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Optimization of TE–TM mode converters on *X*-cut, *Y*-propagation $LiNbO_3$ used for PMD compensation

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ABSTRACT Two-phase and three-phase TE–TM mode converters for integrated optic polarization mode dispersion compensation are compared, and the latter are found to have a slightly better electro-optic efficiency. If a small differential group delay is needed, compensation performance can be drastically improved by a waveguide tilt in the *YZ* plane.

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1 Introduction

Polarization mode dispersion (PMD) is caused by noncircular fiber cores and limits the capacity of optical trunk lines. It can be compensated if appropriately oriented birefringence is added at the receiver side in reversed order. Particularly suited is *X*-cut, *Y*-propagation $LiNbO_3$. Its natural birefringence (0.26 ps/mm) is oriented by electro-optical mode converters and cancels the differential group delay (DGD). This solution combines optimum performance and speed with a high degree of integration and hence low-cost potential. Since the function has already been demonstrated [1], we now concentrate on design optimization.

2 Operational principle

The principle is based on the spatially weighted coupling between two waves with different propagation constants [2]. The phase difference between one mode and the coupled mode therefore depends on the position at which coupling occurs and is periodic with the beat length $\Lambda = \lambda/\Delta n$. The TE–TM refractive index difference in a Ti-indiffused waveguide in *X*-cut, *Y*-propagation $LiNbO_3$ is $\Delta n = 0.0679$ at a free space wavelength $\lambda = 1550$ nm, thereby giving $\Lambda = 22\ \mu$m. Interdigital electrodes are needed for phase matching. The widths of fingers *W* and gaps *G* are equal to $\Lambda/4$. The coupling factor κ is given by

$$\kappa \cong \hat{\Gamma}(\pi/2)n^3(V/G)\lambda^{-1}\,, \tag{1}$$

where $\hat{\Gamma}$ is a weighted field overlap integral factor as defined later and $n = 2.1785$ is the average refraction index of the waveguide. $r_{51} = 28 \times 10^{-12}$ m/V is the relevant electro-optic coefficient and V is the interelectrode voltage.

Figure 1 shows a portion of a chip used in [1]. Voltage V_1 acts on one set of comb electrodes and performs mode conversion in phase. Voltage V_2 acts on another set of comb electrodes which are translated by $\Lambda/4$ with respect to the first and performs mode conversion in quadrature. The resulting complex coupling factor is proportional to $V_1 + \imath V_2$. The need for 2 quadratures at least doubles the necessary chip length. The longitudinal electrode cross-section of one set of electrodes used in [1] are shown in Fig. 2.

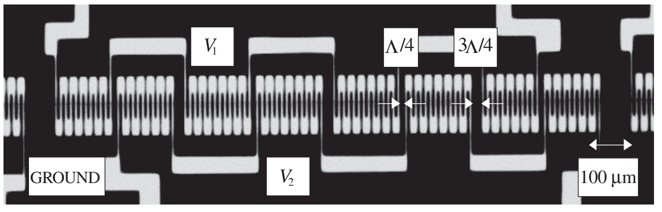

FIGURE 1 Photograph of a mode converter electrode pair (*dark*) on a Ti:$LiNbO_3$ PMD compensator

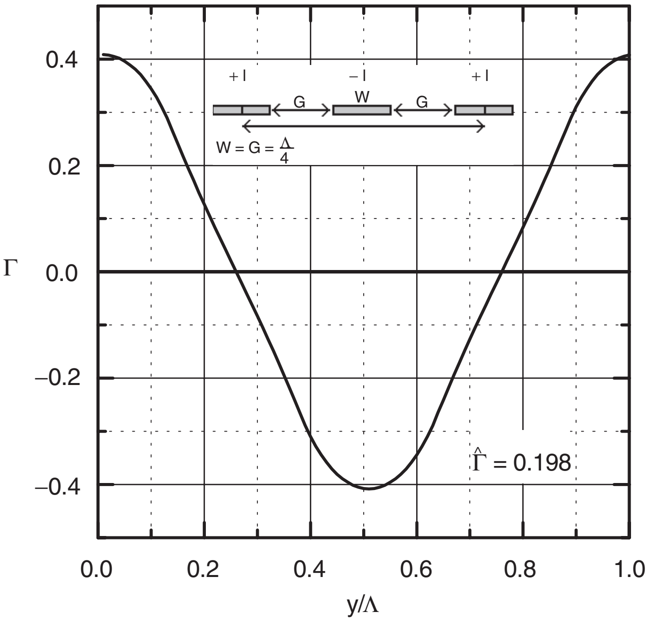

FIGURE 2 Two-phase electrodes with corresponding voltages and local field overlap integral factors vs normalized longitudinal coordinates

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3 Two versus three phases

As has been mentioned in [2], this 2-phase implementation is not the only possible choice. If isolated electrode crossings are available, 3-phase electrodes can be used with electrode widths and gaps equal to $\Lambda/3$. In-phase and quadrature mode conversion can be produced by the linear combinations of the "cosine" and "sine" cases shown in Fig. 3 and Fig. 4. Even and odd voltage distributions are applied by choosing electrode voltages $V_0 f(\hat{y})$, where V_0 serves as a reference voltage and $\hat{y}$ is the longitudinal position of the center of an electrode. Here $f(y) = \cos(2\pi y/\Lambda)$ and $f(y) = \sin(2\pi y/\Lambda)$ are the structure functions needed for the cosine and sine cases, respectively.

The point matching method [3] has been chosen to calculate the electrostatic fields of these periodic electrode structures. The transversal optical field $\boldsymbol{E}_{\mathrm{o}}(x, z)$ is assumed to be Gaussian and Hermite–Gaussian in width and depth of the single-mode Ti-indiffused waveguide in $LiNbO_3$, with mode field diameters matched to our experimental values. The position-dependent overlap integral $\Gamma(y)$ [4] must be multiplied by $f(y)$, integrated over one beat length and normalized to obtain the weighted overlap integral factor

$$\hat{\Gamma} = \frac{2}{\Lambda}\int_0^{\Lambda} \Gamma(y) f(y)\mathrm{d}y \quad \text{with} \tag{2}$$

$$\Gamma(y) = \frac{G}{V}\frac{\int\int |\boldsymbol{E}_{\mathrm{o}}(x,z)|^2 \boldsymbol{E}_{\mathrm{x}}(x,y,z)\mathrm{d}x\mathrm{d}z}{\int\int |\boldsymbol{E}_{\mathrm{o}}(x,z)|^2\mathrm{d}x\mathrm{d}z}.$$

For the cosine and sine cases, $\hat{\Gamma}$ is the real and imaginary part of the spatial Fourier coefficient of $\Gamma(y)$, respectively. $\boldsymbol{E}_x(x, y, z)$ is the vertical component of the electrostatic field in the crystal. In $\Gamma(y)$ we have multiplied by the applicable gap G and divided by the maximum interelectrode voltage V as defined in (2). Figure 2 also shows the local overlap factors $\Gamma(y)$ for one quadrature of the 2-phase TE–TM mode converter. $\Gamma(y)$ is shown in Figs. 3 and 4 for the two cases of the 3-phase TE–TM mode converters. The resulting weighted overlap factors are $\hat{\Gamma} = 0.198$, 0.11 and 0.096, respectively. The 2-phase $\hat{\Gamma}$ is resized to an effective value of $\approx 0.086 - 0.098$, if one takes into account the fact that the two quadratures of the 2-phase design need at least twice the length of the 3-phase design. If the maximum permissible field strength limits the design, the factor V/G in κ as defined in (1) is replaced by a constant. The 3-phase design performs in the worst case (0.096) roughly equal to or slightly better than the 2-phase design. If the output range of the voltage sources is the limiting factor, κ as defined in (1) is obtained through a multiplication by the same V/G ($8/\Lambda$, $9/\Lambda$, $6\sqrt{3}/\Lambda$) as that by which we have divided in calculating $\Gamma(y)$ as defined in (2). This yields equal κ values for both 3-phase cases, and these are $1.26 - 1.44$ as high as κ in the 2-phase case.

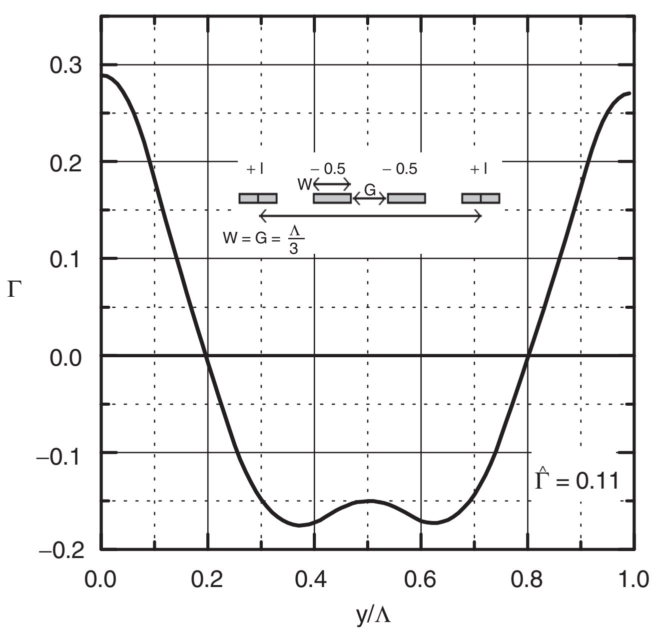

FIGURE 3 Three-phase cosine electrodes with corresponding voltages and local field overlap integral factors vs normalized longitudinal coordinates

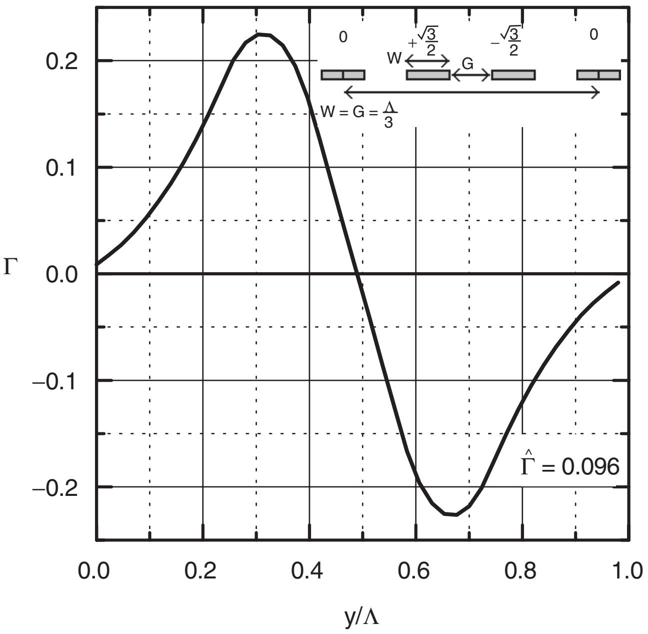

FIGURE 4 Three-phase sine electrodes with corresponding voltages and local field overlap integral factors vs normalized longitudinal coordinates

4 Tilted waveguide

We have also investigated the case of a larger Λ. For large Λ the electrical field reaches deeper into the waveguide and is also more uniform, thereby increasing the overlap integral as shown in Fig. 5. Similar characteristics are found for all three cases. $\hat{\Gamma}$ grows almost $\propto \Lambda$ in the range considered.

A high $\Lambda \propto \cos^{-2}\vartheta$ can be achieved in $LiNbO_3$ if the waveguide is tilted by an angle ϑ in the YZ plane. The coupling factor then becomes $\kappa(\vartheta) = \hat{\Gamma}(\pi/2)n^3(r_{51}\cos\vartheta - r_{22}\sin\vartheta)(V/G)\lambda^{-1}$ with $n = n_{\mathrm{o}} + \frac{1}{2}(n_{\mathrm{e}} - n_{\mathrm{o}})\cos^2\vartheta$ and $r_{22} = 6.8 \times 10^{-12}\,\mathrm{m/V}$. To give an example, for $\vartheta = -\pi/4$ the waveguide runs halfway between the $+Y$ and $+Z$ axes. Λ doubles, $\hat{\Gamma}$ more than doubles, and $\kappa(-\pi/4) \cong 1.75\kappa(0)$ in the field strength limited case. Furthermore, since the DGD per unit length is halved, the DGD spent to implement a full mode conversion is ≈ 3.5-times smaller than for Y-axis propagation. This means a 3.5-times more accurate PMD compensation becomes possible. However, twice a voltage V is required to keep V/G constant. Furthermore, a PMD compensator with a given length can compensate only half as much DGD. A large Λ is particularly advantageous if neither available driving voltages nor total chip length are limiting factors. This is the case at high bit-rates, say ≥ 40 Gbit/s, where truly bad fibers have to be ruled out anyway.

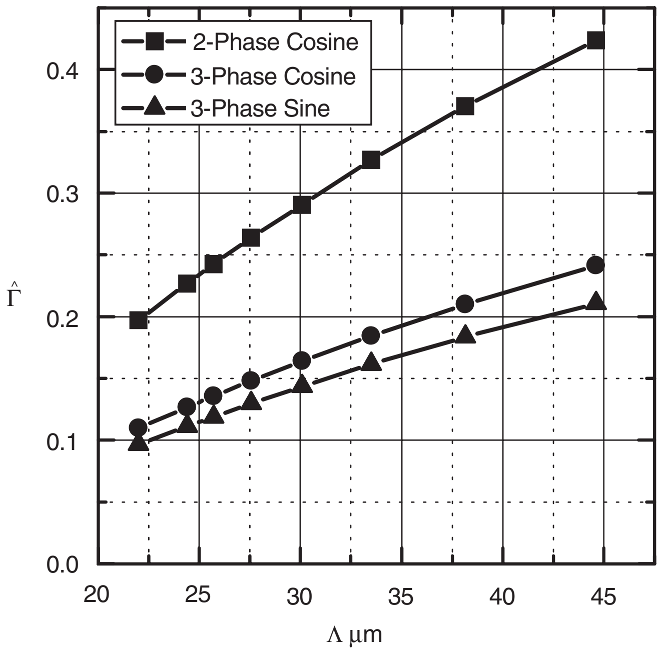

FIGURE 5 Weighted overlap integral factor $\hat{\Gamma}$ as a function of beat length Λ in µm

As a note of caution, it is necessary to consider the substrate radiation modes which may cause high propagation losses in tilted waveguides. They may become significant when the waveguide is tilted in practice, because the propagation constants of the guided mode and the substrate radiation mode may become identical; however, no attempt has been made to verify this possibility.

5 Conclusion

We have found that a 3-phase TE–TM mode converter can (but need not in all cases) outperform a 2-phase one. Tilting the waveguide in the YZ plane can drastically increase the efficiency. This is particularly advisable if just a little DGD needs to be compensated, which may be the case at data rates of ≥ 40 Gbit/s.

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