, a contending ODU can be delayed in a virtual memory until its output is freed; or it can be sent to its destination through an alternative path. In the latter case, the BHP has to be modified accordingly and prior to the arrival of the burst. The delay provided by the virtual memory copes well with the time constraints imposed by the operations related to relaying the contending ODUs.

On the other hand, a contending ODU of the second type arriving at a node may see the following tasks: First, the field containing the information related to the ODU’s path is separated and converted to the electronic domain while the remaining part of the ODU is move to a virtual memory and remains there until a decision is made related to its output path. Second, the information is used to check whether the output wavelength is free or whether another wavelength can be used on the same output port. In that case the ODU is recomposed and outputted. If no wavelength is available the ODU is dropped.

It is worth to notice that the contending ODU to delay can be selected based on a priority scheme associated with the ODLs or based on QoS requirements submitted at the establishment of the connection.

**Quality of service provision:** To provide QoS support for applications with diverse QoS demands, a scalable approach can be proposed based on dynamic QoS parameters. To implement such a scheme, each core node needs to maintain a set of performance parameters statistics, such as the number of ODU arrivals, the number of ODU dropped for an ongoing traffic type, the waiting delay for an ODU (or the average delay for a traffic), and the delay experienced by an ODU before arriving to the node.

Let us suppose that each traffic flow, say T_i, is assumed to require a maximum packet blocking delay, D_{i}^{\text{max}} and a maximum packet loss, L_{i}^{\text{max}} (the unit here the slot period TS). When a core node involved in the transmission receives an ODU_i under the abovementioned constraints, it will operate as follows: It computes two terms. The first, noted D_{i}^{\text{in}} is the buffering duration of the ODU_i. The second term, noted D_{i}^{\text{bf}} is determined only in the case of burst segment. It determines the time spent before arriving on that node. Then, the node computes the difference:

\[ \Delta_{i} = D_{i}^{\text{max}} - D_{i}^{\text{bf}} - D_{i}^{\text{in}} \]

and the average loss observed by the node for the traffic flowing through the node L_i. When contention occurs between two ODUs, say ODU_i and ODU_k, the node compares the differences of acceptable delays \( \Delta_i \) and \( \Delta_k\) and will delay the one who has the highest difference. If \( \Delta_i = \Delta_k \), the ODU with the lowest L_i will be delayed. Another decision rule can combine the two values.

**IV. ALL-OPTICAL SYNCHRONIZER**

We address in this section the design of a synchronizer and discuss an all optical implementation of its control unit.
A. Synchronizer architecture

The design of the optical synchronizer is based on a single loop where signal is enclosed to be delayed. The loop consists of a fiber with a single-sideband modulator and an amplifier, where the optical data unit to synchronize (or delay) will circulate a maximum of \( n \) times. This number \( n \) is such as \( \frac{TS}{n} \) is an acceptable bound for the synchronization error where \( TS \) is the time slot and it is selected for our architecture is closely related to the packet and burst durations. The delay caused by one turn in the loop is the granularity of the synchronizer, \( \epsilon = \frac{TS}{n} \). The duration of this loop is referred to as a minislot. Several recirculations can be done until the signal must be released.

The synchronizer contains four major components (as depicted by Figure 2).

The single-sideband modulator (SSB, [23, 24]): The SSB is used as wavelength shifter. An optical signal circulating in the loop arriving to the SSB with wavelength \( \lambda_i \), is shifted to \( \lambda_{i+1} = \lambda_i - \Delta \lambda \). The recirculations are done until the SSB makes the signal wavelength out of the fiber Bragg grating reflection band. The total delay of a data unit arriving at \( \epsilon \) for \( i=0,..,n-1 \), observes a delay equal to

\[
T_i = i\epsilon
\]

The fiber Bragg grating: The FBG reflects signals having wavelengths in its reflection band. To be delayed, the signal must have a wavelength in the FBG reflection band while turning in the loop. At the delivery moment, the signal wavelength must be out of the FBG reflection band. The band is assumed to contain \( (n-1) \) wavelengths \( \lambda_i \), \( 0 \leq i < n \), such that:

\[
\lambda_{i+1} = \lambda_i - \Delta \lambda, \quad 0 \leq i < n
\]

\[
\lambda_i + \Delta \lambda = \lambda_0 \text{ is out of the reflection band,}
\]

where \( \Delta \lambda \) is a given shift value related to the SSB characteristics and \( \lambda_0 \) is the extracting wavelength.

We can use several FBGs with overlapped reflection band if the FBG reflection band is not large enough to contain \( n \) different wavelengths. As shown in Figure 3, three FBGs with a reflection band of 20 nm each were used. In this case, 19 wavelengths with 3nm separation interval were available.

The selector: This device controls the optical multi-wavelength source which consists of \( n \) laser sources. It activates the \( i^{th} \) laser source during the \( i^{th} \) minislot. When an input signal is detected at \( \epsilon = \frac{TS}{n} \), for \( i=0,..,n-1 \), the running laser \( \lambda_i \) is maintained throughout a duration equal to \( TS \) (which is the period for transmitting a data unit). At the signal end, the next wavelength \( \lambda_{i+1} \) is activated.

The synchronizer control unit: This unit consists of a clock signal, an electronic selector, a multi-wavelength optical source and a semi-conductor optical amplifier (SOA). The clock signal, which period is equal to \( \epsilon = \frac{TS}{n} \), feeds the selector. One single laser source at a wavelength \( \lambda_i \) becomes active depending on the required time delay as illustrated in Figure 3. The cross gain modulation (XGM) effect of the SOA is used to convert the input signal wavelength.

In this part, we describe the whole system functioning. At the signal arrival, the selector collects from the optical detector the information needed to identify the minislot of arrival and keeps the adequate \( \lambda_i \) fixed during a whole slot time to allow the synchronization of the entire packet. The clock signal is included to the control unit to determine the minislot duration. The multi-wavelength source consists of a set of \( n \) lasers controlled by the selector. Its output is used by the SOA to convert the input signal wavelength to \( \lambda_i \). An optical isolator is inserted between the multi-wavelength source and the SOA to limit the \( \lambda_{input} \) propagation. The signal is then led to the loop by a combiner. The three-port circulator transmits the signal to the fiber Bragg grating. If its wavelength is out of the FBG reflection band, it leaves the synchronizer without being delayed. Else, the signal returns to the loop. At each passage in the loop, the signal
wavelength is shifted by $\Delta \lambda$ by the single-sideband modulator (SSB). As the SSB degrades the signal amplitude (about 5.7 dB on each passage), an erbium doped fiber amplifier (EDFA) is used to amplify the signal. However, the EDFA and the SSB decrease the signal to noise ratio (SNR). The turn number cannot exceed a maximum number of times $n$ to preserve from unacceptable degradation of SNR. The envelope of the signal can experience severe changes with loss of information due to the attenuation and dispersion effects.

Every component in the architecture adds some delay to the signal, but the most important delay is induced by the fiber, which length is proportional to the minislot duration, $\varepsilon$. After $i$ recirculations, the signal wavelength becomes $\lambda_0$ which is a wavelength located out of the FBG reflection band. The signal leaves the synchronizer after a delay equal to $T_n$, as given in the Equation 3.

The developed software platform has six main components: combiner, circulator, EDFA, SSB, fiber, FBG. The laser source is modeled as a Pseudo-Random Bit Sequence Generator with a 40 Gbits/s transmission bit rate and 1500 bytes packet size. A Mach-Zehnder modulator is used with a continuous wave laser having a power of 1 mW and a linewidth of 1 MHz, and a NRZ generator. The SSB modulator is represented by a developed Matbab co-simulator component. The FBG array is modeled by a FBG having a wide reflection band of 57 nm to cover 19 wavelengths separated by 3 nm. The optical receiver is modeled as a PIN photodetector and a low pass Bessel filter.

Having chosen the synchronization granularity $\varepsilon$ (minislot duration) as 0.05 $T_S$, the fiber length is calculated to be equal to 3 m. The SSB shift is set to $\Delta \lambda = 3$ nm.

### B. Physical implementation and simulation

The proposed architecture can be implemented using all-optical components.

We build a software platform to simulate the architecture of the all-optical synchronizer and evaluate the impact of turn number on the SNR and $Q$ parameters using Optiwave Optisystem [25] and Matlab 7.1. The equation 2 listed below gives the SNR expression after $i$ turns in the loop, where $P_n$ is the input signal power, $SNR_{SSB}$ is the SSB signal to noise ratio, $\eta_{sp}$ is the ratio of electrons in higher and lower states, $h$ is the Plank's constant, $\nu_f$ is the bandwidth that measures the noise figure and $G_{EDFA}$ is the EDFA gain. The equation 3 gives the logarithmic $Q$ factor expression after $i$ turns in the loop, which is deduced from the $SNR_i$, and where $B_0$ is the optical bandwidth of the photodetector and $B_c$ is the electrical bandwidth of the receiver filter [26].

$$SNR_i = \frac{P_n \cdot SNR_{SSB}}{i\eta_{sp} h \cdot SNR_{SSB} \cdot \nu_f (G_{EDFA} - 1) [2\lambda_0 - \Delta \lambda (i + 1)] + iP_n}$$

$$Q_i (dB) = 20 \log \sqrt{\frac{SNR_i \cdot B_0}{B_c}}$$

Figure 4 illustrates the output eye diagrams after 5, 10 and 19 turns in the loop. Distortions shown on the eye diagram demonstrate that we can synchronize arrived data units with a precision of 45 ns without reaching an unacceptable distortion value to allow signal reconstruction.

Figure 5 illustrates the $Q$ factor curves for three delays; 5, 10 and 19 turns in the loop. We note that the $Q$ factor decreases while the turn number increases. One can conclude that our system offers synchronization with a fine granularity equal to 5% of the data unit duration.
This granularity is limited by the turn number. Other simulations show that a lower granularity may cause the loss of data units.

Figure 5. The output Q factor for different turn number in the synchronizer

V. THE VIRTUAL OPTICAL MEMORY

To provide a tunable delay, the authors of this paper introduced the concept of virtual memory in [27] and proposed an implementation using electronic devices. Our approach organizes the ODL in two loops. The first loop is used to delay the data units and the second amplifies it, when needed. In this section, we discuss the design and implementation of a novel architecture implementing it. We also discuss the design and implementation of the control unit using, in particular, all optical components.

A. Virtual memory architecture

Figure 6 depicts the architecture of the optical memory. At the entering of the first loop (L1), we use one fiber Bragg grating (FBG1), which overlaps a reflection band that forces an optical data unit in loop L1 to keep circulating if the signal carrying the data unit has a wavelength occurring in the reflection band. An input signal entering the virtual memory arrives first at a passive optical coupler that informs the control unit (CU) of the data unit arrival. Then, the input signal is led to the second arm of the first circulator, where it is amplified. It passes across the first FBG, as its wavelength is external to the reflective Bragg band (say \( \lambda_{signal} = \lambda_0 \)). Arriving at the first arm of the second circulator, the signal gets out from the second arm.

The wavelength converter, here a SOA-based system, modulates its input continuous wave and outputs a signal on a different wavelength, [28]. When it detects the signal for the first time, it shifts the signal wavelength to a wavelength in the reflection band of FBG1 and FBG2 (\( \lambda_{signal} = \lambda_1 \)) to confine the signal within the first loop L1. The control unit evaluates whether an amplification operation is needed according to the information previously collected (from the data unit source, for example). When required, the CU shifts the signal wavelength out of the reflection band of the second FBG (\( \lambda_{signal} = \lambda_2 \)) to relay the packet to the second loop (L2) where it is amplified. When the delay duration is elapsed (or more generally, when the QoS of service is decreased), the control unit commands the wavelength converter to shift the signal wavelength to \( \lambda_0 \). As a result, the wavelength signal leaves the first FBG reflection band (\( \lambda_{signal} = \lambda_0 \)) and the signal passes to the third arm of the first circulator.

As the signal is confined in the first loop (L1), some distortions occur caused by the propagation through the various components of the virtual memory (e.g. optical fiber, wavelength converter, and FBGs). The number of rotations that the signal can experience depends obviously on the fiber length and cannot exceed a certain value because of unacceptable signal loss. Each time the signal power reaches a predefined limit, the signal is moved to the second loop, where it is amplified. The number of amplifications cannot, however, exceed a certain number due to unacceptable degradation of the SNR. This will induce the rejection of the signal from the virtual memory.

An optimized good implementation of a virtual memory needs to maximize the recirculation time within the virtual memory and minimize the fiber length in the first loop L1.

Figure 6. Architecture of the all-optical virtual memory
B. Control unit design and implementation

Several works have discussed the use of ODLs in optical communication. However, only few works have been denoted to the study of the control unit. In [29], Yao et al proposed an ODL using a SOA-wavelength converter, which is controlled by a static continuous signal; but do not design any system that can generate a wavelength control signal according to the required delay. In [30], the authors of designed an ODL using a Single-Side Band wavelength converter (SSB-WC). The time delay spent in the loop depend of the SSB-WC, which is controlled in a static manner.

Since wavelength conversion requires a continuous-wave light to be injected into the wavelength converter simultaneously with the data packet, we focus on the design of an autonomic and automated control unit. We assume that the main function of the control unit is to decide the path to be followed by the delayed signal.

i) Control unit design

The control unit is the main component that controls the signal path by delivering a synchronous signal to pilot the wavelength converter. It generates three kinds of wavelengths at different moments. The first wavelength, \( \lambda_1 \), is generated to confine the signal in the first loop and start a delay. When the signal power reaches a critical level, the signal must leave the first loop. The control unit generates the second wavelength, \( \lambda_2 \), to lead the signal to the second loop where it is amplified. The control unit calculates according to the previously received delay information the exit moment and generates a third wavelength, \( \lambda_0 \), to free the signal. Figure 7 shows the design of the control unit. Three main components are important in the architecture:

- **The calculator**: it computes the number of rotations in the first loop (k) and the amplifications number (m). The calculator output is an electronic signal \( S_0 \) shown in Figure 7 (b), where \( k.T_{L1} \) is the time spent by the signal to achieve k rotations (before it gets out from the loop L1) and \( T_{L2} \) is the time spent by the signal in the second loop L2. The shape of the signal \( S_0 \) depends on the specifications of the component used as multi-wavelength laser source.

- **The synchronizer**: when the signal \( S_{\text{input}} \) arrives at the input of the virtual memory, it crosses a passive coupler to inform the control unit of its arrival. The control unit must start generating a continuous wave intended for the wavelength converter. The synchronizer (which is different from the one discussed in the previous section) commands the calculator to start the signal generation at this moment.

- **The multi-wavelength laser source**: this component receives the signal \( S_0 \), which is an electronic control signal, and generates the optical signal \( S_{\text{out}} \), which is a continuous signal carried by a wavelength in \( \{ \lambda_0, \lambda_1, \lambda_2 \} \). The multi-wavelength laser source can be chosen from the existing components. However, its implementation should pay attention to its response time, since it can affect the system performances.

![Figure 7. Design of virtual all-optical memory control unit](image-url)

ii) Multiwavelength laser source implementations

Several technologies can be used to provide a wavelength generator for the control unit [11, 31]. Among the most important technologies, one can distinguish two components; the optical flip-flop memory and the laser neural network.

The optical flip-flop (OFF) consists of two coupled lasers [32]. Laser 1 emits at wavelength \( \lambda_1 \) and laser 2 at \( \lambda_2 \). One laser acts as the master and suppresses lasing action in the other laser, which acts as a slave. The role of master and slave can be interchanged due to system symmetry. The OFF can be in one of two states; in state 1 the output light is \( \lambda_1 \) and in state 2 the output light is \( \lambda_2 \). The input of optical flip-flop contains two ports, set and reset. To change the wavelength, a short pulse is injected in the corresponding port.

Figure 8 depicts how the control unit uses the OFF, where \( \lambda_1 \) is equal to \( \lambda_i \) if the OFF state is 1 and to \( \lambda_i \) if the OFF state is 2. So, the calculator output \( (S_0) \) must be adapted to the OFF input. \( S_0 \) becomes a series of pulses which is put to high level every \( k.T_i+T_{L2} \). Each pulse is divided in two parts. The first part of the pulse is injected directly to the OFF and commands the flip flop to emit \( \lambda_i \). In this case data will be confined in the loop L1. The second part of the pulse is delayed a period of time corresponding to the spent time to cross the loop L1. When this time expires, the second part of the pulse changes the state of the optical flip-flop to emit \( \lambda_2 \). The data will be then sent to the second loop L2.

The switching time of the optical flip-flop between the two states is less than 200ps, allowing good performances to the virtual optical memory. However, since the OFF is capable of managing two wavelengths (\( \lambda_i \) and \( \lambda_{2i} \)), a third wavelength has to be used to extract the signal from the virtual memory, when it is needed. To do so, another WC can be inserted on the output of the original WC to provide the extracting wavelength.
(\lambda_0). The extra WC is only activated as soon as the signal gets out from L2 after m amplifications and the arrival of the last pulse in S_o.

\[ S_o \xrightarrow{\text{Set}} \text{Optical flip flop} \xrightarrow{\lambda_i} \text{WC} \]

Figure 8. Wavelength converter control by an OFF

The second technology that can be used for the wavelength generation is the laser neural network (LNN, [33]). It can be used as a generator of continuous-wave light at a specific wavelength within an all optical control unit. The LNN consists of three coupled ring lasers (laser1, laser2 and laser3). The lasers emits respectively at wavelength \lambda_0, \lambda_2 and \lambda_2. It has two inputs ports and one output. According to the inputs combination, the output signal has three states. If no light is injected in both inputs, the output is at state 1 and the LNN output \lambda_0. If light is injected only in one of the two inputs, the output is at state 2 and the LNN output is \lambda_2. Finally, if we inject light in both inputs, the output signal is at state 3 and the LNN output is \lambda_2. The switching time of the laser neural network between two states is equal to 1.95\mu s.

To use the LNN as tunable wavelength generator, the calculator output (S_o) must have two components: S_{o1} and S_{o2}. The first component S_{o1} is injected to the set input of the OFF, and the second S_{o2} is injected to the reset input. According to the calculator operations, the data units are carried by the LNN based on the following three operations:

- If the calculator selects the state 2 for the LNN, every arriving signal is carried by \lambda_2 into loop L1.
- If the calculator commands the state 3 for the LNN, the circulating signal wavelength is converted to \lambda_2 and will leave loop L1 to L2.
- If the calculator commands the state 1 for the LNN, the circulating signal wavelength is converted to \lambda_0 and the data unit will be extracted.

C. Performance evaluation

The proposed optical virtual memory can be used to allow delaying the optical data units for sufficiently long duration (about five hundred times the packet time). This can be done by having sufficiently large size of fiber length. The virtual memory performances evaluation is realized in two steps. The first step consists in simulating the system effect on the signal quality. The second step aims to comparing the performances of both control unit physical implementations by evaluating their respective loop efficiency.

Using Optisystem, a software platform is built to simulate the architecture of the virtual memory and evaluate the impact of turn number on the SNR and Q parameters. Equation 4 listed below gives the SNR expression after i turns in the loop, where \nu is the total loss after k turns in the first loop L1, P_{in} is the input signal power, SNRSSB is the SSB signal to noise ratio, \eta is the ratio of electrons in higher and lower states, h is the Planck’s constant, V_f is the bandwidth that measures the noise figure and G_{EDFA} is the EDFA gain.

\[ SNR = \frac{B_f}{2 + \frac{P_{in}(A_2 G)}{2 \eta h B_f N (G_{EDFA} + G_{EDFA} - 2)}} \]

(4)

Three other parameters have been evaluated to dimensioning the virtual memory: the total time delay, dispersion and power attenuation. These parameters can be expressed respectively by the equations 5, 6 and 7. T_i is the time spent by the signal in the virtual memory after i turns. T_{FBG}, T_{WC} and T_{EDFA} are respectively the time response of fiber Bragg gratings, wavelength converter and EDFA. m is turn number in the loop L2. The fiber length, refractive index and light celerity are respectively L, n and c. A_2 is the total signal power attenuation after i turns. A_{FBG}, A_{WC}, A_{EDFA} and A_{EDFA} are respectively the attenuation caused by the fiber Bragg gratings, wavelength converter, fiber and circulator. D_i is the total signal dispersion after i turns. D_{FBG} and D_{EDFA} are respectively the dispersion caused by the fiber and circulator. B_{sp} is the signal bandwidth.

\[ T_i = \left(2T_{FBG} + \frac{L_n}{c}\right) + 2(m + 1)T_{WC} + mT_{EDFA} \]

(5)

\[ A_i = i(2A_{FBG} + A_{FBG} + 4\Lambda L + 4A_{EDFA}) + 2A_{WC} \]

(6)

\[ D_i = iLB_{sp}D_{FBG} + (2m + 4\Lambda)D_{EDFA} \]

(7)

The software platform has five main components: the circulator, EDFA, WC, fiber, and FBG. The laser source is modeled as a Pseudo-Random Bit Sequence Generator with a 10Gbits/s transmission bit rate and 1500bytes packet size, a Mach-Zehnder modulator is used with a continuous wave laser having a power of 1mW and a line wide of 1MHz, and a NRZ generator. The WC is represented by a developed Matlab co-simulator component. The used FBGs have overlapped reflection band of 10nm. The optical receiver is modeled as a PIN photodetector and a low pass Bessel filter.

Having chosen the elementary delay of 3.16\mu s, the fiber length is equal to 630m.

Figure 9 illustrates the output eye diagrams after 80, 180 and 196 turns in the virtual memory, respectively. Distortions shown on the eye diagram are due to the SNR degradations caused by the amplifier and distortions caused by the fiber. According to these eye diagrams, one can say that the data units can be buffered for a maximum time delay about 622\mu s without reaching an unacceptable distortion.
Figure 9. The output eye diagram for different turn number in the virtual memory

Figure 10 illustrates the Q factor curves for three delays; 80, 180 and 196 turns in the memory. The maximum Q factor decreases while the turn number increases. One can conclude that our system behaves as a memory where a data unit can be buffered for a relatively important time (622 µs) and it can be delivered after a relatively reduced delay (3.16 µs).

To find the best control unit design for our optical virtual memory, we carried out some comparisons between the proposed technologies for the control unit implementation. Table 1 present the performances and the limits of the two types of control in terms of granularity, minimum loop length, maximum storage time and loop efficiency. The granularity is the minimum time that can take data in the virtual memory. The loop efficiency is the ratio between packet time and the system granularity given by equation 8 where $t_b$ is the time of control in the loop and $g$ is the system granularity. These results are obtained when data rate is 10 Gbit/s with a 1500 bytes packet length. Note that the granularity ($g$) is at minimum equal to 1.22 µs with the OFF based control and minimum equal to 3.16 µs with the LNN based control.

$$L_{eff} = 1 - \frac{t_b}{g}$$

Figure 11. Loop efficiency versus loop length

Optical flip-flop based unit shows the best performances in term of granularity, and loop efficiency. In the other hand, it cannot allow three different wavelengths for control. In this term, LNN based unit is more suitable for our optical virtual memory since it allows a longer delays. However, there is a cost to pay: the efficiency for the loop is reduced.

In order to quantify the limitation for both control unit types, we plot in figure 11 two curves which describe the variation of loop efficiency versus loop length. This figure shows three facts : (a) the loop efficiency of the OFF based control is better, (b) the OFF based control efficiency becomes fastly quasi constant, (c) when the fiber length is important (over than 7 Km), the efficiency becomes comparable.

Optical flip flop memory has the best loop efficiency of 98%. However, LNN has good result (95%) when loop length is very extended (8km). But, when loop length become very short (630m) optical flip flop memory is more suitable than the other method for our optical virtual memory.
VI. CONCLUSIONS

In this paper, we have presented an optical switching node architecture suitable for contention resolution and QoS provisioning. The architecture includes a virtual all-optical memory and an all-optical synchronizer. Both buffering mechanisms are based on the use of fiber Bragg gratings and all optical control units.

The virtual memory allows delaying a data flow for a relatively long period of time. It also allows the data flow to leave the memory almost immediately after a decision is made. We demonstrate that the architecture can be implemented using optical components and its control unit using OFF or LNN technologies. Simulations prove that the memory have good performances.

The synchronizer allows aligning arriving signals on slots to improve the switching node performances. Using simulations, we prove that synchronizer architecture can be implanted using relatively low synchronization errors.

A multi-wavelength memory based on a multiplexer/demultiplexer system is actually under development. This system will enable a single virtual memory to admit traffics arriving over the N ports using the same wavelength. So, the control unit will be less bulky. The proposed node architecture will be improved by introducing multi-wavelength optical synchronizer to improve the contention resolution. Studies will be extended by an evaluation of the node performance with respect to the packet/burst blocking probability, under the discussed contention resolution and QoS schemes.

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Electronic Predistortion for Compensation of Fiber Transmission Impairments - Theory and Complexity Considerations

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Abstract—Electronic predistortion is a powerful technique to combat transmission impairments on fiber-optic communication links. The principle behind electronic predistortion will be revised and different methods to implement this method will be explained. In addition to that, we will focus on the implementation complexity of this scheme and we will highlight how the complexity scales for increasing channel data rates.

Index Terms—Electronic Mitigation, Electronic Compensation, Electronic Predistortion, EPD, Chromatic Dispersion, Fiber Nonlinearity, Complexity.

I. INTRODUCTION

Over the last few years, electronic signal processing methods for the mitigation of transmission impairments in fiber-optic communication systems have attracted a lot of attention. Some particular advantages compared to optical compensation techniques are an increased flexibility and possible adaptation to changing link conditions as well as ease of integration. In addition to that, these techniques are now becoming feasible in practice due to recent advances in highspeed electronics.

Different studies have targeted the compensation of chromatic dispersion (CD) after direct-detection. Some of the proposed techniques are based on digital filtering using feedforward- (FFE) and decision-feedback (DFE) equalizers. This approach, however, is very limited due to the modulus-square operation of the photodetector and the inherent loss of the signal phase. Other techniques subject to research fall into the category of probabilistic estimation algorithms. The most popular approach in this category is Maximum-Likelihood Sequence Estimation (MLSE) [1].

The loss of the signal phase in direct detection has lead to an increased interest in transmitter-side compensation techniques. In this case the optical field is pre-processed in the electrical domain and then converted into the optical domain by using Modulator-structures capable of controlling both amplitude and phase. This technique is referred to as electronic predistortion (EPD) [2] [3].

In the following, we will revise the theory behind EPD. Two different approaches for implementing EPD will be described. One of these approaches is applicable when the transmission can be seen as a linear system with negligible nonlinear effects. In addition to that we will present an alternative method, which can also precompensate nonlinear transmission impairments.

Most of the work reported on electronic predistortion has concentrated on systems operating at 10 Gbit/s. In the meantime 40 Gbit/s system are being deployed and 100GigE is under research. We will therefore discuss the implications that the move towards higher bitrates has on the complexity of electronic predistortion schemes as well as possible upgrade scenarios.

II. PRINCIPLE OF ELECTRONIC PREDISTORTION

When a signal propagates through an optical fiber link, it experiences a certain temporal and spatial evolution. The idea behind EPD is to determine the optical signal \( E_{\text{pre}}(t) \) that will evolve into the undistorted target signal \( E_{\text{tar}}(t) \) while it propagates through a fiber-optic link. This is illustrated in Fig. 1. The link is composed of a sequence of various components \( G_{\text{comp}} = [G_{\text{comp},1}, \ldots, G_{\text{comp},N}] \), each of which has a certain transfer characteristic

\[
E_{\text{out},n}(t) = G_{\text{comp},n}(E_{\text{in},n}(t)).
\]  

(1)

If it is possible to invert the transfer characteristics \( G_{\text{comp},n} \) of all individual components, then it is possible to invert the transfer characteristics of the entire link. The most important component of the link is optical fiber. The nonlinear propagation of the signal \( E(z,t) \) and hence the transfer characteristic are modelled by using the

![Figure 1. Example system with predistorted signal corresponding to a 10 Gbit/s NRZ target signal.](image-url)
Nonlinear Schrödinger-Equation (NLSE) [4],
\[
\frac{\partial}{\partial z} E(z, t) = \left( \frac{\beta_2}{2} \frac{\partial^2}{\partial t^2} + \frac{\beta_3}{6} \frac{\partial^3}{\partial t^3} + \frac{\alpha}{2} \right) E(z, t) - j \gamma |E(z, t)|^2 E(z, t),
\]
where \( \beta_2 \) and \( \beta_3 \) are the 2nd and 3rd order dispersion parameters, \( \alpha \) reflects the attenuation of the fiber and \( \gamma \) is the nonlinear coefficient. Normally, the input signal to the fiber is \( E(z=0, t) = E_{in}(t) \) and the output signal is found by numerically computing the output \( E_{out}(t) = E(z=L_f, t) \). In this case, the transfer function \( G_{\text{fiber}}(\cdot) \) is \( G(\cdot) \). In order to invert the transfer function, i.e. to find \( E_{in}(t) \) for a certain fiber output signal \( E_{out}(t) \) one has to reversely integrate the NLSE by using the transformation
\[
\hat{z} = L_f - z.
\]
As a consequence, the partial derivative also has to be transformed
\[
\frac{\partial}{\partial \hat{z}} = \frac{\partial}{\partial z} \cdot \frac{\partial \hat{z}}{\partial z} = -\frac{\partial}{\partial z}.
\]
The reverse NLSE then is just the forward NLSE with negative parameters, i.e.
\[
\frac{\partial}{\partial \hat{z}} E(\hat{z}, t) = \left( -\frac{\beta_2}{2} \frac{\partial^2}{\partial t^2} - \frac{\beta_3}{6} \frac{\partial^3}{\partial t^3} + \frac{\alpha}{2} \right) E(\hat{z}, t) + j \gamma |E(\hat{z}, t)|^2 E(\hat{z}, t).
\]
In this way, the signal attenuation is transformed into a gain and the operators working on the phase of the signal induce a rotation into the opposite direction. To obtain the input signal \( E_{in}(t) \) for a certain output \( E_{out}(t) \), set \( E(\hat{z} = 0, t) = E_{out}(t) \) and apply the Split-Step Fourier Method [4] to obtain \( E_{in}(t) = E(\hat{z} = L_f, t) \).

The inverse transfer function \( G^{-1}_{\text{filter}}(\cdot) \) is expressed by [5]. Please note, that the previous representation holds for all fiber types, in which signal propagation is described by the scalar NLSE, i.e. if PMD is negligible. The compensation of PMD by using EPD requires a polarization-diversity approach as shown in [5].

In comparison to the optical fiber, the modelling of the other link components is relatively simple. An optical amplifier, for instance, is modelled as multiplication by a constant, \( E_{out}(t) = \sqrt{A_0} \cdot E_{in}(t) \), where it is assumed that the gain profile is constant within this channel. In addition to the gain the amplifier provides, it also adds Amplified Spontaneous Emission noise (ASE) to the signal. Since the process of adding noise cannot be inverted, the inversion \( G^{-1}_{\text{amp}}(\cdot) \) of the optical amplifier is done by using the inverse multiplication factor, \( E_{in}(t) = \left( \sqrt{A_0} \right)^{-1} \cdot E_{out}(t) \). In addition to that, filters in the link, e.g. due to Add-Drop Multiplexing (OADM) may affect the signal if its spectrum is broader than the passband of the filters. An inverse of the filter function \( G_{\text{filter}} \) has to be computed in this case using \( G^{-1}_{\text{filter}} := H_{\text{filter}}^{-1}(\omega) \) if the filter frequency response is \( H_{\text{filter}}(\omega) \).

When the inverse transfer functions of all the link components are available then the predistorted signal can be computed,
\[
E_{\text{in}}(t) = G^{-1}_{\text{comp},1} (E_{\text{out}}(t)) = G_{\text{comp},1}^{-1} \left( \cdots G_{\text{comp},N}^{-1} (E_{\text{out}}(t)) \right).
\]

After the signal has been computed it is necessary to convert the components of the complex field into the optical domain. An arbitrary complex optical field can be created by applying phase- and amplitude-modulation to the output \( E_0 \) of a CW-Laser. The modulator type that has proven most suitable is the Triple Mach-Zehnder modulator (also called I/Q-modulator or Cartesian-modulator), which allows to control the I- and Q-component of the optical field individually by using two driving voltages \( d_I(t) \) and \( d_Q(t) \).

\[
E(t) = \frac{E_0}{2} \left( \cos \left( \frac{d_I(t)\pi}{V_\pi} \right) + j \cos \left( \frac{d_Q(t)\pi}{V_\pi} \right) \right).
\]

One particular advantage of this structure is the approximately linear translation from the electrical into the optical domain when the modulator is operated in the linear regime of the cosine-function. Other structures, such as a single dual-drive Mach-Zehnder modulator or a concatenation of phase and amplitude modulators exhibit highly nonlinear transfer-functions, which can lead to increased electrical bandwidth requirements [6]. The schematic model of the transmitter is shown in Fig. 2a). The digital signal processing block computes the predistorted signal, which is then D/A converted to yield analog driving voltages for the modulator I- and Q-branches. The D/A-conversion plays an important role in the generation of the driving signal. Two parameters that are particularly critical are the resolution, i.e. the number of voltage steps the converter can produce, as well as the conversion-rate. The conversion rate needs to be high enough such that all spectral components of the predistorted signal can
be produced. In case the bandwidth of the predistorted waveform exceeds the Nyquist-frequency $f_{\text{Nyq}} = \frac{1}{2T_{\text{samp}}}$ of the D/A conversion an antialiasing filter has to be applied before the signal is downsampled. So far, the principle behind EPD has been shown by applying the link inversion to a predefined analog target signal. However, in a practical transmission system this link inversion has to be carried out in real time depending on the datastream to be transmitted. In the following we will describe two methods, which can perform the predistortion for a continuous stream of data.

### III. Linear Filtering

If chromatic dispersion (CD) is the dominant transmission impairment that shall be compensated then the fiber model can be significantly reduced from (2) to a linear transfer function accounting for CD only,

$$H(\omega, L) = \exp \left( -j \frac{\beta_2}{2} \omega^2 L \right),$$

which is easily inverted:

$$H^{-1}(\omega, L) = \exp \left( +j \frac{\beta_2}{2} \omega^2 L \right).$$

Please note that the third order dispersion terms have been omitted since this does not cause any significant penalty for the Baudrates considered. Due to the linear transfer function the compensation is a linear filtering operation, which is easily inverted:

$$w[k] = h[k] \cdot w[k],$$

where $w[k]$ is a window of the same length $K_{\text{taps, max}}$ as the FIR-filter $h[k]$. We have tested Hamming, Hanning and Blackman window functions, which are widely used for the design of FIR-filters [8]. However, the evaluation by simulations has shown that these windowing functions may even deteriorate the performance because their tails decay too rapidly. Therefore, we have used another window, which allows to continually modify the rapidness of the cut-off by using a roll-off factor $\alpha$.

The window is described by the following expression:

$$f \leq |kT_{\text{samp}}| < (1-\alpha_w)T',$$

$w[k] = \cos \left( \frac{\pi}{\alpha_w} \left| \frac{kT_{\text{samp}}}{T'} - 1 + \alpha_w \right| \right)$

for $(1-\alpha_w)T' \leq |kT_{\text{samp}}| \leq (1+\alpha_w)T'$,

$0$ for $|kT_{\text{samp}}| > (1+\alpha_w)T'$,

where $T' = k_{\text{max}}T_{\text{samp}}/2(1+\alpha_w)$. This window leads to a significant reduction of the ripple in the magnitude and group delay response, which can be seen in Fig. 3.

The truncation of the impulse response corresponds to a multiplication by a rectangular window. This leads to a distortion of the ideal inverse fiber frequency response as shown in Fig. [3] The frequency response in this case exhibits ripples both in the magnitude and group delay response. One common approach to solve this problem is to multiply the tap-vector by a window-function [8],

$$h_{\text{w}}[k] = h[k] \cdot w[k],$$

where $w[k]$ is a window of the same length $K_{\text{taps, max}}$ as the FIR-filter $h[k]$. The taps of the FIR-filter are set to the impulse response sampled at times $kT_{\text{samp}}$, 

$$h[k] = h(kT_{\text{samp}}, L)$$

$$= -\frac{1}{\sqrt{2\pi\beta_2 L}} \cdot \exp \left( -j 2\pi \left( \frac{t^2}{4\pi\beta_2 L} \right) \right),$$

where $\mathcal{F}^{-1}\{\cdot\}$ denotes the inverse Fourier transform. The taps of the FIR-filter are set to the impulse response $h[k] = h(kT_{\text{samp}}, L)$.

This leads to

$$k_{\text{max}} = \left\lfloor \frac{\pi \beta_2 L}{T_{\text{samp}}^2} \right\rfloor$$

and consequently the maximum number of taps that fulfills this criterion is

$$K_{\text{taps, max}} = 2 \cdot k_{\text{max}} + 1.$$
ability to represent pulse shapes by such a small number of samples per symbol is rather limited, the complex amplitude of the target signal is usually kept constant for all samples within one symbol interval. For an On-Off keying (OOK) target modulation all samples for a bitslot that corresponds to a bit “1” to be transmitted are set to \(A_0\) and all samples in a bitslot that corresponds to a “0” are set to 0.

\[
e_{\text{target}}[nM+l] = \begin{cases} A_0 & \text{for } u[n] = 1 \\ 0 & \text{for } u[n] = 0 \end{cases}, \quad (17)
\]

where \(l = 0, \ldots, M\) and \(u = [u_0, \ldots, u_N]\) is the data vector to be transmitted. The extension to other modulation formats is straightforward. For higher order modulation format the previous assignment has to be carried out on a per symbol basis, where one symbol consists of a bit-tupel.

We have so far only concentrated on the compensation of chromatic dispersion. However, compensation of all other components with a linear transfer function can be included in the tap weights as well (unless there are spectral zeros). In this way, for example, improved transmission was demonstrated for a system, in which the

ISI keying (OOK) target modulation all samples for a bitslot for all substrings \(u_n\) in \(u\) of length \(K_{\text{ISI}}\),

\[
u_n = [u_{n-K_{\text{ISI}}}, \ldots, u_n, u_{n+K_{\text{ISI}}+1}].
\]

All bits in the bitstream \(u\) that are separated further than \((K_{\text{ISI}} - 1)/2\) from the bit \(u_n\) under consideration do not influence the required predistorted signal in the corresponding symbol slot \([n - 1]T_{\text{symb}}, nT_{\text{symb}}]\).

In this way, it is sufficient to compute the inverse propagation of the target signal for all \(2^{K_{\text{ISI}}}\) possible bit-tupels of length \(K_{\text{ISI}}\). In a practical implementation, the inversely propagated signal \(E_{u_n}(t)\) can be stored in a look-up table (LUT) at the address corresponding to the bit tupel \(u_n\). We assume that the interference is symmetric around the center symbol. Therefore only odd \(K_{\text{ISI}}\) is considered. In order to compute all entries of the LUT at once a PRBS-sequence of order \(K_{\text{ISI}}\) can be used. The LUT is then filled using a sliding window on the signal and storing the signal parts at the respective LUT-addresses.

An example of a LUT for the case of \(K_{\text{ISI}} = 3\) is depicted in Fig. 4. You can see the inversely calculated input signals for the target output signals corresponding to all \(2^{3}\) possible bit combinations in NRZ-OOK format. When the transmission of a stream of bits is carried out by using this LUT, only the part of the signal \(E_{u_n}(t)\) corresponding to the central time-slot of width \(T_{\text{symb}}\) is used. The bits in the stream are shifted into transmitter buffer of length \(K_{\text{ISI}}\) and the buffer content is then used to address the LUT. In this way, different pieces of predistorted optical signals are concatenated to generate the optical transmission signal of the entire bit-stream, i.e.

\[
E_{\text{prec}}(t) = \begin{cases} E_{000}(\tau), u_n = [0, 0, 0] \\ E_{001}(\tau), u_n = [0, 0, 1] \\ \vdots \\ E_{111}(\tau), u_n = [1, 1, 1] \end{cases}, \quad (18)
\]

where \(\tau \in [-T_{\text{symb}}/2, T_{\text{symb}}/2]\) and \(t = (n - \frac{1}{2}) T_{\text{symb}} + \tau\).

For illustrative purposes the signals stored in the LUT so far have been described as continuous waveforms. For practical implementation the driving voltages correspond-

<table>
<thead>
<tr>
<th>LUT Address</th>
<th>Predistorted optical signals</th>
</tr>
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<tbody>
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<td>![Predistorted optical signals](0 0 0)</td>
</tr>
<tr>
<td>1 0 0</td>
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<td>1 1 1</td>
<td>![Predistorted optical signals](1 1 1)</td>
</tr>
</tbody>
</table>

Figure 4. Example for a 3-bit LUT. The dashed box marks the part of the signals that is actually transmitted for a specific bit tuple in the transmitter buffer.
ing to \( u_k \) have to be sampled and quantized to match the rate and resolution of the available D/A converter before they are stored in the LUT.

Contrary to the linear approach described in the previous section, the LUT based approach is very generic and can be applied to linear as well as nonlinear effects. An additional difference to the linear filtering method is that the target waveform can actually be specified by using a better time resolution. Therefore it is possible to predefine the pulse shape of the target signal.

V. COMPLEXITY OF ELECTRONIC PREDISTORTION

In principle, the EPD-transmitter that has just been explained can compensate for an arbitrary amount of CD or combination of CD and intrachannel nonlinearity. There have been a range of investigations showing that the performance of EPD greatly depends on the specifications of the hardware that is used in the EPD transmitter. The most crucial elements are highlighted in Fig. 2.

The signal processing block is one of the critical components with respect to complexity. In a practical system this can be either implemented as an application-specific integrated circuit (ASIC) or by using a Field-Programmable Gate-Array (FPGA). For the first method we have described, the linear precompensation, the complexity of this device is mainly driven by the number of complex taps necessary for the FIR-filter to compensate for the chromatic dispersion. One complex tap can be realized by four real-valued filters as shown in Fig. 2c). The upper bound on the number of taps is given by \( \binom{2N}{N} \). However, it is also possible to further reduce the number of taps such that the response of the predistortion filter is matched to the spectral range, which contains the major portion of the target signal’s power.

We have carried out numerical simulations to determine the minimum required number of taps for a certain link distance. In these simulations an EPD transmitter was used to compensate for 1000 km of standard single-mode fiber (SSMF) with \( D = 16 \text{ps/nm/km} \). We have used a De Bruijn sequence of order 12. A D/A converter rate of 21.4 GSamp/s and \( m_Q = 6 \) bit resolution was assumed and the bandwidth of the Mach-Zehnder modulator was modelled using a 5th-order Bessel-filter with a cut-off frequency equal to the symbolrate, \( B_{\text{MZM}} = R_s \).

The receiver consisted of an optical 2nd-order Super-Gaussian-filter with a 3dB-bandwidth of \( 3R_{\text{symb}} \) and a PIN-photodetector followed by an electrical 5th-order Bessel-filter with a cut-off frequency of 0.75\( R_{\text{symb}} \). The target modulation formats that the predistortion transmitter was set to were OOK, DPSK and DQPSK. For the differential phase shift keying modulation formats the receiver included the appropriate Mach-Zehnder delay-interferometer structures for detection.

The results of the simulations are depicted in Fig. 5. One observation that can be made is that the predistortion without windowing of the filter taps results in significant fluctuations of the results. This effect is particularly visible for OOK and we attribute it to the large power concentration of the signal power on the center frequency for OOK. When the magnitude response of the FIR-taps exhibits a notch at the center frequency then the performance is degraded significantly. By using windowing (in this case with \( \alpha = 0.3 \)) the ripples are reduced, hence improving performance. Due to the windowing the bandwidth of the filter is slightly reduced, which has the effect that in the absence of windowing the number of taps for OOK can be even further reduced.

The performance for DPSK and DQPSK does not fluctuate as much as for OOK. The reason for this is that the power spectral density of DPSK and DQPSK is more spread out around the center frequency.

One common observation that can be made for all modulation formats is that a significant reduction in the number of taps from the maximum number \( k_{\text{max}} \) is possible at moderate penalty. The results for OOK target format show that a reduction to 39 taps is possible for a penalty that remains within 1dB of the Back-to-Back req. OSNR. For the case of DPSK and DQPSK the number of taps can be reduced to 43. Contrary to the result for OOK,
the required OSNR for DPSK and DQPSK is degraded steadily once the number of taps is reduced from $k_{\text{max}}$. In Fig. 6 the number of required taps for higher Baudrates is shown. Due to the previous results we assume that the number of taps can be reduced to $\approx 0.7k_{\text{max}}$. The number of equalizer taps grows $\sim 1/T_s^2\Delta f_{\text{amp}}$. Hence it also increases quadratically with the Baudrate. In addition to the increase in the number of taps for higher baudrates the computation of the filter output also needs to be performed within a a shorter time interval. Doubling of the Baudrate results in an increase of the complexity by a factor of 8 due to a fourfold increase in the number of taps and a reduction of the computation interval by a factor of 2. Therefore in order to increase the bitrate of the system it seems to be a better solution with respect to complexity to use a higher order modulation format.

We have carried out further simulations to evaluate the potential of using DQPSK to double the bitrate compared to a binary transmission. The simulation was carried out for OOK, DPSK and DQPSK modulation and the baudrate for each modulation format was chosen such that the resulting bitrate was 21.4 Gbit/s. The receiver parameters were not changed from the simulation described above. The predistortion transmitter was simulated using a fixed maximum number of taps $k' = 55$. The linear transmission was carried out over increasing distances up 1600 km of SSMF. For distances, in which 55 taps exceed the limit given by the aliasing criterion (14) the corresponding tap-weights were set to zero to fulfill the criterion.

The simulation results are shown in the upper part of Fig. 7. Again, the results for using a windowing function exhibit less fluctuation and will therefore be used for the comparison. An OSNR penalty of 1 dB for OOK and DPSK is observed at $\approx 350$ km, which approximately corresponds to the limited number of taps $k' = 55$ (350 km of SSMF in (14) leads to 83 taps, which can be reduced by a factor of 0.7 as shown above to obtain 58 taps). The results for DQPSK indicate that 55 taps are sufficient for compensation of 1350 km while maintaining an OSNR penalty below 1 dB.

For comparison, the results for OOK and DPSK are given for 10.7 Gbit/s in the bottom part of Fig. 7. It can be seen that the 1 dB penalty occurs at 1350 km for DPSK as well. With OOK target format the 1 dB penalty point is shifted by about 100 km. Therefore the CD-tolerance is preserved for the same number of complex taps while the bitrate can be doubled by using DQPSK. Note, that using DQPSK also increases complexity because the target waveform contains an imaginary part. Since OOK and DPSK formats do not contain an imaginary part two of the real filters in Fig. 2 can be omitted. However, compared to a realization using a binary modulation format for a bitrate of 21.4 Gbit/s this still offers a reduction of the number of real taps by a factor of two and the processing speed is also reduced by a factor of two.

The previously discussed approach only holds for links, which do not exhibit significant nonlinear effects. In the presence of nonlinear effects nonlinear filtering has to be used. When a LUT is used for filtering then the number of entries is determined by the number of symbols that significantly interfere with each other as described in Section IV. The number of LUT-entries grows exponentially according to $Q^{K_{\text{int}}}$, where $Q$ is the number of bits per symbol. When dispersion is compensated optically and only intrachannel-nonlinearity is compensated by EPD than the number of LUT-entries remains relatively small. However, for the joint compensation of CD and intrachannel-nonlinearity it is necessary to use a more complex nonlinear filter as described above.

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nel nonlinearity the number of states can get impractical rapidly, especially at Baudrates beyond 10 GBaud. This does not only lead to increased storage requirements but also increases the effort that needs to be spent on computation of predistorted waveforms. As an estimate of the number of LUT-entries required for compensation of both CD and intrachannel nonlinearity we have used $K_{ISI}$ from CD only because CD contributes the dominant part to the ISI. The resulting number of LUT-entries for a binary target modulation format and propagation on SSMF is shown in Fig. 8 for 10 Gbit/s and 40 Gbit/s. While the compensation of a few hundred kilometers at 10 Gbit/s is still possible with a reasonable number of LUT-entries, the approach becomes impractical at 40 Gbit/s because the amount of CD that can be compensated with reasonable effort only corresponds to a fiber link much shorter than 100km.

A combination of LUT and linear filter was proposed to reduce the size of the LUT [3] for joint compensation of CD and fiber nonlinearities. In this system the samples read out from the LUT are filtered by a filter corresponding to the inverse fiber transfer function before they are converted into modulator driving voltages. We have carried out simulations for this type of predistortion system with NRZ OOK target format and a link consisting of 10 spans of SSMF, each of length 80km. The span-loss was compensated after each span and there was no inline dispersion compensation. The precompensation filter was assumed ideal and $M = 2$ samples per bit were used. The receiver parameters are the same as described above.

The simulation results are shown in Fig. 9. By using the combination of LUT and linear filter the tolerance towards nonlinearities could be increased for 10.7 Gbit/s. The results for a bitrate of 21.4 Gbit/s, however, do not show much improvement by applying the combination of LUT and linear filter. The pure electronic compensation of chromatic dispersion and fiber nonlinearity together still poses a significant challenge. One of the possible approaches to solve this problem is to implement a coarse-version of the split-step algorithm to compute an approximation to the inverted link response in real time. This approach is shown for example in [11] for application in coherent receivers.

Besides the signal processing another crucial component for EPD is the D/A converter. The rate of the D/A converter determines the oversampling factor $M$ and consequently the number of samples per target symbol. In this way, the bandwidth of the signal, which can be produced is restricted to $B < M/(2T_{symb})$. Several investigations [6] [12] have shown that $\geq 2$ samples per symbol interval should be used to avoid large penalties, in particular if the predistorted signal has a broadened spectrum compared to the target signal. This can be the case when a significant amount of fiber nonlinearity is compensated. However, there have also been studies demonstrating that symbol rate processing at the cost of moderate penalties is possible for compensation of CD when a narrowband target modulation format such as Duobinary [10] [13] is used. The symbolrate that can be supported by a predistortion system is therefore fundamentally limited by the D/A converter. The fastest currently available D/A converters offer a conversion rate of up to 22 Gsamp/s [14]. It was also shown [15] that the output ports of a high speed FPGA can be used to generate a D/A converter of up to 20 Gsamp/s. One way to increase the bitrate of a predistortion system without the requirement for a faster D/A converter is to use higher order modulation formats instead of binary modulation. In this way, the bitrate can be doubled while the conversion rate remains constant. Apart from the conversion rate the resolution of the D/A converter plays an important role.

![Figure 8](image1.png)

Figure 8. Estimated number of LUT-entries required to compensate for a certain amount of chromatic dispersion at 10 Gbit/s (top) and 40 Gbit/s (bottom).

![Figure 9](image2.png)

Figure 9. Simulation results for a link consisting of 10x80km spans of SSMF. Predistortion is applied based on a combination of LUT and linear filtering [3].
in EPD. In order to avoid significant penalties, \( m \geq 4 \) bit quantization should be used, where \( 2^m \) is the number of quantization levels.

VI. CONCLUSIONS

We have revised the concept of electronic predistortion for the compensation transmission impairments in fiber optic communication system. The compensation of purely linear impairments by using FIR-filters was explained for systems with negligible nonlinearity. An upper limit on the number of filter taps was given and the benefits from using a window function were highlighted. In addition to that, filtering based on look-up tables was described as a way to produce a nonlinear predistortion. In addition to that, the complexity of electronic predistortion was discussed. For predistortion using linear filtering the complexity scales linearly with the amount of accumulated dispersion to be compensated. It was shown that the complexity grows quadratically with an increasing baudrate. Moving to a higher order modulation format is an elegant way to increase bit-rate with a modest increase in complexity. Contrary to the filtering approach the complexity of predistortion based on look-up tables scales exponentially with increasing amount of ISI. Therefore this approach is limited to scenarios with short ISI. It remains an open topic to find an efficient method to jointly compensate for linear and nonlinear transmission impairments.

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Photonic Routing Systems Using All-optical, Hybrid Integrated Wavelength Converter Arrays

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Abstract—The integration of a new generation of all-optical wavelength converters within European project IST-MUFINS has enabled the development of compact and multi-functional photonic processing systems. Here we present the realization of demanding functionalities required in high-capacity photonic routers using these highly integrated components including: Clock recovery, data/label recovery, wavelength routing and contention resolution; all implemented with multi-signal processing using a single photonic chip – a quadruple array of SOA-MZI wavelength converters which occupies a chip area of only 15 x 58 mm². In addition, we present the capability of the technology to build WDM signal processing systems with the simultaneous operation of four quad devices in a four wavelength burst-mode regenerator. Finally, the potential of the technology to provide photonic systems-on-chip is demonstrated with the first hybrid integrated all-optical burst-mode receiver prototype.

Index Terms—photonic routers, optical packet switching, all-optical wavelength converters, photonic integration, silicon optical bench, silica-on-silicon, functional integration

I. INTRODUCTION

The potential of photonic routing as a solution for the capacity scaling of electronic routers is the driver for intensive research efforts and investment worldwide. The role of optics in high-capacity routing has been thoroughly examined and experts are claiming that optical technologies show an excellent potential for inclusion in high capacity routers [1]. Now the focus is on photonic integration and the development of components with high integration density, intelligence and capability to deliver the functionalities of a photonic router - the “machine” that will take over the switching and routing of data, promising low power consumption and footprint. The pieces of the puzzle are being gathered with the demonstration of the first photonic systems-on-chip: indium phosphide (InP) chips that perform packet forwarding [2], packet envelope detection (PED) [3] and wavelength switching [4], as well as integrated buffers [5] and integrated Arrayed Waveguide Grating (AWG) switching fabrics [6]. The target is to integrate more and more optical processing elements into the same chip in order not only to bring down cost (which is dominated by the use of discrete devices) but also to enable the implementation of key functionalities in the optical domain in a straightforward manner. In this context, a photonic integration platform, which will allow higher levels of flexible, functional and cost-effective integration, is expected to lead to the development of compact, high-speed “optical line cards” and thus open the possibility for scalable and fully integrated high capacity photonic routers.

The silica-on-silicon hybrid integration developed by the Center for Integrated Photonics (CIP UK) is a technique that enables flip-chip bonding of pre-fabricated InP components, including SOAs and modulators on silicon boards with low loss waveguides [7]. The approach has been successfully used for the development of single element, all-optical wavelength converters and due to the passive assembly process it shows great potential for the development of advanced photonic components. Seeking a scalable solution for wavelength conversion - required for the regeneration or wavelength routing process in a photonic router - the silica-on-silicon integration technology was expanded to develop quadruple arrays of 40 Gb/s wavelength converters. As a result, the cost per wavelength converter has been reduced due to the sharing of one common photonic package, while the processing capability of the photonic chip is enhanced. In what follows we present the application of the “MUFINS” all-optical wavelength converter array platform to demonstrate a number of key functionalities required in high-capacity photonic routers including: WDM wavelength conversion, PED-based contention resolution, label/payload separation, clock recovery and 3R burst-mode regeneration. The first step for the transition from single element to multi-element hybrid integrated photonic devices has been made.

II. HYBRID INTEGRATED ARRAYS OF SOA-MZI WAVELENGTH CONVERTERS

The hybrid integration technology of CIP relies on the unique combination of high-speed InP monolithic
elements, precision-machined silicon (Si) submounts and low-loss silica-on-silicon planar lightwave circuits (PLC) for achieving characteristics such as high yield, small footprint, device scaling and low power consumption. A planar silica-on-silicon waveguide platform acts as a motherboard to host both active and passive devices. Integration is achieved by using precision-machined silicon submounts or “daughterboards” carrying individual optical components into the motherboard. These components have precision cleaved features for accurate mechanical positioning on the daughterboard. For the quad array device, two monolithic arrays of precision-cleaved SOAs are mounted on two micromachined silicon submounts, by pushing the monolithic chip against end stops on the submount and reflowing solder. Apart from the passive waveguide network, the motherboard employs the mechanical end stops and landing site features to enable the passive, flip-chip assembly of the two daughterboards. Fig. 1 shows the basic fabrication process, whereas fig. 2 illustrates a packaged and pigtailed single and quadruple array of SOA-MZIs. Table 1 summarizes the specifications of both devices, showing in numbers the advancements achieved by the expansion of the hybrid photonic platform. The quad array with 160 Gb/s aggregate chip throughput (4 times higher than the single element) hosts 8 SOAs and 8 phase shifters in a slightly increased chip area. The reduction of packaging and pigtailing costs comes with a significant increase in chip processing capability while consuming the same board space.

III. PHOTONIC ROUTING FUNCTIONALITIES

Fig. 3 illustrates a simplified block diagram of a

![Figure 3](image-url)  
Figure 3. Draft sketch of photonic router indicating key functionalities

### Table 1.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Single SOA-MZI</th>
<th>Quad SOA-MZI array</th>
</tr>
</thead>
<tbody>
<tr>
<td>chip throughput</td>
<td>40 Gb/s</td>
<td>160 Gb/s</td>
</tr>
<tr>
<td>number of SOAs</td>
<td>2</td>
<td>8</td>
</tr>
<tr>
<td>number of phase shifters</td>
<td>2</td>
<td>8</td>
</tr>
<tr>
<td>chip size (W x L)</td>
<td>11 mm x 44 mm</td>
<td>15 mm x 58 mm</td>
</tr>
<tr>
<td>total size of device (L x W x D)</td>
<td>72 mm x 30 mm x 14 mm</td>
<td>90 mm x 32 mm x 12 mm</td>
</tr>
</tbody>
</table>
A photonic router. The router consists of an optical routing and an electronic control plane. The routing plane is typically constructed using optical technologies and components and includes:

- A WDM channel de-multiplexing stage at the input.
- The wavelength routing stage: this consists of arrays of wavelength converters and AWG-based wavelength-routed optical switch fabrics.
- The contention resolution stage: this typically includes optical wavelength converters and space switches for resolving contention in three dimensions: time (buffering), wavelength and space (deflection).
- The regeneration stage: 2R and/or 3R regeneration is employed in routers to compensate transmission and switching impairments.
- A WDM channel multiplexing stage at the output.

In this scenario the label processing and routing decisions for correct intra-node routing are undertaken by FPGA electronic controllers. The interface between data routing and control plane includes two functionalities that can be performed all-optically: 1) label separation/recovery and 2) packet envelope detection for synchronization of both planes. In what follows, we will show how the integrated wavelength converters have been used to implement the functionalities required for photonic routing, with a view of reducing the number of optical cards in a photonic routing platform. Specifically, we present the implementation of multi-channel wavelength conversion as part of the wavelength routing stage, optical contention resolution in the wavelength domain as part of the contention resolution stage, 3R regeneration and finally all-optical label extraction and packet envelope detection as part of the interface between photonic routing and electronic planes.

A. Multi-channel wavelength conversion

The first milestone that has been set to prove the reliability and multi-functionality of the integrated arrays of SOA-MZI switches is to prove that the technique can be employed to implement a scalable and compact wavelength routing stage of a photonic router. This means the demonstration of multi-channel wavelength conversion required for the routing of optical packets through the AWGR as well as for multi-casting purposes. Four CW lasers (wavelengths 1554, 1556, 1558 and 1559 nm) have been used to power each one of the four wavelength converters of the quad device and convert the wavelength of a 10Gb/s NRZ packet stream. Fig. 4a) shows the eye diagrams of the incoming packet-mode and wavelength converted signals. Fig. 4b) depicts bit-error-rate curves that were obtained for the input and output signals. The circuit performed error-free for all four outputs with a maximum power penalty of 1.68 dB.
In this section we present an all-optical sub-system performing on-the-fly contention resolution implemented with the hybrid integrated SOA-MZIs. The sub-system is capable of operating for both Non-Return and Return-to-Zero (NRZ and RZ) modulation formats by utilizing an all-optical packet envelope detection circuit and additional optical gates for performing deflection and wavelength conversion of contenting packets. Fig. 5 shows the schematic diagram of the system that is capable of resolving contention for two incoming packet streams that are on the same wavelength ($\lambda_1$). The circuit has two input ports and a single output port and employs packet detection, optical space switching and wavelength conversion. The PED circuit consists of a passive filter in combination with a SOA-MZI wavelength converter operated as a low-bandwidth 2R regenerator. The PED circuit generates a packet envelope, indicating the presence of a packet at the specific timeslot (P1 in Fig. 5). Contention resolution in space is achieved by triggering the 1x2 switch with the PED signals generated by stream S1 to control packet stream S2. As such, the packets of stream S2 are spatially separated (switched) at the 1x2 optical switch. The contenting part of this packet stream can either enter a recirculating buffer or be deflected to a different output port or be wavelength converted, depending on the contention resolution strategy and the architecture of the router. In this case, we demonstrate contention resolution in the wavelength domain, using the same PED signal to wavelength convert the deflected packets. Hence, the packets of stream S2 contenting with the packets of stream S1 are wavelength converted onto the packet envelope ($\lambda_2$) generated by stream S1. Fig. 6 shows the experimental setups and fig. 7 shows time-domain results for contention resolution of 10 Gb/s NRZ optical packets. Two cases of contention between streams A and B are indicated as well as the presence of an empty slot.
Figure 9. BER results for optical contention resolution of 10Gb/s NRZ and 40Gb/s RZ

Figure 10. Oscilloscope traces for multi-functional circuit and eye diagrams for a) incoming data, b) wavelength converted data, c) recovered clock, d) regenerated data, e) payload and f) recovered label slot in stream A. Stream A is launched in the PED circuit which extracts the envelopes of packets 1, 2 and 3. When Stream B reaches the 1x2 switch, it is triggered by the PED signal so that packets 4 and 6 appear at the switched port, whereas packet 5 is spatially separated. Packet 5 is directly forwarded to the output together with stream A and fills the empty slot after combination of the two signals in a passive coupler. As a result, a time multiplexed packet stream consisting of packets 1, 5, 2 and 3 appears at the output, whereas packets 4 and 6 are wavelength converted by the PEDs and forwarded through the third port of the coupler avoiding any collision. The Optical Spectrum Analyzer (OSA) trace indicates the presence of two wavelengths at the output of the system, namely 1558nm for packets 1, 5, 2 and 3 and 1555nm for packets 4 and 6. Fig. 8 depicts the results recorded for 40Gb/s RZ data packets. Fig. 9 shows the BER curves obtained for the deflected, time multiplexed and wavelength converted packets for both 10 Gb/s NRZ and 40 Gb/s RZ operation. The power penalties are less than 1 dB in the case of 10 Gb/s operation and less than 2 dB at 40 Gb/s.

C. High-speed, all-optical wavelength conversion, label recovery and data regeneration

Apart from a scalable solution for multi-channel wavelength conversion, the hybrid integrated wavelength converter arrays present a unique opportunity for multi-functionality, i.e. simultaneous performance of functionalities using multi-signal processing on a single chip. Here we demonstrate the processing power and multi-functionality of the quadruple array device by performing wavelength conversion, clock recovery, data regeneration and label/payload separation using a single photonic integrated device. Figure 10 shows the basic setup and the time-domain results. Each SOA-MZI of the array is assigned a different role operating as: wavelength converter (wavelength converter 1), amplitude equalizer (wavelength converter 2) in the clock recovery unit, regenerator (wavelength converter 3) and AND gate (wavelength converter 4) for label/payload separation. After the wavelength conversion process, the signal is fed into the clock recovery circuit which employs a low-Q FPF filter and SOA-MZI1 operating as an equalizer. The recovered clock has a double role: 1) it is used as the input signal to the 3R decision gate, where retiming and reshaping is achieved by triggering the incoming data with the retimed optical recovered clock pulses and 2) it...
Figure 12: Experimental set-up of the 4\(\lambda\) Burst Mode Regenerator circuit implemented with three quadruple HMZI switch arrays. Letters show connected points.

is delayed and used to perform an AND operation with the incoming data to perform label recovery. Figure 10 a) – f) shows the eye diagrams of the input, wavelength converted data, recovered clock, regenerated data, payload and recovered label, indicating a clear eye opening for all cases. The BER measurements of fig. 11 reveal a negative power penalty of up to -1.9 dB for the regenerated data. Negative power penalties of -0.7 and -0.2 dB were also measured for the extracted payload and label channels respectively.

D. Four-Wavelength Burst Mode Regeneration

The wavelength converter arrays present a solution for the implementation of WDM systems that require a high degree of parallel processing. In this section, we present the simultaneous operation of three, hybrid integrated quad SOA-MZI wavelength converters, comprising a total amount of twelve integrated photonic optical gates and demonstrate an all-optical, four-wavelength burst mode regenerator (4\(\lambda\)-BMR). The BMR system performs power equalization and 3R regeneration of variable length RZ data packets. Each one of the three SOA-MZI quad arrays performs a different functionality in the 4\(\lambda\)-BMR architecture: a) power equalization, b) clock recovery and c) regeneration of the incoming packets. The experimental set-up of the 4\(\lambda\)-BMR is shown in Figure 12. It consists of the 10 Gb/s optical packet generator, the power equalization unit, implemented with the first quad SOA-MZI array, the clock recovery (CR) circuit which employs a Fabry-Perot filter (FPF) in combination with the second quad SOA-MZI array and the data recovery circuit, implemented with the third quad SOA-MZI array. The packet generator consists of 4 DFB lasers, an Electro-absorption modulator (EAM) for pulse carving and two Ti: LiNbO\(_3\) electro-optic modulators for data and packet generation. The data exhibit 6 dB packet-per-packet power fluctuations. The 40 Gb/s (4 x 10 Gb/s) multi-wavelength packet stream is demultiplexed using an AWG and is injected into the power equalization unit of the 4\(\lambda\)-BMR circuit. The SOA-MZIs of the first quad had unequal splitting ratios (60/40) and were configured to self-switch the incoming packet traffic, by also using different current values for their two SOAs [8]. With this arrangement the saturation properties of the two SOAs resulted in high gain for the low power packets and low gain for the high power packets, delivering nearly power equalized packets at the outputs of the SOA-MZIs. The power equalized data packets at the output of the first array were multiplexed and split into two parts. One part was fed to the clock recovery circuit (CR), whereas the other part was fed as a control signal into the third quad SOA-MZI switch for the data reception. The CR employed a low-Q fiber FPF with free spectral range (FSR) equal to the line rate (10.0229 GHz) and a finesse of 47, followed by the SOA-MZIs of the second quad array [9]. Each SOA-MZI of the second quad array was powered by a CW signal. The four wavelengths at the output of the FPF were separated with the use of a 1x4 AWG and each one served as a control signal to one of the four SOA-MZIs of the second quad. In order to reduce the switching window of the SOA-MZIs, both control ports of each gate were used. By properly adjusting the relative time delay and powers of the two control signals (push-pull configuration), a reduction of the gates’ switching window was achieved, resulting in the extraction of 8 ps clock pulses at their outputs. Data reception was completed in the SOA-MZI switches of the third quad array, by using the equalized packets as switching signals and their corresponding recovered clock packets as inputs, reducing in this way the timing jitter, and the amplitude modulation of the input signal [9]. The timing jitter of the input data has been mainly
due to the timing jitter of the signal generator used to drive the EAM. Push-pull control configuration was also adopted in the HMZIs of the third array. Finally BER measurements were performed for all four data channels.

Fig. 13 illustrates the evolution of the Burst Mode Regeneration process through temporal oscilloscope traces and eye diagrams obtained at each stage of the 4\(\lambda\)-BMR circuit. Fig. 13(a) shows two incoming data packets at 10 Gb/s having a length of 119 and 60 bits, respectively, and exhibiting power fluctuation of 6 dB. Fig. 13(b) shows the respective power equalized packet stream obtained at the outputs of the SOA-MZIs of the first quad array. The 6 dB power fluctuation between the incoming packets has been reduced into roughly 1 dB amplitude modulation between the pulses within the power equalized packets. Fig. 13(c) depicts the recovered clock packets obtained at the output of the CR stage. They persist for time duration that equals that of the corresponding input data packet, extended on its leading edge by a 4-bit rising time and on its trailing edge by a 16-bit decay time. The former value indicates the time required by the CR to lock to the line-rate of the incoming data packet, while the latter value determines the time required by the CR signal to decay to 1/e after each packet. The guardband bits depend on the finesse of the FPF, which is dictated by the order of the PRBS [8]. Fig. 13(d) illustrates the received equalized data packets at the output of the 4\(\lambda\)-BMR circuit, indicating timing jitter and amplitude modulation reduction, with respect to the corresponding signal at the output of the power equalization unit (Quad 1). Jitter reduction was confirmed after conducting measurements on the incoming data signal, the recovered packet clocks and the regenerated output. In particular, the root- mean-square (rms) values were 1 ps for the input, 600 fs for the packet clock and 730 fs for the regenerated signal. Fig. 14 shows the BER curves obtained for the input packets with 6 dB power fluctuation, the power equalized data packets and the regenerated data packets. Error-free operation was obtained for all wavelengths with similar results. The recovered data channels exhibit a negative power penalty of up to 2.5 dB and 1.3 dB with regard to the corresponding input data channels with 6 dB power fluctuation and the power equalizer output respectively. Error-rate measurements were also taken for the input packets with 0 dB power fluctuation that serves as the baseline for the complete BER measurement procedure. This BER curve lies between the BMR output and the power equalizer output curve and has a 1 dB positive power penalty offset with respect to the BMR output revealing the 3R capability of the 4\(\lambda\)-BMR circuit.

IV. DISCUSSION

The expansion of traffic growth in broadband core networks is driving the development and deployment of new transmission and switching equipment: the hardware for the new information age. The driver for implementing the photonic equipment for next generation broadband networks is the integration of photonic processing systems-on-chip with the development of multi-element, functional integration. In this context EU project IST-MUFINS has managed a major step towards the establishment of a photonic platform capable to generate such multi-element photonic chips with the demonstration of the first arrays of all-optical wavelength converters. Here we have presented how the enhanced processing power and capacity of these devices can be exploited to implement key functionalities required in next generation photonic routers. All-optical processing has been realized either by interconnecting externally the SOA-MZI wavelength converters of a single quad chip or by using multiple quads as arrays of photonic gates for...
parallel processing. It is clear that the next step towards photonic systems-on-chip involves the integration and on-chip interconnection of multiple SOA-MZIs. This development will enable the realization of compact and fully functional optical circuits that avoiding bulk fiber interconnections and implementing all the processing on the chip-level. This next step towards photonic systems-in-a-package has already been made: Fig. 15 depicts the first photonic ASIC - a hybrid integrated burst-mode receiver prototype. The chip employs three serially interconnected, hybrid integrated optical switches and a periodic optical filter and can realize single channel all-optical burst mode reception on-chip, i.e. packet equalization, clock recovery and data recovery using a single photonic module. This device is an ideal expression of the roadmap envisaged for the hybrid technology. The integration “toolkit” is now available for further exploitation and for developing customized, fully integrated photonic components, opening the possibility for compact and fully functional photonic routing platforms.

V. CONCLUSION

We have presented high-level system applications using the first quadruple arrays of hybrid integrated all-optical wavelength converters. The functionalities that have been demonstrated include: multi-channel wavelength conversion, packet envelope detection, data/label recovery, contention resolution and WDM burst-mode regeneration. This wide range of applications reveals the versatility and multi-functionality of the silica-on-silicon wavelength conversion platform. The roadmap now involves the integration of photonic ASICs: integrated circuits that can deliver networking functionalities with on-chip signal processing.

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Efficient Signal Processing in OFDM-based Indoor Optical Wireless Links

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Abstract—We propose a rate-adaptive optical wireless transmission system based on orthogonal frequency division multiplexing for indoor communications. The investigations rely on realistic parameters of the key system components and focus on throughput maximization. We will show that a dynamically adaptive system can greatly enhance performance when compared to static operation, and how a loading algorithm, which optimally performs in power-limited systems, needs to be adjusted if the specific terms of the optical wireless channel are to be rigorously obeyed. Our investigations include scenarios in which the non-negativity constraint on the optical source driving signal is strictly met and in which a certain amount of symmetric clipping is tolerated. In the latter case, the system can be regarded as power-limited and conventional loading algorithms are hence the most suitable. We will show that the transmission rate can be significantly improved even further by accepting a minor increase in the error rate as a result of controlled clipping, and we will compare our results with the upper system capacity limit.

Index Terms—optical wireless, OFDM, adaptive transmission, loading algorithm, clipping, channel capacity

I. INTRODUCTION

The transmission of information via infrared (IR) radiation is considered to be attractive for high-speed indoor communication, especially where electromagnetic interference with existing radio systems must be avoided or wireless transmission with enhanced security is desired (like in medical facilities, airplanes, military objects, banks, etc.).

According to many authors who provide an overview of indoor IR communication (e.g., [1]), the directed line-of-sight (LOS) and the non-directed link are the two basic link designs. Directed links have been shown to achieve speeds of more than 100 Mbit/s. However, user mobility can only be supported with complex and costly pointing and tracking mechanisms (mechanical [2] and electrical [2]–[5]). Even then, blocking of the LOS results in link failures. In a non-directed link configuration, communication relies on numerous signal reflections off the surfaces in the room, instead of, or in addition to the LOS. The so-called diffuse link is made to operate completely without LOS. Even though it inherently provides user mobility (no transceiver alignment necessary), a diffuse link requires relatively large transmission power and is severely speed-limited due to multipath propagation effects. Bit rates ranging up to 50 Mbit/s have been achieved in experiments using equalization [6].

Practical applications often call for a combination of the advantageous features of the two generic designs (user mobility of the diffuse and high-speed capability of the LOS link). The multi-spot diffusing approach was considered in [7]–[9] to improve the power efficiency of diffuse links by creating multiple LOSs between diffusing spots on the ceiling and the angle-diversity receiver. In [10], an additional diffuse transceiver mode was proposed to secure the reliability of an electronically tracked LOS link. However, this requires complex, multiple-element transmitters and receivers.

To benefit from both basic link types, we consider a non-directed LOS link that encompasses both LOS and diffuse signal components at the receiver. Both measurements [11], and analytical models (e.g., [12]) indicate high dynamics in bandwidth and gain of the resulting compound channel, depending primarily on LOS prominence. In order to exploit the varying channel capacity with reasonable transmitter power levels (limited by safety regulations), yet to provide reliable operation and full coverage, we consider a rate-adaptive system concept based on multiple-subcarrier (MS) transmission. In such a system, the adaptive behavior is implemented by the processing of the electrical base-band signal. This means that it is possible to use simple and low-cost optical components. The frequency-domain approach is implemented with multicarrier modulation techniques (Discrete Multi-Tone, DMT or Orthogonal Frequency Division Multiplex, OFDM) which inherently deal with multipath distortion and allow for frequency-domain equalization with reduced complexity at the receiver. Rate-adaptive transmission, as proposed in [13]–[19], considers adaptive coding and modulation for serial transmission and, thus, requires complex time-domain equalization.

The need for relatively high DC levels, i.e., a poor power efficiency, is a major disadvantage of optical wireless (OW) multiple subcarrier systems [20]–[22]. Approaches towards improving this situation involved block-coding [23]–[25], variation of the DC component on symbol interval basis, or peak-to-average power reduction techniques used in OFDM [26], [27] which resulted in a lower transmission rate or/and in high implementation complexity, thus limiting the number of subcarriers to just
In this paper, we will focus on throughput maximization in an OW OFDM-based system in a typical WLAN application. As often implemented in practical systems to improve power efficiency, we will adopt the simplest approach and allow a certain controllable amount of clipping at the transmitter, assuming that the additional errors can be tolerated. A special OFDM scheme, which foresees clipping, was suggested, e.g. in [28], for a narrowband OW channel. This scheme, however, exploits only half of the channel bandwidth since it assumes modulation of odd subcarriers only. We will show how a loading algorithm, which performs best in power-limited systems, needs to be adjusted under the specific constraints that the OW channel poses on the transmit signal waveform. We will also show that a dynamically adaptive system can hugely enhance transmission rates compared to a statically designed one, even if we put a conservative constraint on the electrical signal waveform by avoiding any clipping. In addition, we will investigate the influence of controlled symmetric clipping on system performance in terms of both error and transmission rate. The results show that the system capacity can be exploited even more by tolerating a minor increase in error rate as a result of clipping.

The paper is organized as follows. A communication scenario and channel model are introduced in Chapter 2. Realistic system parameters, which are necessary for our investigations, are derived in Chapter 3, while the properties of the proposed rate-adaptive system are discussed in the subsequent section. Chapters 5 and 6 provide the performance analysis in the absence and presence of both error and transmission rate. The results show that the system capacity can be exploited even more by tolerating a minor increase in error rate as a result of clipping, respectively, while conclusions are drawn in the final chapter.

II. COMMUNICATION SCENARIO AND CHANNEL MODEL

A. Indoor Communication Scenario

Figure 1 shows a typical example of a scenario where a non-directed LOS link could be of use. It represents communication between a fixed access point (Tx) and a mobile terminal unit, e.g. a laptop (Rx) in a moderate-size room. Numerical simulations performed throughout this paper assume the model room depicted in Fig. 1 with an Rx which may be anywhere at the desk-top surface (distance, \( r \) to the Tx) and which has a free orientation (parameter \( \theta \)) but the work is applicable to WLAN scenarios in general. Tx is assumed to be positioned at the center of the ceiling and pointing downwards. Tx and Rx are assumed to have wide-beam radiation characteristics and a wide Field-Of-View (FOV), respectively, so that both LOS and diffuse signals are simultaneously present at the receiver. It should also be noted that even though the access point at the ceiling is regarded in this paper as Tx and the mobile end-terminal as Rx, the communication scenario foreseen is bidirectional and the conclusions are valid for both directions.

B. Noiseless Channel Model

An analytical model for the OW channel of a non-directed LOS link was developed by Pohl et al. [29]. This model provides a good rule-of-thumb prediction of essential channel properties, such as path loss and bandwidth. It is simple and yet takes into account an infinite number of reflections unlike some the other models proposed.

Unless the LOS is blocked, the impulse response of the non-directed LOS channel consists of two distinctive components - a discrete Dirac-like pulse (LOS contribution) and a continuous signal, arriving some time later (contribution of diffuse reflections) at the receiver.

It was, in fact, recognized in [12] that the response of the diffuse channel is similar to that of an integrating sphere and that it can be well modeled with an exponentially decaying function. Accordingly, the channel frequency response is modeled as a superposition of a flat and a first-order low-pass transfer function

\[
H(f) = \eta_{\text{LOS}} + H_{\text{DIFF}}(f) e^{-j2\pi f \Delta T} = \eta_{\text{LOS}} + \eta_{\text{DIFF}} \frac{e^{-j2\pi f \Delta T}}{1 + j f f_0},
\]

where \( f_0 = (2\pi)^{-1} \) represents the 3 dB cut-off frequency of the diffuse channel. The delay of the diffuse component \( \Delta T \) causes a frequency dependent phase offset \(-2\pi f \Delta T\) in (1). Whereas the LOS-path amplitude gain \( \eta_{\text{LOS}} \) depends on the Tx-Rx geometry (radiation characteristic of Tx, FOV of Rx, orientation of transceivers with respect to the direct LOS direction), the two parameters describing the contribution of the reflections in the room, i.e. the diffuse-path amplitude gain \( \eta_{\text{DIFF}} \) and the cut-off frequency \( f_0 \) both depend on the room properties (room dimensions, reflection coefficients of the surfaces, etc.).

When the LOS component exists, large dynamics are present in the channel, depending on the ratio of the LOS and diffuse signal powers, i.e. on the so-called \( K \)-factor

\[
K[\text{dB}] = 20 \log_{10} \frac{\eta_{\text{LOS}}}{\eta_{\text{DIFF}}}. \tag{2}
\]

In cases where LOS is blocked, \( K[\text{dB}] = -\infty \). The total channel frequency response magnitude \(|H(f)|\) for different \( K \)-factors is shown in Fig. 2. The channel gain is calculated over a 300 MHz frequency range for some illustrative \( K \)-factor values between -40 and +20 dB. It is clear that the channel response depends heavily...
on LOS prominence (described by the $K$-factor). Where the LOS is blocked or very weak, the channel response is approximately low-pass and its bandwidth is quite small. The bandwidth corresponds in fact to the diffuse channel cut-off frequency, which is $f_0 \approx 9$ MHz for the room parameters considered. As the LOS becomes more pronounced, the channel response varies until it becomes almost flat for sufficiently large $K$-factors which means that bandwidths become, by an order of magnitude, greater than in the diffuse case. The notches in the characteristic appear at certain frequencies where both components destructively interfere.

Further on, the $K$-factor is used as a single parameter to describe the channel state. To obtain the span of $K$-factors of interest for the subsequent investigations, $K$-factor distribution is calculated across the Rx-surface (desk-top height) in the room from Fig. 1, and for the Tx with a Lambert radiation characteristic of index $m = 3$. The distribution is presented depending on the receiver horizontal distance from Tx, $r / \sin \phi$, and the angle $\theta$ (angle between the LOS arrival direction and Rx orientation) when the LOS is free. In the event of a more directed Tx characteristic (larger $m$), higher $K$ values would be obtained for adequately oriented Rx and short Tx-Rx horizontal distances. However, with a growing distance or poor orientation, $K$-factors decrease more rapidly (there is a smaller "hot-spot" area). In the following sections, the channel will be roughly characterized using a $K$-factor span of $[-20, +25]$ dB.

III. PARAMETERS OF SIMPLE OPTO-ELECTRICAL FRONT-ENDS

Since the improvement in performance in the system considered is implemented with electronic signal processing, a simple optical system can be deployed (no angle diversity or pointing and tracking mechanisms). In such non-directed LOS optical links, a wide beam transmitter and a wide FOV receiver are used. As a prerequisite for investigation of the transmission method in question, this chapter will briefly discuss the opto-electrical front-ends of the system, whilst aiming to keep them simple as possible, and at the same time, however, to allow broadband transmission with a sufficient link budget.

IR wireless systems, where optical power propagates openly through the air, generally comply with safety regulations. These limit the average optical power which is allowed at the Tx, depending on the viewing time, beam shape, apparent size and wavelength of the optical source [10]. Because of their wide modulation bandwidth, laser diodes present the optical source of choice for high speed links. However, due to the narrow semi-angle of their emission beam, lasers are classified as point sources which need to be diffused (with an additional component) if they are to be considered eye-safe. A diffuser destroys the spatial coherence of laser output and spreads the beam over a sufficiently extended aperture and emission angle, allowing thus for much higher power levels. In [10], it was shown that a diffused-beam source of extended diameter, emitting at 900 nm, has a much more relaxed power limit (at least several hundred mW due to the skin-safety limit) than a source working under the same conditions in the 1550 nm window. Moreover, low-cost and mature components are available for the 900 nm window. Even though ambient light noise has less influence at longer wavelengths, we assume that this is diminished by filtering at the Rx. For these reasons, in our present investigations, we consider transmission at 900 nm and a mean optical power of 400 mW for the numerical examples.

At the transmission link receive-site, the IR receiver must be carefully designed. Typical receivers consist of an optical front-end (an optical filter to reject out-of-band ambient light plus an optical concentrator to collect and concentrate the incoming radiation), a photodiode and electrical front-end (preamplifier). Because of the unguided wireless propagation of optical signals, large area receivers are needed in order to receive sufficient power. Unfortunately, increasing the detector surface not only increases the cost, it also leads to an increase of inner capacitance and this has a negative impact on the receiver bandwidth. This is due to the fact that optical
detectors are usually designed for optical fiber systems where detector areas are small and the capacitance per unit area is less important. The following section outlines the arguments that lead to the choice of Rx parameters assumed in our further investigations and simulations.

An optical concentrator is used to enlarge the receiver collection area. The constant radiance theorem limits the collection area of a given detector for a certain FOV, imposing a trade-off between gain and acceptance angle. We assume a hemispherical, non-imaging concentrator with a full FOV (180°), achieving a gain of \( n^2 \), where \( n \) is the refractive index of the concentrator [20]. A hemispherical band-pass filter is assumed to reject out-of-band ambient illumination because wide FOV and narrow bandwidth (tens of nm) can be obtained at the same time [20]. It is well-known that APDs have a better sensitivity than PIN photodiodes. However, using APDs increases the shot noise component at the receiver. PINs are cheaper and do not enhance shot noise, however, their poor sensitivity can limit the range for a given Tx power to an insufficient value in a desired system scenario.

The relevant parameters of simple and available receiver components assumed for the numerical simulations in the next chapters are presented in Tab. 1. It is assumed that such an APD-based receiver is shot-noise limited and that the SNR is given by

\[
\text{SNR} = \frac{\eta^2 P_{\text{sig}}^2}{2qM^2I_{\text{BG}}B},
\]

where \( P_{\text{sig}} \) is the optical signal power, and \( I_{\text{BG}} = \eta p_{\text{BG}} \Delta \lambda (d^2/4) \pi n^2 T_0 \) is the photocurrent induced by the background light. In our numerical examples, a bright sky background light irradiance of \( p_{\text{BG}} = 5.8 \mu W/cm^2 \) is considered.

### IV. RATE-ADAPTIVE OW SYSTEM

#### A. OFDM-Based Rate-Adaptive OW System

As introduced in Chapter 2, depending on the Tx, Rx and room characteristics, the channel frequency response varies from low-pass (about 10 MHz bandwidth) to approximately flat over the frequency range of hundreds of MHz. It is, therefore, difficult to maintain the SNR sufficiently for high bit rates under all channel conditions using reasonable transmitter power levels. If, on the one hand, the system is designed to achieve sufficient SNR on all subcarriers according to the worst-case (purely diffuse) channel characteristic, the baseband bandwidth offered under favorable channel conditions will not be efficiently used, and the transmission rates achieved would then be too conservative. On the other hand, if the system is designed for transmission rates requiring bandwidths larger than those offered by the diffuse channel, error performance would be prohibitively poor and likely to lead to system outages.

In order to efficiently exploit the channel bandwidth (i.e. to maximize the transmission rate, while maintaining reliable operation and full coverage), it makes sense to consider dynamically adjusting the set of subcarriers used along with the modulation orders deployed on them to the current channel condition. In such a scenario, a good subcarrier (faced with a favorable channel gain and relatively small noise enhancement) would use a higher-order modulation scheme and carry more information, whereas a poorer one would carry less or even no information, but transmission with a required bit-error-rate (BER) could still be achieved despite the impairments in the channel. In this way, such a rate-adaptive system is able to transmit at high speeds under favorable channel conditions and to reduce throughput (until the desired BER is achieved) as the channel degrades. The following sections present the main features of a modulation-adaptive system which is able to dynamically adjust the transmission rate to the OW channel and to efficiently exploit system capacity.

Figure 4 illustrates the most important operational blocks and signal flows of this concept. Data, mapped to M-QAM symbols, is sent over multiple orthogonal subcarriers which together directly modulate the optical source. In an adaptive system, the order of the modulation and the symbol amplitude (i.e. power) are determined by a loading algorithm which is based on the Cannel State Information
(CSI) obtained from the receiver via a feed-back link. An IM/DD OW system requires a real-valued OFDM signal to be generated by digital signal processing as well as a DC component to help properly drive the source. With N independent subcarriers foreseen in the system (i.e. subcarriers carrying independent information), a 2N-IFFT block is required to generate a real-valued OFDM symbol because conjugate symmetry is required on the input vector at the IFFT input \( \mathbf{X} = [X_0 X_1 \ldots X_{N-1}] \) by setting
\[
X_{2N-n} = X(n)^*, \quad n = 1, 2, \ldots, N-1, \\
X_N = 0, X_0 \in \mathbb{R}.
\]  
(4)
The first input \( X_0 \), corresponding to zero frequency, must be real-valued and is generally left unmodulated. Obviously, the need for Hermitian symmetry at the input requires doubling the size of the IFFT block (which is not an implementation issue for a moderate number of subcarriers considered in OW), however, it does allow digital implementation of multicarrier modulation (without prohibitive analogue filter banks). Such an approach to real-valued IFFT output is used in DSL technology where it is referred to as Discrete Multitone (DMT) [30].

The element \( X_0 \) can be used to set the DC level of the output signal. Alternatively, it can be set to zero and DC can be added directly in front of the source. If the other \( N-1 \) elements of the input vector (corresponding to the independent subcarriers) are given by \( X_n = a_n + jb_n \), time samples at the output of \( 2N \)-IFFT (assuming (4)) are
\[
x(k) = \frac{1}{2N} \sum_{n=0}^{2N-1} X_ne^{j2\pi \frac{nk}{2N}} \\
= \frac{X_0}{2N} + \frac{1}{N} \sum_{n=1}^{N-1} \sqrt{a_n^2 + b_n^2} \cos \left( \frac{2\pi nk}{2N} + \arctan \frac{b_n}{a_n} \right),
\]  
(5)
for \( k = 0, 1, \ldots, 2N-1 \). As in conventional OFDM systems, a guard interval in the form of a cyclic prefix (CP) is inserted between consecutive OFDM symbols at the Tx in order to eliminate inter-symbol interference and preserve subcarrier orthogonality.

The signal reaches the receiver corrupted by the influence of the OW channel and ambient light (enhancing the shot-noise component at the Rx). A channel estimation is obtained using a training sequence, so that equalization and demapping of valid data can be performed and the information needed at Tx can also be obtained. For instance, in order to follow channel variations, adaptive transmission requires accurate channel estimates at the receiver and a reliable feed-back link between Rx and Tx to provide CSI (or more particular, the noise enhancement vector, NEV) to the transmitter. The feedback link required can be made reliable if it is set to occupy the lowest subcarrier(s) of the uplink which are always suitable for transmission (low frequencies in Fig. 2). In this way, the link operates even under worst-case conditions where LOS is shaded and only diffuse light is received. At the Tx, each subcarrier is consequently modulated with the most suitable modulation format. The data rate is therefore chosen carrier by carrier, and the system is able to adapt to a large span of channel conditions, thus efficiently exploiting its capacity while keeping the transmission reliable.

B. Constraints on the Transmit Signal Waveform

The digital-to-analog converter (DAC) in front of the optical source generates a modulating signal \( x(t) \) which presents the sum of cosine waves, with amplitudes \( A_n \) and initial phases \( \varphi_n \) determined by the symbols of input vector \( \mathbf{X} \)
\[
x(t) = \frac{X_0}{2N} + \frac{1}{N} \sum_{n=1}^{N-1} A_n \cos (2\pi f_n t + \varphi_n),
\]  
(6)
\[-T_{CP} \leq t < T_{FFT},
\]
where \( T_{CP} \) is the length of CP and \( T_{FFT} \) is the OFDM symbol interval without CP. Values \( A_n = \sqrt{a_n^2 + b_n^2} \) and \( \varphi_n = \arctan b_n/a_n \) are constant over one OFDM symbol period.

The OW channel places specific constraints on the waveform \( x(t) \). Whereas the mean electrical power of the modulating signal is limited in conventional RF or DSL channels, in OW, safety regulations limit the mean optical power at the transmitter. This means that the mean value of \( x(t) \) (i.e. the DC value) is constrained. Additionally, a non-negative electrical waveform signal is needed (since this signal is directly translated into optical power). The conditions can therefore be written as
\[
\left( x(t) = \frac{X_0}{2N} \leq P_O \right) \land (x(t) \geq 0).
\]  
(7)
We hereby assume that the data symbols appear with equal probability, so that the insertion of the CP does not influence the mean value of \( x(t) \) and the conversion factor of the optical source is set to 1 W/A without loss of generality.

Since \( x(t) \) is characterized with relatively high peaks, the DC figure needed is relatively high. Nevertheless, since these high peaks occur rarely, setting the system constraints as (7) results in a conservative limit for the transmission rate as will be shown in the following chapters.

V. THROUGHPUT MAXIMIZATION UNDER CONSERVATIVE CONSTRAINT

The throughput achievable in a given system depends on many parameters. In Chapter 3, a set of suitable parameters was chosen for a realistic system with simple optoelectronic front-ends. A baseband signal bandwidth of \( B = 100 \) MHz was assumed for broadband transmission. This leads to a sampling interval of \( T_{sam} = 5 \) ns, so that a typical indoor OW channel length can be handled with guard intervals \( L = 16 \) samples long (cyclic prefix).

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$N - 1 = 63$ independent subcarriers are considered over the bandwidth $B$. With such values for bandwidth and the FFT size, the potential for throughput maximization using efficient signal processing is investigated in this section.

One of the main design parameters limiting the bit-rate performance of a practical system is the maximal bit-error ratio (BER) allowed which can be interpreted as the guaranteed quality-of-service (QoS) on the physical layer.

With the CSI available at the transmitter, the transmission scheme can be adapted to the current channel state. In rate-adaptive systems, the amount of information transmitted can be maximized by means of loading algorithms (LA). Based on the CSI, these schemes vary the transmitted power level, constellation size and/or code rate or type of transmitted signals. This means that maximum throughput is achieved while keeping the system constraints satisfied regarding the BER and transmit signal. Thus, without increasing the error probability, these algorithms provide high average spectral efficiency by assigning a lot of information to be transmitted under favorable channel conditions and reducing the throughput as the channel degrades.

For further evaluations in this paper, we adopted the loading algorithm proposed by Krongold [31] for power-limited systems and considering QAM modulation.

### A. Review of the Krongold Algorithm

The Krongold algorithm uses the Lagrange multiplier method to maximize the discrete function of the total system throughput (i.e. spectral efficiency) under the total signal power constraint

$$\text{max}(R_{\text{TOT}} = \sum_{n=1}^{N} R_n) \sum_{n=1}^{N} P_n \leq P_{\text{TOT}},$$

where $N$ denotes the number of subcarriers in the system, $R_n$ represents the (integer) number of bits allocated to the $n^{th}$ subcarrier and $P_n$ represents the power which is needed for symbol complexity $R_n$ to be detected with a desired BER.

The fact is that, depending on the BER required, each MF requires a certain SNR, and depending on the noise in the related channel, this SNR requires a certain power of the transmitted signal, thus rendering a relation between the rate $R_n$ and the power $P_n$ needed for transmission. This means that each variation of bit allocation represents a point in a total-rate-vs-total-power ($R_{\text{TOT}}$-$P_{\text{TOT}}$) plane as shown in blue in Fig. 5a. This figure presents an illustrative example where allocation over 4 subchannels is assumed with possibilities for 0, 1, 2, 3 and 4 bits.

The Lagrange method with multiplier $\lambda \in \mathbb{R}$ finds the maximum of the function

$$J(\lambda) = \sum_{n=1}^{N} R_n - \lambda \sum_{n=1}^{N} P_n,$$

as

$$\frac{\partial R_n}{\partial P_n} = \lambda \ (\forall n = 1..N).$$

In other words, the rates and powers for each subchannel of the optimal solution correspond to the "optimal operating point" on the convex hull of the $R_{\text{TOT}}$-$P_{\text{TOT}}$ diagram. The optimal operating point is the first point on the convex hull which is met by a line of slope $\lambda$. The optimal solution is always on the convex hull because this curve connects solutions (bit combinations) that require minimal total power for each possible total rate (see Fig 5a). All other bit combinations are located on the right side of these best solutions, i.e. they require more power. Moreover, since only discrete rates are possible, each point on the convex hull has a continuous range of optimal $\lambda$.

Since the SNR on subcarrier $n$ is given as $\text{SNR}_n = P_n/N_n$, the optimal solution (10) can be written over the so-called operational characteristic (OPC) of each channel

$$\frac{\partial R_n}{\partial \text{SNR}_n} \frac{1}{N_n} = \lambda \rightarrow \frac{\partial R_n}{\partial \text{SNR}_n} = \beta_n,$$

where $\beta_n = N_n \lambda$ is the optimal slope on the OPC of subchannel $n = 1, \ldots, N$. The OPC is given as the rate-vs-power characteristic with unity equivalent noise (i.e. rate-vs-SNR). This characteristic (illustrated in Fig. 5b) is the same for all the subchannels and the $R_n$-$P_n$ characteristics all have the same shape; the only difference being the scaling of the OPCs $x$-axis with $N_n$.

Conveniently, since the OPCs are the same for all subcarriers, the ranges of optimal $\beta$ slopes for each operating point on the OPC are constant, and can be

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placed in a look-up-table (LUT). For a particular \( \lambda, \beta_n \) is calculated for each subchannel and due to its value, a corresponding point \( (R_n, SNR_n) \) is read from the LUT. It was proposed that the optimal \( \lambda \) value be found by way of iteration using a bisection method.

B. Conservative Loading in the OW System

Since the optical wireless channel is not power-limited in a conventional manner, no loading algorithm developed for power-limited systems can be applied quite in its original form. The reason for this is the fact that due to the constraints on the signal in front of the optical source (7), the electrical AC power of the signal cannot be uniquely determined.

The only way to ensure that the conditions concerning signal non-negativity and mean value (mean optical power) are fulfilled is to set the maximum amplitude of the signal below or equal to the DC value (which is best set to the optical power constraint). Without a loss of generality, scaling of the FFT blocks is omitted. In a system of \( N - 1 \) independent subcarriers, it can be written

\[
2 \sum_{n=1}^{N-1} A_{n,\text{max}} \leq P_O, \tag{12}
\]

where \( A_{n,\text{max}} \) presents the maximal amplitude of a signal constellation on the \( n \)th subcarrier. Originally, the OPC presents \( R_n \), \( P_n \) relationship with unity equivalent noise. In the case where the limit is set on the sum of maximum amplitudes of the constellations on different subcarriers, an \( OPC=\text{Convex hull} \) with unity \( N_n \) is needed.

The relationship between the SNR on a subcarrier modulated by \( MF_n \) and its maximum amplitude is given by

\[
SNR_n = f(MF_n)A_{n,\text{max}}^2 / N_n, \tag{13}
\]

where the factor reflecting the ratio between the mean constellation power and the square of its maximum amplitude is

\[
f(MF_n) = 1, \quad \text{BPSK}, \quad \frac{1}{3\sqrt{M+1}}, \quad \text{square-M-QAM}. \tag{14}
\]

In the case of M-QAM constellations where \( M \) is an odd integer, there is no common relationship like (14). \( f(MF) \) depends on the constellation layout. For the purpose of OPC illustration, factors for 8-QAM, 32-QAM, and 128-QAM have been found to be 0.599, 0.588 and 0.428, respectively. Because of the quadratic relationship of \( SNR_n \) and \( A_{n,\text{max}} \), it is convenient to define a normalized \( SNR^n \) for each \( MF_n \) as \( SNR^n = SNR/f(MF) \) and obtain the OPC using the relation

\[
\frac{\partial R_n}{\partial A_{n,\text{max}}} = \frac{\partial R_n}{\partial \sqrt{SNR^*_n}} \frac{1}{\sqrt{N_n}}. \tag{15}
\]

Therefore, the OPC \( R_n \), \( A_{n,\text{max}} \) with unity equivalent noise is given by \( R_n = \sqrt{SNR^*_n} \) relation. Such OPC is presented in Fig. 6a for \( R_n = 0, 1, \ldots, 8 \) and \( \text{BER}= 10^{-4} \) as an example. It is clear that the OPC no longer has a concave shape, i.e. that there are some points which lie below the convex hull (\( R_n = 1, 6 \) are obvious). A first consequence of such an adverse shape is that the algorithm cannot be reliably implemented. This means that it can happen that a slope \( \beta_n \) obtained at some point in the algorithm corresponds to more than one pair \( (R_n, A_{n,\text{max}}) \) because the ranges for \( \beta \) are not disjunct. As this is not foreseen by the implementation, the algorithm would not converge to a unique solution. A second consequence is that, even if the algorithm converges, a non-concave OPC would mean that the result may not be the optimal solution. In the case of an OPC, like in Fig. 6a, the total-rate-vs-sum-of-amplitudes diagram (analogous to the diagram from Fig. 5a) would also not have a concave shape, and some of the operating points (for some total rates) would be located below the convex hull of this diagram. Since the Lagrange method converges to a point on the convex hull, the LA will reliably converge to the optimal solution only if all the (best) operating points

\[\text{Figure 6. (a) The OPC, i.e. } R_n \cdot A_{n,\text{max}} | N_n = 1 \text{ plot on each individual subchannel. (b) The OPC, i.e. } R_n \cdot (A_{n}) | N_n = 1 \text{ plot on each individual subchannel. Changing the BER will not affect the shape of the characteristics.}\]
possible are located on the convex hull of the total \( R_{TOT} = \sum_n A_{n,\text{max}} \) diagram. Since this is not the case, it may happen that that the LA “oversees” the optimal solution and converges to the first solution below it on the convex hull, as illustrated in the inset in Fig. 6a. Figure 6b presents a similar plot, the difference here being that the SNR is not represented by the maximum constellation amplitude but by a mean amplitude. The OPC defined in this way will be of interest in the next chapter. It is rather important to note here that the same conclusion can be reached for such OPC with respect to the choice of MF set as for the OPC in Fig. 5a.

This condition of the concave total-rate-vs-sum-of-amplitudes diagram translates into two conditions for the subchannel OPC: it must be concave (i.e. it must be equivalent to its convex hull) and it must have uniform steps on the \( y \)-axis. The term “uniformity of the \( y \)-axis” means that the difference between two adjacent symbol complexities considered is constant. The necessity for this second condition can be easily shown by a counter-example (assuming \( R_n = 0, 2, 3, 4 \) also results in a “non-concave” total diagram).

Based on the previous discussion, only square QAM is taken in consideration \( (R_n = 0, 2, 4, 6, \ldots) \) for application of the Krongold algorithm to the OW channel with the limit on the sum of maximal constellation amplitudes. This ensures reliable convergence of the algorithm to the unique optimal solution under the given conditions.

Figure 7 presents the rates obtained by the LA over different channel realizations where the bits are distributed over \( N - 1 = 63 \) subcarriers and under the limit on the sum of maximal constellation amplitudes (12). The total bit rate \( R_{TOT} \) [bit/s] is obtained by

\[
R_{TOT} = \frac{B}{N} \sum_{n=1}^{N-1} R_n,
\]

which represents the gross rate (without the influence of CP or other overhead components). In the figure, three plots are shown, representing achievable bit rates in a dynamically adaptive system and reflecting different target BER values. Clearly, the rates achievable can be increased by allowing for a higher error tolerance. With deployment of sophisticated error correction techniques or automatic repeat request (ARQ) algorithms, for instance, the BER can be set as high as \( 10^{-3} \ldots 10^{-2} \), however, their complexity will influence the system overhead needed.

Curves for a statically designed multiple subcarrier system, which aim guarantee the corresponding transmission performances over the whole area of interest (in this paper, the whole desk-top plane in the model room of Fig. 1), are also shown as a reference. These plots are referred to in Fig. 7 as a non-adaptive solution and are obtained also by (16), in the case of a purely diffuse channel (worst case channel realization) where only a small portion of the bandwidth is available for transmission (i.e. the first few subcarriers).

Fig. 7 clearly shows that an adaptive system would be more than adequate for indoor IR applications. An adaptive system can recognize and exploit the channel knowledge efficiently and thus has the potential for much higher transmission rates than a non-adaptive system. The benefits (with respect to a non-adaptive system) increase with the \( K \)-factor and become significant earlier (lower \( K \)-factor) if higher BERs are permitted. The increase in achievable rates can be seen to be slow to start, however, this becomes much faster from the point where the link becomes transparent (i.e. all subcarriers are modulated).

VI. PERFORMANCE IN THE PRESENCE OF CLIPPING

The investigations in the previous chapter showed that the dynamically adaptive system design is very attractive for the communication scenario considered, and that it enhances the transmission rates achievable with respect to the non-adaptive system for many channel states of interest, even under the strict non-negativity condition. While this approach ensures the non-negativity of the signal transmitted, setting the sum of amplitudes equal to DC (thus, assuming all subcarriers in phase) at the same time leads to rather conservative bit rates. In order to increase the achievable transmission rates further, it is assumed in this chapter that symmetric clipping is permitted at the transmitter. It is assumed that the additional clipping errors could be post-compensated or even tolerated since very high peaks occur rather rarely in the transmission signal. Hence, in a system where DC is limited, if somewhat greater tolerance of error performance is accepted by allowing for clipping, significant improvements in transmission rate can be expected. Similarly, for the same targeted transmission rate, a significantly lower DC is needed. Such approach thus also enhances system power efficiency. This chapter investigates the influence of controlled clipping on the performance of the OFDM-based OW system in terms of both error and transmission rate.

As already mentioned, if clipping is permitted, the AC power of the transmission signal can be increased and this leads to higher transmission rates (assuming,
of course, that the error performance remains below the
tolerance boundary). There are two ways to approach this.
The total AC power can be either directly increased or
it can be indirectly increased by increasing the sum of
subcarrier amplitudes until error performance reaches
a certain threshold. Since the investigations in the previous
section regarding the limit on the sum of amplitudes and the
rate when no clipping is permitted can be used as a
reference, the same approach is adopted in order to qual-
itative evaluate and describe the effects of clipping on
the BER. Both approaches are considered and compared,
however, for bit rate maximization. In order to be able
to easily distinguish between the two approaches, we refer
to them as amplitude distributing LA (AD-LA) and power
distributing LA (PD-LA).

A. Simulations Scenario

The effect of clipping on system performance is in-
vestigated by means of Monte-Carlo (MC) simulations.
For a certain assumed channel state (described by the
K-factor within the range of $-10\ldots +25$ dB) and a
certain assumed DC limit (eye safety limit of 400 mW
optical power), the sum of amplitudes (the amplitude
budget, $A_{\text{bud}}$) is increased in steps over DC and the
resultant total BER is observed.\footnote{Because of simpler implementation, MC simulations were performed to render a symbol error ratio. Additional investigations were performed (not shown in the paper) which showed that for the targeted BERs up to an order of magnitude of $10^{-2}$, BER=SER/\log_2 M is a good approximation if Gray coding is assumed (i.e. most erroneous symbols due to clipping have only one falsely detected bit).} For each new $A_{\text{bud}}$, a new result of the AD-LA is obtained. It is assumed that each subcarrier can carry $R_n = \{0, 2, \ldots \}$ bits. The AD-LA itself distributes bits and amplitudes yielding to a chosen BER, assuming only the presence of noise in the channel (no clipping). This error rate will be later referred to as $\text{BER}_{\text{LA}}$, as opposed to the total error rate as $\text{BER}_{\text{TOT}}$, which includes errors due to clipping.

AD-LA is practically the same as in the previous chapter
with only a minor change. It was, namely, necessary
to define the amplitude budget in the previous section
as the sum of maximal amplitudes of the constellations
associated to the subcarriers in order to fully ensure no
clipping. Nevertheless, the amplitude budget is redefined in the following as the sum of mean amplitudes of the constellations associated to the subcarriers

$$2 \sum_{n=1}^{N-1}(A_n) \leq P_0,$$

and the corresponding OPC is given in Fig. 6b. The reason
for this is that, even though clipping can theoretically
occur with (17), its influence is statistically too low to
cause a significant error increase (as can be seen
in the simulation results), yet, the rates provided are less
conservative. This will be seen in the examples illustrated
later in this paper. The relationship hereby between the
SNR on a subcarrier modulated by $MF_n$ and its mean
amplitude is given by

$$\text{SNR}_n = f(MF_n)\langle A_n \rangle^2 / N_n,$$ (18)

where the factor reflecting the ratio between mean (square
M-QAM) constellation power and the square of its mean
amplitude is derived as

$$f(MF_n) = \frac{\frac{2}{3}(M-1)}{\left(\sum_{p=1}^{M-1} \sqrt{\sum_{q=1}^{M-1} \sqrt{p^2 + q^2}}\right)}.$$ (19)

B. Influence on Error Performance

Symmetric clipping is assumed to occur in the digital
domain (at the FFT output, in front of DAC), so that
for the assumed DC, given by $P_0$ and set by $X_0$, clipping
deep peaks is enforced on the digital signal $x(k)$ as

$$x_c(k) = \begin{cases}
0, & x_c(k) < 0, \\
2X_0, & x_c(k) > 2X_0, \\
x(k), & \text{otherwise}.
\end{cases} \quad (20)$$

Clipping introduces a random shift in each symbol sent,
depending on amplitudes and phases of the symbols sent
on all subcarriers. It therefore presents additional interfer-
ence which can significantly affect the decision process
at the Rx. Unless the variance of clipping distortion is
negligible with respect to background Gaussian noise
variance, BER$_{\text{TOT}}$ will rise due to clipping errors.

Figure 8 presents the influence of clipping errors on
BER performance for several illustrative channel states.
The bit-error rates shown in Fig. 8 are given as mean val-
ues (averaged over all active subcarriers). Since BER$_{\text{LA}}$
of $10^{-5}$ is assumed, in the case where clipping is not
implemented (imaginary scenario of a perfectly linear,
infinitely long laser characteristic), the result of the MC
simulations reflects the system performance only under
disturbance from background noise and corresponding to
the error rates predicted by the loading algorithm.

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Figure 8. Influence of clipping on the system error performance for different channel realizations. The plot with non-filled markers reflects the case where no clipping is enforced and corresponds to the error rates predicted by the loading algorithm.
(non-filled markers) therefore serves as a reference and also indicates that simulations are properly performed. From the plots reflecting clipping, it can be concluded that for a certain K-factor the amplitude budget can be increased up to a certain point without a noticeable influence on the total BER which remains in the range of \(10^{-5}\) as set by the loading algorithm. At the same time, with a larger amplitude budget, the amount of information transmitted grows. In this range, the effect of clipping is still negligible with respect to the background noise component. As the signal power is further increased, errors due to clipping unavoidably become more significant and occur more often, evoking a steep increase of \(\text{BER}_{\text{TOT}}\) which asymptotically approaches 0.5.

Clearly, the increase of the BER due to the increase of signal power depends on the channel state. Increasing the signal power generally results in two effects:

(a) two bits are allotted to the subcarriers so far unused (4-QAM, where all symbols have the same amplitude) which has a minor influence on BER or/and,

(b) amplitudes of already active subcarriers are increased by increasing the modulation order and this has a strong influence on the resultant BER growth.

When effect (a) is dominant over effect (b), the increase of the BER is slow and steady, whereas, in the opposite case, the BER increases rapidly. Figure 8 shows that in case of \(K = 0 \text{ dB}\), effect (b) is dominant from relatively low \(A_{\text{bud}}\). In the case of relatively flat channels, however, there is a significant range of \(A_{\text{bud}}\) for which the effect (a) is dominant before (b) takes over. It can be concluded that with the larger K-factor, as long as new bits are predominantly allocated, the increase of the BER is slower. At the same time, however, the effect (b) becomes dominant for lower \(A_{\text{bud}}\) (compare cases \(K = 15 \text{ dB}\) and \(K = 20 \text{ dB}\) in Fig. 8). This can be explained by the fact that the same increase of \(A_{\text{bud}}\) causes more subcarriers to be activated when \(K\) is higher. It is, therefore, important to realize that for a certain tolerated BER in the system (which allows for clipping), the maximum amplitude budget permitted (indirect power budget) differs for different channel realizations. Which BER can be accepted in a system depends finally on the error correction mechanism.

Further investigations show how allowing some clipping can improve system performance in terms of the rates achievable. In the simulations, a DC of 400 mW is assumed and the amplitude budget is increased in steps of 25 mW. This can lead to some inaccuracy in the resulting maximum budget, but has no significant implications for the rate achieved. In this case, instead of bit-error-rates, certain symbol-error-rates are targeted in simulations because of slightly better accuracy. Since the conclusions have a rather qualitative character, they are just as valid for practical systems where the BER is targeted. In order to have a sound basis for MC simulations, a maximal \(\text{SER}_{\text{TOT}} \leq 2 \cdot 10^{-3}\) is targeted. When clipping is permitted in the system, different ratios between noise and clipping error rates are investigated in order to determine the influence on the improvement of transmission rate. The results are presented in Fig. 9 regarding the following cases:

1) \(\text{SER}_{\text{LA}} = 1.9 \cdot 10^{-3}\), so that the permitted \(\text{SER}_{\text{CL}} \leq 10^{-4}\); the ratio between the errors due to noise and due to clipping is well in favor of noise (green).

2) \(\text{SER}_{\text{LA}} = 10^{-3}\), so that the permitted \(\text{SER}_{\text{CL}} \leq 10^{-3}\); noise and clipping contributions are approximately the same (red).

3) \(\text{SER}_{\text{LA}} = 10^{-4}\), so that \(\text{SER}_{\text{CL}} \leq 1.9 \cdot 10^{-3}\); the ratio between the errors due to noise and due to clipping is well in favor of clipping (blue).

---

2 In practical systems, part of the budget, left over after distribution, is often equally divided among the active subcarriers in order to enhance the SNR and minimize errors (below \(\text{BER}_{\text{LA}}\)). In order to be able to distinguish the clipping from noise effects more easily, such a step is not included in the simulation results presented in this work. Nevertheless, additional simulations with extra budget distribution were performed, supplying practically the same results. This led to the conclusion that the budget is usually well exhausted.

3 The maximum rates achievable can be slightly higher.
states. The black line represents the reference case when no clipping is permitted in the system and the loading algorithm itself aims at $\text{SER}_{\text{LA}} = 2 \cdot 10^{-3} = \text{SER}_{\text{TOT}}$. No clipping in the system is achieved by deploying the conservative AD-LA which distributes the maximal constellation amplitudes, with $A_{\text{bud}} = \sum A_{n,\text{max}} = X_0/2 = \text{DC}$ (as in the previous chapter). Three stair-like plots reflect the three cases when clipping is permitted. It can be seen that the maximal $A_{\text{bud}}$ first increases with channel quality, however, it decreases again for very large $K$-factors. The reason for this is that only a moderate or small budget increase is enough for such particularly good channels to drive the system into the state where all subcarriers are active. This means that any further increase of $A_{\text{bud}}$ results only in an increase of modulation orders (effect (b)), thus causing a rapid deterioration of the BER. For poorer channel realizations a certain increase in budget over DC is possible, however not a large one, since in such low-pass-like channels, the extra power conveys in increase of the amplitudes of the few subcarriers active. Nevertheless, in the case of most channel states, with the small decrease of performance (from $\text{SER}_{\text{LA}}$ to $\text{SER}_{\text{TOT}}$), the system can operate with much more modulating power and expect to provide significantly higher transmission rates. This can be seen in Fig. 9b, where the rates achievable are compared for the same cases considered.

Looking at the plots, it can be concluded that operation with clipping is very beneficial. The rates achieved in case $\text{SER}_{\text{LA}} = 2 \cdot 10^{-3}$ are clearly below the other cases where clipping is foreseen. Moreover, targeting a rather significant portion of clipping errors is recommendable. For instance, in the case $\text{SER}_{\text{LA}} = 1.9 \cdot 10^{-3}$, the rates achieved are not as high as in the other two cases with clipping since there is practically no buffer for clipping and the $A_{\text{bud}}$ values permitted are closer to the figure of $\frac{1}{2}$ DC (see Fig. 9a). Cases 2 and 3 show very similar performances in terms of achievable rates. Setting the $\text{SER}_{\text{LA}}$ target too conservatively (case 3), however, requires deployment of higher signal powers (see Fig. 10a). In addition to this, the transmission rate $R$ achieved for a given channel state increases when $A_{\text{bud}}$ increases and/or when $\text{SER}_{\text{LA}}$ increases as $R = f(A_{\text{bud}}, \text{SER}_{\text{LA}})$. This means that if the LA target is too conservative (case 3), it can occur that the increase in amplitude budget (with respect to DC), which is gained by higher clipping tolerance, is not sufficient to cope with the decrease of the noise contribution permitted. No significant improvement in rate can then be achieved, or a decrease in the maximum rate may even occur. It can hence be concluded that it makes most sense to target the total SER (or BER) with approximately equal contributions due to noise and clipping (case 2).

If the LA has a finite set of modulation formats, once all the channels are filled with a maximum number of bits, a further increase of the $K$-factor then leads to no increase in bit rate. However, since less power is needed for the transmission of these bits, error performance improves and approaches error performance without clipping. Since a maximum of 12 bits were assumed in the LA foreseen, $K$-factors by which this occurs are beyond the range shown in the figure.

C. Bit Rate Maximization

As stated at the beginning of this chapter, the effect of clipping on error and rate performance in the system considered has been investigated and qualitatively described up to now by varying the sum of subcarrier amplitudes since it is easier to compare the results to the “no clipping” case. Now, once clipping is permitted in the system and if the total BER limit is the only constraint, one can directly increase the electrical power budget in the similar manner and distribute the electrical power over the subcarriers using the conventional Krongold loading algorithm, denoted as PD-LA. Figure 10a presents the rates achievable when PD-LA is implemented (blue plots) and when AD-LA assumed up to now is considered (red plots). The results for $\text{BER}_{\text{TOT}} \leq 2 \cdot 10^{-5}$ and $\text{BER}_{\text{TOT}} \leq 2 \cdot 10^{-3}$ are shown with corresponding loading algorithms aiming at $\text{BER}_{\text{LA}} = 10^{-5}$ and $\text{BER}_{\text{LA}} = 10^{-3}$, respectively. The increase of power budget $P_{\text{bud}}$ is performed in steps of 0.625 mW. Figure 10a shows that for a significant range of channel realizations, the distribution of powers results in better rates. The most striking differences appear for the intermediate $K$ values (neither too low and nor too high). This is due to the fact that for a particular $K$ within this range, the permitted power budget $P_{\text{bud}}$, which still satisfies the limit on $\text{BER}_{\text{TOT}}$, is larger than the equivalent power $P_{\text{eq}}$ obtained from amplitude distribution (with permitted $A_{\text{bud}}$ for this $K$), as can be seen in Fig. 10b. More precisely, if AD-LA, distributing the mean constellation amplitudes $\langle A \rangle$ and modulation formats $M_n$ to each particular subcarrier $n$, the total equivalent signal power is

$$P_{\text{eq}} = \sum_{n=1}^{N-1} f(M_n)\langle A_n \rangle^2.$$  \hspace{1cm} (21)

In Fig. 10b, these equivalent total powers are depicted by dashed red plots, presenting the power of the signal when the LA with amplitude distribution is used and compared to the permitted $P_{\text{bud}}$ which results from the condition on $\text{BER}_{\text{TOT}}$ when the PD-LA is used (blue solid plots). It can be seen that for these channels $P_{\text{bud}} > P_{\text{eq}}$. This can be explained by the fact that in the system considered, once clipping is permitted, apart from maximal DC value, the limit on $\text{BER}_{\text{TOT}}$ is the only constraint. Since the limit on the BER indirectly poses the limit on the signal power (rather than amplitude), it can be expected that the PD-LA maximizes the rate. The AD-LA achieves lower rates because it complies with an additional constraint on the sum of amplitudes. Moreover, the equivalent sum of amplitudes obtained from the results of the PD-LA with
the fact that the PD-LA results in more active subcarriers
ence in rate for $K$ system.

amount of clipping is of interest; this can be particularly

 Nevertheless, having such an ad-

well exceeds the maximal $A_{bud}$ permitted as the sum of
amplitudes under the same BER condition. This is shown
in Fig. 10c which compares the maximum permitted
amplitude budgets (red solid plots) and the equivalent sum
of amplitudes resulting from the corresponding power
budgets from Fig. 10b (blue dashed plots). This means
that the advantage in rate is accomplished by disregarding
the sum of amplitudes. Nevertheless, having such an ad-
tional constraint may be necessary if the crest-factor or
amount of clipping is of interest; this can be particularly
important if asymmetric clipping is considered in the
system.

According to the result of the simulations, the differ-
ence in rate for $K$-factors up to about 15 dB is due to
the fact that the PD-LA results in more active subcarriers
(than the AD-LA) which can then have larger powers
(or amplitudes) for the same BER limit (by the sum of
more sinusoids, there is a smaller clipping probability).
For channels where both PD-LA and AD-LA activate
all subcarriers, their performance is quite similar, with
$A_{eq} \approx A_{bud}$ and $P_{eq} \approx P_{bud}$.

D. Capacity Estimate by Water-Filling

The previous investigations showed the rates achievable
when square-QAM modulations are used on condition that
a certain error rate is not exceeded. The overall best rate
performance was seen with the conventional LA when the
total signal (AC) power was chosen so that the number of
errors due to background noise and due to clipping is
of the same order of magnitude.

In conventional coherent systems, maximum rates ob-
tained by an LA are upper bounded by the system
capacity obtained by the water-filling (WF) solution. The
following figures present a similar comparison in the
optical wireless system considered. Figure 11a shows the
rates already introduced in Fig. 10a for $BER_{TOT} \leq
2 \cdot 10^{-5}$ in case of symmetric clipping (solid plots). The two dashed plots represent
the capacity estimate obtained by the classical water-
filling solution where the corresponding $P_{bud}$ as found in
Fig. 10b is taken for each channel realization. There are
two capacity plots because the power budgets permitted
for $BER_{TOT} \leq 2 \cdot 10^{-5}$ are larger than for $BER_{TOT} \leq
2 \cdot 10^{-5}$. Figure 11a shows the capacity estimates signif-
ically above the achieved transmission rate curves. This
is because the WF solution assumes a sum of Gaussian
distributions which superimpose in front of the optical
source and result in a Gaussian distribution $\mathcal{N}(DC, P_{bud})$.
The ratios between standard deviation and mean value of
this distribution $\sqrt{P_{bud}}/DC$ are presented in Fig. 11b (left
y-axis). As symmetric clipping is considered (at levels of
0 and 2DC), these ratios correspond to equivalent clipping
probabilities of several per mill to several percent (on
both sides together). Once all the subcarriers are active,
a further increase of the $K$-factor results in a smaller
AC power increase with respect to DC, leading to a
decrease of the permitted clipping probabilities for the
same final BER. The WF solution provides optimistic
capacity estimates (especially for lower $K$-factors) since
it assumes transmission through a channel without loss of
energy due to clipping. It can be said that WF results
in an upper bound on the achievable rates of a system
based on discrete square-QAM which is tighter as the
clipping probability decreases. From a practical point
of view, the WF estimate can be used for upper bounding
the capacity of OFDM-based systems when such relatively
low clipping probabilities are considered.

E. Implementation of Power Control

It is obvious that in order to maximize the transmission
rate, the system must recognize the channel state and
adaptively set the permitted $P_{bud}$. Some form of power
control algorithm is hence needed. One possibility would be to use an LUT which would consist of the permitted $P_{bud}$ values for all relevant channel realizations and BER$_{TOT}$ targets. Each time the LA is performed, it is then carried out in 2 steps:

1) Reference loading is performed where a fixed $P_{bud,ref}$, $P_{bud,ref}$ is chosen to render a different number of subcarriers used, or a different total number of bits, which they carry for each channel realization; this virtual loading is hence an indicator of the present channel state.

2) The actual $P_{bud}$ is extracted from the LUT (for the channel state found in step 1) and the LA is performed again to provide the parameters for transmission.

An example is shown in Fig. 12. Such implementation does not depend on the number of subcarriers, and only requires that the execution time of the loading algorithm be doubled (since it needs to be performed twice) which is not an issue in the channel considered.

**F. Summary of Main Results**

Finally, Fig. 13 presents the improvement of transmission rates which can be achieved by designing the system adaptively (i.e. implementing a loading algorithm) and by implementing a power control mechanism. Rate curves of the same color reflect the cases where BER$_{TOT} \leq 2 \times 10^{-3}$ (triangle marker) and BER$_{TOT} \leq 2 \times 10^{-5}$ (circle marker) are considered. It can be seen how beneficial it is to use the channel information and dynamically adapt the transmission rate with respect to a system without such a feature where the necessary link reliability (BER) limits the system throughput to the worst-case scenario (i.e. a purely diffuse channel). Even under most conservative clipping conditions (no clipping permitted), the rates achievable substantially increase for many channel realizations of interest. By adding an additional power control mechanism, controlled clipping can be permitted in the system. This is the simplest method for enhancing power efficiency and results in further significant transmission rate improvements as well. As a reference marker, the estimate for system capacity obtained by water-filling is included in the figure. All bit rates of practical systems are obtained by (16) which does not account for transmission overhead. The overhead includes the effects of guard interval (cyclic prefix) length, training sequences, error correction coding, etc. and decreases the net transmission rate, depending on the implementation design.

Note that by combining the information from Fig. 13 with the information regarding the distribution of $K$-factors in our model room from Fig. 3 (depending on transmitter radiation index, Tx-Rx distance and Rx orientation), the rates achievable throughout the room can be derived showing that with the adaptive modulation technique significant improvements in rate performance are possible under many channel conditions encountered in the communication scenario considered.

**VII. Conclusion**

In this paper, we have focused on throughput maximization in an optical wireless system for a typical indoor WLAN application by means of efficient signal processing. In order to exploit the dynamics present in the channel considered, we investigated the potential of a modulation adaptive OFDM technique. Since the main
complexity of such an approach lies in digital signal processing, simple key components of the optical link front-ends can be assumed.

We have shown that an optimal loading algorithm, proposed for power-limited systems, needs to be adapted to strictly fulfill the specific constraints of the OW channel on the transmit signal waveform. It was concluded that the dynamical adaptive system can provide considerable transmission rate improvements compared to a statically operating system even with a conservative constraint on the electrical signal waveform (i.e. no clipping).

In order to increase the transmission rates and enhance the system power efficiency further in the simplest manner possible, controlled symmetric clipping of the transmit signal was introduced. The effect of clipping was investigated in terms of both error and transmission rate performance. It was shown that once clipping is permitted (relaxed non-negativity constraint), the system can be regarded as power-limited with the conventional loading algorithms rendering the best performance. We have also shown that further significant rate improvements can be achieved by accepting a minor increase of the error rate as a result of clipping and that the rates achievable approach the upper bound of the system capacity much more.

The investigations reported in this paper considered performance of the adaptive modulation technique assuming the simplest possible optical front-ends in a single-input single-output (SISO) link. Such approach was taken in order to be able to recognize the contributions of the chosen transmission method. It can be imagined that further improvements in rate performance could be obtained if additionally optical front ends were optimized, either in a way to be able to lift the safety limit, or to optimize the photodetector further (to obtain larger photodetectors with enhanced bandwidth, or to achieve better receiver sensitivity). Moreover, angle or space diversity could be realized in the system as well as the true multiple-input multiple-output (MIMO) approach. Additional improvements in both power efficiency and transmission rate can be expected also if the errors introduced by clipping are corrected using suitable procedures at the cost of an increased overhead and complexity in signal processing. Related investigations are, however, beyond the scope of this paper.

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Characterisation of the PMD distribution along optical fibres and improvement of the backbone fibre infrastructure by a POTDR

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Abstract - Fibre links in optical networks generally comprise several relatively short fibre segments which have been spliced together in cables. These fibre segments or “sections” are assembled with optical connectors and have a typical length of some tens of km. The important characteristic parameters of the fibre sections are attenuation, chromatic dispersion (CD) and polarisation mode dispersion (PMD). However, the PMD of the optical fibres can hamper the upgrade of optical backbone networks towards higher data rates of 40 Gbit/s and beyond. The PMD distribution along a buried fibre link is not constant and can also significantly vary between the concatenated fibres of the same optical cable. In the absence of such spatial information, the whole cable with higher PMD-values may have to be replaced in order to transmit 40 Gbit/s transparently over long distances. But investigations have shown that frequently localized pieces within a section are the major contributors to the overall high PMD value of the whole fibre, rendering the link unsuitable for higher data rates. A new random-scrambling polarization optical time domain reflectometry (POTDR) measurement technique is used to investigate the spatial distribution of the cumulative PMD in deployed fibres. Results help to identify high-PMD fibre pieces or sections which need to be replaced to enable 40 Gbit/s transmission and beyond, rather than substitution of a whole fibre link. Techno-economical investigations show the high economic potential of this method leading to significant reduction of expenses for infrastructure improvements. These improvements will enable network operators to transmit high data rates without limitation given by PMD.

Index Terms - PMD, OTDR, 40 Gbit/s transmission, field trial, optical fibres, backbone network, network upgrade

I. INTRODUCTION

When the first single mode fibre links were installed the most important fibre parameter was the attenuation measured in dB/km. Data rates up to 2.5 Gbit/s had to be transmitted over distances of up to 80 km. Attenuation and chromatic dispersion were the limiting effects determining data rates and link lengths. However, with the advent of optical amplifiers and techniques for compensation of the chromatic dispersion optical network links could be lengthened significantly and/or the transmitted data rates markedly increased. For optical single channel transmission at 40 Gbit/s, and even higher bitrates, PMD is a huge problem for network carriers. In particular, legacy fibres, installed when PMD was not of concern for the prevalent bitrates, often exhibit very high PMD values that are not suitable for high-speed transmission in today’s networks. The occurrence of link PMD can be divided into two major categories: intrinsic and extrinsic reasons. The intrinsic reason mainly arises in the production process and is based on fibre birefringence due to non-ideal roundness of the fibre and other imperfections, causing such a fibre to possess inherently a high PMD value. The extrinsic reason is caused by asymmetric pressure on the fibres and cables during or after the cable burying process. Such external influences can increase the PMD value of the fibre links. During the last few years Deutsche Telekom has reviewed its network and characterised the PMD value of large number of fibres (Fig. 1). The fibre links where the PMD value is too high for 40 Gbit/s transmission were identified. One high-PMD section can be sufficient to render the entire 400 km link “bad”. In order to avoid limitations of data rate and transmission length, PMD-compensators [1] can be used for each channel, an approach that can become quite costly and unwieldy. Alternatively, or in addition, advanced modulation scheme (e.g., 40G RZ-DQPSK) [2] can be applied to partially mitigate the sensitivity to PMD [3]. The third possibility for reducing the link PMD is to identify and replace the “bad” fibre section [4].

In fact, the latter approach may be the most economically viable solution, because replacement of the fibres along an entire link is extremely costly and time consuming. Fortunately, it is generally the case that the PMD of a
fibre link is not uniformly distributed but tends to be localized in a few distinct pieces or sections accounting for the large majority of the total link PMD [5]. A possible distribution of the PMD value over a fibre link is shown in Fig. 2. There are two fibre pieces having high PMD value - one “bad” fibre section between L₁ and L₂ and another very short piece fibre located at L₃ that, for example, may be a short but high birefringence fibre which is the primary contributor to the link PMD. The other portions of the link exhibit acceptable PMD behaviour, suitable for high bitrate transmission. As a result, such replacement is much more cost effective and could be economically competitive to the deployment of PMD tolerant transmission systems.

In this paper, a prototype random-scrambling tunable POTDR [6] was used in a field trial to measure the cumulative PMD along buried fibres in the network of the fixed network division of Deutsche Telekom group (T-Home). Results show the distributed PMD of the fibres and enable those fibre sections which are the primary contributors to the link PMD to be identified.

In fact, for each centre wavelength \( \lambda \), and I/O-SOP (input/output-State of polarisation; i.e., both launched-SOP and detected polarisation component), and transmissions \( T_j(z_n) \) are computed from the traces. The \( J \) traces are acquired in pairs of traces observed at closely-spaced wavelengths with the same I/O-SOP, from which two transmissions are computed to yield one local difference \( \Delta T_k(z_n) \) at each point \( z_n \). K differences (\( K >10 \)) are obtained for a random set of K independent couples \( (\lambda, I/O-SOP) \), where \( \lambda_k \) is the centre wavelength defined as the mean of the \( k^{th} \) pair. "Closely-spaced" in practice means PMD \( \delta \nu < 0.1 \) to 0.15, where \( \delta \nu \) is the optical-frequency difference between the two wavelengths of a pair.

In fact, for each centre wavelength \( \lambda_k \), one can also concurrently perform measurements using other wavelength pairs centred about \( \lambda_k \), whose optical-frequency spacing is different. In this way, a good relative (%) accuracy is attained for all of the large range of cumulative PMD values over the fibre length, even with a single scan sequence.

II. POTDR MEASUREMENT PRINCIPLE

A classical OTDR measures the fibre attenuation along an optical fibre in order to identify the fibre, splice and connector quality in terms of light attenuation. This reflectometer is based on detecting the Rayleigh backscattering along the fibre length. A fibre link can be characterised and bad fibre sections in terms of high attenuation and connectors with high reflection or loss can be found and in the case of unsuitable values the defective elements can be replaced in order to guarantee a good quality of transmitted signals corresponding to the physical limits. However, as mentioned above, such measurements are not sufficient for links carrying high data rates over long-haul distances, since the signal quality is also very much limited by the link PMD. Overall, the PMD can be characterised for the whole link with a number of different established measurement techniques. By using randomly polarisation scrambling POTDR PMD measurement technique a cumulative PMD curve can be obtained and high PMD fibre pieces or sections then can be located [5]. A screenshot of a typical measured cumulative PMD curve for a test fibre is shown in Fig. 3. An OTDR trace corresponding cumulative PMD curve is combined. The “bad” fibre pieces with high PMD values can be identified, for example, see PMD distribution of the selected fibre sections shown in right of figs. 5 and 6.

The detailed optical setup of the RS-POTDR prototype used for this work is presented in Fig. 1 of [5]. A number \( J \) of OTDR traces, i.e. back reflected power \( P_j(z_n) \) as a function of \( z \), are obtained for different settings of the wavelength, \( \lambda \), and I/O-SOP (input/output-State of polarisation; i.e., both launched-SOP and detected polarisation component), and transmissions \( T_j(z_n) \) are computed from the traces. The \( J \) traces are acquired in pairs of traces observed at closely-spaced wavelengths with the same I/O-SOP, from which two transmissions are computed to yield one local difference \( \Delta T_k(z_n) \) at each point \( z_n \). K differences (\( K >10 \)) are obtained for a random set of K independent couples \( (\lambda, I/O-SOP) \), where \( \lambda_k \) is the centre wavelength defined as the mean of the \( k^{th} \) pair. "Closely-spaced" in practice means PMD \( \delta \nu < 0.1 \) to 0.15, where \( \delta \nu \) is the optical-frequency difference between the two wavelengths of a pair.

In fact, for each centre wavelength \( \lambda_k \), one can also concurrently perform measurements using other wavelength pairs centred about \( \lambda_k \), whose optical-frequency spacing is different. In this way, a good relative (%) accuracy is attained for all of the large range of cumulative PMD values over the fibre length, even with a single scan sequence.

The RT-PMD (roundtrip-PMD, i.e. RMS value of roundtrip-DGD) at \( z_n \) is deduced from the mean-square value of the K random differences divided by the relative variance of the traces computed at the same point over the K I/O-SOPs (this compensates for depolarization).
However, one generally wishes to know the "one-way" PMD (from 0 to z) rather than RT-PMD. This value, henceforth termed cumulative PMD, is obtained by first multiplying RT-PMD by the statistical, average roundtrip factor \( \alpha_{\text{RT}}^2 = 3/8 \) (using rms-DGD definition of PMD). However, over a finite wavelength range, the actual RT-PMD at a given time and distance inherently fluctuates as a function of z even if cumulative PMD is constant. Fortunately, this stochastic factor converges toward this average value if the RT-PMD is averaged over some length \( \Delta z \), thereby permitting determination of the cumulative PMD even from a single scan sequence.

Resolution: In contrast with more traditional POTDR techniques using a single wavelength and measuring local birefringence instead of PMD as such, the resolution is not determined by the pulse length. For instance, as shown in [5], the PMD of even a short 1-2 m PMD emulator (beat-length \( \sim 2-3 \) mm) can be detected and measured correctly. Under other circumstances, for instance where there are two, relatively long (e.g. \( \sim 2 \) km each), joined segments having similar PMD-coefficients, it may not be possible to accurately quantify separately the PMD contribution of each of the two from one scan at a given time without more measurements corresponding to different wavelength steps (\( N_i > 3 \)) being used in the default operating mode of this prototype. (However, the PMD contribution of the two together may be accurately determined, provided that other adjacent segments have significantly different PMD coefficients.) Hence, in the context of the POTDR, "resolution" and accuracy over different parts of the whole span is a complicated issue that will be discussed in more detail in [5].

III. MEASUREMENT PROCEDURE

The field trial was performed in southern Germany on two fibre links within the network of the fixed network division of Deutsche Telekom group (T-Home). The tested fibre cable segments have approximate lengths of 31 (Link 1) and 38 km (Link 2). Each cable contains 16 standard single mode fibres (ITU G.652) although not all of them are accessible due to commercial traffic load. A total of 15 fibres were investigated. We first measured the PMD of both links with an EXFO FTB5500 interferometric analyzer (post processed using GINTY equations), and then the OTDR traces with an EXFO FTB7400. PMD was measured to be between 3.8 and 16.6 ps for Link 1 and between 2.7 and 7.2 ps for Link 2, showing that both are not suitable for 40 Gbit/s transmission in general.

The distributed PMD of each fibre was then measured with the POTDR prototype. The pulse duration was set to 100 ns (i.e. a pulse length of 20 m), and the number of I/O-SOPs set to \( K = 100 \) and 200 for the individual scans of Links 1 and 2, respectively, with the corresponding K centre wavelengths uniformly distributed between \( 1530 \) to \( 1570 \) nm. Data was acquired from both the forward and backward directions at two well-spaced times (e.g. 1 day apart) and subsequently post processed and analyzed on a PC. For verification of the results the fibre section has to be measured from both ends as explained in the next chapter.

IV. MEASUREMENT RESULTS

To illustrate the measurement data analysis needed to obtain the PMD distribution of each fibre, we start from the measured RT-PMD and cumulative PMD curves obtained in the forward and backward directions for one fibre of Link 1, as shown in Fig. 4. The result of the two-sided measurement, obtained by double mirroring of the backward square-PMD curve and stitching to the forward square-PMD curve at \( z = 21 \) km, is shown in Fig. 5 (above). A two-sided measurement is notably suited for longer links because the uncertainty on cumulative PMD increases at the end of the fibre and, in particular, downstream from high PMD sections. Also shown in the graph are the end points of the fibre sections corresponding to the splice locations observed from the standard OTDR trace, which also match the actual splice map. It is generally the case that the end points of sections having different slopes are strongly correlated with the splice locations, although this need not always be true. Finally, the PMD of each individual section, calculated from the square-PMD curve, is shown in Fig. 5 (down) where the sections that contribute the most to the overall link PMD are clearly identified. Recall that according to the sum-of-squares law of PMD, the relative contributions are in fact proportional to the square-PMD. An additional result, for a fibre of Link 2 with the stitching point at \( z = 24 \) km is shown in Fig. 6.
This prototype instrument was used to acquire a limited number of (temporally well-spaced) scans taken at a limited number of optical-frequency spacings, resulting in an approximately constant slope (of cumulative PMD vs. distance). As mentioned earlier, the “resolution” over this extended region (e.g. a few km) may not be sufficient to provide reliable results for the relative PMD contributions of short sections within this region. Such is likely the case in Fig. 5, between $z = 9$ to $14$ km, leading to additional uncertainty on the three individual section PMD-values shown in the section PMD graph. However, the total link PMD contribution arising from the three adjacent sections is nevertheless very reliable.

Overall, the measurement results show that high PMD sections in a link can be localized and quantified. Since the shown sections are in general the shortest accessible units of a spliced fibre network, the spatial resolution of this prototype instrument using the POTDR method is well adapted to carriers’ needs. The limitation of the prototype to accurately measure short adjacent fibre sections of similar PMD in the basic operating mode of the technique does not prevent network operators from correctly identifying those fibre sections predominately responsible for the link PMD, in the context of a focused and limited, thus cost efficient, fibre replacement strategy. The replacement of only these high-PMD sections would likely enable network operators to roll-out complete 40 Gbit/s-able backbone networks on legacy fibre without the need for sophisticated and costly PMD compensation technology.

V. TECHNO-ECONOMICAL CONSIDERATIONS

Usually the fibre segments or “sections” are assembled with optical connectors and have lengths of some tens of km. The sections begin and end at patch panels with connector pigtailed which are located in cable housing stations of the network operator. For comparison of the costs for the improvement of the fibre infrastructure by mitigation of the influence of PMD we assumed a typical fibre section length of 30 km between the accessible stations and considered 2 possible scenarios for the fibre exchange. Such a fibre section is a concatenation of short cable pieces with a mean length of 2 km which are spliced together (Fig. 7). It is necessary to divide such a cable section into those short cable pieces due to technological reasons of the technology for cable burying which is based on placement of the cables in subterranean cable conduits. A typical optical cable between two fibre network access stations is shown in Fig. 7. The cable consists of cable pieces connected by splice boxes.
Cables and splice boxes are buried and usually only in the case of a fibre break the splice boxes have to be opened. Therefore it is necessary to have access to the boxes which can be associated with high costs in dependence on the local particularities. In the case of the replacement of a whole cable section, all splice boxes of the section must be opened. In Fig. 8 two possible scenarios for the improvement of the cable parameters in terms of PMD are shown. Scenario 1 of the PMD mitigation is the classical scenario based on the replacement of the whole cable section as can be seen in Fig. 9. In the case of a lower number of “bad” cable pieces the cost relation between scenario 1 and scenario 2 becomes significantly better in favour of scenario 2. Furthermore a transparent link between the optical terminals consists of many sections of 30 km cable length. If there are e.g. 2 or 3 sections which are not suitable for high bitrates then also the “bad” cable pieces of these sections can be identified and replaced afterwards instead of the replacement of the whole sections 2 and 3. In our techno-economical considerations we assumed the same number of cable pieces to be substituted in each cable section to simplify the illustration in Fig. 9. The accumulated costs show that the replacement of four cable pieces in all three sections requires lower costs than the replacement of one single cable section. In the case of three cable sections and five cable pieces per section in our model the expenses for a particular replacement of the cable pieces are identical to the costs for the replacement of only one whole cable section. The used model is a simplification of the real network in order to have an opportunity to compare the different scenarios. The link lengths in reality are different and depend on the local conditions but nevertheless the results of comparison show a good agreement to the reality in the field.

VI. CONCLUSION

In this field trial a new POTDR prototype which enables the single-sided measurement of the cumulative PMD of optical fibres was used to investigate the PMD distribution in deployed fibre sections. The presented results show the individual PMD of different fibre sections. Hence, sections with very high PMD can be identified with sufficient spatial resolution, and can be subsequently replaced. The measurement technique can not only be applied to old fibres in order to replace bad pieces within them, but can be also applied for newly installed cables directly after installation to check the quality and the adherence of the fibres to the physical limits. This information is important for network operators who want to improve their networks in order to install 40 Gbit/s systems and beyond. Two scenarios for fibre replacement are investigated. The corresponding techno-economical considerations show that the replacement of short pieces of cables with increased PMD-values instead of replacement of whole cable sections can significantly reduce the expenses for the improvement of the fibre infrastructure.
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REFERENCES


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System impact of cascaded all-optical wavelength conversion of D(Q)PSK signals in transparent optical networks

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Abstract—We will compare techniques for all-optical wavelength conversion of differentially phase-modulated signals using four-wave mixing and super-continuum generation. For the super-continuum generation, a relation between the conversion efficiency and the nonlinear phase distortion will be derived and it will be shown that this technique is not suitable for the conversion of phase-modulated signals. For the four-wave mixing, techniques for the improvement of the conversion efficiency will be studied. Mainly the suppression of Brillouin scattering and its impact on phase-distortions will be discussed. A detailed discussion of its cascadability in transparent optical networks will conclude the contribution. The introduction of a maximum outage probability can significantly relax the OSNR requirements.

Index Terms—Nonlinear optics, wavelength conversion, differential phase-shift keying, differential quadrature phase-shift keying, stimulated Brillouin scattering, super continuum generation.

I. INTRODUCTION

For transparent optical burst or packet-switching networks, wavelength converters are commonly accepted as key components [1]. They are envisioned to fulfill multiple tasks inside the switching nodes and supposed to be involved in routing, contention resolution and signal regeneration. Among others, tuneability, cascadibility and transparency to bitrate and modulation formats are important characteristics of all-optical wavelength conversion to compete with optoelectronic solutions. Fiber-based components seem very promising to match in particular the two latter requirements. Solutions based on Super-Continuum Generation (SCG) are an interesting approach due to their simple setups with few components [2]. Wavelength conversion based on Four-Wave Mixing (FWM) of 80 Gb/s DQPSK [3] and multi-wavelength conversion of up to 40 Gb/s per channel DPSK signals have been shown [4]. We recently presented all-optical wavelength conversion of 320 Gb/s differentially phase-modulated signals based on FWM in a highly nonlinear fibre [5],[6] where we used CW pumping to provide modulation format transparency.

In this paper, we will compare the two approaches of all-optical wavelength conversion based on SCG and FWM [7]. We investigate the limits of SCG concerning nonlinear phase noise. For the FWM-based approach, we theoretically describe the impact of the induced phase distortions on the performance of DPSK and DQPSK signals. Quantitative results for a single conversion are compared to our recent experiments and extended to multiple conversions to study the cascadibility. Our results give a lower bound of the achievable OSNR penalty and will identify the induced phase distortions as a major degradation source when dealing with differentially phase-modulated signals.

II. WAVELENGTH CONVERSION BY SUPER-CONTINUUM GENERATION

Super-continuum generation in HNLF has been used successfully for wavelength conversion and signal regeneration of amplitude-modulated signals. Thus, it is considered to be also a possible conversion scheme for phase-modulated signals mainly attractive due to its simplicity. However, since it relies on self-phase modulation, it inevitably introduces phase distortions which may degrade the signal quality. For a Gaussian pulse shape the broadening factor describing the ratio between the output and the input spectral width can be expressed as follows:

$$\frac{\Delta \omega_{\text{rms}}}{\Delta \omega_0} = \left(1 + \frac{4}{3\sqrt{3}}\phi_{NL,\text{max}}\right)^\frac{1}{2}.$$ \hspace{1cm} (1)

Fig. 1(a) shows the spectral broadening and the corresponding nonlinear phase shift for 40 Gb/s and 160 Gb/s signals with a pulse width TFWHM = 2 ps, corresponding to the experimental value. The values of the fibre parameters have been chosen to $\gamma = 10\text{ (W km)}^{-1}$ and $L=1.1\text{ km}$. The spectral broadening seen in Fig. 1(a) for the 40 Gb/s signal is higher since, for the same average power, the pulse peak power is higher. This can also be observed in Fig. 1(b) showing the non-linear phase shift. From Fig. 1, it can be concluded that a significant spectral broadening, and therefore a reasonable conversion bandwidth within the C-band, can be obtained for 160 Gb/s signals only with very high average powers. But still, for 40 Gb/s, the feasible conversion bandwidths seem to be acceptable. However, the nonlinear phase shift required for e.g. 16 nm output spectral width is already about 10 rad. For large nonlinear phase shifts, the noise generated by the optical amplifiers can result in considerable phase-noise, which results from the strong...
amplitude-phase conversion by the Gordon-Mollenauer effect [8]. We can estimate the variances of the nonlinear phase noise from the total nonlinear phase shift, while the linear noise described by the signal-to-noise ratio (SNR) is equally distributed in phase and amplitude noise. With these assumptions, considering matched filters, and using a Gaussian approximation, the bit-error rate (BER) can be evaluated:

$$\sigma_{NL}^2 \approx \frac{2\phi_{NL,\text{max}}^2}{\text{SNR}}, \quad \sigma_L^2 \approx \frac{1}{2 \cdot \text{SNR}}$$  \hspace{1cm} (2)

$$\Rightarrow \text{BER} = \frac{1}{2} \text{erfc} \left( \frac{\pi}{4\sqrt{\sigma_L^2 + \sigma_{NL}^2}} \right)$$ \hspace{1cm} (3)

The thus estimated OSNR penalties at $\text{BER} = 10^{-9}$ for 40 Gb/s and 160 Gb/s DPSK signals are shown in Fig. 1(c) as a function of the non-linear phase shift, which is directly connected to the conversion bandwidth. Already for comparatively small non-linear phase shifts, the OSNR penalty for both data rates grows very fast beyond any acceptable value. Thus we assume that wavelength conversion based on SCG will not be a preferable solution.

III. FOUR-WAVE MIXING AND PHASE DISTORTIONS BY SBS SUPPRESSION

One of the most important features of the all-optical wavelength converter (AOWC) is low loss, which increases the cascadability of the converter. This implies large input powers, which are limited by the stimulated Brillouin scattering (SBS). Several strategies to bypass the SBS limit can be found in the literature. One of the most effective and simple is to phase modulate the pump wave. The spectral distribution into sidebands decreases the amount of power within the Brillouin bandwidth $\Delta \nu_B$ and consequently enhances the Brillouin threshold. However, a phase modulation of the pump will be transferred to the converted signal. For differentially phase-modulated signals like D(Q)PSK, a time-dependent pump phase directly changes the phase difference between consecutive pulses and leads to a phase distortion in the idler wave given by [9]

$$\Delta \phi_{\text{mod}}(t) = 2 \left[ \phi_P(t + T_{\text{Bit}}) - \phi_P(t) \right],$$ \hspace{1cm} (4)

with $T_{\text{Bit}}$ the length of a bit slot. In the literature, the most commonly used modulation signals are, firstly, binary phase-shift keying signals (BPSK) with a pseudorandom bit sequence (e.g. [10]) and, secondly, a multi-frequency scheme using several sinusoidals [11]. With regard to Eq. 4, the first scheme has the problem to keep the BPSK phase transitions out of the input signal bit slot otherwise destroying the phase information. For these reasons, the BPSK scheme is not applicable for the conversion of differentially phase-modulated signals and will not be considered further. To account for the multi-frequency scheme, we assume at first a modulation with two sinusoidals (dual-tone scheme). Then, the complex amplitude of the pump signal at the fibre input is given by

$$A_P(t) = \sqrt{P_{\text{in}}} \exp \left[ im_1 \cos \left( 2\pi f_1 t \right) + im_2 \cos \left( 2\pi f_2 t \right) \right]$$ \hspace{1cm} (5)

and is characterized by its power $P_{\text{in}}$, the two modulation frequencies $f_1 > f_2$ and the two modulation indices $m_1$, $m_2$. For the case that the line rate $B = 1/T_{\text{Bit}}$ is much higher than the maximal modulation frequency, $B \gg f_1$, we can approximately calculate the maximal phase distortion introduced by the pump phase modulation,

$$\Delta \phi_{\text{mod}}^{\text{max}} \approx 4\pi \left( m_1 \frac{f_1}{B} + m_2 \frac{f_2}{B} \right).$$ \hspace{1cm} (6)

Having identified the phase distortions introduced by the pump-phase modulation, we now want to evaluate the corresponding OSNR penalty as the most important parameter for the system design. Following the line of thought above, we can interpret $\Delta \phi_{\text{mod}}(t)$ given in Eq. 6 as a time-dependent additional interferometer phase error. The average BER is then given as a summation over all possible BER within one pump phase modulation period $T_{\text{PM}}$. Now, we can use the analytical formulas proposed by [12] to obtain the OSNR penalty for DPSK and for DQPSK signals. The results for single-tone modulation with $m_1 = 2.5 \text{ rad}$ and dual-tone modulation with $m_1 = m_2 = 1.25 \text{ rad}$ is given in Fig. 2 (a). In Fig. 2 (b), the experimentally obtained power penalty for a DQPSK

Figure 1. (a) Output spectral width and (b) corresponding nonlinear phase shift as functions of the average input powers of 40 Gb/s and 160 Gb/s phase-modulated signals. (c) Estimated OSNR penalty at BER = $10^{-9}$ for 40 Gb/s and 160 Gb/s DPSK signals.
signal and a single-tone scheme with $P_{in}^{+} = 15$ dBm, $m_1 = 2.5$ rad and $f_{mod} = 100$ to 400 MHz and the calculated OSNR penalty for the same case are shown, for better comparison both penalties are given relative to their value at $f_{mod} = 100$ MHz. For modulation frequencies below 100 MHz (not shown here) the measured penalty is also growing due to SBS-induced distortions. This shows that the modulation frequency has to be chosen to about twice the Brillouin bandwidth for a reasonable trade-off between good SBS suppression and low induced OSNR penalty.

IV. CASCADABILITY OF THE FWM-BASED SETUP

In optical networks, signals pass several nodes. Therefore, several wavelength conversions should be possible without severely degrading the signal. This feature is addressed by the question for cascadability. In this section, we will investigate how the phase distortion induced by the pump modulation and the resulting OSNR penalty develops for several wavelength conversions. Let us first assume a signal passing two identical wavelength converters inducing the same phase distortion, e.g. within a single node. Its envelope is given by

$$A_2(t) = \sqrt{P_{in}(t)} \eta \exp[t(2\phi_p(t) - (2\phi_p(t) - \phi_s(t) - \frac{\pi}{2}) - \frac{\pi}{2})]$$

$$= \sqrt{P_{in}(t)} \eta \exp(i\phi_s(t)).$$

We see that, in principle, the pump phase contributions $\phi_p$ of the subsequent AOWCs can cancel out each other keeping an undistorted data signal phase. However, in this case the phase modulation signals must be synchronized by a phase-locked loop to keep up a constant phase relation between them. In a general case, with the AOWCs separated by hundreds of kilometers, we have to assume that the pump-phase contributions of different AOWCs are distributed randomly. Let us assume a signal that has passed N wavelength converters with identical conversion efficiency $\eta_c$ but different pump phase contributions $\phi_{p,n}$:

$$A_N(t) = \sqrt{P_{in}(t)} \eta_c \exp[i(-(1)^N(\phi_s(t) + \frac{\pi}{2}) + \sum_{n=1}^{N}((-1)^{n+1}2\phi_{p,n}(t)))] \Phi_{p,N}(t).$$

We see that the pump phase contributions simply add in the signal phase. For simplicity, we want to restrict the following discussion to the single-tone modulation scheme. Then, the explicit form of $\phi_{p,n}(t)$ is given by

$$\phi_{p,n}(t) = m \cos(2\pi f_{mod} t + \theta_n)$$

with the randomly distributed phases of the modulation functions $\theta_n$. The overall pump phase contribution can then be written as

$$\Phi_{p,N}(t) = 2m \sum_{n=1}^{N}((-1)^{n+1} \cos(2\pi f_{mod} t + \theta_n))$$

$$= 2m \Re \left\{ \sum_{n=1}^{N}((-1)^{n+1} \exp(2\pi i f_{mod} t + i\theta_n)) \right\}$$

$$= 2m \cos(2\pi f_{mod} t + \xi)$$

Figure 2. (a) Calculated OSNR penalty for 40 Gb/s DPSK and 80 Gb/s DQPSK. Single-tone scheme: $m_1 = 2.5$ rad, dual-tone scheme: $m_2 = 1.25$ rad, $f_2/f_1 = 2/3$. (b) Comparison of measured power penalty and calculated OSNR penalty for 80 Gb/s DQPSK (both relative to penalties at $f_{mod} = 100$ MHz).
The last factor acts as a multiplier for the modulation index \( m \) and will be called \( m' \), and \( \xi \) represents the sum’s complex phase. Since, according to Eq. 6, the phase distortion due to the pump-phase modulation is proportional to the modulation index, \( m' > 1 \) will lead to a further signal degradation. We assume that the \( \theta_n \) are uniformly distributed in the interval \([0, 2\pi]\). The best case is given by \( m'_{\text{BC}} = 0 \) for \( \theta_1 = \theta_2 = \ldots = \theta_n \) and the worst case is given by \( m'_{\text{WC}} = N \) for \( |\theta_1| = |\theta_2| = \ldots = |\theta_n| \) but alternating signs.

To investigate the general case, we use the Monte-Carlo Method. For evaluation, we defined the worst case multiplier \( m'_{\text{WC}} \) as above and a mean multiplier \( \bar{m}' \). Additionally, we can establish a multiplier \( m'_{0.999} \) that is implicitly defined as

\[
\int_0^{m'_{0.999}} p(m')dm' = 0.999
\]

with \( p(m') \) the probability density function for \( m' \). Its interpretation is the following: If we interpret \( P(m' > m'_{0.999}) = 1 - 0.999 = 10^{-3} \) as an maximum allowed outage probability, \( m'_{0.999} \) is the maximum multiplier that must be covered by the system tolerances. We will see that this can significantly relax the OSNR requirements.

The results for a calculation with 100000 trials are given in Fig. 3 (a). Note that, unlike in Fig. 2, the OSNR penalties in this section are calculated at a bit-error rate of \( 10^{-3} \) since this represents the error threshold for forward-error correction. Also, all OSNR penalties calculated in this section originate only from the pump-phase modulation, no other degrading effect is included. While the worst case multiplier grows linearly with the number of conversions, the mean multiplier is growing more slowly, leading to only \( m' = 3 \) compared to \( m'_{\text{WC}} = 10 \) for 10 conversions. The most interesting parameter for the system design is \( m'_{0.999} \) because it allows to lower the OSNR requirements by defining maximum outage probabilities. For a small number of conversions, \( m'_{0.999} \) is very close to the worst case. However, for conversion
numbers above 5, \( m_{0.999} \) also starts to saturate. The same tendencies can be seen in Fig. 3 (b), where the corresponding OSNR penalty for DPSK signals after [12] is shown. The difference in penalty between the worst case and the 99.9% case is even more significant: The OSNR corresponding to \( m_{WC} \) grows exponentially with the number of conversions while the one corresponding to \( m_{0.999} \) is growing linearly above 5 conversions. For 10 conversions, the difference is about 3dB. So, the definition of a maximum outage probability can relax the OSNR requirements significantly.

In Fig. 4, the OSNR penalties corresponding to the 99.9% case for different modulation indices \( m \) are shown. It is clear from Eq. 6 that the OSNR penalty will decrease with decreasing \( m \). For 10 conversions, the difference between the curves corresponding to \( m = 1 \text{rad} \) and \( m = 2.5\text{rad} \) is about 3dB. However, since the modulation index also determines the conversion efficiency, we have to pay for the penalty reduction with a decrease in FWM efficiency. Therefore, a careful compromise for the modulation index \( m \) has to be found depending on the position of the wavelength converter and the number of the conversions. Generally, the penalties obtained for DPSK signals are within a reasonable range for more than 10 conversions.

In Fig. 5, the OSNR penalties due to the pump-phase modulation for DQPSK are shown. The tendencies are the same as for DPSK, but the magnitude of the penalties is generally higher. This is due to the smaller Euclidian distance of the signal states which results in a higher sensitivity to phase distortions compared to DPSK. Note that we calculated the penalties for a smaller number of conversions in comparison to the DPSK case. As shown in Fig. 5 (a), for \( m = 2.5 \), the OSNR penalty grows too fast to be tolerable. Therefore, also the definition of a maximum outage probability does not yield relaxed OSNR requirements since the multiplier \( m_{0.999} \) shown in Fig. 3 a) does not yet saturate. However, for \( m = 1 \) the OSNR penalty grows much slower allowing more cascaded conversions and taking an advantage from the definition of a maximum outage probability. The corresponding OSNR penalty in Fig. 5 b) shows acceptable values for about 10 conversions.

V. CONCLUSION

We have investigated two different approaches for all-optical wavelength conversion based on SCG and FWM. In the first method, nonlinear phase noise limits the maximum allowable nonlinear phase shift and thus the conversion range, which makes this scheme unattractive. The latter method must be combined with a SBS-suppression method leading to phase distortions. Quantitative results for a single conversion of signals with a line rate of 40 Gbaud have shown negligible penalties for DPSK and more 1 dB for DQPSK. If cascaded, the phase distortions of different converters will statistically add up in the signal phase. Referring to the worst case, DPSK wavelength converters may be cascaded up to 10 times, while for DQPSK cascading is critical and restricted to only a few conversions. However, if we introduce a maximum outage probability, the OSNR requirements are significantly relaxed. Then, for DPSK, up to 20 conversions seem possible, and for DQPSK, we can reach up to 10 conversions if the conversion efficiency is reduced.
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On local CAC schemes for scalability of high-speed networks

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Abstract—Next generation networks are required to provide bandwidth on-demand for sessions with fine-time granularity. In this sense, centralized CAC (Connection Admission Control) approaches could suffer from scalability problems if the number of requests for connections is excessive. In this paper we investigate local CAC schemes where the admission decisions are performed at the network edges, based on pre-calculated admission quotas.

Index Terms—Connection Admission Control; Distributed control; High-speed networks.

I. INTRODUCTION

Call Admission Control (CAC) is an essential functionality for networks supporting Quality of Service (QoS). In this paper, we study distributed CAC schemes that serve to alleviate the scalability and responsiveness problems observed in their centralized counterpart. Actually, currently proposed Call Admission Control methods for most next generation networks are centralized, thus complicating matters for scalability and robustness. On the other hand, distributed CAC techniques show a better behaviour in terms of reliability. However, the CAC becomes more sophisticated and difficult to implement and manage, primarily because resource brokerage is now carried out on an individual basis on many different CAC units. Consequently, such CAC units either need to be notified about the availability of resources network-wide, so as to make a decision in a coordinated, yet decoupled, manner or must be allocated a certain share of the resources to be managed in a completely independent fashion. In this paper we follow the latter approach, initially proposed in [1] by the MUSE IST project1, and later followed in RUBENS 4, in which the CAC units are assigned a certain share of the resources beforehand, which are updated on-demand. In what follows, we shall refer to this technique as “local” CAC.

1Multiservice Access Everywhere Project: Website: http://www.ist-muse.org/

This study further provides a performance evaluation of such a local CAC strategy in terms of capacity planning for the network control plane and bandwidth efficiency. On the other hand, we also assess the responsiveness of such local CAC to the input traffic demands. It follows immediately that the more frequent the resource allocation updates occur, the more closely the CAC controller follows the input demand. However, this may lead to undesired oscillations of network resources, together with a significant signalling overhead. In order to tackle these issues we first evaluate the dynamic behaviour of a single CAC controller. Our aim is to find the rate of the resource allocation updates such that the input demand is satisfied with a certain probability.

A. Application scenarios

Local CAC is originally thought to be used at the border of multi-service networks that must deal with a huge number of service requests that require QoS guarantees. As a centralized approach for handling these requests, it may show important scalability problems.

The MUSE IST project (see [1]) proposes the usage of local CAC in the (aggregation network) access nodes, such as the Digital Subscriber Line Access Multiplexers (DSLAMs) or the Optical Line Terminals (OLTs) for a subset of services that have stringent QoS requirements such as IPTV traffic. Note that the implementation of a local CAC requires being locally able of handling service signalling in order to be able to make the user aware of the CAC decision result (acceptance or rejection). In this way, local CAC can easily handle multicast traffic (e.g. IP Television), as it is possible to be locally aware of Internet Group Management Protocol (IGMP) messages in most of current IP DSLAMs.

The proposed technique is also suitable to be employed in the delivery of VoD services. In this scenario, there are unicast connections established between the video server and each one of the end users. So, with the application of the local CAC in the video server, it is possible to make a local decision about the network resource availability for the desired route according the assigned quota for the corresponding destination, stored in the CAC module of the video server.
For other services such as VoIP, local CAC may be used provided that the access nodes were able to participate in the signalling process. In this way, the proposal of distributing Session Border Control (SBC) capabilities [2] into the access nodes would allow to locally take the CAC decisions.

RUBENS proposes to take into account several parameters to accept the request of a new application (it can be a game session, VoD service, etc). One of these inputs to make such decision is the availability of bandwidth for a session of such application, which may be performed in a scalable way by means of the local CAC.

A long term application scenario consists of local CAC at the edge of an Optical Burst Switching (OBS) Network. In this light, edge nodes would have pre-established quotas per destination and class of service, assuring that the load inside the network is controlled, avoiding harmful congestion.

II. ANALYSIS

We provide capacity planning rules for the local CAC controller, based on the input demand (traffic matrix). Second, we analyze the timescale for updates. This is the timescale at which the capacity allocation (quota) per CAC controller should be updated. Third, a reliability analysis is presented. Finally, we perform an experiment using real traffic traces from the Spanish National Research and Education Network, in order to show the suitability of the results in a real network scenario.

A. Capacity planning for the local CAC controller (packet case)

Let us first assume a packet network case in which traffic is expressed as a Gaussian fluid over the graph \((V, E)\) where \(V\) is the set of nodes and \(E\) the set of links. The input demand is expressed as a matrix \(T\), with \(V\) rows and \(V\) columns and the uplink and downlink traffic to a certain node \(i\) is expressed as the sum of the entries of row \(i\) and column \(i\) respectively. The goal is to derive the capacity planning rule (quota assignment) for local CAC i assuming that the average traffic intensity is \(I\) in bits per second and its standard deviation is given by \(\sigma\). The capacity planning problem is stated as follows, find \(C\) such that

\[
P(X > C) < \epsilon
\]

where \(X\) is the offered traffic, resulting in the rule \(C = I + \sigma\), where is the percentile of a standard Gaussian random variable. The following figure shows the value of \(C\), \(n\) times the average input rate, for a coefficient of variation \(\sigma/I\) equal to 1 (bursty traffic) and equal to 1/3 (highly multiplexed traffic).

It turns out that in a bursty case (exponential variability) the local CAC quota should be three times the input demand, for a desired quality of service level larger than 97% approximately. On the other hand, the above results

\[
P(X(t_0 + t_b) = N|X(t_0) = 0.9N) < 0.01
\]

show that the increase of the quota is linear with the demand increase, if the coefficient of variation remains constant. For example, if the ratio (standard deviation)/mean is constant and the traffic demand doubles, then the quota simply doubles. If, on the other hand, the standard deviation is left constant, then the quota decreases, as the traffic becomes less bursty. The following graph shows a case for constant standard deviation, and a traffic intensity that doubles.

B. Timescale for updates

As a first approximation, we assume that the input demands follow a Poisson arrival process with rate \(\lambda\) arrivals per unit time and consume a resource unit (for example, a lightpath). Such resource unit is held for an exponential time, with mean \(1/\mu\) in time units per resource unit. The local CAC controller is provided with \(N\) resource units. We further assume that there is no queueing of demands, i.e. if the \(N\) resource units are already occupied, the CAC controller does not take any subsequent demand arrival, which are consequently dropped. Note that the CAC controller can be modelled by a \(M/M/N/N\) birth-death process. Let \(X(t)\) denote the number of busy resource units. Provided that the CAC controller initial state is empty, namely \(X(0) = 0\), we wish to find the value of \(\epsilon\), such that

\[
P(X(t_0 + t_b) = N|X(t_0) = 0.9N) < 0.01
\]
The equation can be explained as follows. Let us assume that $t_b$ is the hitting time to state $0.9N$ of the $X(t)$ process. Namely, the CAC controller boots at $t = 0$ and it is empty. Then, at time $t = t_b$ the CAC reaches an occupancy of 90%. This is an indication that the resources should be updated at the CAC. The point now is How much can the CAC wait for an update?. Note that the time $t_b$ represents such maximum waiting time for an update, because at time $t_0 + t_b$ the CAC will be fully booked and new demand arrivals will thus be blocked.

In order to evaluate such a conditional probability we consider the infinitesimal generator of the birth-death process of eq. 3, where state $N + 1$ is absorbent and let $Q(t) = e^{-\lambda t}$. The parameter $I = \lambda/\mu$ is the traffic intensity in Erlang.

Then,

$$P(X(t_0 + t_b) = N | X(t_0) = 0.9N) = Q(t_0)[0.9N, N + 1]$$

where $[a, b]$ means selection of the row $a$ and column $b$ in the matrix. For simplicity, we assume that 0.9N is an integer. There is a direct eigenvalue decomposition for hitting times of this sort [3]. However, we provide a numerical solution to the matrix exponential. The computation is performed by first block-diagonalizing the matrix and then applying a Padé approximation on each block.

In what follows, $t_b$ is denoted by “Time to saturation”. The following figures compares the distribution of the time to saturation from state zero and state nine for a local CAC controller with ten resource units, and 60% utilization (traffic intensity of 6 Erlang offered to the 10 circuits). The x-axis represents “average session duration”. As shown, average session durations in the range of minutes demand more frequent resource updates than session durations in the range of the days.

We note that the time to saturation is smaller in the loaded CAC, which matches our intuition. Clearly, a loaded system is more likely to saturate than an empty system, in the same time interval.

On the other hand, it is worth pointing out that the timescale for saturation can be relatively small, especially if the session duration is short. The following figure shows the lower percentiles of the distribution of the time to saturation, for the loaded CAC controller.

It turns out that in order to achieve the targeted 1% probability of not reaching saturation the updates must be as frequent as 0.001 times the average session duration. Let us assume that a worst case of a resource unit per video clip, with an average duration of 2.7 minutes. Then, the CAC controller requires an update every 0.162 seconds. Assuming that such updates comprise 1000-bit management packets, then the control plane requires 1620 bps per CAC controller.

However, this is not comparable to the centralized CAC case in terms of signalling load in the control plane, which

\[\text{http://www.itworld.com/Tech/2987/070913onlinevid/}\]
\[ P = \begin{pmatrix} -I & I & 0 & 0 & 0 & \ldots & 0 & 0 & 0 & 0 \\ 1 & -(1 + I) & I & 0 & 0 & \ldots & 0 & 0 & 0 & 0 \\ 0 & 2 & -(2 + I) & I & 0 & \ldots & 0 & 0 & 0 & 0 \\ 0 & 0 & 3 & -(3 + I) & I & 0 & \ldots & 0 & 0 & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & 0 & 0 & 0 & \ldots & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & \ldots & 0 & 0 & 0 & 0 \end{pmatrix} \] (3)

is much higher. In that case, we have a resource allocation request per session which implies a much higher rate.

C. Reliability analysis

In what follows, let us assume that the failure probability of a CAC, either local or centralized is \( p \), which can be obtained as the ratio between the MTBF and MTTR.

Let us now compare the reliability ratios obtained by the local and centralized CAC. We consider the probability that resources are unavailable, i.e., no bandwidth allocation request can be satisfied in the network. This implies failure of all CACs in the local case and failure of the centralized CAC in the centralized case, which can be calculated using a Binomial distribution.

The following graph shows the ratio between such probabilities (resource unavailability) versus the number of local CAC units, for a very large failure probability of \( 10\% \).

The results show that the unavailability ratio is as high as 10000 times as much in the centralized case, for a small network of 3 CAC controllers and very high failure rate. For smaller CAC failure rates, the unavailability ratio is even higher.

III. A CASE STUDY WITH REAL DATA

In this section, we consider a case study with real data from the Spanish academic network. We collect the busy hour traffic in the access links of four universities, two of them being small and the other two large. The following figure shows the topology of the Spanish academic network, along with the measurement collection infrastructure.

We consider that each of the access links is controlled by a local CAC and proceed with the Gaussian capacity planning rule. The results are shown in the following table.

<table>
<thead>
<tr>
<th>University</th>
<th>Mean (Mbps)</th>
<th>Std. deviation (Mbps)</th>
<th>Quota</th>
</tr>
</thead>
<tbody>
<tr>
<td>U1</td>
<td>32</td>
<td>5.6</td>
<td>45.02</td>
</tr>
<tr>
<td>U2</td>
<td>8</td>
<td>2.6</td>
<td>14.04</td>
</tr>
<tr>
<td>U3</td>
<td>206</td>
<td>7.6</td>
<td>223.68</td>
</tr>
<tr>
<td>U4</td>
<td>206</td>
<td>35.6</td>
<td>288.81</td>
</tr>
</tbody>
</table>

TABLE I.
RESULTS IN A CASE STUDY WITH REAL DATA

It turns out that the variability of the busy hour is relatively small in a real case, thus making the local CAC scheme amenable for use, as the resulting quotas are not too large in comparison with the real demand.
IV. CONCLUSIONS

In this paper we have analyzed a case of local CAC. The analysis has been performed both from an analytical and from an experimental point of view, including real traffic traces from the Spanish NREN. Our results show the advantages of the local CAC mechanism, with resulting quotas that are close to the average traffic demand per CAC controller.

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Storage and Mirroring in Single and Dual Section Metro WDM Rings under Different Traffic Scenarios

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Abstract — This paper introduces a novel data mirroring technique for storage area networks (SANs) in a metropolitan wavelength division multiplexing (WDM) ring scenario. Sectioning links are introduced to the ring to help deal with the hot node (SAN node on ring) scenarios created by the SANs and their mirrors. Two network architectures are studied: The first architecture accommodates conventional access nodes and a single SAN and its mirror, the latter connected through a sectioning link. The other architecture accommodates conventional access nodes and two pairs of SANs and their mirrors with a sectioning link connecting each pair. Simulation is carried out to evaluate the performance of both architectures under the proposed mirroring technique for a 24 node architecture with 1 Gb/s access node rate and 5 Gb/s SAN node rate and under two different traffic models—Poisson and self-similar. In addition to the fixed-size (FS) slot scheme, performance is evaluated under two different slot schemes accommodating variable size packet traffic — variable-size (VS) and super-size (SS) slot schemes. Simulation Results of average node throughput and queuing delay are presented and analyzed.

I. INTRODUCTION

The days when storage systems were expected only to store and retrieve randomly accessible data are long gone. Today storage systems are expected to play an integral role in supporting high levels of flexibility, scalability and data availability. Storage area networks (SANs) [1], [2] are emerging as the storage management structure to meet these requirements. SANs were initially designed to work within distance limited environments such as a campus. As the effect of natural disasters such as earthquakes, fires and floods, power outage, and terrorist attacks can be severely destructive in a limited distance environment; the need for extending SANs over large distances has become essential to protect data against loss or damage and to share storage resources among a larger number of users over large geographic areas. Most of the existing literature covering SAN extension is mainly concerned with long-haul overlay. The proposed solutions include optical-based extension solutions and IP-based extension solutions. The optical-based extension solutions include extending SAN over the synchronous optical network (SONET) and over wavelength division multiplexing (WDM). IP based extension solutions encapsulate data units of SAN traffic into standard IP frames to be transported over core networks. In [3], models were developed to compare the reliability of SONET-based extensions with IP-based extensions. It was found that SONET-based solutions are more able to satisfy customers while IP-based solutions have service interruptions due to hardware/software failure. In this work, a WDM-based SAN extension is considered in a metropolitan sectioned ring scenario.

In earlier work [4-6], a WDM slotted ring architecture with a single SAN node was proposed and evaluated. This work is an extension of the work presented in [7] where a sectioning link is added to the ring to help deal with traffic asymmetry and hot node (SAN node on ring) scenario. Also a novel technique is introduced to mirror the SAN node to another node (having identical capacity) considered as a secondary SAN. The presence of two SANs and their associated mirrors is also examined in an architecture with an additional sectioning link.

An assumption was made in [8-10] that the packet size is fixed. However, in reality the packet size in data communication traffic is variable. In this work, in addition to the fixed-size (FS) slot scheme, two schemes accommodating variable size packets are evaluated — variable-size (VS) slot and super-size (SS) slot schemes [11].

The remaining of this paper is organized as follows: Section II compares different optical switching techniques. In Section III, SANs, their protocols and mirroring techniques are reviewed. In Section IV, the network architecture and MAC protocol under the different slot schemes are presented. Section V introduces the mirroring technique. In Section VI the simulation results are presented and analyzed. Finally, the paper is concluded in Section VII.

II. STORAGE AREA NETWORKS (SANS)

SANs are emerging as alternatives to traditional direct-attached storage. SANs take storage devices away from servers and connect them directly to the network, simplifying the management of large and complex storage systems.

The fiber channel protocol (FCP) [12] has been considered for years as the premier SAN protocol to transport the SCSI commands used to deliver block storage. FCP provides a reliable, fast, low latency, and high throughput transport mechanism. However, FC was designed to work within environments limited to a few hundred meters however natural disasters, power outage, and terrorist attacks can be severely destructive. In addition, installing FC networks requires separate physical infrastructure and new network management skills. To remedy the distance limitation two extensions of the FCP were developed — FC over TCP/IP (FCIP) and Internet
FCP (iFCP) [3]. FCIP is a tunneled solution that interconnects FC SAN islands by encapsulating FC block data and subsequently transporting it over a TCP socket. On the other hand, iFCP is a routed gateway-to-gateway protocol that enables the attachment of existing FC devices to an IP network by transporting FC frames over TCP/IP switching and routing elements. However, FC extensions are still associated with high cost as they assume that the user has already invested in FC components. The IETF developed a new SAN protocol, internet SCSI (iSCSI) [13], to overcome the drawbacks of FC protocols. iSCSI transports SCSI data using already existing networks by encapsulating it in TCP/IP packets which makes iSCSI versatile and affordable even for small businesses.

Most of the existing literature covering SAN extension is mainly concerned with long-haul overlay. The proposed solutions include optical-based extension solutions and IP-based extension solutions. The optical-based extension solutions include extending SANs over the synchronous optical networks (SONET) and over wavelength division multiplexing (WDM). IP-based extension solutions encapsulate data units of SAN traffic into standard IP frames to be transported over core networks. In [12], models were developed to compare the reliability of SONET-based extensions with IP-based extensions. It was found that SONET-based solutions are more able to satisfy customers while IP-based solutions have service interruptions due to hardware/software failure. In this work, a WDM-based SAN extension is considered.

Backup methods for SANs are usually based on data mirroring [14, 15], where exact replicas of the original data are created and sent to secondary storage systems in far locations. SANs extensions facilitate automatic performance of backups and disaster recovery functions across the MAN or WAN. Data mirroring is usually implemented by one of two strategies — synchronous and asynchronous. In synchronous mirroring data is transmitted from the transmitting node to the two SANs simultaneously. However, in addition to the high bandwidth requirements, synchronization can introduce significant delays for large distances. On the other hand, under asynchronous mirroring data is initially transmitted to the primary SAN, and then the primary SAN replicates it to the secondary SAN. Usually asynchronous mirroring is scheduled to run after peak hours to save peak hour bandwidth. Therefore it is efficient and cost effective. However it is unsuitable for critical applications as the state of the storage locations is not synchronized.

### III. PROPOSED NETWORK ARCHITECTURE

The two network architectures considered are illustrated in Fig.1. Both architectures are metropolitan WDM ring networks with a unidirectional (clock wise) multi-channel slotted fiber. The networks connect a number of access nodes within a circumference of 138 kilometers. There are two types of nodes: access and SAN nodes operating at 1 Gb/s and 5 Gb/s, respectively.

For the first architecture, shown in Fig.1-a, the ring is sectioned using a 44 km point-to-point link passing through the center of the ring. The link consists of a pair of fibers with opposite propagation directions and directly connects two nodes, a primary SAN and its secondary SAN. The sectioning link provides a shorter path for some source-destination pairs to communicate through instead of going through the entire ring and helps deal with traffic asymmetry and the hot-node scenario created by the SANs. The SANs are located at the sectioning points of the link to make use of the links and to ensure that they are separated by the maximum distance to survive disasters in a limited distance scenario.

For the second architecture, shown in Fig.1-b, the ring contains two sectioning links and two pairs of SANs and their mirrors. In [8-10], an assumption was made that the packet size is fixed and equal to the slot size (Ethernet maximum transfer unit (MTU), defined as 1500 bytes). However, in reality the packet size in data communication traffic is not fixed. According to measurements on the Sprint IP backbone [16], there are mainly five major sizes of packets in data traffic, i.e., 40, 211, 572, 820 and 1500 bytes. The 40 bytes size is for TCP ACKs. The 572 bytes and 1500 bytes are the most common default MTUs. The 211 bytes packets correspond to a content distribution network (CDN) proprietary user datagram protocol (UDP) application that uses an unregistered port and carries a single 211 bytes packet. The packets of around 820 bytes are generated by media streaming applications. Obviously, the original architecture is not suitable for this situation, in which a huge percentage of slot space will be wasted. In this paper, in addition to the FS slot scheme, the performance of the sectioned ring is evaluated under two different schemes that accommodate variable size packet traffic — variable-size (VS) slot and super-size (SS) slot schemes. For the VS slot scheme, slots of five different sizes circulate around the ring. The slots sizes correspond to the five different sizes of data traffic packets from the access links, which are 40, 211, 572, 820 and 1500 bytes, with probabilities of 0.1, 0.2, 0.1, 0.2 and 0.4, respectively [16]. The total number of slots on the ring is 240. The number of slots for each size (related to the probability distribution of packet size on the access links) is 24, 48, 24, 48 and 96, respectively. The average size for a slot is 870.44 bytes. In the simulation, the different size slots are generated with a fixed sequence. The 240 slots are divided into 24 groups, in each group there are 10 slots with one slot of 40 bytes, two slots of 211 bytes, one slot of 572 bytes, two slots of 820 bytes and four slots of 1500 bytes. In order to simplify the calculation and rotation, the length of the ring is changed to 133 kilometers. In the SS slot scheme, the ring is divided into several super-size slot (24 super slots), which are much larger than the packets in the traffic (9000 bytes). Obviously, this approach is more realistic and suitable for a general situation than the VS slot scheme as it is not based on the packet size distribution. In this scheme, the length of the ring does not change.

Fig.2 illustrates the logical topology of the two architectures. The two architectures can be considered as a number of logical rings: three for the single section ring and five for the two section ring. The logical ring for each source-destination pair over the time slotted ring is chosen according to a shortest path algorithm. To demonstrate the effects of network loading,
asymmetric traffic, hot-node scenario and the impact of sectioning, the number of wavelength is limited to 4 for the single section ring which is the optimum number needed given this architecture. The minimum number of wavelength can be decided by considering the three logical rings. Considering a 24 node network, there are 11 nodes common to rings A and B each operating at 1 Gb/s. If the wavelength rate is 2.5 Gb/s, then 4 wavelengths are needed, ignoring the statistical multiplexing gain achieved by the ring, i.e. 2 wavelengths per ring. Rings B and C can use the same two wavelengths as the sectioning link is made up of two counter propagating fibers. Therefore a total (minimum) of 4 wavelengths is needed in this architecture offering each of the three rings 2 wavelengths. A minimum number of 6 wavelengths can be calculated in the same way for the two section ring architecture.

Under the different slot schemes, packets arriving from the access networks or the SANs are sorted according to the logical ring they need to be transmitted through. In addition, under the VS slot scheme packets also to be sorted according to their size so the node can choose the packet with a size matching the current slot size.

The MAC protocol associated with the FS slot scheme is illustrated in Fig. 3. Each node monitors slots on the wavelength assigned to it on all the logical rings. If an empty slot is found, the node uses it to transmit packets queuing in the buffers associated with that logical ring.

Fig. 4 shows the MAC protocol algorithm under the VS slot scheme. Having variable size slots complicates the process of packet to slot allocation in two distinct ways. Firstly, only packets from the appropriate buffer (different buffers for different packet sizes) can be allocated to a given empty slot. Secondly, unlike the FS slot scheme, there is no common time unit (slot size unit) for ring rotation, rather the MAC allows rotation by the smallest time unit that enables one of the nodes round the ring to point to the start of a slot. In a practical

![Network architecture](image1)

**Fig. 1** Network architecture

![Logical topology of the architecture](image2)

**Fig. 2** Logical topology of the architecture
system this is feasible as the process of packet insertion can be
controlled through clock with bit period increments.
Note that the time increase in the FS slot scheme is equal to the
fixed slot time. Under the VS slot scheme, the variable
Min_Time is set to compare the time increases needed for each
node. After finding the node with the minimum time
(Current_Acting_Node), it is allowed to point to the next slot,
the current_acting_node then checks its current slot. If it is
empty, the packet waiting in the buffer associated with the slot
size is transmitted.

The SS slot scheme MAC protocol algorithm is shown in
Fig. 5. Under SS slot scheme, each node checks if the super
slot it is currently pointing to has enough space to
accommodate the packet waiting to be transmitted. Slots
circulation under SS slot scheme is simpler compared to the
VS slot scheme as the time increase needed to make each node
point to the next slot is fixed and equal to the super slot time.

Under the single section architecture, each access node is
equipped with two fixed transmitters to connect it to the two
logical rings available to it. SAN nodes placed at the
sectioning points require an additional transmitter as each has
access to the three logical rings. As the number of nodes is
greater than the number of wavelengths, each wavelength is
shared by a number of nodes for transmission. Four fixed
receivers are used to allow nodes to share wavelengths for
reception which results in higher scalability compared to the
single fixed receiver architecture where the number of
wavelengths is equal to the number of nodes. The use of a
tunable receiver is possible but results in receiver collisions
where multiple wavelengths carry packets in a given time slot
all destined to the same node. Receiver collision mitigation by
receiving one packet and allowing the others to circulate the
ring until received will be considered in future work. Tunable
transmitters can be also introduced, however tuning latencies
can degrade the performance and are more expensive that three
fixed transmitters.

Under the two section ring architecture, each node is
equipped with three fixed transmitters and six fixed receivers.
The SAN nodes also require an additional transmitter as they
have access to four logical rings. As nodes in this architecture
can receive packets on any wavelength, each node is assigned a different
subcarrier multiplexed tone. The Destination’s subcarrier tone,
representing the packet’s destination address, is multiplexed by
the source node into the packet. At the same time, nodes
constantly monitor all wavelengths in parallel. If a node
detects its own subcarrier tone, it receives the packet. A
physical implementation of such a mechanism was proposed in
[17] where sub-carrier multiplexed tones are transmitted over
the wavelength along with the data to identify the destination.

Destination stripping is applied, i.e. marking the slot
empty after receiving the packet is the responsibility of the
destination node. This empty slot can be used by the same
node if the node has a packet to send. We also introduced and
evaluated in [18] a fairness mechanism where the node that
marks the slot empty is not able to use the slot immediately.

The results indicated that this restriction reduces the
bandwidth utilization efficiency of the ring while introducing
little change in the throughput difference between nodes.
Therefore this restriction is not used here.

### MAC Protocol Fixed Size Slot

```plaintext
MAC_Protocol_Fixed_Size_Slot()
Begin
Global_Time = time since simulation started
for (each node) do {
  for (each logical ring) do {
    if (Node_Buffer is not empty AND Current_Slot is empty) {
      Move packet to the Current_Slot;
      Log Queuing_Delay;
      Transmitted_Packets ++;
    }
    Global_Time = +Fixed_Slot_Time;
  }
}
End
```

### MAC Protocol Variable Size Slot

```plaintext
MAC_Protocol_Variable_Size_Slot()
Begin
Global_Time = time since simulation started
Current_Acting_Node = node chosen to transmit
Node_Next_Time = Global_Time + time increase to make
the node point to the next slot
Min_Time = minimum increase to the Global_Time
Min_Time = random number > Node_Next_Time;
for (each logical ring) do {
  if (Node_Next_Time < Min_Time) {
    Min_Time = Node_Next_Time;
    Current_Acting_Node = Node_Id;
  }
  Node_Next_Time =+ Slot_Time;
  Global_Time = Min_Time; /*Set Global_Time to the least
Node_Next_Time*/
if (Current_Acting_Node_Current_Slot is empty) {
  Check the size of Current_Acting_Node_Current_Slot;
  Determine the Node_Buffer according to the Current_Slot_Size;
  if (Node_Buffer is not empty) {
    Move packet to the slot;
    Log Queuing_Delay;
    Transmitted_Packets ++;
  } else
    Wasted_Slots ++;
}
End
```

Fig. 3 Algorithm of the MAC protocol of the fixed size slot scheme

Fig. 4 Algorithm of the MAC protocol of the variable size slot scheme
A simple technique is used to mirror each primary SAN node to its corresponding secondary SAN node. Under this mirroring technique, the secondary SAN nodes do not send any traffic to the primary SAN nodes and ordinary nodes do not send any traffic to the secondary SAN nodes. However, the secondary SAN nodes ultimately receive all the traffic addressed to the primary SAN nodes as the primary SAN node remove a packet from a slot upon reception only if its corresponding secondary SAN node has already received this packet. Otherwise it will let the packet remain in the ring to go to the secondary SAN node. Therefore those packets passing by the primary SAN node first will travel further in the network to be mirrored in the corresponding secondary SAN node which means extra bandwidth is used. However, on average this proposed mirroring scheme saves bandwidth and introduces efficiency in that separate transmissions are not needed to synchronise the SAN and its mirror. The Two remain synchronised at all time subject to the ring propagation delay. The receiving algorithm under the mirroring technique is shown in Fig. 6. To indicate whither the packet is received by the other corresponding SAN node, a flag, initially set to zero, is attached to each packet destined to one of the SAN nodes. As it receives a packet, each SAN node checks this flag. If it has been changed to 1 (indicating that the packet has been received by the other corresponding SAN) the packet will be removed from the ring. Otherwise it remains in the ring to continue its way to the other corresponding SAN.

According to the original MAC protocol, for the single section architecture, nodes in the upper part of the network can use either ring B or ring A to send to the primary SAN node. Although both rings result in the same distance to the primary SAN node, ring A results in a longer distance to the secondary SAN node. This extra distance increases mirroring time and leads to inefficient bandwidth usage. To reduce bandwidth usage and the mirroring time for the upper nodes, a modification can be introduced to the MAC protocol. In this modified version of the protocol, the upper nodes have to use ring B to send to the primary SAN node. To overcome the extra load introduced to ring B, a wavelength can be taken from ring A and assigned to rings B and C. Ring A can accommodate its traffic in a single wavelength as less traffic travels through it (40% to 60% is assumed to be destined to the SAN node in the asymmetric scenario). The mirroring technique with the modified MAC protocol is applied to the two section ring. Fig. 7 illustrates the modified mirroring technique with the modified MAC protocol for both architectures.
Simulation is carried out to evaluate the node throughput and queuing delay of the two architectures under the proposed mirroring technique. Two different traffic models are used – Poisson and self-similar. It has been shown in [19] that LAN and WAN traffic is better modeled using statistically self-similar processes. However, Poisson models are still used because they are analytically tractable and can be modeled easily compared to self-similar processes; see e.g. [20, 21].

Networks of 24 nodes are simulated. This is a typical node count in metro settings, however some of our previous work has also considered 16 nodes [9]. The performance of the network is evaluated under varying levels of traffic loads. The presence of the SANs creates traffic asymmetry and hot node scenarios i.e. the access nodes send to the primary SANs with a relatively higher probability while they send to each other with equal probabilities. For the single section ring, the primary SAN receives 40% or 60% of the total traffic. For the two section ring this amount is divided equally between the two primary SANs.

V. PERFORMANCE EVALUATION

Fig. 7 Mirroring Technique

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A. Single Section Ring Architecture

Results of the single section ring emphasize the performance difference introduced by the mirroring technique under the fixed-size slot scheme. Performance of the mirroring technique with the FS slot scheme is also evaluated under self-similar traffic. The results also compare the performance of the VS slot and SS slot schemes. All these results are shown under both 40% and 60% asymmetric traffic.

A maximum aggregated rate of 28 Gb/s (5 Gb/s from the SAN and 23 Gb/s from the access nodes) is generated by nodes without the mirroring technique, and 32 Gb/s (10 Gb/s from the SANs and 22 Gb/s from the access nodes) with the mirroring technique. This maximum traffic represents a normalized load of 1, L=1. Note that several choices exist and the two extreme cases (mirror does not transmit data, and mirror transmits to all nodes at the same rate as the primary SAN) are considered. Intermediate scenarios exist where for example nodes choose to retrieve data from the SAN or mirror according to proximity. In this case the SAN and the mirror each transmits only to a subset of the total nodes.

As mentioned before 4 is the minimum number of wavelengths giving a total bandwidth around 15 Gb/s (2.5 Gb/s × 6, where for rough guidance only we assume that there are 3 rings with 2 wavelengths per ring, however in reality the rings are not strictly separated) and therefore a normalized load of 1 will create more traffic on the ring than the total carrying capacity of an unslotted WDM ring network, however the slotted regime introduces a further spatial multiplexing gain. Therefore the network is stressed at L=1 allowing the performance of the protocol introduced to be examined.

To reflect the effect of traffic asymmetry and mirroring on the performance of the network, average results relating to access nodes in the upper part of the network are shown separately from those of access nodes in the lower part of the network. Also results of the primary SAN are shown separately from those of the secondary SAN.

Fig. 8-a shows the node throughput of the upper access nodes. The maximum throughput is achieved without the mirroring technique under 40% asymmetric traffic. Under 60% asymmetric traffic the throughput is slightly less than the maximum under L=1 as more traffic is destined to the primary SAN which heavily loads ring B. It can be seen that applying the mirroring technique reduces the throughput. This is due to the increase in the transmission rate of the secondary SAN which reduces the bandwidth available to the upper access nodes. The reduction in throughput is significant under 40% compared to 60% asymmetric traffic as under 40% asymmetric traffic ring A, which has a single wavelength, is more loaded. Fig. 8-a also shows that the use of self-similar traffic reduces the throughput significantly. This is expected due to the burstiness of the traffic produced by self-similar sources. The significant improvement obtained by applying the VS slot scheme is also clear in Fig. 8-a. This improvement is due to the increase in slot space utilization, which is defined as the ratio of used slot space to the total slot space. As mentioned before, in the FS slot scheme, the size of the slot is 1500 bytes. However, according to the packet sizes and packet size...
distributions used (see section IV), the average size of the packets is equal to 867.4 bytes, which is about 58% of 1500 bytes. This means that only 58% of the slot space is used on average and the rest is wasted. In the VS slot scheme, the slot space is fully used. Therefore the maximum throughput of the FS slot scheme is predicted to be around 58% of that of the VS slot scheme. However the throughputs obtained in this case do not match these predictions as the maximum throughput is achieved with the VS slot schemes. However results in the two section ring will match the predictions. The difference between VS and SS slot schemes will be clear in the queuing delay results.

Without the mirroring strategy, the lower access nodes (Fig.8-b) achieve lower throughput than that achieved by the upper access nodes. This is understood as packets from upper access nodes destined to the primary SAN can be transmitted through either ring A or B. However this is not the case for lower access nodes where these packets have to be transmitted through ring C. Lower access nodes perform better under 40% asymmetric traffic compared to 60% asymmetric traffic as higher proportion of traffic sent to the primary SAN means more load on ring C. Applying the mirroring technique results in reducing the achieved throughput under 40% asymmetric traffic. However, under 60% asymmetric traffic the throughput increases. This is a result of removing a wavelength from ring A and adding it to ring C as under 60% asymmetric traffic, traffic going through ring A can be accommodated in one wavelength i.e. giving more bandwidth to ring C traffic without affecting traffic on ring A. It can also be seen that self-similar traffic reduces the throughput as with the upper access nodes. The same trends of the VS and SS slot schemes for upper access nodes can be noticed for the lower access nodes.

For the primary SAN (Fig.8-c), it can be seen that without mirroring the network performs better under 60% compared to 40% asymmetric traffic as increasing the proportion of traffic going to the primary SAN means more slots will be emptied and possibly reused by it. The mirroring technique reduces these slots as half of the packets (those from the upper access nodes) destined to the primary SAN continue their way to the secondary SAN. Therefore the bandwidth available to the primary SAN decreases and the bandwidth available to the secondary SAN increases. Similar to the “without mirroring” case, 60% asymmetric traffic outperforms 40% asymmetric traffic under the mirroring technique. The figure also shows that with the mirroring technique the throughput decreases under self-similar traffic. It can also be noticed that the SS slot scheme has resulted in further improvement in the throughput under loads less than or equal to 0.9. This can be understood if we remember that in the VS slot scheme, the different size slots are only allowed to carry packets with the corresponding size. However, if an empty slot arrives and the buffer, which contains its corresponding packets, is empty, the slot will be released to the downstream nodes which will affect the slot utilization probability (the ratio of the number of empty slots which have been used to the number of total empty slots). On the other hand, the SS slot scheme achieves a better slot utilization probability as the super slot can accommodate all packet sizes. However, it can be seen from the figures that under extremely high loads (L> 0.9) the VS slot scheme outperforms the SS slot scheme. It is noticed from the figure that for the mirroring cases the network becomes heavily loaded at loads higher than 0.5 where the throughput is almost constant up to a load of 0.8. For higher loads, the throughput starts to decrease.

For the secondary SAN (Fig.8-d), without the mirroring technique the throughput under 40% asymmetric traffic is higher than 60% asymmetric traffic as ring C is more loaded under 60% asymmetric traffic. As mentioned before, under the mirroring technique the maximum transmission rate of the secondary SAN increases to 5 Gb/s. Other trends in Fig.8-d are similar to those of the primary SAN.

Fig.9-a shows the upper access nodes average queuing delay. It can be seen that with the mirroring technique under 40% asymmetric traffic after a certain load the queuing delay appears to be almost constant as the buffers become full. All the trends noticed for the throughput of FS slot scheme can be noticed for the queuing delay. The figure also shows that the use of self-similar traffic gives higher queuing delay than with Poisson traffic. The difference between the VS and SS slot schemes is clear. For the lower access nodes (Fig.9-b) similar trends are noticed.

Similar trends to those of node throughput are observed for the queuing delay of the primary SAN (Fig.9-c) under Poisson traffic. It can be seen from the figure that without the mirroring technique under 60% asymmetric traffic that after a certain load the queuing delay appears to be almost constant as the buffers become full. The figure also shows that the use of self-similar traffic gives higher queuing delay than with Poisson traffic when the load is below 0.8. For higher loads, the performance under self-similar traffic is better than under Poisson traffic. This is understood from the nature of Poisson traffic whose transmission rate, at high loads, becomes more constant as the packet interarrival duration decreases. Under L= 1, the congestion state becomes severe due to the almost nonvariable, constant arrival of packets when the nodes transmission buffer are constantly filled with packets. Therefore, the queuing delay also increases rapidly when the congestion state is reached. On the other hand, under self-similar traffic, packet bursts tend to frequently increase the buffer loads and the average queuing delay value is seen to be higher than with Poisson traffic when L<0.8. However, as the maximum transmission capacity of the network is reached, the buffers are filled in an intermittent way and the congested state does not induce an abrupt decline in networking performance. Therefore, the queuing delay appears to be almost constant for higher loads.

Fig.9-d shows the queuing delay of the secondary SAN. Due to the higher transmission rate, the queuing delay increases with mirroring at loads less than 0.9. For higher load the queuing delay for the secondary SAN without the mirroring technique is higher as ring C becomes highly loaded and fewer packets are emptied by the node compared to the mirroring technique. Applying self-similar traffic results in reducing the queuing delay for loads higher than 0.6 as with the primary SAN. Also the queuing delay after a certain load appears to be almost constant as with the primary SAN. The
trends of VS and SS slot schemes are similar to those of the node throughput.

B. Two Section Ring Architecture

Simulation results of the two section ring compare the performance of the different slot schemes. All the results are also shown under both 40% and 60% asymmetric Poison traffic. For comparison reasons results of the FS slot scheme are shown also under self-similar traffic.

For the two section ring architecture, the aggregated data rate is 40 Gb/s (20 Gb/s from the SANs and 20 Gb/s from the access nodes). As mentioned before 6 is the minimum number of wavelengths for this architecture giving a total bandwidth around 25 Gb/s (2.5 Gb/s × 10, where we assume that there are 5 rings with 2 wavelengths per ring). To reflect the effect of traffic asymmetry and mirroring on the performance of the network, average results of the access nodes are shown separately from those of the SANs (introducing the two section create symmetry in the performance of different parts of the network).

Fig.10-a shows the average throughput of the access nodes. As with the single section architecture the significant improvement obtained by applying the VS slot scheme is clear. Also it can be seen that while the maximum average throughput achieved under the VS slot scheme reaches 980 and 930 Mb/s for 40% and 60% asymmetric traffic, respectively, it was limited under the FS slot scheme to 590 and 540 Mb/s for 40% and 60% asymmetric traffic, respectively, which is around the theoretical predictions mentioned previously. It is also noticed from the figure that under the three different slot schemes, the throughput achieved under 40% asymmetric traffic is higher than that under 60% asymmetric traffic. This is understood as higher proportion of traffic sent to the SANs unbalances traffic between logical rings.

Fig.10-b presents the average node throughput for the SANs. The significant increase in the throughput achieved by the VS slot scheme compared to the FS slot scheme is noticed. The average node throughput under L=1 increased from 2390 to 4000 Mb/s under 40% asymmetric traffic, and from 2550 to 4330 Mb/s under 60% asymmetric traffic. These values are not far from the predictions. Unlike the performance of the access nodes, it is noticed that 60% asymmetric traffic outperforms 40% asymmetric traffic as higher proportion of traffic sent to the SANs means more packets are emptied and possibly reused by them. The difference in performance between the VS and SS scheme is not clear for the node throughput. It is noticeable for the queuing delay as discussed below.

The average queuing delay for the access nodes and the SANs are shown in Fig.11-a and Fig.11-b, respectively. In both cases, it is clear that applying the VS slot scheme has significantly reduced the queuing delay. As with the single section ring, the SS slot scheme has resulted in further reduction in the queuing delay under loads less or equal to 0.9. Other trends in Fig.11 are similar to the throughput. From Fig.10 and Fig.11, it is clear that applying self-similar traffic results in a worse performance as with the single section ring.
a) The upper access nodes
b) The lower access nodes
c) The primary SAN node
d) The Secondary SAN node

Fig. 8 Node Throughput of the single section ring architecture
Fig. 9 Queuing delay of the single section ring architecture
Fig. 10 Node Throughput of the two section ring architecture

Fig. 11 Queuing Delay of the two section ring architecture
VI. Conclusions

In this paper, the performance of a novel mirroring technique was evaluated. Simulation was carried out for two metropolitan WDM ring architectures of 24 nodes under both Poisson and self-similar traffic. In addition to the FS slot scheme, the performance was evaluated under two different slot schemes accommodating variable size packets — VS slot and SS slot schemes. Results of node throughput and queuing delay were presented and analyzed. For the single section architecture, the results showed that applying the proposed mirroring technique has different effects on different parts of the ring. For the upper access nodes mirroring degraded the performance. The deterioration was significant under 40% compared to 60% asymmetric traffic. For the lower access nodes, while applying the mirroring technique resulted in worse performance under 40% asymmetric traffic, under 60% asymmetric traffic the performance improved. The primary SAN performance was impacted under the mirroring technique and the maximum transmission rate of the secondary SAN increased to 5 Gb/s.

For the two section ring, the results showed that the access nodes achieved good performance under the FS slot scheme. Because of their high transmission rate, the SAN nodes suffered from more performance degradation compared to the access nodes. Significant improvements in the performance of both architectures were obtained under VS slot and SS slot scheme.

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Optical Code Processing System, Device, and its Application

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Abstract—Recent progress of optical code processing technology is explained. Ultra-high speed time domain, spectral domain, hybrid domain, and multiple optical code processing devices and systems are shown. As application of these technologies, OCDMA-PON, OPS network, and ultra high-speed optical clock generation will be demonstrated.

Index Terms—optical code and label, optical processing, optical code division multiplexing (OCDM), optical packet switching (OPS), ultra high-speed optical clock, photonic network

I. INTRODUCTION

Recently, in spite of immaturity of optical technology, many optical code generation and processing technologies and related technologies have been developed because of their potential abilities of ultra high-speed and/or ultra wideband processing and many applications in future photonic network. In optical code division multiple access (OCDMA) and optical packet switching (OPS) system, optical code (i.e. optical label in OPS systems) generation and processing are key technologies [1, 2]. Many different types of optical code (OC) processing technologies have been proposed and demonstrated.

In 1997, we have proposed and experimentally demonstrated 200G chip/s, 8 chip long, binary phase shift keying (BPSK), code processing technology using transversal filter (TVF) type devices made by planer lightwave circuit (PLC) [3, 4] as time domain code processing technology. Then, time domain optical code processing technologies based on super structure fiber Bragg grating (SS-FBG) devices are developed by many gropes and applied to OCDMA and OPS systems [5, 6].

Spectral domain optical code processing has been studied for a long time and may be the most popular technology to generate and recognize optical code [7]. This method is introduced in many OCDMA and OPS system demonstrators [8, 9].

Time-spread/wavelength-hopping, hybrid domain, optical code processing technologies based on multi-section fiber Bragg grating (FBG) devices have been developed and experimentally demonstrated. This type methods are also used OCDMA [5] and OPS [6] demonstrations.

In all these previous method, a device can process a code only. Therefore, large number of processing device is required in the real system. Recently, multiple optical-code processing method has been proposed [10], and multiple code processor is developed and experimentally demonstrated [11, 12]. A multiple code processor can generate and recognize many optical codes simultaneously by a device.

In this paper, optical code processing technologies are explained as examples of recent progress. Ultra-high speed time domain, spectral domain, hybrid domain, and multiple optical code processing devices and systems are shown. As application of these technologies, OCDMA-PON, OPS network, and ultra high-speed optical clock generation are demonstrated.

II. ALL-OPTICAL CODE PROCESSING

A. All optical Hierarchical Code Processing

The principle of hierarchical optical code is introduced in the work of Ref. [13, 14]. The hierarchical optical code (OC) is determined by a series of convolution operation of code functions with different chip intervals. In Fig.1(a), three stage BPSK hierarchical code $H_{3}(t)$ is given by convolution of three code functions, $a(t), A(t), and AA(t)$. $\theta$ and $\pi$ stand for optical phase. Here, 200G chip/s, 128-chip length two stage hierarchical optical BPSK codes are introduced. Processing (en/decoding) methods of 128-chip hierarchical optical codes are shown in Fig.1(c) and (d). Here, a 25G chip/s, 16-chip encoders, $A(t) and B(t)$, are used together with 200G chip/s, 8 chip encoders $a(t) and b(t)$. $A(t)$ consists of 1x16 splitter, tuneable attenuators, delay lines, phase shifters, and a combiner (Fig.1(b)). In $a(t)$, tuneable taps are used instead of splitters and attenuators in $A(t)$ (Fig.1(b)). Both encoders are integrated by PLC technology. Decoders have same structures with encoders. Encoder $A(t)$ divide a pulse (Fig.2(a)) into, 40ps interval, 16-chip pulses (Fig.2(b)) and give phase shifts either $0 or \pi$ to each chip pulses. Immediately, each of 16-chip pulses is divided again into, 5ps interval, 8-chip pulses to generate 128-chip pulses by encoder $a(t)$ (Fig.1(c)). Encoder $a(t)$ also gives phase shifts either $0 or \pi$ to re-divided chip pulses and outputs a 200G chip/s, 128-chip length optical BPSK code (Fig.2(c)).
Hierarchical code recognition is performed by correlation operation (i.e. matched filtering) between generated hierarchical codes and decoders $A^*(t)$, $B^*(t)$, $a^*(t)$, and/or $b^*(t)$ (Fig.1(d)). For example, Fig.2(d) and (e) represent auto-correlation outputs of decoder $A^*(t)$ to input labels $A(t)a(t)$ and $A(t)b(t)$, respectively. Steep camera traces of these central peaks are also represented. Both central peaks keep waveforms of 8 chip labels $a(t)$ and $b(t)$. On the other hand, Fig.2(f) shows cross-correlation output of decoder $B^*(t)$ to input label $A(t)a(t)$. Therefore, 16-chip decoders $A^*(t)$ and $B^*(t)$ can recognize only a 25G chip/s component in 128-chip hierarchical labels. 200G chip/s, 8-chip labels kept in auto-correlation peaks (Fig.2(d) and (e)) are recognized by 8-chip decoders $a^*(t)$ and $b^*(t)$. Figure 2 (g) and (h) are auto-correlation outputs of decoders $a^*(t)$ and $b^*(t)$ to signals shown in Fig.2 (d) and (e), respectively. In these cases, central peak has same waveform with conventional matched filtering with 8-chip PLC en/decoder [4]. Fig.2 (i) shows cross-correlation output of decoder $b^*(t)$ to input signal $A(t)a(t)A(t)$ shown in Fig.2 (d). In order to recognize 128-chip hierarchical labels, a set of 16- and 8-chip decoders is required. Consequently, 16- and 8-chip en/decoders enable hierarchical label generation and recognition [13, 14].

**B. Spectral ain ultiple Optical Code Processin**

Although AWG based spectral domain code processing device has been proposed [15], it dose not equip tune-ability. Therefore, we have proposed single port type fully-tuneable optical spectrum synthesizer (OSS) [16]. In order to increase the scalability and functionality of previous OSS, we have developed multi-port OSS based on high resolution cyclic AWG and PLC [17].
shifters (VOPS), mirror, and circulators. All components are fully integrated by PLC technology except circulators. The VOAs were fabricated as Mach-Zehnder interferometers having heaters on a planar waveguide circuit. The VOPSSs were fabricated by placing heaters on linear waveguides. The heaters attached to the VOAs and VOPSSs could be controlled by electrical currents supplied from a controller with PC. The refractive index of the waveguides changes depending on the amount of heat generated by the heater, allowing the amplitude and phase to be arbitrarily controlled. Therefore, OSS can control amplitude and phase of each spectrum components. The time required to control the VOAs and VOPSSs from one state to another was about 3 ms. Channel spacing and central wavelength of AWG is 20GHz and 1552nm, respectively. The optical pulse input from the in/out port of the AWG was separated into spectral components. The amplitudes and phases of the spectral components were controlled by the VOAs and VOPSSs. These components were reflected at the mirror, and recombined by the AWG. This recombined optical spectrum is output from the in/out port as a controlled optical pulse [16].

Figure 5. Configuration and operation principle of free space device for spectral domain optical signal processing.

Therefore, we introduce free space optics techniques and developed a variable-bandwidth spectrum shaper (VBS) as a high-resolution optical spectrum control system [18]. This system controls optical signals in the spectral domain with 10 GHz resolution for the entire C-band. The amplitude and phase of each spectral component can be controlled independently. Figure 5 shows the configuration of the VBS and the principle of pulse processing in the spectral domain. The VBS can process spectral components in the entire C-band. The VBS consists of a collimating lens, a grating serving as a functional dispersion device, a lens, a reflector, a polarizer, and spatial light modulators (SLM). The specifications are as follows. The channel-spacing of the VBS is 10 GHz, and the insertion loss is less than 6 dB. The VBS can be controlled in the wavelength range between 1535 nm and 1565 nm. The number of control channels is 340. The amplitude control range and resolution are 20dB and 0.1dB, respectively. Phase control range and resolution are 2π and 2π/50, respectively.

In contrast, it is possible to process ultrahigh-speed devices. Figure 5 shows also the principle of pulse processing in the spectral domain. The pulse repetition rate corresponds to the inverse of the frequency spacing of the spectral components. The mode spacing of an ultra high-speed optical pulse train is so broad that it could easily be processed in the spectral domain. The THz-rate optical clock is generated by suppressing the spectral components wider than 8 nm.

C. Free Space device or Spectral domain Optical Signal Processing

OSS based on cyclic AWG shows good performance to control optical signals in spectral domain. However, the device has problems such as limitation of port number and covered bandwidth, limitation of controllability and tunability, and signal distortion due to Gaussian pass band.

Figure 4. Configuration and operation principle of multi-port OSS

We have also developed a multi-port type OSS in order to perform multiple optical code processing [17]. The multi-port type OSS consists of 20x20 ports cyclic AWG, variable optical attenuators (VOA), variable optical phase shifters (VOPSS), mirror, and circulators as shown in Fig.4. Channel spacing and central wavelength of AWG are 10GHz and 1552nm, respectively. The spectrum focused to an output waveguides shifts by changing the input port when cyclic AWG with 20x20 in/out ports are employed in the OSS. The proposed OSS enables multiple optical codes generation and processing. This OSS performs spectral en/decoding in all wavelength band matching with free spectral range of cyclic AWG. Spectral domain code recognition is performed by same device [17].
In principle, it is possible to control the waveform of the pulse train by processing the amplitude and the phase of the spectral components [19].

Figure 6. Concept of time domain multiple optical code processing and photo of 16x16 multiple encoder/decoder.

In general, the all-optical recognition of OC-label is performed by the optical correlation between an incoming OC-label and the OC-label entries [2]. Optical encoder/decoder, which is basically the same passive optical device, generates and recognizes an OC-label. The main limitation of all-optical code generation and recognition is the hardware complexity. The code generation is performed by a set of different devices, one for each codeword, and accordingly an array of optical correlators have to be prepared to perform the code recognition as shown in Fig.2 (a). We have designed an innovative passive planar encoder/decoder (E/D) that is able to generate and to process a set of optical labels simultaneously as shown in Fig.6. We have fabricated a full E/D with 16 input/output ports, that is able to process/generate 16 optical codes in parallel [11-12]. Although this device has similar structure as the arrayed waveguide grating (AWG), it does not work as a wavelength de-multiplexer. It behaves like a transversal filter (TVF) for simultaneous multiple label generation and processing of 16-chips optical phase shift keying (PSK) codes. Figures 7(a) and (b) represent multiple code generation and recognition, respectively. The code chip interval is Δτ=5 ps [12]. The theoretical details of this device are expressed in Ref. [10, 11].

Figure 7 shows the experimental setup for characterisation of the device and parallel label recognition. To generate 16 different labels at the device outputs, a 10GHz Gaussian laser pulse of 2.5 ps width (i.e. FWHM: full width half maximum) from a mode locked laser diode (MLLD) is fed into one of the device input ports. Signals from output ports are examined with a streak camera. Figures 7 (a) shows simultaneously generated optical codes at output ports #1 to #16, respectively, when a pulse train fed into input port #1.

When a pulse train fed into input port #1 of encoder, the label generated at the output port #16 is amplified and forwarded to input port #16 of a second device that acts as a decoder. Decoded waveforms are also measured by a streak camera. Figures 7 (b) shows simultaneously processed (correlated) optical codes at output ports #1 to #16 of the decoder. Each generated label is a 16-chips PSK optical code with chips that have different phases. The amplitude variation of the chips is caused by the slab diffraction effect due to the mode profile in the input waveguides. In addition, the loss non-uniformity of the AWG like structure causes labels generated at different outputs to have different intensities.

Figure 7. Experimental demonstration of multiple optical code processing.
The full E/D processes all the 16 OC labels simultaneously, in parallel: if a codeword is forwarded to one of the device inputs, at the device outputs we measure the correlation signals between the incoming label and all the labels the device is able to generate. The autocorrelation peak (ACP), detected at one of the device outputs unequivocally identifies the incoming label. We measure the autocorrelation waveform at the output port #1 of the decoder (second device), whereas cross-correlation signals of much lower intensity are detected at all the other output ports (see Figs. 7(b)). The intensities at outputs #3 to #15 are practically zero, whereas at outputs #2 and #16, we can discern a cross-correlation peak (CCP). The code detection parameter, evaluated as the ratio between the ACP and the maximum CCP, is about 5. These results show this novel planar lightwave device, that is able to generate or recognize a large number of optical codes simultaneously.

We have also developed 50-chip, 500Gchip/s multiple OC encoder/decoders with arrayed waveguide configuration [20]. It can generate and recognize simultaneously 50 different optical phase shift keying (PSK) codes with low latency.

III. SYSTEM APPLICATION

A. All optical Hierarchical Code Processing or OPS Network

We have also proposed optical code/code-band routing and photonic UNI for optical label switched networks as an application of all-optical hierarchical code processing. The proof-of-principle experiments of these concepts are shown. Figure 8 (a) represents conventional wavelength path and wave-band routing. As an analogy with it, we have proposed a novel concept of OC-path and OC-band routing (Fig.8 (b)) [13]. OC/OC-band routing can be realized by using hierarchical label processing.

Figure 8. Network application of hierarchical optical code processing

As an application of OC/OC-band routing, we have proposed the photonic UNI for optical label switched networks. In the actual data communication network, user networks are interconnected with a provider network via UNI. UNI allows user and provider network to use their own address space independently. Figure 9 represents a set-up of proof-of-principle experiments of proposed concept. It consists of source node (Fig.9 (a)), transit node in provider network (Fig.9 (b)), and destination node in user network (Fig.9 (c)). Source node generate two optical data packets with same 16-chip label $A(t)$ and different 8-chip labels $a(t)$ and $b(t)$ as shown in Fig.10 (a).

Figure 9. OPS application of hierarchical optical code processing

Transit node switches packets according to matched filtering by 16-chip decoders $A*(t)$ and $B*(t)$ (Fig.10 (b)). Waveforms of switched packets are shown Fig.10 (c). Both packet route to upper port due to matched label $A(t)$. In user network, destination node switches packets according to matched filtering by both 16- and 8-chip decoders $A*(t)a*(t)$ and $A*(t)b*(t)$ (Fig.10 (d)). Packets with label $A(t)a(t)$ and $A(t)b(t)$ are switched to lower and upper ports, respectively (Fig.10 (e)). Here, in order to measure BER, only a data part of packet with label $A(t)a(t)$ is switched to lower port. Switched data show good BER property (Fig.10 (f)).

These results guarantee the high chip rate, long chip length, and reconfigurable OC labels processing technology and its networking application. The hierarchical OC label processing technology can provide ultra-fast new optical processing functions in photonic networks.

Figure 10. Experimental demonstration of OPS based on hierarchical optical code processing

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B. Spectral ain Optical Processin or ltra Hi h Speed Pulse train (Optical Clock) Generation and OC Syste

20GHz to 160GHz variable rate return to zero (RZ) and carrier suppressed return to zero (CS-RZ) pulse trains (optical clock) generation and spectrum domain matched filtering for OCDM system have been proposed and are experimentally demonstrated as applications of optical spectrum synthesizer (OSS) [16].

Figure 11 represents an experimental setup of pulse train generation.

Figure 11. Experimental set-up for pulse train generation

Figure 12. Experimental results of pulse train generation

In order to verify the CS-RZ pulse train condition, we have also done an additional experiment performed by 1 bit shift interferometer as shown in Fig. 13. Generated RZ and CS-RZ 160GHz pulse trains are fed into this 1 bit shift interferometer. Spectrum and autocorrelation of these output signals from this interferometer are measured and shown in Fig. 13 (a) and (b), respectively. RZ pulse train keeps its spectrum and waveform after 1 bit shift interferometer. On the other hand, CS-RZ pulse train has almost zero energy after this interferometer. These results represents the generated CS-RZ pulse has correct phase condition “0” and “π”.

Figure 13. Verification of CS-RZ pulse

Figure 14. Experimental setup of matched filtering
In addition, all-optical spectrum domain matched filtering and OCDM system are shown. Fig.14 represents an experimental setup. It consists of 10GHz MLLD, LiNbO3 intensity modulator (LN-IM) to reduce repetition rate to 1/64, pulse pattern generator (PPG), OTDM-MU, two AWG based optical pulse synthesizers as pulse encoders, and an optical pulse synthesizer as a pulse decoder. In this experiment, though Prim code of prim number 5 is used, all-kind amplitude and phase coding are realized.

Fig.15 (a) is source pulse train and its streak camera trace. Figs.15 (b) and (c) are encoded pulses by encoders #1 and #2. Their spectra are represented in Figs.15 (d) and (e), respectively. Waveforms of pulses spread into time-domain. Figs.15 (f) and (g) are waveform and spectrum of the merged signal, respectively. Figs.15 (h) and (i) are waveforms of decoded pulses by decoding code #1 and #2. Their spectra are represented in Figs.15 (j) and (k), respectively. In both code cases, only matched signal is extracted. Unmatched signal is well suppressed. These results guarantee all-optical processing technology by using OSS.

**C. Application of Spectral Signal Processing Performed by Free Space Device**

Free space devices for all-optical spectral domain signal processing shown in Section II-C are very powerful. They can perform almost all kind spectral processing done by OSS (see Section III-B). In addition, these free space devices have wider spectral range and flexibility in pass band shape. For example, we can generate ultra high-speed optical pulse train with Tera Hz repetition rate by using these free space device and wideband coherent light source such as supercontinuum (SC) light [18, 19]. Generation method is basically same with it shown in previous Section III-B [18]. The application of spectral domain signal processing performed by free space device could be wider than that by OSS.

**. Application of Multiple Optical Code Generation and Recognition**

The impact of time domain multiple optical code generation and recognition method to the OPS and OCDM system research is immense. In last 3 years, we have done many demonstrations on OPS and OCDM with this multiple optical code processing technologies. For example, asynchronous OCDM transmission capacity increased from “1 Gbps x 10 user” to “10 Gbps x 25 user”. In “1 Gbps x 10 user” demonstration, 10 conventional encoder devices are used to generate 10 different optical codes signal. On the other hand, we have used one device to generate 25 optical codes in the demonstration of “10 Gbps x 25 user”. Addition, in latter demonstration, we have also show the “10 Gbps x 25 OCDM-user x 5 WDM” transmission by using one encoding device and another one decoding device only. Total transmission capacity is 1.24Tbps [20]. This multiple optical code processing technology can accelerate the performance of OCDM and OPS systems.

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IV. CONCLUSIONS

All-optical code processing technologies have been explained as examples of recent progress. Ultra-high speed time domain, spectral domain, hybrid domain, and multiple optical code processing devices and systems have been shown. As application of these technologies, OCDMA-PON, OPS network, and ultra high-speed optical clock generation have been also demonstrated.

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SLA-Aware Survivability

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Abstract— The paper discusses the provision of differentiated guarantees to a population of users who share a network with different requirements for their connections. The basic concept underlying the proposed solutions is the required availability of the connections, both in the long term and during the period covered by the Service Level Agreement. An adequate metric for the latter is provided by the interval availability. The paper discusses how Markov chains may be used to model interval availability during the SLA period.

Index Terms—optical networks, availability, survivability, protection, restoration

I. INTRODUCTION

Optical networks emerged in the 1990’s as multi-client, transparent transport infrastructures aimed at providing on demand linkage capacity to assigned source-destination client node pairs. Some client networks had their own resilience capabilities, which could be overlaid in their own upper layer. Among them, SONET/SDH networks outranked all others in their fast recovery mechanisms and guaranteed uniform protection against any single failure. For this reason, early attempts to provide survivability in optical networks were based on layering a SONET/SDH network on top of the optical network, or on mimicking some of its protection mechanisms in the optical layer (without reaching its superior recovery speed).

During the current decade, IP traffic has become the dominant component of traffic in communication networks, and MultiProtocol Label Switching (MPLS) now provides a means of differentiating traffic according with QoS requirements and routing. New approaches are then needed to take advantage of MPLS features to provide QoS awareness in survivability mechanisms, in the benefit of efficiency in the allocation of network resources to meet a heterogeneous traffic demand.

The new approach implies a necessary shift in the design objectives to be met by the survivable network. Such objectives must now concern directly the client requirements, usually expressed in Service Level Agreements (SLA’s), and not some operational requirements of the shared infrastructure. This leads naturally to differentiated levels of connection survivability.

From the client perspective, the main connection survivability feature is the availability during the period covered by the SLA. Section III discusses the distinction between this metric, called interval availability, and long-term availability, usually known as availability. Section II reviews the evolution of resilience mechanisms in optical networks. Section IV shows how Markov modeling may be used to estimate availability of connections in Shared Backup Path Protection (SBPP) protected networks. Finally, Section V discusses some challenging issues regarding the extension of these results to the estimation of interval availability.

I. RESILIENCE MECHANISMS

When a network element fails, protected traffic that was assigned to the failed element must be re-routed so that the corresponding connections survive. Restoration time cannot exceed a threshold given by application requirements or by the triggering mechanisms of upper layer protection schemes. SONET/SDH – based protection mechanisms may activate path protection or line protection. Path protection is evoked at path terminations, and switches the end-to-end path traffic to a backup path. Line protection is evoked at the ends of a failed link, and re-routes all traffic that was assigned to it. For this reason, QoS-aware differentiation is more naturally performed under path protection.

For fastest possible recovery, dedicated protection must be provided. However, dedicated path protection is extremely inefficient in the allocation of network resources. Moreover, most applications from the IP world do not require the recovery speeds that are provided by automatic path protection switching mechanisms. For these reasons, current optical networks may safely share backup capacity resources among two or more paths (connections). In this scheme, called Shared Backup Path Protection (SBPP), the backup paths are pre-computed, but cannot be pre-configured because switching must wait for failure identification, thus leading to (usually acceptable) longer recovery times.

In SBPP, it is important to notice that practical optical networks bundle wavelengths into fibers, fibers into cables, cables into ducts, etc., and all such higher

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granularity resources are subject to cuts, leading to concurrent failures of the bundled resources. The set of resources that share the risk of any such failure is called a Shared Risk Link Group (SRLG). In order to provide survivability under any single failure, including SRLG failures, SBPP must conform to the SRLG constraint, which states that the backup paths of any two paths that share the same SRLG must not share any backup capacity. If the design objective is only the survivability under single failures, backup resource sharing in SBPP is limited only by the SRLG constraint, leading to high gains in resource allocation efficiency, especially in closely meshed topologies. However, backup capacity sharing degrades the survivability of connections when multiple failures are accounted for. Sharing must then be limited by the protected connections availability requirements.

III. CONNECTION AVAILABILITY

Availability is the average probability that a system is found operative in a period that tends to infinity [1-5]. Interval availability is the fraction of an observation period in which the connection is operative. Notice that the interval availability is essentially different from the availability: the first is an observable random variable, while the second is a steady-state, asymptotic probability value. Service Level Agreements (SLAs) require usually a maximum connection downtime [6], which can be in turn translated to a corresponding minimum interval availability.

The interval availability is a random variable with a relatively high dispersion around the mean for the typical SLA duration of 1 year or less. Thus, the distribution of the interval availability is a meaningful resource for planning and operating networks with SLA-aware survivability. The interval availability cumulative distribution function can be interpreted as the risk that an specific guaranteed interval availability is not honored. Fig. 1 shows an illustrative scenario in which two nodes are connected by three distinct paths. Each path supports a single wavelength.

<table>
<thead>
<tr>
<th>Mechanism</th>
<th>Availability</th>
</tr>
</thead>
<tbody>
<tr>
<td>Unprotected</td>
<td>0.996</td>
</tr>
<tr>
<td>Dedicated (DPP)</td>
<td>0.999984</td>
</tr>
<tr>
<td>Shared (SBPP)</td>
<td>0.999976</td>
</tr>
</tbody>
</table>

Figure 1. Simple network scenario.

IV. MARKOV MODELLING

The availability of connections protected by DPP can be easily estimated using the traditional equations for series/parallel systems. But shared-protection schemes are considerably more complex. We have proposed an analytical approximation [4] and analytical bounds [5] which estimate the availability of connections protected by shared schemes. Both methods model the network failure state by Markov chains such as in Fig. 3. For the sake of simplicity, at most double link failures are considered. In state $S_{0i}$ all links are operating. In state $S_{Fi}$ the network has a single failure in fiber link i. In state $S_{Fi,Fj}$ the network has a double failure in fiber links i and j, and fiber link i failed before fiber link j.

After the network is modeled by the Markov chain, the next step is to identify which are the states of the chain in which the connection is inoperative. As for an unprotected connection, these are all the states which involve a failure in the working path. As for a DPP protected connection, these are all the states which involve double failures in the working path and in the backup path. As for a SBPP protected connection, these are all the states which involve double failures in the working path and in the backup path, or first in the working path of other connection (part of the sharing...
group), and then in the working path of the connection whose availability is being evaluated. The connection unavailability can be estimated by adding up the balance probabilities of the states in which the connection is inoperative. As an example, Fig. 3 shows the availability calculation for connection 1 of the scenario presented in Fig. 1.

The analytical estimation of the interval availability distribution, however, is a complex task, even for the most elementary systems. One could resort to simulations, but this would be too time-consuming for common planning tools. A plausible alternative, which is relatively complex but still quicker than simulations, is the method described in [7]. The method first models the network failure state by the Markov chain described above, and then applies an existing numerical method proposed in [8].

In the scenario shown in Fig. 1, link failure rates are equal to \( \alpha = (200 \text{ FIT/km}) \cdot (1000 \text{ km}) = 2 \cdot 10^{-4} \text{ hour}^{-1} \), and link repair rates are given by \( \beta = (1/20) \text{ hour}^{-1} \). All such uniform failure and repair rates generate a “failure traffic” with “intensity” \((\alpha/\beta) = 0.004 \text{ Erlang in the network. The existence of a common value } \beta \text{ for all link repair rates, as in Fig. 1, makes it possible to lump all single-failure states of the network model of Fig. 3 into one state 1, all double-failure states into one state 2, and so on, as shown in Fig. 4a.}

Identical links, such as in Fig. 1, will make all \( k \)-failure states in Fig. 3 to be equally likely, so the probability of each (ordered) \( k \)-failure state is given by failure rates and a common value \( \beta \) for all link repair rates, when the network has no failure (state 0), is in any of 2 of the three possible single-failures that put the network at inoperative. As an example, Fig. 3 shows the availability probabilities of the states in which the connection is unoperative. The connection 1 of the scenario presented in Fig. 1.

Identical links, such as in Fig. 1, will make all \( k \)-failure states in Fig. 3 to be equally likely, so the probability of each (ordered) \( k \)-failure state is given by failure rates and a common value \( \beta \) for all link repair rates, when the network has no failure (state 0), is in any of 2 of the three possible single-failures that put the network at inoperative. As an example, Fig. 3 shows the availability probabilities of the states in which the connection is unoperative. The connection 1 of the scenario presented in Fig. 1.

Equations (4) yield the following solution for the steady-state probabilities:

\[
\begin{align*}
\pi_0 &= \left[1 + 3 \frac{\alpha}{\beta} + 3 \left(\frac{\alpha}{\beta}\right)^3\right]^{-1} \equiv 1 - 3 \frac{\alpha}{\beta} = 0.988; \\
\pi_1 &= 3 \frac{\alpha}{\beta} \pi_0 \equiv 3 \frac{\alpha}{\beta} = 0.012; \\
\pi_2 &= 3 \left(\frac{\alpha}{\beta}\right)^2 \pi_0 \equiv \left(\frac{\alpha}{\beta}\right)^2 = 4.8 \times 10^{-5}; \\
\pi_3 &= \left(\frac{\alpha}{\beta}\right)^3 \pi_0 \equiv \left(\frac{\alpha}{\beta}\right)^3 = 6.4 \times 10^{-8}. 
\end{align*}
\]

If connection \( c_1 \) is unprotected, it will be available when the network has no failure (state 0), is in any of 2 of the three possible single-failures that put the network at state 1 (\( S_{F2} \) and \( S_{F3} \)), or is in any of 2 of the six possible double-failure states (\( S_{F1F3} \) and \( S_{F3F2} \)). Therefore, the availability of an unprotected connection is:

\[
A_{\text{unprotected}} = p_0 + \frac{2}{3} p_1 + \frac{1}{3} p_2 \equiv 1 - \frac{1}{3} p_1 = 0.996. 
\]

If connection \( c_1 \) has dedicated protection, it will be available if and only if the network has no failure (state 0), has any single-failure (state 1), or is in any of 4 of the six possible double-failure states (\( S_{F1F3} \), \( S_{F3F1} \), \( S_{F2F3} \), and...
Therefore, the availability of a connection with dedicated path protection is:

\[ A_{dpp} = p_0 + p_1 + \frac{2}{3} p_2 \equiv 1 - \frac{1}{3} p_2 = \]

\[ = 1 - \left( \frac{\alpha}{\beta} \right)^2 = 0.999984. \]  \hfill (7)

Finally, if \( c_1 \) shares the backup capacity with \( c_2 \), it will be available if and only if the network has no failure (state 0), has any single-failure (state 1), or is in any of 3 of the six possible double-failure states \( \{S_{2F2}, S_{2F3}, \text{and } S_{3F2}\} \). Therefore, the availability of a connection under shared backup path protection is:

\[ A_{spp} = p_0 + p_1 + \frac{1}{2} p_2 \equiv 1 - \frac{1}{2} p_2 = \]

\[ = 1 - \frac{3}{2} \left( \frac{\alpha}{\beta} \right)^2 = 0.999976. \]  \hfill (8)

The results obtained above are referenced to a canonical protection model in which each link is assumed to have its own private maintenance resources. Hence, upon the occurrence of multiple failures, say a \( k \)-failure, \( k \) maintenance teams are assumed to work simultaneously, multiplying by \( k \) the effort that would be spent for a single-failure. If network maintenance resources are limited, so that only one repair team is available for the whole network, then the Markov model of Fig. 4b would be closer to reality. In this case, the balance equations would be:

\[ 3\alpha p_0 = \beta p_1; \]  \hfill (9a)

\[ 2\alpha p_1 = \beta p_2; \]  \hfill (9b)

\[ \alpha p_2 = \beta p_3. \]  \hfill (9c)

Combined with ever-holding (4d), (9) yield:

\[ p_0 = \left[ 1 + \frac{3\alpha}{\beta} + 6 \left( \frac{\alpha}{\beta} \right)^2 + 6 \left( \frac{\alpha}{\beta} \right)^3 \right]^{-1} \equiv \]

\[ \equiv 1 - \frac{3\alpha}{\beta} = 0.988; \] \hfill (10a)

\[ p_1 \equiv \frac{3\alpha}{\beta} = 0.012; \] \hfill (10b)

\[ p_2 \equiv 6 \left( \frac{\alpha}{\beta} \right)^2 = 9.6 \times 10^{-5}; \] \hfill (10c)

\[ p_3 \equiv 6 \left( \frac{\alpha}{\beta} \right)^3 = 3.84 \times 10^{-7}. \] \hfill (10d)

Equations (6) to (8) would then yield numerical values 0.996 (unchanged) for \( A_{unprotected} \) but 0.999968 for \( A_{dpp} \), and 0.999952 for \( A_{spp} \). The new numbers show that keeping only one repair team for the whole network will double the unavailabilities of both dedicated and shared backup path protection, while generating negligible degradation in the availability of unprotected connection. This example illustrates the fact that canonical models must be applied with care to any real life environment: a network, just like people, is also a network and its circumstances!

\section{V. CHALLENGING ISSUES}

SLA-aware survivability is clearly a technoeconomical issue. The optimization problem involves the allocation of the least possible network resources constrained by conditions determined in SLAs. Price, maximum connection downtime, and non-conformity penalties are typical SLA figures. This section presents a case study which aims at raising some discussion about the challenging issue of solving such optimization problem.

Let the utility curve of a connectivity customer be a logarithmic function of the guaranteed interval availability \( IA \) (see the black curve in Fig. 5a):

\[ u[IA] = \ln[100,000(IA - 0.998) + 1], \]

\[ \hspace{1em} 0.998 \leq IA \leq 0.99999 \]

Utility functions are usually concave, representing the maximum price the customer accepts to pay for a specific good. Suppose now that the costs of providing shared (\( C_{sp}[IA] \)) and dedicated (\( C_{dp}[IA] \)) protection are given by:

\[ C_{sp}[IA] = k_{sp} + pr_{sp}[IA]; \]

\[ C_{dp}[IA] = k_{dp} + pr_{dp}[IA]; \]

where \( k_{sp} \) and \( k_{dp} \) are the provisioning costs, \( r_{sp} \) and \( r_{dp} \) are the risks of SLA violation, and \( p \) is the non-compliance SLA penalty agreed by contract. Constants \( k_{sp} \) and \( k_{dp} \) depend on the amount of network resources allocated for shared or dedicated protection \((k_{sp} < k_{dp})\), and the risks \( r_{sp}[IA] \) and \( r_{dp}[IA] \) are properties of the network estimated by methods such as in Reference [7] \((r_{sp} > r_{dp})\).

Fig. 5a shows \( C_{sp}[IA] \) and \( C_{dp}[IA] \) when \( k_{sp} = 0.5 \), \( k_{dp} = 1 \), and \( r_{sp}[IA] \) and \( r_{dp}[IA] \) are calculated as in Fig. 2. The SLA penalty is set to \( p[IA] = 20 \cdot u[IA] \), meaning that the network operator pays 20 times the period income if the agreed maximum downtime is not honored. It is interesting to notice that for \( IA < 0.9995 \) the cost of shared protection is lower than of dedicated protection, whereas for \( IA > 0.9995 \) the curves switch positions. Fig. 5b shows the expected profit (difference between utility and cost) for shared or dedicated protection. Clearly, shared protection is the best choice if \( IA < 0.9995 \), and dedicated protection otherwise.
VI. CONCLUSION

The interval availability is the most common survivability metric specified in SLAs. The cumulative distribution function of the interval availability is therefore an important resource for planning networks with SLA-aware survivability. Markov modeling can be used to analytically derive the connection availability, and numerically derive the interval availability distribution. The whole optimization problem of efficiently allocating capacity for connections with guaranteed survivability is actually a techno-economical problem. As an example, we presented in this paper a simple, illustrative case study on how to choose a protection scheme in order to maximize the profit.

REFERENCES


JAVOBS: A Flexible Simulator for OBS Network Architectures

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Abstract—Since the OBS paradigm has become a potential candidate to cope with the needs of the future all optical networks, it has really caught the attention from both academia and industry worldwide. In this direction, OBS networks have been investigated under many different scenarios comprising numerous architectures and strategies. This heterogeneous context encouraged the development of various simulation tools. In this paper we present our novel Java-based OBS network simulator called JAVOBS. We discuss its architecture, study its performance and provide some exemplary results that point out its remarkable flexibility. This flexibility should permit an easy integration of upcoming new network protocol designs but also support changing and evolving research goals.

Index Terms—Optical burst switching (OBS), simulation tool, flexibility, performance evaluation.

I. INTRODUCTION

To move towards IP-over-WDM architectures, various optical switching techniques have been under intensive research. Among them, three switching paradigms appeared as potential candidates. First, Optical Circuit Switching (OCS) [1] pursues a wavelength routed networking architecture with a whole wavelength as finest granularity. However, it lacks both the flexibility and efficiency required to cope with the needs of current traffic patterns. Second, in the Optical Packet Switching (OPS) approach [2], each packet is sent into the network together with its own header. This header is either electronically (or even roughly all-optically [3]) processed at each intermediate node while the packet is optically buffered. Although OPS may be seen as both the natural choice and conceptually ideal for the future all-optical networks, current optical technology is still immature and not able to overcome its exigencies. Finally, in order to provide optical switching for next-generation Internet traffic in a flexible yet feasible way, the Optical Burst Switching (OBS) paradigm was proposed in [4][5]. In OBS networks, burst control packets (BCPs) are sent out-of-band both to reserve all resources and to set up the path for their associated data bursts, which will be sent optically after an offset time in a cut through manner. In this way, OBS allows for an efficient use of resources without the need of optical buffering at any intermediate node. Although it can be seen as an intermediate step of the migration from OCS to OPS, OBS has emerged as a more competitive choice for the transmission of data traffic in the near future. In essence, OBS combines the best from both OCS and OPS while avoiding their shortcomings. Consequently, OBS has received an increasing amount of attention from the optical research community and has become, nowadays, a research field of its own.

OBS networks display a complex structure and the design of their constituent elements offers several degrees of freedom. So far, much of the research on OBS networks has been conducted through theoretical analysis. Undeniably, the analytical approach can provide valuable insights in reduced complexity scenarios but might scarcely cope with the multiple factors that hide behind a complete network schema. Simulation tools have become essentials to evaluate complex OBS network scenarios. Indeed, simulators solve many difficulties such as the need to build a real system, but more important, they allow for the reproducibility of results, which is the basis for scientific advance [25].

In this paper, we present our Java-based OBS network simulator (JAVOBS), which was firstly presented in [24]. Considering how rapidly new strategies are engineered to improve the performance of OBS networks, it is our objective to demonstrate how versatile a simulation tool should be in order to be able to provide reliable results in a relatively fast yet straightforward way. Section II gives an insight of the wide variety of OBS schemes proposed so far as well as it reviews OBS network simulation tools presented in the literature. Section III presents the architecture of the JAVOBS simulator. Section IV provides some numerical results that both validate the simulator and show its flexibility when implementing different OBS protocols and algorithms. Section V summarizes the flexibility and extendibility of JAVOBS. We conclude this paper in Section VI.
II. OBS REVIEW

An OBS network is made up of two types of nodes, namely edge and core nodes. Edge nodes are in charge of both assembling input packets coming from different sources (e.g., IP, Ethernet) into outgoing bursts and of disassembling incoming bursts. For each outgoing burst, edge nodes emit a separate BCP in advance, to reserve resources (i.e., bandwidth on a desired output channel) along the way from the ingress node to an egress node. Core nodes in OBS are responsible for switching individual bursts and for reading, processing, and updating burst control packets. Core nodes are generally assumed wavelength conversion capable.

The BCP carries, among other information, the remaining offset time at the next hop (i.e., the time separating the arrival of the BCP from the arrival of the burst), and the burst length.

A. Burst Reservation Protocols

In order to transmit bursts over an OBS network, a resource reservation protocol must be put in place to ensure the allocation of resources and to properly configure the optical switch before the corresponding data burst arrives at the node. Two different approaches were designed. A wavelength-routed OBS reservation protocol was proposed in [7] as a two-way reservation scheme (i.e., a burst cannot be sent without the successful reception of an acknowledgement). Nevertheless, much of the research has been devoted to the one-way reservation scheme aiming to reduce the light-path setup time and consequently increase the resource utilization in OBS networks. The just-in-time (JIT) [8], Horizon [5] and just-enough-time (JET) [4] resource reservation protocols are the most well-known one-way reservation schemes. More recently, JIT+ [9] and E-JIT [10] protocols have also been proposed. The main difference between all one-way reservation schemes stems from the manner in which output wavelengths (i.e., channels assuming wavelength conversion) are reserved for bursts. These schemes include: (a) immediate reservation (JIT, E-JIT); (b) delayed reservation with void filling (JET); (c) delayed reservation without void filling (Horizon); (d) modified immediate reservation (JIT+).

A comparison of the JIT, JIT+, JET and Horizon protocols can be found in [9]. Delayed schemes produce a more efficient use of resources, especially when void filling is applied, and perform better in terms of burst loss probability. However, the sophisticated scheduling algorithms that they require increase the processing times of BCPs at intermediate nodes. Thus, in such a scenario, the simplicity of JIT may balance its relative poor performance [9]. Indeed, in contrast to the other protocols, hardware implementations of the JIT signaling protocol have already been realized and published [11].

B. Burst Scheduling

When a core node receives a BCP, it must decide which output channel should be reserved to later forward the burst corresponding to this BCP. Scheduling algorithms aim to transfer efficiently the input traffic to the desired output while configuring the switching matrix adequately.

To date, several algorithms have been proposed to solve the wavelength scheduling problem in OBS networks. They can be divided into two sets depending whether they perform void filling (a) or not (b). Algorithms belonging to group (b) pursue simplicity when searching an available wavelength. They are not aimed to maximize the use of resources but to generate low processing times. A simple scheduling algorithm based on the Horizon reservation protocol and called latest available unused channel (LAUC), was proposed in [8]. Another example is the first fit unscheduled channel (FFUC) algorithm [13].

More advanced scheduling algorithms belong to group (a). These algorithms are designed both to provide efficient use of resources and to reduce blocking probabilities. However, void filling algorithms are more complex, hence difficult to implement and imply high processing times. Among the void filling algorithms one finds: (1) latest available unused channel with void filling (LAUC-VF) [12]; (2) first fit unscheduled channel with void filling (FFUC-VF) [12]. More recently, the minimum starting void (Min-SV) and the minimum ending void (Min-EV) scheduling algorithms were presented in [14]. Min-SV and Min-EV algorithms improve significantly the processing time over LAUC-VF. However, Min-SV/EV algorithms involve time-consuming memory accesses. Therefore, the void filling algorithms are still considered too slow to provide a viable solution to the problem [15]. Table 1 summarizes the comparison between the algorithms based on the study in [16]. It uses the following notation: (w) number of wavelengths at each output port; (N_b) number of bursts currently scheduled on every wavelength.

C. OBS Simulation Tools

OBS networks are still in a phase where several options may have their own opportunity. Therefore, there is a strong need to mimic the behavior of real OBS networks. That is precisely the task of simulation tools. Since OBS is a relatively young field, much of the studies that can be found in the literature use quite simple simulation models. For instance, several proposals have been applied only to a single node [9][21]. In general, these simulation models were developed in purpose for a specific situation and are not suitable to study complete
OBS scenarios. On the other hand, some well-known simulators such as the widely known ns-2 [17] or the IKR Simulation Library [18], have or allow extensions for the study of OBS networks. A comparison of some existent OBS simulator tools can be found in [19]. To our best knowledge, none of them was specifically developed for the study of OBS networks, and thus do not provide support to the full set of OBS representations. Besides, given their divergence of perception of the OBS scenario it is not possible to compare their results [19].

In consequence, other tools exclusively aimed to analyze OBS networks have been proposed. Two new simulation models are presented in [20][21]. Both exploit the object-oriented approach using either the C++ in the former case or the Java programming language in the latter. The common goal of these new models is to reach the flexibility degree that simulation of OBS networks requires. Following a modular construction process, a high degree of flexibility is exhibited. At the same time, the introduction of further developments is facilitated.

Yet another OBS network simulator (ADOBS) has been developed in C++ [6]. Formerly, ADOBS served to study routing algorithms in OPS networks. Lately, it has been modified to become an ad-hoc event-driven simulator for OBS networks. ADOBS has been basically used to study the performance of the OBS network layer. Since C++ is a low level programming language, the developer deals with concepts and operations strictly connected with computer hardware. Hence, speed and efficiency are achieved at the cost of complexity.

III. JAVOBS ARCHITECTURE AND FEATURES

The JAVOBS simulator is a Java-based application that has been exclusively built to simulate OBS networks on top of the JAVANCO framework [22].

A. The JAVANCO Framework

The JAVANCO framework is programmed within the Java 1.6.0 platform, using the popular Java programming language. It has been conceived to provide a coherent object oriented structure that is able to properly represent graph and network topologies in a compelling yet versatile way. Over this fundamental structure, several packages offer a variety of features including graphical visualization, support for disk serialization of topologies and execution of common graph algorithms. It is thanks to these core packages that the user can rapidly develop and test network planning procedures through the construction of simulation models.

By its nature, Java is an interpreted language. This means that user code is temporarily compiled into "Java byte code", and does not become executable code until the program is actually run. Consequently, C++ runtime performance is better than that of Java. Nevertheless, Java has been selected both to avoid the complexity of building a simulator completely from scratch using C++ and to benefit from the many advantages provided by the Java environment. In particular, Java being a garbage-collected language, the procedures of memory handling are greatly simplified.

Figure 1 shows a general representation depicting the architecture of JAVANCO. The cornerstone of its architecture is the NetworkHandler object, in charge of both organizing the references towards each object composing the graph (i.e. layers, links and nodes) and providing access to several managers and engines (e.g. user interface manager, serialization manager, script engine). JAVANCO permits to load and save files that describe network topologies and their components. This functionality makes use of the Multilayer Network Description (MND) proposed in [23], which is based on the XML standard. Taking advantage of this description format, it is easy to associate several attributes to any element present in a network topology (e.g. the capacity in a link).

JAVANCO embeds a script engine which allows calling any functionality of the framework and dispenses the user to write complete Java classes. It also supports different kinds of user interfaces.

B. JAVOBS Features

Two general models exist to conduct discrete simulations: next-event time progression and fixed-
increment time progression [21]. The JAVOBS core simulation engine implements both. Each simulation is thus separated in fixed length steps, but events can occur within a step with fine time granularity. Arrivals of bursts at either intermediate or egress nodes are implemented with events. Other activities such as traffic generation and reservation list updates are executed at the beginning or at the end of each step. This hybrid simulation scheme mainly permits to generate the traffic progressively and thus keep the memory usage moderate. Furthermore, it also permits, under a minor constraint, to execute concurrently all the events expiring in one step, and therefore to take advantage of the parallel computing functionalities provided by recent CPUs. The constraint is not restrictive at all. Only the step time length should be shorter than the propagation delay of the shortest link.

JAVOBS offers a basic OBS model which can be adapted and modified in many ways. Using the functionality offered by JAVANCO, for example multiple OBS network topologies can be constructed over various NetworkHandler objects. These topologies can either be dynamically constructed, graphically designed, or loaded from MND files. Also, many of the aforementioned reservation protocols and scheduling algorithms have been implemented within JAVOBS to enhance the basic OBS model.

It has to be mentioned that the JAVOBS model offers a shortest path routing scheme, which uses link lengths defined in the JAVANCO NetworkHandler to compute shortest paths. This minimal routing logic can however be replaced by a more sophisticated one. JAVOBS offers two options to define the routing logic: (a) one unique routing element is defined, this element being then responsible for all routing operation during the whole simulation; (b) one routing element per node, each one supporting independent configuration.

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<th>TABLE II. ADOBS/ JAVOBS FEATURES COMPARISON</th>
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<td>Programming Language</td>
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JAVOBS offers another degree of flexibility with respect to the traffic generation. In many case studies, traffic characteristics are supposed to be independent of the source or destination node. In these cases, a unique traffic generator can be used to generate all burst sizes and departure times. However, JAVOBS also allows equipping each edge node with an independent burst generator, permitting in this way studies involving source and/or destination dependent traffic flows.

Eventually, JAVOBS natively supports simulations of the emulated offset time control architecture (E-OBS) [4] along with the conventional OBS control architecture (C-OBS). In Table 2 we summarize the abilities of JAVOBS and compare them with the aforementioned ADOBS simulator.

IV. SIMULATION RESULTS

The purpose of this section is to evaluate the performance and the flexibility of the JAVOBS simulator. We first provide results related to simulator validation and runtime tests. Second, we focus on four different case-studies: (1) Performance comparison of reservation protocols under both the C-OBS and the E-OBS control architectures supported in JAVOBS; (2) Comparison of the Horizon and the Constant Time Burst Resequencing (CTBR) [15] schedulers under the single node topology; (3) Analysis of the network topology flexibility using different degrees of meshed-rings; (4) Evaluation of the network-wide burst loss performance with different burst traffic statistics. Apart from presenting performance results we take the opportunity to discuss some OBS specific issues and, in particular, compare different OBS architecture and protocol proposals.

A. Validation and Benchmarking of JAVOBS

In order to assess its credibility, the JAVOBS simulator has been validated by means of an analytical model for the calculation of the network-wide burst loss probability and by comparison with results obtained with the ADOBS simulator. The analytical results are based on a reduced link load model for OBS networks presented in [26].

We use within both ADOBS and JAVOBS simulators a network topology called SIMPLE [6] with 6 nodes and 8 links, and compute an identical shortest path routing. Each node is an edge node generating 25.6 Erlangs (0.8, when normalized to the link capacity) and each link has a capacity of 32 channels. Bursts have exponential distributed arrival time and length. To keep relation with the real world, we set the channel capacity to 10 Gbit/s and the burst mean size to 1 Mb. In obtaining the simulation results, we estimated 99% confidence intervals. Since the confidence intervals found are very narrow, we do not plot them in order to improve readability. As it can be seen from Figure 2, the results obtained by JAVOBS match both the analytical results and the results obtained with ADOBS. Hence, in this case, we consider the simulator validated.
A test measuring the running time of both simulators has also been performed. Simulations were run according to the number of bursts generated and prompted more than one hundred hours of simulation (on an Intel Core 2 Quad 2.4 GHz desktop computer). In this case, we consider two network topologies: (1) NSFNET [27] (US network); (2) EON [28] (a pan-European network defined in European COST 266 action) with 15 and 28 nodes, and 23 and 39 links respectively. JET signaling and LAUC-VF scheduling are used.

The results obtained are shown in Figure 3. ADOBS performs better at low values of generated bursts, probably taking advantage of the C++ performance. However, tendency changes at about 1 million bursts. From this point on, the ADOBS curves exhibit an exponential increase which finally creates gaps of up to 96 hours between both simulators. This gap is apparently due to unoptimized memory utilization in the ADOBS simulator which obliges the operating system to use the hard disk as RAM extension, and thus, drastically reduces the simulator throughput. Although JAVOBS is outperformed in short simulations, we observe a constant growth of the running times for all time scales which exhibits its robustness. Thus, in this case, the benefits of using a garbage-collected language, which dispenses the user to take care of many memory management related operations, become apparent.

B. Evaluation of the E-OBS and C-OBS Architectures.

Considering that fiber delay line (FDL) buffers are not used, it has been proved in [29] that the best worst-case performance of an online best-effort scheduling algorithm is achieved when all bursts have the same offset time and the same length. One of the benefits of E-OBS comes from the fact that offset times are introduced at each core node by means of additional fiber delay coils inserted in the data path at the input port of the node. As a result, E-OBS does not experience offset variation inside the network. In such scenario, scheduling algorithms do not need to implement any void filling technique. Therefore, in an E-OBS network, JIT and Horizon reservation mechanisms seem to be the most appropriate ones due to its low complexity compared to JET. Indeed, the overprovisioning of resources that characterizes JIT is substantially reduced using E-OBS due to smaller offset times. Figures 4 and 5 present the results obtained in both control architectures under the different signaling protocols supported by JAVOBS.
We observe that using E-OBS, the performance of the five different signaling protocols is very similar, thus, the possibility of reducing the network complexity by using low complexity techniques such as JIT is not unfounded. On the contrary, in C-OBS becomes clear the advantage of using complex reservation mechanisms due to the variable offsets.

C. Implementation of the CTBR Algorithm.

Since OBS has ultra high speed requirements, the bandwidth efficient scheduling algorithms proposed so far are not considered a viable solution to the problem due to their large processing times. Recently, in [15], a hardware implementation of an optimal wavelength scheduler that can produce burst schedules in a time complexity of O(1) was presented. The idea consists of producing schedules by bursts arrivals rather than BCPs arrivals. The optimal wavelength scheduler consists of two components: (a) the CTBR block; (b) the horizon scheduler. It is important to notice that the driving force behind this technique is the simplicity of horizon and its ability to operate at high speed.

We developed a set of classes implementing this alternative OBS model. To perform the simulation, we used the parameters specified in [15] with the aim of comparing the results obtained. Since the topology utilized for the simulation is not mentioned, we assumed the single node implementation. The performance of both the Horizon and CTBR scheduler is compared. The offset times of all bursts are generated according to a lognormal distribution with mean 100µs. Figure 6 shows the results obtained. We observe a clear match with the results presented. The burst loss probability of the horizon scheduler increases when the ratio between the offset time standard deviation and the burst length increases. On the other hand, in the CTBR scheduler, the curves remain flat regardless of the ratio variation.

D. Flexible topology simulations.

The flexibility of JAVOBS has been tested in respect of the topologies with the aim of demonstrating that JAVOBS allows topological modifications in a straightforward way. We evaluate the simulator adaptability performing a set of simulations over different degrees of meshed-ring topologies.

The study begins with an 8 node ring topology with 32 wavelengths per link and ends with 28 links (full-mesh) and 9 wavelengths per link. At each step of the study, a bench of simulations is conducted on the topology. Then, topology is extended with additional links. However, in order to keep constant the network capacity, the number of wavelengths per link is recomputed at each step.

Figure 7 shows the results of the simulations conducted on each intermediate topology. A shortest-path routing algorithm has been used. The BCP arrival rate $\lambda$ of BCPs is maintained constant for all scenarios. As expected, the blocking probability is evidently reduced as more direct links between each source-destination pair become available.
D. Impact of Burst Traffic Statistics on Burst Loss Performance

Eventually, in order to study the network-wide burst loss performance, we used the JAVOBS capability of synthesizing diverse burst traffic statistics. Several statistical distributions are available for creating either burst arrival times or lengths, and consequently, different combinations of burst traffic statistics.

The analytical model that we used for the validation of our simulator does not depend on the burst length distribution chosen; however, it only holds for burst arrivals following a Poisson process. We thus evaluated the difference between the analytical model and the simulation results when burst arrivals follow statistical distributions different from the abovementioned Poisson. Concretely, we run two different sets of simulations.

In the first case, we preset the burst length distribution to Exponential with mean burst length equal to 1 Mb, and alternatively, we used Exponential, Uniform, Gaussian and Deterministic burst arrival distributions with their corresponding parameters set according to the load generated. Afterwards, we performed a second set of simulations; however, this time, we preset the burst arrival distribution to Exponential, and alternatively, the burst length distribution is replaced. In this experiment we used the NSFNET network topology and JET signaling together with LAUC-VF scheduling.

The results shown in Figure 8 agree with the analytical model. While the second set of simulations exhibits negligible differences with respect to the analytical values, the first set of results presents gaps of up to 28% between both. Since the analytical model assumes exponential burst arrival times to compute the burst loss probability, these results verify, again, the right performance of the JAVOBS simulator.

V. SUMMARIZING JAVOBS FLEXIBILITY

The JAVOBS simulator we presented all through this contribution provides the user with a dedicated OBS simulation framework. This can be adapted, modified or extended in many ways, implying different level of confidence with the Java programming language and with the JAVOBS API. This section recapitulates what can be changed and adapted referring to the examples presented in the previous section.

Very fundamental parameters such as switching time or offered traffic rate are given at the beginning as input values. Results displayed on Figure 2 are thus straightforward to reproduce. To configure the reservation protocol and the scheduling algorithm is also a straightforward operation, which makes Figures 4 and 5 easily reproducible, too.

Changing the routing logic and the traffic generators (Figure 7) is slightly more complex. Basic knowledge of the Java language is required. Similarly, studies involving topological (Figure 8) modification at runtime require knowledge of the JAVANCO API and of Java.

To setup studies where traffic is generated according to specific rules (e.g. constant flows, self similar traffic or aggregation of finer granularity traffic), additional implementations of burst generators are required. In the same way, prototyping of more complex scheduling algorithms is also possible by implementing new classes. Additional classes have to comply with well defined rules and implement strictly defined methods.

Eventually, several changes have been required in the core simulation engine to generate the results of Figure 6 (CTBR). However, in obtaining these results, most of the functionalities developed for conventional OBS were reused, and thus, they did not involve a whole reimplementation of the simulator.

Table III summarizes the configurable or extendable parts of JAVOBS.

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<tr>
<th>Parameter or functionality</th>
<th>Requirements</th>
<th>Difficulty</th>
</tr>
</thead>
<tbody>
<tr>
<td>Processing time,</td>
<td>none</td>
<td>Very low</td>
</tr>
<tr>
<td>Switching time,</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Mean burst size,</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Offered rate,</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Simulation time,</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Number of simulations</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Scheduling Algorithms,</td>
<td>OBS fundamentals</td>
<td>Very low</td>
</tr>
<tr>
<td>Reservation Protocols</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Topology (from file)</td>
<td>XML and MND knowledge</td>
<td>Low</td>
</tr>
<tr>
<td>Topology (dynamic</td>
<td>Basic Java and Javanco knowledge</td>
<td>Low</td>
</tr>
<tr>
<td>modification)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Routing logic,</td>
<td>Basic Java knowledge</td>
<td>Low</td>
</tr>
<tr>
<td>Traffic generators</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Alternative scheduling</td>
<td>Java knowledge</td>
<td>Moderate</td>
</tr>
<tr>
<td>algorithms, reservation</td>
<td></td>
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</tr>
<tr>
<td>protocols, traffic</td>
<td></td>
<td></td>
</tr>
<tr>
<td>generator or routing</td>
<td></td>
<td></td>
</tr>
<tr>
<td>logics</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Alternative simulation</td>
<td>JAVOBS knowledge, Java knowledge</td>
<td>High</td>
</tr>
<tr>
<td>schemes or OBS models</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
VI. CONCLUSIONS

We have presented our novel Java-based simulation tool JAVOBS, which has been exclusively developed for the study of OBS networks. We have also given a recent overview of the existent simulation tools for OBS networks. We have verified that comparisons between simulators were impossible due to their heterogeneity. The JAVOBS simulator has been described, validated and compared with an ad hoc C++ based simulator.

The flexibility of our simulator has been highlighted through a series of experiments that exhibit its performance. From the results of these experiments, it is concluded that: (1) as OBS networks are still undergoing intense research and development, its study requires simulation tools that facilitate the introduction of enhancements and new techniques, (2) as long as the simulation model is valid, flexible simulation tools such as JAVOBS can save time and computational resources.

ACKNOWLEDGMENT

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REFERENCES


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- Getting submissions, arranging review process, making decisions, and carrying out all correspondence with the authors. Authors should be informed the Instructions for Authors.
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