Novel design method for generalized lattice filters

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3. Equivalent Noise Caused by the Inband Crosstalk

Neglecting the possible delay arising at each optical switch, a packet experiences a delay of $d_k$ (0$s < d_k < 2^n$) when it passes through the PDLM. Each of the $2^n$ order crosstalk follows a different path and hence has a different delay. Let $d_{k-1}$ be the delay of the $2^n$-order crosstalk, where $(0$s < d_{k-1} < 2^n$). It can be shown that, by ignoring the higher order terms, the intensity noise caused by the interference between the main signal and the $2^n$ order inband crosstalk at the output of the PDLM can be expressed as:

$$i(t) = \frac{1}{2}E_p \left[ \frac{2}{\gamma d_{k-1} (t - d_{k-1})} \right] \coth \left( \frac{\gamma d_{k-1} (t - d_{k-1})}{2} \right)$$

(1)

where $E_p$ is the input electrical field, $b_k$ and $b_{k-1}$ are the corresponding "mark" or "space" modulation states of the signal and crosstalk, $\omega_t$ is the carrier frequency and $\phi(t)$ represents the laser phase fluctuation.

The phase fluctuation of a DFB laser is commonly modeled as a Wiener-Levy random process [5-7]. According to the Wiener-Levy model, the DFB laser's phase variance is a nonstationary zero-mean Gaussian random process. However, its phase difference $\Delta \phi$ between two times is stationary and independent zero-mean Gaussian random processes with a variance given by $\sigma_{\Delta \phi}^2 = \frac{2 \Delta \phi}{\omega_t}$ for a laser with its power spectral density function of Lorentzian line shape, where $d_{k-1}$ is the linewidth of the laser source and is the time difference [7].

Applying the Wiener-Levy model, the variance of the phase difference of the beating items in eq. (1) varies from $\sigma_{\Delta \phi}^2 = \frac{2 \Delta \phi}{\omega_t}$ to $\sigma_{\Delta \phi}^2 = \frac{2 \Delta \phi}{\omega_t}$. As an example, for a time slot duration $T = 1.5$ ms as in the KEOPS project [9], and a laser linewidth of $1$ MHz, the variance of the phase difference varies from 0.0368 to 0.0351 for $n = 8$, covering both the coherent and incoherent regimes [5].

We now examine each cosine item in eq. (1). The variance of the phase difference $\Delta \phi = \phi_{d_{k-1}} - \phi_{d_k}$ between the signal and the $2^n$ crosstalk is $\sigma_{\Delta \phi}^2 = \frac{2 \Delta \phi}{\omega_t}$. Let $\gamma \equiv \frac{\Delta \phi}{\omega_t}$, the probability density function of $\gamma$ is given by eq. (2) [6]:

$$f(\gamma) = (2\pi \sigma_{\Delta \phi}^2 (1 + 2\gamma^2))^{-1/2} \exp \left( -\frac{3(\gamma^2 - 1)}{2(1 + 2\gamma^2)} \right)$$

(2)

The mean and variance of each cosine item can be obtained as in eq. (3):

$$m = \int_0^{\infty} \gamma f(\gamma) d\gamma = 1/\sqrt{2}$$

(3)

It is reasonable to assume that each cosine item to be independent because the laser phase noise difference is considered to be a stationary and independent zero-mean Gaussian random process. However, according to the central limited theory, with a reasonable large value of $N$, when the main signal is of "mark" state, the intensity noise in eq. (1) will be a Gaussian distribution with the mean and variance given by:

$$m_{\alpha} = m \sqrt{\frac{1}{2N} \sigma_{\alpha}}$$

(4)

where $m_{\alpha} = \frac{1}{2}E_p \left[ \frac{2}{\gamma d_{k-1} (t - d_{k-1})} \right] \coth \left( \frac{\gamma d_{k-1} (t - d_{k-1})}{2} \right)$

(5)

where $m_{\alpha}$ is the main signal, $m_{\alpha}$ is the mean value, $\sigma_{\alpha}$ is the variance of the noise caused by the receiver noise. The resultant bit error rate (BER) $P_{e,\alpha}$ for the main signal delayed by $d_k$ (0$s < d_k < 2^n$) can be obtained as following the analysis of BER caused by Gaussian noise [12]. Since a packet is likely to experience any of the $2^n$ possible different delays in the PDLM, the average BER can be expressed by eq. (3), assuming an equal probability for each packet delay.

$$P_{e,\alpha} = \frac{1}{2} \left( 1 - \frac{1}{2} E_p \left[ \frac{2}{\gamma d_{k-1} (t - d_{k-1})} \right] \coth \left( \frac{\gamma d_{k-1} (t - d_{k-1})}{2} \right) \right)$$

(6)

We have also estimated the power penalty caused by the 2$n$ order crosstalk. The power penalty is calculated at the BER of $10^{-12}$ when the crosstalk level is less than -25 dB. The power penalty is negligible for crosstalk levels less than -30 dB. However, when the crosstalk approaches to -20 dB, the power penalty increases as $n$ increases. A power penalty of 1dB only allows for $n=6$ with the crosstalk level of -20 dB. When the crosstalk level is reduced to -25 dB, $n$ is allowed to be increased to 10 for a power penalty of 1dB.

4. Numerical Results of BER and Power Penalty

In order to get a better understanding of how the BER is affected by the 2$n$ order inband crosstalk, the BER and power penalty of the signal are numerically calculated, based on the analysis given above. In the following discussions, we have assumed that BER=10$^{-12}$, when there is no crosstalk (i.e. BER caused by the receiver's noise alone). Fig. 2 shows the numerical results of the BER versus different crosstalk level for various values of $n$. Note that $\Delta \phi=1MHz$, and $T=1.5$s.

For a given value of $n$, the BER is significantly increased when the crosstalk level increases from -30 dB to -20 dB, but the BER is almost not affected when the crosstalk level is less than -30 dB. The BER rapidly deteriorates as the value of $n$ increases, because the overall 2$n$ order inband crosstalk increases in proportion to $n$. Fig. 3 shows the relation between the BER and the laser linewidth for various values of $n$ with crosstalk level of -25 dB. This figure clearly shows that, the BER reaches its maximum at a certain laser linewidth. Further increase or decrease in the linewidth will reduce the BER. For a large value of $n$ (e.g. $n=10$), a large linewidth will significantly reduce the BER.

5. Conclusion

We have analyzed the inband crosstalk caused by the synchronization module of photonic packet switches. It has been found that the 2$n$ order inband crosstalk is dominant but its contribution increases with the number of delay lines, which can cause significant impact on the BER performance. The numerical study has shown that the BER is significantly affected as the number of delay lines increases. For a reasonable time resolution (e.g. $n=10$), the crosstalk has to be less than -25 dB so as to ensure the power penalty is below 1 dB.

6. References


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Novel Design Method for Generalized Lattice Filters

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A new design method for wide-band gain equalizers using cascaded MZI's is proposed. The method achieves a greater accuracy with fewer stages than previously described methods. Furthermore, the method is capable of minimizing PDL.

Introduction

Long haul transmission in the backbone network and functionalities in the metro- and access WDM networks require amplification, which is normally provided by erbium doped amplifiers. Unfortunately, the spectral gain of an erbium amplifier is non-flat over the C-band, providing different gain at the individual channels, which decreases the channel SNR and eventually leads to errors in the transmission. Gain Equalizing Filters (GEF), having the inverse amplifier transmission, are attractive components for compensating this gain variation. A possible way of realizing a GEF is to cascade N Mach Zehnder Interferometers (MZI) to obtain a transmission gain $N$ order Fourier expansion. Previously, the filter parameters were extracted by a recursive method [1].
However, the method assumes wavelength independent coupling constants and integer delay ratios, which limits the viability. As a consequence, the method is only usable for narrow-band filters. The necessity of including the coupler wavelength dependence for wide-band multiplexers has also been discussed in [2]. In the present paper a new method is proposed that includes coupler wavelength dependence, non-integer delay ratios and a dual fitting scheme that effectively minimizes Polarization Dependent Loss (PDL). It is shown that the new method significantly reduces the approximation error between the desired and the actual transmission.

Theory

The basic element in the filter is shown in figure 1.

![Fig. 1: First order MZI consisting of two couplers and one delay line. The framed part is the basic lattice element.](image)

It consists of a delay line and a directional coupler. The first directional coupler is treated separately and a given filter is realized by cascading basic lattice elements to the first coupler. The transfer matrix (in field amplitudes) is found by multiplication of the transmission matrices of each individual basic lattice element. The k'th transfer matrix is related to the (k-1) as follows:

\[
T^k = \begin{bmatrix}
    c_k & \alpha_k & -\beta_k \\
    -\beta_k & \alpha_k & c_k \\
    \alpha_k & c_k & -\beta_k
\end{bmatrix} T^{k-1}
\]

where \(c_{k}(\omega_o)\) is the through-coupling constant, \(s_{k}(\omega_o)\) the cross-coupling constant, \(\beta_{k}\) is the ratio of the k'th lattice elements Free Spectral Range (FSR) and the FSR of the entire filter. The variable \(\omega_o = \exp(-j\lambda_0 k_{ip} n_a f_{opt})\) is the harmonic in the normalized frequency \(\omega_o = 2\pi f(1552 \text{THz})\), where \(f\) is the optical frequency, FSR the Free Spectral Range and \(T = \text{FSR}\) is the delay difference between the arms of the interferometer. If \(n_{o} = 0\), then the delay is in the upper arm and \(i = 0\), the delay is in the lower arm. In practice, the phase \(\theta_{k}\) is applied with thermo-optic heaters placed along the longer arm of the interferometer.

A non-linear Levenberg-Marquardt algorithm [4] iteratively searches the amplitude of the basic lattice elements. Free Spectral Range and \(T = \text{FSR}\) is the delay difference between the arms of the interferometer. If \(n_{o} = 0\), then the delay is in the upper arm and \(i = 0\), the delay is in the lower arm. In practice, the phase \(\theta_{k}\) is applied with thermo-optic heaters placed along the longer arm of the interferometer.

The phase shift \(\theta_{k}\) is an exponential, which is then given by \(\theta_{k} = \exp(-j\lambda_0 k_{ip} n_a f_{opt})\). A birefringence of \(\Delta n \approx 0.001\) gives a phase shift of \(\Delta \phi = 0.013\) at \(f_0 = 26.1\) 1545 nm and \(L = 20.1\) μm corresponding to a FSR of 40 nm of the basic lattice elements.

In the figures 5 and 6, the single- and dual fitting scheme are compared for a 6th order lattice filter with birefringent couplers (PDW of 5.1 μm first crosspoint) and \(\Delta \phi = 0.013\).

![Fig. 3: \(\chi^2\) as a function of physical delay number (lattice number) for the linear- and iterative method. The iterative method shows superior performance due to non-integer delays and coupler wavelength dependency. Notice that the ideal linear fit refers to the case of wavelength constant coupling.](image)

The iterative method shows superior performance already at low lattice numbers in comparison to the ideal constant coefficient linear filter. At two delay stages, the iterative method gives error estimate as good as that of a 4th order linear method. For all the considered lattice numbers, the iterative method has an error estimate \(\chi^2 < 0.024\), which corresponds to a maximum transmission error of less than 0.5 dB over the entire C-band. The linear method only achieves this for lattice numbers larger than 4. Including coupler wavelength dependency and non-integer delay ratios is seen to have the effect of enhancing the fit performance. The iterative method shows remarkable improvements, especially for low delay numbers.

![Fig. 4: Principle of the dual fitting scheme. Two different functions representing two orthogonal polarization states fitted to two identical inverse amplifier responses. The fitting routine selects one of the functions in the left half and the other function in the right half.](image)

The desired filter response is artificially duplicated. In one iteration step, the fitting routine is given the response of the first polarization state, corresponding to the left half in figure and the other polarization eigenstate, corresponding to the right half in figure. The result is a simultaneous least square minimization of both responses, thereby squeezing the responses of arbitrary polarization states.

A scalar 3D-BPM modal analysis with appropriately adjusted grid parameters have been applied to obtain the difference in the supermode propagation constants of the coupler for the two polarizations. The waveguide parameters were adjusted to obtain best agreement with polarization requirements.

Birefringence in the delay section introduces a polarization dependent phase shift of magnitude \(\Delta \phi_k = \Delta \phi_{o} + \Delta \phi_{\text{PD}}\), where \(k_{ip}\) is the center wavelength, \(L\) is the relative delay length and \(\Delta \phi_{\text{PD}}\) is the difference in effective indices of the two polarizations. The phase shift is added to the argument of the exponential, which is then given by \(\exp(-j\lambda_0 k_{ip} n_a f_{opt})\). A birefringence of \(\Delta n \approx 0.001\) gives a phase shift of \(\Delta \phi = 0.013\) at \(f_0 = 26.1\) 1545 nm and \(L = 20.1\) μm corresponding to a FSR of 40 nm of the basic lattice elements.

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Multi-Channel Chromatic Dispersion Compensator Consisting of Modified Interleave Filter

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We have successfully demonstrated a low loss dispersion slope compensator consisting of a PLC-type modified interleave filter. It has different chromatic dispersion to meet the demand of each 50 GHz-spacing WDM channel in the C-band.

Introduction

The dispersion compensation of transmission fiber is indispensable in high-bit-rate (e.g. 40Gb/s) wavelength division multiplexing (WDM) transmission systems. Of the various possible approaches, distributed fiber management is one of the most likely candidates, especially for long-haul systems. With this method, we can easily design the chromatic dispersion so that it is zero at a given wavelength. However, it is still difficult to compensate for the dispersion differences between the WDM channels over a wide wavelength range. Numerous devices have been developed to remove these residual dispersions, which are mainly due to the dispersion slope of optical fiber. These devices include arrayed waveguide gratings, and multi-channel compensators with a channel selector [4,5]. Recently another method has been reported that employs a pair of filters with opposite dispersion slopes indicating the possibility of compensating for dispersions in multi-channels simultaneously as described below. The fundamental unit (filter Fig. 1(a)) is a low insertion loss, high selectivity, and linear response. This type of filter is known to be one of the most flexible devices for compensating for dispersion slopes.

Fig. 1 (a) Schematic configuration of a fundamental filter unit. (b) Simulated transmittances and delay time responses for each set of input and output ports in the filter unit.

Operating principle and basic design

Here we describe the basic principle we used for designing the proposed dispersion slope compensator. Figure 1(a) shows the schematic configuration of the fundamental unit that we used for constructing the compensator and Fig. 1(b) shows its simulated transmittance and relative group-delay responses. The compensator consists of two of the above fundamental units with different center wavelengths, as described below. The fundamental unit is composed of a lattice-form filter, and is almost the same as that previously reported as an interleave filter [7]. The unit is designed to have a rectangular transmittance and a parabolic group-delay response simultaneously. This type of filter generally has a special feature in that it can either compensate for the dispersion effects of the channel from 1530 to 1565 µm, which corresponds to the compensation of the dispersion slope of the filters.

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