Coarse WDM/CDM/TDM Concept for Optical Packet Transmission in Metropolitan and Access Networks Supporting 400 Channels at 2.5 Gb/s Peak Rate

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Abstract—To improve the networking flexibility in the metropolitan and access area, the granularity in the optical domain has to be increased above that in the core network requiring more channels at lower bit rates. Pure dense wavelength division multiplexing (DWDM) as it is applied in the core does not seem to meet this requirement at affordable cost. We propose and analyze a network based on hybrid optical multiplexing techniques including wavelength, code, and time division multiplexing. Applied to optical packet transmission this approach enables several 100 all-optical channels between end users and headend with average bit rates up to 100 Mb/s per channel while keeping the installation and maintenance cost at a minimum.

Index Terms—Code division multiplexing, multiaccess communication, network reliability, optical crosstalk, optical noise, packet switching, time division multiplexing, wavelength division multiplexing.

I. INTRODUCTION

In metropolitan area networks (MAN) the required number of independent optical channels is likely to increase to >100 with channel bit rates up to the Gb/s range within the near future. This fine granularity enables flexible utilization of fiber bandwidth and simple network reconfiguration. Network operators will be able to lease single optical channels or groups of optical channels to service providers on demand. Even in access networks with multiple optical channels, the increase of total throughput to the Gb/s range is anticipated [1]. It is, however, questionable whether the straight forward scaling of today’s dense wavelength division multiplexing (DWDM) technology, as it is presently used in the core network, to the specific requirements of MAN and access networks is an affordable option for the realization of >100 optical channels. Specially for packet-based services like Internet with average data rates per user that are up to two orders of magnitude lower than the peak rate [2] the implementation of DWDM in network parts close to the end user will not be a realistic option. The spectral efficiency of such networks would be extremely poor making this approach very expensive.

With increasing channel number and taking into account the optical bandwidth of fiber networks that is mostly limited by the gain bandwidth of optical amplifiers the channel spacing will rapidly drop below the current 100 GHz and 50 GHz spacings. The technological and economical effort that has to be spent to realize even smaller optical channel spacings increases dramatically and is hardly justified by the required channel capacity. The main issues to be encountered at the optical transmitters and receivers, when the channel spacing drops down to the 10 GHz range, are selection, tuning and control of the laser center frequency, the requirement for external modulation even for low channel bit rates and the fabrication and control of narrowband optical filters. Even without considering additional impairments from fiber transmission (e.g., channel crosstalk due to nonlinear effects) and optical amplification (gain flatness, transient effects) these systems are expected to become extremely demanding with respect to component specifications and network management.

For packet-based transmission systems the additional application of time and/or code division multiplexing (TDM, CDM) techniques in the network is a favorable way to increase the channel number. TDMA (time division multiple access) is a viable option for increasing the channel number in simple fiber network structures with limited capacity to several 100 [3]. CDM techniques can also be applied to more complex network architectures, since some optical CDM approaches can be realized to support asynchronous operation of channels. Moreover, with asynchronous CDM a statistical multiplexing gain can be achieved in packet transmission networks enabling a large number of channels to be allocated. The bursty nature of the traffic helps to reduce the performance degradation due to multiple access interferences. Alternatively, optical TDMA can be applied to a limited part of the network, typically in the access part, whereas in the MAN part WDM and CDM techniques are implemented.

There have been different proposals how to implement optical CDM applying time domain [4], [5] or wavelength domain [6], [7] coding. Also, the combination of WDM and CDM techniques has been proposed in some papers [8]–[10]. We favor spectrally encoded CDM due to the fact that, in contrast to most time domain encoded CDM approaches, the different channels do not have to be synchronized to each other to achieve full orthogonality between codes. This guarantees independence of the optical multiplexing scheme from the network architecture. Moreover, in time domain CDM the optical pulses would have
to be much shorter than the bit period, so that the optical components must satisfy Gb/s specifications even for Mb/s channel bit rates.

In this paper, we present a detailed system proposal for the implementation of CDM enhanced WDM based on previous experimental and theoretical studies that we have performed on a special form of spectrally encoded optical CDM. Our CDM approach applies periodic spectral encoding of broadband optical sources like light emitting diodes (LEDs) [11]. This approach has some advantageous features like stability and insensitivity of system performance with respect to component drift and specifications making it very attractive for MAN and access applications. On the networking side the benefits are asynchronous operation of different optical channels, support of different signal formats and ease of redistributing electrical bandwidth among optical channels. After a short review of the operating principle these aspects will be discussed in the next section with reference to experiments that have been performed in multichannel system demonstrators transmitting continuous data streams. Then the basic system limitations, namely channel crosstalk and optical intensity noise, will be discussed yielding some simple rules of thumb to estimate the achievable channel number and bit rate. A full numerical simulation is used for system evaluation when using different transmitter and receiver configurations. From this analysis two network scenarios for optical multichannel transmission are derived that also include time domain features like statistical multiplexing and synchronous TDMA within limited network areas. At the end impacts from transmission of broadband optical signals are discussed, specially from chromatic dispersion induced signal distortions.

II. CONCEPT AND FEATURES OF PERIODIC SPECTRAL ENCODING

The concept of periodic spectral encoding as proposed by Möller [11] is a generalization of coherence multiplexing [6] in the sense that all kind of optical filters having periodic power transmission characteristics are considered for encoding and decoding. The optical output power of a thermal light source, like an LED, is intensity modulated by the electrical data. After emission the broadband spectrum is optically shaped using passive filters with periodic transmission functions like Mach–Zehnder (MZ) or Fabry–Perot (FP) filters (Fig. 1). The periodicity of the transmission function for these type of filters is given in terms of the free spectral range (FSR) which is defined by the filter round trip time $\tau_{\text{FSR}}$ for every pair of transmitters $j$ and $k$ in the order of the source coherence time $\tau_{c}$ larger. In practice the gain spectra of optical amplifiers define the coherence time, since usually they are narrower (<30 nm) than the source spectra (>50 nm). This has been experimentally verified in an 8 × 155 Mb/s system where the transmitter LEDs were taken from two different suppliers (MRV and Anritsu) with different spectral width [FWHM (full width at half maximum) = 53–68 nm] and center wavelengths (1539–1551 nm) [12]. The filter FSR in this system (MZ at the transmitters, FP with finesse 4 at the receiver) were chosen in the range 10–20 GHz, the power extinction of the transmitter filters was 10–13 dB and slightly polarization dependent, the FP receiver filter extinction ratio was only <9 dB. Nevertheless the system operated with bit error rate (BER) BER ≤ 10^{-10}. The second benefit of periodic encoding is that only the difference in filter roundtrip times $|\tau_{\text{TR}_{k}} - \tau_{\text{TR}_{j}}|$ for any two transmitters needs to be larger than some picoseconds. The absolute value of the $\tau_{\text{TR}}$ is not important as long as they are much larger than $\tau_{c}$. From that it follows that even temperature changes of more than 100 °C do not affect the system performance as proven by experiment [14]. The reason for this insensitivity to temperature changes is the low temperature coefficient of the refractive index of glass (about 10^{-5}/°C for fused silica [15]). Our MZ encoding filters were made from standard 3 dB fiber couplers, so that $\Delta FSR/FSR \approx 10^{-3}$ for the above temperature range. The fine tuning of the receiver filter to drifting transmitter filter FSR was accomplished by applying a local low-frequency dithering technique in the receiver [16], but no control of the transmitter filter FSR was required to keep BER < 10^{-9} [14].

The robustness of the system with respect to component specifications and drift as demonstrated above is unique to the approach applying periodic spectral encoding with broad source spectra and small filter FSR. Other code multiplex systems like
**III. Basic Limits on System Performance**

After having reviewed the basic features of the optical CDM approach applying periodic spectral encoding we will now dis-
other works must be obtained from the IEEE

Discussions about crosstalk have been discussed in [18] (note: the symbols used here slightly differ from those in [18]). Here, we derive the number of optical channels that could be allocated for the channel with \( k \neq k' \). This results in a Gaussian distribution of the crosstalk current \( \tilde{i}_{\text{t}}(\Delta \tau) \) as

\[
i_{\text{d}}(\Delta \tau) = S_0 \frac{\alpha_1}{2} \exp \left( -2(\pi \delta \nu \Delta \tau)^2 \right)
\]

with FWHM given by \( \text{FWHM}_{\Delta \tau} = (4 \ln(2)/\pi) / \Delta \nu \). The parameter \( \alpha_1 \) in (1) is the second coefficient of the transmitter cos-series \( H_{\text{Tx}}(\nu) = \sum_{m=1}^{\infty} \alpha_m \cos(2\pi m \nu \tau_{\text{Tx}}) \) with the value \( \alpha_1 = 1/2 \) for MZ or \( \alpha_2 = \pi/F \times \exp(-\pi/F) \) for FP with finesse \( F \). In the following, it is assumed that all transmitters continuously send NRZ signals (non return to zero) with equal probability \( p_0 = p_1 = 0.5 \) for “0” and “1.” This results in a binomial probability distribution of “1” signals at the receiver which for many interfering transmitters is approximated by a Gaussian distribution of the crosstalk current \( p(\tilde{i}_{\text{c}, \text{talk}}) = 1/\sqrt{2\pi \sigma_{\text{c}, \text{talk}}^2} \exp(-\tilde{i}_{\text{c}, \text{talk}}^2 / 2 \sigma_{\text{c}, \text{talk}}^2) \) with the mean value \( \langle \tilde{i}_{\text{c}, \text{talk}} \rangle = 1/2 \times \sum_{j \neq k} \langle \tilde{i}_{\text{c}, \text{talk}} \rangle \) where \( \langle \tilde{i}_{\text{c}, \text{talk}} \rangle \) is calculated from (1) for all transmitters \( j \neq k \) with \( \Delta \tau = \tau_{\text{Rx}} - \tau_{\text{Tx}} \). The standard deviation \( \sigma_{\text{c}, \text{talk}} \) is given by

\[
\sigma_{\text{c}, \text{talk}}^2 = \varepsilon \sum_{j \neq k} \langle \tilde{i}_{\text{c}, \text{talk}} \rangle^2
\]

with \( \varepsilon = 1/4 \) for channels transmitting synchronously to each other and \( \varepsilon = 1/6 \) in the asynchronous case [19]. With these results the crosstalk-limited Q-factor for the channel with \( k = [N/2] \) (\( \cdots \) denotes the integer part of its argument) can now be expressed as

\[
Q_{\text{c, talk}} = \left( \frac{\Delta \nu}{\text{FWHM}_{\text{c, talk}}} \right)^{0.5} \frac{R}{\text{SNR}_{\text{ave}}} \frac{1}{\sqrt{2 \ln(2) \delta \nu}} \sum_{j \neq k} \left( \frac{\langle \tilde{i}_{\text{c}, \text{talk}} \rangle^2}{\text{SNR}_{\text{ave}}} \right)
\]

with source FWHM \( \Delta \nu = \sqrt{2 \ln(2) \delta \nu} \) and PSR_{\text{max}} = 1/\tau_{\text{R}}.

The number of addressable codes is very large (cf. Fig. 5) if transmitter or receiver filter or both are MZ like filters. In case of an FP/FP combination the crosstalk around points
where $\tau_{R,\text{eff}}/\tau_{T,\text{source}}$ is a rational number strongly reduces the useful number of codes. However, even with MZ filters not all codes can be simultaneously used, since the intensity noise at the receiver imposes a more restrictive limit to the achievable BER. For single ended detection, i.e., no differential receiver, the intensity noise of a broadband thermal light source with coherence time $\tau_c (\tau_c = \sqrt{2} \ln(2) / \pi / \Delta \nu$ for Gaussian and $\tau_c = 1 / \Delta \nu$ for rectangular spectra) is given by $\sigma^2 = 2I_0^2 \tau_c B_{\text{d}}$ [20]. Here $I_0$ is the photocurrent and $B_{\text{d}} \ll \Delta \nu$ is the noise equivalent bandwidth defined via the normalized electrical receiver transfer function $h_{\text{d}}(f)$ as $B_{\text{d}} = \int |h_{\text{d}}(f)|^2 df$. It was assumed that the detected light is linearly polarized. For partly polarized light the electrical noise power is reduced by $(1 + P^2)/2$ where $P$ denotes the degree of polarization with $P = 1$ and $P = 0$ for polarized and unpolarized light, respectively [13]. If the source spectrum is passively filtered by a periodic optical filter with transmission function $H(\nu)$ and FSR $\leq \Delta \nu$ the relative intensity noise is enhanced by a factor $\Psi = (H^2)/\langle H \rangle^2$ (for MZ) and $\Psi = \cos[\mathcal{H}(\pi / F)]$ for FP which can be regarded as the noise figure of the optical filter for thermal light. The effective coherence time of $N$ independent passively filtered thermal light sources is $\tau_{\text{c,eff}} = \sum_{j=1}^{N} S_j / \langle S_j \rangle$ for identical source spectra $S(\nu)$ with coherence time $\tau_{\text{c,source}}$. For identical source spectra $S(\nu)$ with coherence time $\tau_{\text{c,source}}$, this reduces to

$$\tau_{\text{c,eff}} = \tau_{\text{c,source}} \left( \sum_{j=1}^{N} H_j \right)^2 \quad (3)$$

where again $\langle \cdots \rangle$ denotes the average over the optical frequency $\nu$. This spectrum is filtered at the receiver and differentially detected. With the effective receiver filter function $H_{\text{eff}} = \sum_{j=1}^{N} S_j \langle S_j \rangle / \langle S_j \rangle^2$ the electrical noise power at the output of the differential receiver is

$$\sigma^2_{\text{noise}} = 2I_0^2 \tau_{\text{c,eff}} \langle \Delta H^2_{\text{R,source}} \rangle B_{\text{d},\text{eff}}. \quad (4)$$

$I_0$ denotes the photocurrent that would be measured if the total spectrum $\tilde{P}(\nu) = \sum_{j=1}^{N} S_j(\nu) H_j(\nu)$ was directly detected with a single photodiode: $I_0 = S_0 \sum_{j=1}^{N} \langle H_j \rangle$. With the signal current for transmitter $\#k$, given by $I_k = S_0 \langle H_k \Delta H_{\text{R,source}} \rangle$, the optical noise limited $Q$-factor for transmitter $\#k$ in case of identical Gaussian source spectra with and MZ receiver filter is

$$Q_{\text{noise}} = \sqrt{\frac{\pi}{8 \ln(2)}} \cdot \frac{1}{\sqrt{N + \sqrt{2N} \cdot \sqrt{\Delta \nu / B_{\text{d},\text{eff}}}}} \cdot \frac{1}{\sqrt{\Delta \nu / B_{\text{d},\text{eff}}}}. \quad (5)$$

It was assumed that the noise originating from all transmitters is identical although the photocurrent from $\#k$ is slightly larger than the others. In practice, however, the above approximation is valid as supported from experiments. Furthermore, the optical power $S_0$ has only entered the formula by 50% to account for the data statistics [19]. The $Q$-factor in (5) can be optimized by the choice of the transmitter filter type and specifications. For FP transmitter filters with finesse $F \{\alpha_0 = \pi/(2F), \alpha_n = (\pi/F) \cdot \exp(-n \pi/F) \text{ for } n > 0\}$ the term in the denominator is minimized with the optimum finesse

$$F_{\text{opt}} = \frac{2\pi}{\ln(1 + \frac{1}{N})} \approx \pi \left(1 + \sqrt{2N}\right). \quad (6)$$

This relation is depicted in Fig. 4 showing optimum finesse values that can be easily realized from a technological point of view. The reason for the existence of an optimum finesse lies in the fact that with increasing $F$ the differential signal current degrades and the noise figure of the transmitter filters increases, but on the other side the interchannel beat noise is reduced. With the optimum finesse the optical noise limited $Q$-factor in the floor is related to the number $N$ of active channels by

$$Q_{\text{noise}} = \sqrt{\frac{\pi}{8 \ln(2)}} \cdot \frac{1}{\sqrt{N + \sqrt{2N} \cdot \sqrt{\Delta \nu / B_{\text{d},\text{eff}}}}} \cdot \frac{1}{\sqrt{\Delta \nu / B_{\text{d},\text{eff}}}} \approx 0.54 \cdot \frac{\sqrt{\Delta \nu / B_{\text{d},\text{eff}}}}{N^{0.63}}. \quad (7)$$

The last approximation is well suited up to $N = 100$. Equation (7) indicates that systems, where $N^{0.63} B_{\text{d}} / \Delta \nu = \text{const}$, is valid, show the same error floor under the above assumptions, specially that the filter finesse is optimized to the number of active transmitters. This is, of course, not feasible in practice. However, the dependence of the $Q$-factor on filter finesse is weak at large $N$ so that $F = 10–20$ is a good choice in many cases. Equation (7) indicates that the system employing FP/MZ filters performs slightly better as compared to MZ/MZ systems like coherence multiplexed systems [6]. The electrical signal-to-noise ratio (SNR) with respect to MZ/MZ systems is improved by about 3 dB for six channels and depends more weakly on channel number as compared to MZ/MZ systems. This improvement originates from the use of FP transmitter filters which reduce the optical interchannel beat noise.

The number of channels $N_{\text{active}}$ that are allowed to be active at the same time is reduced considerably as compared to the number of addressable codes $N_{\text{code}}$. The noise limited $Q$-factor for given $N_{\text{active}}$ is determined by the ratio $\Delta \nu / B_{\text{d}}$ [see (7)].
For a given number of channels $N_{\text{code}}$ to be addressed, on the other hand, the crosstalk limited $Q$-factor is determined by the ratio $\Delta \nu/\text{FSR}_{\text{code}}$ [see (2)]. The number of addressable codes and the number of active channels are compared as function of the source spectral width in Fig. 5 for FP transmitter filters with FSR = 10–20 GHz and optimized finesse. The requirement to be met was $Q_{\text{crosstalk}}/Q_{\text{noise}} \geq 6$ in the floor ($\text{BER} \leq 10^{-9}$) at 155, 622, and 2500 Mb/s (noise equivalent electrical bandwidth 122, 488, 1961 MHz with fourth-order Bessel filters). It was assumed that the channels were synchronized and that the detected light was linearly polarized, both assumptions yielding lower numbers as compared to the opposite cases.

There are many simplifications made in the calculations shown in Fig. 5, as indicated in the text. However, comparison with results from a full numerical model show that the results presented in the figure can serve as a guideline to the performance to be expected with such kind of systems. The calculations presented in the next section are based on a more complex mathematical description.

**IV. SYSTEM DESIGNS FOR OPTICAL PACKET MODE TRANSMISSION**

In this section, we discuss how an optical CDM network must be dimensioned to allow large channel numbers. Using a similar but more complete numerical model [19]–[21] as described in the last section we evaluated the number of channels that can be supported with $\text{BER} \leq 10^{-9}$, if the source spectra and the optical coding filters are optimized. Throughout this section, the source spectra are assumed to be unpolarized and the different CDM channels are assumed to be unsynchronized to each other. For reference (Fig. 6) the optical channel number is evaluated with Super-Gaussian spectra and $\text{BER} \leq 10^{-9}$. If the available optical bandwidth is divided into several WDM slices and optical CDM is applied within each of those slices. As discussed in the last section with $\varphi$MZ combinations large numbers of codes can be addressed, while the number of simultaneously active transmitters must be considerably lower. Such systems are attractive for optical packet mode transmission, taking advantage from statistical multiplexing to increase the optical channel number. In Fig. 7, nearly rectangular optical spectra are considered (mathematically described by Super-Gaussian spectra $\approx \exp\left(-x^{2m}\right)$ with $m = 16$) with spectral width of 10 nm and 1.6 nm (200 GHz). In the former case, 84 codes can be addressed (FSR = 10–20 GHz), in the latter case, there are 27 codes (FSR = 5–10 GHz) allowed before crosstalk increases the BER to $>10^{-9}$. The number of transmitters that may be active at the same time is limited to $\leq 34$ and $\leq 11$, respectively, due to optical noise limits. The 10-nm slice can be used to transmit optical packets with, e.g., 622 Mb/s peak rate ($N_{\text{active}} = 7$), the 1.6 nm slice, with, e.g., 155 Mb/s ($N_{\text{active}} = 5$). The channel spacing required in an equivalent DWDM system would be 190 GHz and 40 GHz, respectively, if the number of simultaneously active channels is taken as reference. However, in packet mode operation the required channel spacing for an equivalent DWDM system would be 16 GHz and 7.4 GHz for $N_{\text{code}} = 84$ and $N_{\text{code}} = 27$. In a packet mode CDM system either the packet statistics or the network management must assure that no more than $N_{\text{active}}$ channels are received at the same time. Violations of these limits might be corrected applying forward error correction in the receivers.

If the available optical bandwidth is, for example, 50 nm, it can be divided into five (10 nm) or 31 (1.6 nm) WDM slices. These system proposals then enable $N_{\text{code}} = 420$ and $N_{\text{code}} = 837$ optical packet mode channels within that bandwidth, not...
problems are to be expected with the optical power budget in that case. In the above simulations, it was assumed that the optical source power was ≥1 mW within the considered bandwidth before the optical encoding filter at the transmitters.

In the network scenario developed above there is one differential receiver for each optical channel. Since mechanically tunable filters are used, this requires much floor space and will be an expensive solution as long as the average channel capacity is well below 100 Mb/s. In this section, we propose an approach where the number of optical CDM channel is again reduced, but now each CDM channel carries many optical TDM signals. The proposed network comprising metropolitan and access fiber networks is depicted in Fig. 8 [22]. The optical CDM technique is supposed to be only applied in the upstream direction from the optical network units (ONU) to the headend, so the discussion is restricted to transmission in that direction. For downstream transmission a combination of DWDM and TDM seems better suited [22]. There are three cascaded optical multiplexing techniques in the network. Starting from the headend, there is a coarse WDM on the primary ring and optical CDM on the secondary ring. The link between both rings is realized by the indicated a coarse WDM coupler. The combined WDM/CDM operation is the same as discussed above, with the only exception that now the CDM channels are made up of sequential optical packets from a TDM network (cf. below) and are quasi-continuously active. So the CDM system must be optimized for continuous data transmission in contrast to the situation discussed in the preceding section. The best performance is achieved, if we use FP filters both at the transmitters and at the receiver (FP/FP combination). The finesse is set to \( F = 20 \). To reduce crosstalk and optical noise originating from the overlap of multiple integers of the filter FSR in terms like \( \langle H_j \Delta H_{RD} \rangle \) and \( \langle H_j H_j^* \rangle \) the FSR are allocated within the semi-octave 15–20 GHz using ratios of large prime numbers. The number of continuously active CDM channels that can now be supported within broad WDM slices is depicted in Fig. 9. Now transmission even at 2.5 Gb/s channel bit rate is possible with six (10) CDM channels within 10 nm (16 nm) WDM slices. However, the modulation transfer function of the FP filters is reduced to bandwidths in the order of 0.75–1.0 GHz leading to signal distortions at 2.5 Gb/s. Using an electrical decision feedback equalizer (DFE) after detection these distortions can be compensated. DFEs have been successfully applied to compensate for even larger ratios of electrical signal bandwidth to equivalent electrical filter bandwidth in dispersive LED transmission [23].

Each optical CDM channel in Fig. 8 carries optical packets from a TDM PON (passive optical network). The all optical conversion from TDM to CDM applies a technique that has been recently demonstrated for bit rates up to 10 Gb/s [24]. It is based on the optical modulation of the backward ASE (amplified spontaneous emission) output power of an SOA (semiconductor optical amplifier) or an LED by means of an intensity modulated laser diode (Fig. 10). The logic of the modulated ASE signals is inverted with respect to the input. This is, however, easily reverted in the electrical domain after detection. The modulated ASE is optically encoded using FP filters as above and now behaves like the other optical CDM channels discussed in this paper. Since the laser diode emission wavelengths may
Fig. 8. Optical network incorporating different cascaded optical multiplexing techniques (WDM denotes a wavelength selective coupler).

Fig. 9. Allowed number of simultaneously active channels in different slice widths for optimized FP/FP filter combinations.

V. Impacts from Transmission of Broadband Optical Signals

We will now highlight two issues arising from the broad spectral width of the optical signals: 1) optical amplification and 2) transmission over dispersive fibers.

Since both crosstalk and noise depend critically on the shape and width of the source spectrum as seen by the receiver, any deformation of the spectrum during transmission may degrade the system performance. However, the spectral gain shape of the optical amplifiers is not as critical as it might seem from first sight. With the exception of one fluoride-based EDFA we used conventional silica-based EDFAs with the well-known...
ASE peak around 1530 nm for low input signals. At high input power (>−10 dBm) the corresponding gain peak vanishes leaving a sufficiently flat gain response over almost 30 nm width. If the signal input power was so low that the gain peak around 1530 nm became too large we added a small piece of unpumped erbium-doped fiber after the silica-based EDFAs, thus equalizing the gain shape in a very simple yet effective way. This measure was accompanied by a 2–3 dB loss in total optical signal power. With narrowband optical signal spectra as proposed in the last section the requirement of gain flatness is more related to power equalization among the WDM slices, rather than to CDM specific crosstalk and noise issues. So the impact on system performance is not CDM specific in that case.

In the remainder of this section, we will concentrate on chromatic dispersion induced signal distortions.

The chromatic dispersion of the fiber link has a potential impact both on the signal and on the noise in systems operating with thermal light sources [25]. However, for broad source spectra the electrical noise power in the receiver is not significantly affected by the chromatic dispersion, even if transmission takes place far from the wavelength of zero fiber dispersion [19]. The degradation of the SNR is primarily due to distortions of the signal shape. In our experiments we used matched lengths of DCF to restore the optical signal shape before detection.

The question with optical dispersion compensation is, how exactly the DCF dispersion value must match the SMF dispersion to enable error free transmission, or in other words, how large is the length tolerance of optical compensation schemes. To answer that question we investigated the optical power penalty (eye opening penalty, EOP) that is induced, when high-speed thermal light pulses around 1550 nm are transmitted over uncompensated SMF links. The EOP is the additional optical power \( \Delta P_{opt} \) that is required to maintain a given BER in the presence of signal distortions. The value \( \Delta P_{opt} \) after transmission over fiber length \( L \) can be estimated from the relative pulse peak amplitude \( A \) for a single isolated “1” (losses neglected) [26]. It is calculated from

\[
\Delta P_{opt} = -10 \log (2A - 1).
\]
Assuming an electrical Gaussian pulse shape after low pass filtering in the receiver the peak amplitude $A$ of the detected optical pulse, relative to the input value at $L = 0$, can be expressed in terms of the parameter $\gamma = BDL\Delta \lambda$. $\gamma$ is a normalized dispersion and yields the asymptotic behavior of the relative broadening of the width $\Delta(L \to \infty)/\Delta L(0)$ of NRZ pulses at bit rate $B$. Here $D$ is the fiber dispersion parameter (usually given in ps/nm km) and $\Delta \lambda$ is the FWHM of the optical signal spectrum. $D$ is taken to be constant over the entire optical spectrum considered which is a valid assumption for transmission far from the wavelength of zero dispersion. The factor $A$ is given as

$$A_{\text{Gauss}} = \frac{1}{\sqrt{1 + \gamma^2}}$$

for Gaussian source spectrum

$$A_{\text{Rect}} = \frac{\sqrt{\pi/\ln(2)}}{2\gamma} \text{erf}\left(\gamma\sqrt{\ln(2)}\right)$$

for rectangular source spectrum

where $\text{erf}(x)$ denotes the error function. The corresponding power penalties as calculated from (8) are valid as long as $A > 1/2$, or equivalently as long as $\gamma < \sqrt{3}$ and $\gamma < 2.1$ for Gaussian and rectangular spectra, respectively. After these $\gamma$ values are reached a BER floor will be measured.

In Fig. 11 the calculated penalties are depicted for Gaussian and rectangular source spectra as function of the normalized dispersion $\gamma$. At small $\gamma$ values ($\gamma < 1.5$) the calculated curves are verified by experimental values that were extracted from transmission measurements of $B = 155\times2$ Mb/s NRZ signals in the range around 1550 nm over different SMF lengths using $\Delta \lambda = 71$ nm wide Gaussian and $\Delta \lambda = 8.6$ nm wide rectangular optical spectra. These types of spectra correspond to situations discussed in the section about the optical CDM system designs.

For large $\gamma$ values, approaching the limits of uncompensated transmission $\gamma = \sqrt{3}$ and $\gamma = 2.1$, the experiments showed higher penalties than predicted by the simple model. The measured fiber length limits, where a BER floor set in, were somewhat smaller than calculated (around 8 km and 77 km instead of 9.3 km and 93 km). This may be due to not taking into account electrical noise in the calculations or that the Gaussian pulse shape approximation is not good enough. Nevertheless, the model is useful for a quick estimation of the expected system performance in terms of chromatic dispersion induced degradations.

If the fiber lengths are larger than the above given limits or if the penalties are too large to be tolerated, an electrical postcompensation applying a DFE can be used to improve the performance. Using a two stage DFE a normalized dispersion of $\gamma = 1.59$ ($L = 8.6$ km SMF, $\Delta \lambda = 70$ nm with Gaussian spectrum), i.e., beyond the practically measured limit (cf. Fig. 11), has been successfully compensated with only 4.0 dB residual penalty [23]. Approaches applying linear electrical filters to compensate the low pass characteristic of the fiber-LED system have been found not to be as effective as the DFE approach in that the power penalty easily exceeds 10 dB [27].

In practical system implementations the larger part of the transmission fiber dispersion should be compensated using appropriate lengths of DCF. The residual dispersion with normalized values up to around $\gamma = 1.0$ can then either be tolerated or the induced power penalty may be reduced by applying the DFE approach. To give a number that is relevant for the proposed systems of the preceding section the theoretical limit for uncompensated transmission of 2.5 Gb/s signals with 10 nm rectangular spectra lies at 4.9 km. This is the order of magnitude to which the length of the DCF must be matched to the SMF length.

The dispersion slope of commercially available DCF is well enough matched to the slope of the SMF to enable compensation over broad spectral ranges for long fiber links. In 8 $\times$ 155 Mb/s system experiments with about $\Delta \lambda = 30$ nm wide spectra we compensated the dispersion of up to 111 km field installed SMF [17]. With a dispersion slope of 0.09 ps/nm$^2$/km for the SMF ($D = +17$ ps/nm$^2$/km) and $-0.2$ ps/nm$^2$/km for the DCF ($D = 102$ ps/nm$^2$/km) the maximum differential group delay due to mismatched dispersion slope alone for the wavelength components of $\Delta \lambda = 30$ nm and $\Delta \lambda = 10$ nm wide spectra amounts to 6.4 ps and 0.7 ps per kilometer SMF, respectively. So if the dispersion is perfectly compensated at the center wavelength of the spectra, severe signal degradation due to dispersion slope mismatch would only be measured with signals at 2.5 Gb/s with 30 nm spectra after about 30 km SMF. In the case of narrow rectangular spectra as they have been proposed in the system design section of this paper the residual mismatch in total dispersion is irrelevant even at 2.5 Gb/s over 200 km SMF.

VI. CONCLUSION

We have proposed and analyzed a hybrid optical multiplexing approach for metropolitan and access networks. Based on a mix of wavelength, optical code and time division multiplexing a very flexible and robust implementation of multiuser optical networks can be realized. The basic technologies applied have been demonstrated experimentally. The network was analyzed using a detailed theoretical approach that takes into account crosstalk and intensity noise in wavelength sliced optical CDM
networks. By taking advantage of the fact that optical CDMA supports more codes than can be used simultaneously an optical packet transmission system with more than 400 optical channels was proposed that relies on statistical multiplexing gain. Alternatively, using a novel optical TDM-to-CDM conversion technique the implementation of a 400-channel system with average channel bit rates up to 100 Mb/s was discussed. In the latter approach, the optical packet transmission is realized sequentially within each CDM channel. The applied optical CDMA technology as well as the TDM-to-CDM conversion has been shown to be very insensitive to poor component specifications and drift, thus keeping the cost low for components and network supervision in the physical layer.

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