

DESIGNS OF COMPACT OPTICAL DEVICES BASED ON PERIODIC META-STRUCTURES

A THESIS

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By

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I certify that I have read this thesis and that in my opinion it is fully adequate, in scope and in quality, as a thesis for the degree of Master of Science.

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ABSTRACT

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Manipulation of the flow of light is demanded for several applications such as communication, data storage, sensor, photovoltaic cells, microscopy, lasers and light emitting diodes for the purpose of designing compact, high-throughput and high efficiency optical devices. Nevertheless, the control of the propagation of the light becomes much harder in devices with smaller geometries mostly because of diffractions, loss mechanisms and fabrication difficulties. Furthermore, materials that are already available in the nature do not provide unprecedented optical properties for nanoscale optical applications. Due to this fact that fabrication of artificial materials is needed for utilizing novel and intriguing optical devices. For this purpose, some relatively new research fields have emerged like photonic crystals, metamaterials and high contrast gratings. We propose several designs based on aforementioned meta-structures to achieve compact and practically realizable optical devices. We presented compact optical demultiplexer, diode-like device and electro-optic modulator designs that are based on photonic crystals. We also proposed two circular polarizer designs based on metamaterials and high contrast gratings. Further, we investigated unidirectional transmission and polarization manipulation properties in chiral metamaterials. For most of the proposed designs, we also experimentally verified the numerical and theoretical findings. In conclusion, we can claim that the utilization of artificially structured materials give opportunity to realize the control of light much more easily in nanoscale designs.

Keywords: Photonic Crystals, Metamaterials, High Contrast Gratings, Slow Light, Optical Demultiplexing, Asymmetric Transmission, Circular Polarization.

ÖZET

PERİYODİK META-YAPILAR TABANLI OPTİK AYGIT TASARIMLARI

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İletişim, veri depolama, sensör, fotovoltaik hücreler, mikroskopi, laser ve led gibi uygulamalarda kullanılmak üzere kompakt, üretimi kolay ve yüksek verimli optik aygıtların tasarlanabilmesi için ışığın hareketinin manipülasyonu önem arz etmektedir. Ancak, ışığın daha küçük boyutlarda yapı elemanlarına sahip malzemelerde ilerlemesinin kontrolü saçılımlar, kayıplar ve fabrikasyon hataları nedeniyle oldukça zor bir hal almaktadır. Ayrıca, nano boyutlarda, doğada hazır bulunan malzemeler de bizlere beklenmedik ve üstün optik özellikler sunmamaktadır. Bu nedenle, ilginç ve üstün optik özelliklere ulaşabilmek için yapay olarak üretilen malzemelere ihtiyaç duyulmaktadır. Bu amaçla, fotonik kristaller, metamalzemeler ve yüksek kontrastlı ızgaralar gibi daha yeni sayılabilecek araştırma alanları doğmuştur. Biz de yukarıda bahsedilen yapay malzemeleri kullanarak kompakt ve rahat üretilebilir optik aygıt tasarımları sunmaktayız. Fotonik kristaller kullanarak optik çoklayıcı, diyot ve elektro-optik modülatör tasarımları oluşturduk. Metamalzemeler ve yüksek kontrastlı ızgaralar tabanlı dairesel polarizasyon üretimi yapan optik aygıt tasarımı çalışmalarında bulunduk. Son olarak ise, kiral metamalzemelerde tek yönlü iletim ve polarizasyon manipülasyonu fenomenlerini inceledik. Çoğu çalışmamız için, elde ettiğimiz teorik ve nümerik bulguları deneysel sonuçlarla doğrulamayı başardık. Sonuç olarak ise, yapay olarak üretilmiş yapıların nano boyut tasarımlarda ışığın kontrol edilmesi için bizlere önemli fırsatlar doğurabileceği bilgisine ulaştık.

Anahtar sözcükler: Fotonik Kristaller, Metamalzemeler, Yüksek Kontrastlı Iızgaralar, Yavaş Işık, Optik Çoklama, Asimetrik İletim, Dairesel Polarizasyon.

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Contents

1	Introduction	2
2	Compact Optical Devices Based on Photonic Crystals	6
2.1	Introduction	6
2.2	Slow Light in Photonic Crystals & Tight Binding Mechanism . . .	10
2.3	Optical De-multiplexer Design by Utilizing Slow Light in Photonic Crystals	16
2.3.1	Optical De-multiplexing Methodology	17
2.3.2	Single Wavelength Extraction	18
2.3.3	Frequency Division Demultiplexing	29
2.4	Optical Diode-like Devices based on Graded Photonic Crystals . .	38
2.4.1	Asymmetric Light Propagation Methodology	39
2.4.2	Numerical Results of the Proposed Device & Experimental Verification	41
2.5	Compact Photonic Crystal based Fabry-Perot Cavity Design for Electrooptic Modulation	51

2.5.1	Proposed Cavity and Sensitivity to Changes in Dielectric Permittivity	53
3	Circular Polarization Generation and Asymmetric Transmission Utilizing Chiral Metamaterials	64
3.1	Introduction	64
3.2	Circular Polarizer based on Chiral Metamaterials	67
3.3	Polarization Angle Dependent Rotation	73
3.3.1	Proposed Geometry	74
3.3.2	Numerical Results	75
3.3.3	Experimental Results	78
3.3.4	Formulation	80
3.3.5	Surface Currents	83
3.4	Diodelike Asymmetric Transmission	84
3.4.1	General Idea	85
3.4.2	Method	87
3.4.3	Results and Discussion	88
3.4.4	Eigenstate Interpretation	92
4	Circular Polarizer Design based on High-Contrast Gratings	94
4.1	Introduction	94
4.2	Theoretical and Numerical Investigations of the Proposed Structure	95

4.3	Experimental Verification	101
4.3.1	Proposed Geometry	101
4.3.2	Numerical Results	103
4.3.3	Fabrication	106
4.3.4	Experimental Setup	106
4.3.5	Experimental Results	108
5	Conclusions	110
	Bibliography	115
A	Publication List	132

List of Figures

2.1	(a) The dispersion relation under transverse magnetic (TM) field excitation of the two dimensional PC waveguide structure; the black curve is the waveguide mode. (b) The super-cell of the PC coupled cavity configuration. The dashed lines enclose a coupled cavity waveguide, which has smaller coupled cavities (c) The dispersion relation of the PC coupled cavity structure, the studied waveguide bands are given bold, which are observed to be approximately flat bands. (d) The complete PC coupled cavity configuration.	13
2.2	(a) Single PC cavity formation. (b) Double PC cavity formation to observe eigen-mode splitting. (c) The transmission spectra for the single and double cavity structures. (d) The comparison of dispersion relations of a selected waveguide band of the PC coupled cavity structure, which is constructed by the demonstrated cavities. (e) The comparison of the numerical and theoretical group index values.	14
2.3	(a) The steady-state electric field distribution when the wavelength of incident light corresponds to a propagating mode within the PC waveguide. (b) The field distribution for a different frequency, which is in the vicinity of the slow light regime. The same PCW is used in both cases.	19

2.4	The schematic of the device that is designed for the wavelength demultiplexing. The different wavelengths are spatially separated at different locations along the x-direction. The complete structure is composed of three PCWs of different dielectric filling factors.	20
2.5	The relation between the waveguide band cut-off frequency and the dielectric filling factor. The highlighted frequencies correspond to the selected PCW sections.	21
2.6	(a) The evolution of the waveguide bands due to the different dielectric filling factors, (b) the enlarged view of the vicinity of the slow light region, where the dashed lines correspond to the operating frequencies.	22
2.7	The output frequency spectra of the channels of each PCW section that has a r/a ratio of (a) 0.22, (b) 0.2, (c) 0.18, respectively.	24
2.8	The steady-state electric field profiles of the three de-multiplexed frequencies for each waveguide channel where the input beam is placed $5a$ away from the PC structure, (a) $a/\lambda = 0.3086$, (b) $a/\lambda = 0.3158$, (c) $a/\lambda = 0.3247$, (d) the steady-state electric field profile of an unselected frequency, which has a normalized frequency of $a/\lambda = 0.34$	28
2.9	The spatial distributions of the operating frequencies at three output channels in x-direction. The output levels are calculated nearly $5a$ away from the output channels. The green dashed lines show the positions of the output channels in x-direction.	29
2.10	The geometry of the proposed DEMUX structure. The design consists of two cascaded PCWs with unequal waveguide widths w_1 and w_2 , respectively. The radii of the dielectric rods and the lattice periodicities are identical.	32

2.11	(a) The dispersion diagram of a two-dimensional classical PC with a r/a ratio of 0.24 and a dielectric constant of 12. (b) The evolution of the waveguide band with respect to the waveguide width. The blue and red lines correspond to the widths for the DEMUX design. The PCW widths are altered linearly from $1.6a$ to $2.4a$ with an increment of $0.5a$	33
2.12	The transmission spectra of the output channels. The inset shows the enlarged view of the de-multiplexed frequency intervals.	34
2.13	The spatial distributions of the pulses with normalized frequencies of (a) $a/\lambda = 0.2905$ and (b) $a/\lambda = 0.3105$ that correspond to the centers of the de-multiplexed frequency bands at PC1 and PC2 sections, respectively.	36
2.14	The spatial distributions of the center frequencies of the de-multiplexed frequency bands at the output channels along the x -direction. The vertical green dashed lines correspond to the positions of the output channels.	37
2.15	Schematic representation of a chirped photonic crystal waveguide that enables lights asymmetric propagation is provided. (a) and (b) correspond to left-to-right and right-to-left propagation, respectively. Yellow boxes in (a) show unit cell variations along the propagation direction.	41
2.16	The detailed presentations of the intensity distributions in linearly chirped waveguide configuration for the forward and backward light propagations are shown in (a) and (b), respectively. . .	42
2.17	(a) Schematic of a regular photonic crystal waveguide. (b) The dispersion diagram of regular photonic crystal waveguide. (c) Incident light propagation from left-to-right in the waveguide structure. . .	43

2.18	(a) A photonic crystal waveguide that has linearly increased the distance separation between each neighboring column of rods. (b) The spectral content of the contrast ratio. The blue rectangle designates the selected frequency value. (c) Time-domain snapshot of propagation of light from left to right. (d) The transverse field profile across the end face of the structure. (e) The transverse field profile across the front side of the structure. (f) Time-domain snapshot of light propagation from right to left.	45
2.19	(a) Another chirped photonic crystal waveguide is shown. The distance between each neighbouring column of rods increased in a parabolic pattern. (b) The spectral content of the contrast ratio. The blue rectangle designates the selected frequency value. (c) Time-domain snapshot of propagation of light from left to right. (d) The transverse field profile across the end face of the structure. (e) The transverse field profile across the front side of the structure. (f) Time-domain snapshot of propagation of light from right to left.	47
2.20	The simulated intensity distributions are provided at the outside of the structure for microwave region. (a) The transverse field profile across the end face of the linearly chirped waveguide. (b) The transverse field profile across the front face of the same structure. (c) The transverse field profile across the end face of the parabolically chirped waveguide. (d) The transverse field profile across the front face of the same structure.	48
2.21	The measured intensity distributions are provided at the outside of the structure. (a) The transverse field profile across the end face of the linearly chirped waveguide. (b) The transverse field profile across the front face of the same structure. (c) The transverse field profile across the end face of the parabolically chirped waveguide. (d) The transverse field profile across the front face of the same structure.	50

2.22	The cross-sectional view of the proposed Fabry-Perot cavity geometry. The PC structure is assumed to be periodic and infinitely long along the $\pm y$ and $\pm x$ directions, respectively. The incident wave is assumed to be a plane wave propagating in the $+z$ direction. The transmission spectrum is obtained with the aid of the transmission monitor.	54
2.23	(a) The dispersion diagram and (b) power transmission spectrum of the bulk PC structure. The red and the black curves represent the upper and the lower band of the PBG region, respectively. . .	56
2.24	(a) Dispersion diagram and (b) power transmission spectrum of the proposed Fabry-Perot type cavity. (c) The enlarged view of the cavity resonance that is enclosed by the dashed rectangle. . .	58
2.25	Power transmission spectra of the Fabry-Perot cavity for the cases of (a) $N=5$, (b) $N=6$, and (c) $N=7$. The insets show the enlarged view of the regions enclosed by the dashed rectangles.	59
2.26	Electrical field distributions of the Fabry-Perot cavities for the cases of (a) $N=5$, (b) $N=6$ and (c) $N=7$. The white dashed circles denote the positions of the holes.	60
2.27	Evolution of the spectral position of the cavity resonance with respect to the change in the refractive index of the cavity region, Δn	61
2.28	Power transmission spectrum of the proposed Fabry-Perot type cavity for the dual-polarization operation regime.	62
2.29	(a) Dispersion diagram and (b) power transmission spectrum of the dual-mode operation regime of the proposed Fabry-Perot type cavity.	63

3.1	Geometry of the simulated and fabricated chiral metamaterial unit cell.	69
3.2	(a) Ratio of magnitudes of T_{yx} and T_{xx} , and (b) phase difference between T_{yx} and T_{xx}	70
3.3	(a) Numerical and (b) experimental results for the circular transformation coefficients.	71
3.4	Directions of the surface currents on the SRRs due to the incident electric field polarized in the x -direction, at 5.1 GHz and 6.4 GHz	72
3.5	Geometry of the simulated and fabricated chiral metamaterial unit cell.	75
3.6	Magnitudes of the linear transmission coefficients when the CMM is illuminated by (a) x -polarized and (b) y -polarized incident waves. (c) The mutual phase differences between the transmission coefficients.	76
3.7	(a) Ellipticities and (b) polarization azimuth rotation angles of the transmitted waves for x -polarized and y -polarized illumination. . .	78
3.8	(a) Experimental magnitudes of T_{xx} and T_{yx} , and (b) T_{xy} and T_{yy} . (c) Mutual phase differences obtained from the experiments. . . .	79
3.9	(a) Experimental ellipticities and (b) polarization rotations of the transmitted waves due to x - and y -polarized incident waves. . . .	79
3.10	(a) Polarization angles of the transmitted linearly polarized waves with respect to the polarization angle of the incident wave, at 6.2 GHz, for the $-z$ and $+z$ propagating waves. (b) Introduced polarization rotation to the $-z$ and $+z$ propagating waves with respect to the incident polarization angle, at 6.2 GHz.	82
3.11	Directions of the induced surface currents due to x -polarized plane waves propagating in the $-z$ and $+z$ directions at 6.2 GHz.	83

3.12	Geometries of (a) layer A and (b) layer C. (c) Arrangement of A and C with respect to each other along with the mesh geometry and the stacking scheme.	87
3.13	$ T_{xx} $ for (a) separate and (c) double layers, and $ T_{yx} $ for the same (b) separate and (d) double layers.	89
3.14	Numerical and experimental transmission spectra for the ABS stack, for x -polarized (a), (b) forward and (c), (d) backward propagating waves; (e), (f) numerical and experimental asymmetry factor.	90
4.1	Geometrical description of the HCG circular polarizer. The blue regions indicate the presence of Si, whereas the white regions are SiO_2	97
4.2	The normalized magnetic field distributions at $\lambda_0 = 1.55 \mu\text{m}$ inside region II, (a) $ H_y / H_0 $ and (b) $ H_x / H_0 $ for TM and TE waves, respectively. The distributions are obtained under the two-mode approximation. Grooves and ridges are separated by the white dashed lines and denoted by G and R, respectively.	98
4.3	(a) Field transmission coefficients for TM and TE waves and (b) the phase difference between the transmission coefficients.	99
4.4	(a) Circular conversion coefficients obtained using the RCWA and (b) the conversion efficiencies obtained from the RCWA and FDTD. The wavelength range for the FDTD result satisfying the 0.9 efficiency threshold is denoted by $\Delta\lambda$	100

4.5	Illustration of the proposed quarter-wave plate geometry. The dashed square box on the left denotes one period. The geometrical parameters are given by $r = 220$ nm, $g = 350$ nm, $h_g = 320$ nm, and $\Lambda = 570$ nm. In the theoretical and numerical consideration, for the sake of simplicity, regions I and III are assumed to be infinite in the $-z$ and $+z$ directions, respectively. The materials constituting region III and the ridges in region II are sapphire and silicon, respectively, whereas region I is free-space. n_0 , n_{si} , and n_s represent the refractive indices of free-space, silicon, and sapphire, respectively. <i>Grating direction</i> is defined such that it corresponds to the y direction.	102
4.6	Numerical (a) TM and TE transmitted intensities, (b) circular conversion coefficients, and (c) conversion efficiency spectrum obtained via FDTD simulations. The wavelength interval of operation is denoted by $\Delta\lambda$. The numerical conversion efficiency spectrum yields a percent bandwidth of 42%.	104
4.7	(a) Zoomed out and (b) zoomed in top view SEM micrographs of the fabricated HCG structure. In (b), the legends V1, V2, and V3 denote the geometrical parameters g , r , and Λ , respectively. . . .	105
4.8	Visual illustrations of the experimental setups that are utilized for the measurement of (a) linear transmission coefficients and (b) circular conversion coefficients. The arrows, which lie inside the HCG samples and point upwards, denote the grating direction that is defined in the caption of Fig. 4.5.	107
4.9	Experimentally obtained (a) TM and TE transmitted intensities, (b) circular conversion coefficients, and (c) conversion efficiency spectrum. The wavelength interval of operation is denoted by $\Delta\lambda$. The experimental conversion efficiency spectrum yields a percent bandwidth of 33%.	108

Chapter 1

Introduction

Photonics is the research area that can beat the technical performances of the conventional electronic devices due to the ability of high speed data transmission. Nevertheless, the manipulation of the properties and flow of light is less elucidated than the electrons for nanoscale device applications. Besides being a rather new research field, the conventional and already available materials in the nature also does not exhibit intriguing optical properties for the purpose of realizing optical devices and circuit elements. Some applications that the optical devices are needed for such as communication, data storage, sensor, photovoltaic cells, microscopy, lasers and light emitting diodes also require compact, high-throughput and high-efficiency designs. To meet the needs of this area, several rather new research areas have emerged like photonic crystals, metamaterials and high contrast gratings. The basic advantage of these periodic meta-structures is the opportunity of providing control of light at high frequencies in smaller scale.

Photonic crystals are one, two or three dimensionally and periodically distributed composite structures that are generally formed of dielectric or metallic constituents. The photonic band gap effect that inhibits the propagation of light at a certain frequency interval makes photonic crystals more intriguing [1–3]. In other words, the photonic band gap effect can be viewed as the analogous of the electronic band gap in semiconductor crystal. Moreover, several optical devices can be realized by means of the PBG effect such as waveguides with high

confinement and resonators with high quality factors.

In Chapter 2, we investigated and proposed several optical devices based on photonic crystals. First of all, the theoretical approach to the exhibition of slow light in photonic crystal waveguides are performed and the tight-binding mechanism is applied to the coupled photonic crystal waveguides for the incoming wave that has low group velocities. After the characteristics of the photonic crystal waveguides and the slow light regime, a novel optical demultiplexing methodology and design by utilizing the slow light regime in cascaded photonic crystal waveguides are proposed. The extraction of the targeted frequencies due to the spatial widening of the slow light modes in the waveguide, is both numerically and experimentally shown. Thereafter, a frequency division demultiplexer which extracts an interval of targeted frequencies instead of single frequency demultiplexing is investigated. The numerical results of the proposed device also verified with the experimental efforts. Furthermore, the technical performance of the demultiplexer designs are compared with the conventional ones.

The second proposed optical device based on photonic crystals is a diode-like device that works as an analogue of an electronic diode for photons. Instead of regular photonic crystals, a graded photonic crystal waveguide is utilized for the purpose of achieving a spatially dynamic photonic band gap region and waveguide band. The guiding mechanism is unidirectionally provided by this utilization of graded waveguide. The experimental observations of the asymmetric deflection of the incoming wave for backward and forward illuminations are compared with the numerical results.

The last photonic crystal based proposal is a highly wavelength selective cavity design for electro-optic modulators. The photonic crystal cavity is formed by two bulk photonic crystal sections with a cavity region which is perpendicularly placed to the direction of the propagation. In other words, a classical Fabry-Perot type cavity is constructed by two side mirrors which are photonic crystal sections. Because of the utilization of Lithium Niobate for the substrate and the highly confined resonance modes in the substrate at the cavity region, we claimed that the effective dielectric permittivity of the substrate around the cavity can be

increased more than a bulk substrate under DC bias. The theoretical approach of the effective change in the permittivity is also investigated.

The third chapter is based on devices that are formed by chiral metamaterials. Metamaterials are periodically distributed metallic elements that are in sub-wavelength scale which also makes them diffraction-free. Moreover, with the optimization of the geometrical parameters of the metallic constituents, negative refraction can be realized. Besides the classical metamaterials, Pendry claimed that negative refraction can be possible by using a chiral route in metamaterial based applications [4]. The most significant electromagnetic property of chiral metamaterials can be explained as the existence of the cross coupling between the electric and magnetic fields of the wave. This cross coupling also leads to different transmission coefficients under right hand circularly polarized and left hand circularly polarized wave illumination [5].

Firstly, we study the potential of an electrically thin, asymmetric chiral circular polarizer structure, which is composed of isotropic materials, in obtaining LCP and RCP waves using a linearly polarized incident wave. The second proposed design is a chiral metamaterial based polarization rotator which also has the similar operation mechanism with the circular polarizer which can also be explained with the difference in the transmission coefficients and the cross coupling between the electrical and magnetic fields. Similarly, our last study on chiral metamaterials is a unidirectional transmission device which is the optimized mechanism of the aforementioned asymmetric transmission effect. The optimization of the contrast ratio between the forward and backward waves is provided with the utilization of an additional sub wavelength metallic mesh between the metamaterial layers which are forming the chiral metamaterial structure. All of these studies and effects based on chiral metamaterials are also experimentally observed and verified with the numerical results.

Chapter 4 is dedicated to the studies based on high contrast gratings. High contrast gratings are periodic structures with subwavelength periodicities. Due to exhibiting broadband and diffraction-free high reflectivity regimes, high-contrast gratings have attracted numerous researchers [6, 7]. One should also note that

high refractive index difference between the consecutive constituents are required for the realization of broadband reflectivity regimes. Lastly, we propose a novel broadband circular polarizer design based on binary high contrast gratings. The numerical and theoretical analysis of the proposed device is investigated for a stand-alone structure. Additionally, a more realistic design on a substrate is proposed and the experimental verification of the proposed circular polarization generation operation is performed around the wavelength of $\lambda = 1.55\mu m$.

The last chapter summarizes the achievements in the thesis.

Chapter 2

Compact Optical Devices Based on Photonic Crystals

2.1 Introduction

Photonic crystals (PCs) are one, two or three dimensionally and periodically distributed artificial composite structures that are usually consist of dielectric or metallic materials [8, 9]. The most intriguing property for the PCs can be seen as the photonic band gap (PBG) effect [10]. PBG effect can be explained as the inhibition of the propagation of light inside the PC structures at a specific frequency interval [1]. This inhibition can be provided for any direction in the PC structures that is called as full PBG region. This PBG idea has spawned the field of PCs, which allows the manipulation of the flow of light in unexpected and new ways.

PBG effect in PC structures can be seen as the analogous of the electron band gap effect in semiconductors. The change in the properties of the photons in PC structures is in much the same way that ordinary semiconductor crystals effect the electrons [2, 3]. Basically, the manipulation of the light has been performed by the general mechanism of total internal reflection. Total internal reflection occurs at the interface of two media with different refractive indices where the input wave is illuminated at the medium that has higher refractive index. However, this

mechanism can only be fully controllable when the surfaces are perfectly smooth. Thus, the practical realization of optical components is strictly limited for the utilization of total internal reflection mechanism. Nonetheless, PC structures offer a different way to manipulate the flow of the light that is based on the PBG effect.

For semiconductors, the periodic atomic lattice provides a periodic potential to the electrons that are propagating through the crystal. By means of the Bragg-like diffractions from the atoms in the crystal, a gap can open up to inhibit the propagation of the electrons in allowed energies. However, this inhibition is strongly dependent to the geometrical parameters of the periodic lattices and the strength of the periodic potential [3]. Likewise, this artificial potential is provided by the periodic distribution of dielectric media in macroscopic scales. If the ratio between the refractive indices of periodic constituents is high enough, then the Bragg scatterings at the interfaces provides the similar inhibition effect for the photons. One should note that the ideal PBG effect can only be observed in a perfect PC crystal. If a defect or mistake in the periodicity leads to localized states in these areas which makes the structure behaves as microcavity [11].

As it can be seen in the above paragraph, this manipulation of the crystal can also lead to new application areas. Considering the fact that introducing defects to the perfect PC crystal leads to the occuring of localized states, the pattern of the inclusions or extractions can be utilized in new optical phenomena. As it previously mentioned, creating any type of defect inside the bulk PC structure can be utilized with the necessary geometrical optimizations for microcavity and resonator applications [12]. In addition, extraction of a full line of constituents from the bulk PC structure which also looks like a classical waveguide with sidelines formed with PC sections, can be utilized as an optical guiding mechanism with high confinement [12,13]. Likewise, if this line defect is constructed as planar, this new structure can be utilized as a perfect mirror [14]. One should also note that all this effects can only occur if the periodicity of the structure is comparable to the wavelength of the light, which is also similar with the semiconductor crystals.

Naturally, the theoretical description of the waves inside the PC structure

involves the solution of Maxwell's equations in a periodic dielectric medium. If the constituents of the PC structure are purely dielectric, then Maxwell equations also indicate the solutions are scalable for the length and the frequency scale. In other words, a PC structure designed with pure dielectric materials exhibits the same fractional band gap for any length scale. Therefore, it is advantageous for the purpose of realizing the similar observations at microwave frequencies with millimeter scale feature sizes instead of using nanometer scale designs. Further, the properties of the PC structure with nanometer scale designs can easily be deduced by constructing the similar PC structure with scaling up all of the geometrical parameters and utilizing in lower frequencies.

If there is not any existence of external sources and currents, the simple form of the Maxwell equations can be written as [15]

$$\{\nabla \times \frac{1}{\epsilon(r)} \times\} \mathbf{H}(r) = \frac{\omega^2}{c^2} \mathbf{H}(r), \quad (2.1)$$

where $\mathbf{H}(r)$ represents the magnetic field, ω is the frequency of the photon and $\epsilon(r)$ is the dielectric function of the media. If the dielectric function is perfectly periodic, then the possible solutions can be characterized by a wavevector $\mathbf{k}$, and band index n . The k -space region of all allowed wavevectors is named as Brillouin zone and the collection of all solutions is formed as a band structure. If a region of frequencies with no solutions for any wavevector inside the Brillouin zone indicates that the PBG effect is observed at that frequency interval.

As the presence of mirror symmetry plane perpendicular to the constituents for two dimensional photonic crystals, it is possible to define the characteristics of the photons for both different polarizations with respect to the plane normal. These polarizations are called as transverse electric (TE) and transverse magnetic (TM) where the electric fields and the magnetic fields are perpendicular to the rods, respectively [15].

As it can deduced from the geometry of the two dimensional PC structure, the behaviours of the photons with different polarizations changes inside the structure. For a general rule of thumb, TE band gaps are favoured in the connected

lattices and TM band gaps are favoured in a lattice of isolated high dielectric permittivity regions.

More information on the basics of PC structures can be found in several references [1–3, 10, 15]. In addition, more theoretical information is given in the subsections of this chapter below. The basic operation mechanisms of PC waveguides and cavities are investigated and explained in sections 2.2, 2.3.1 and 2.5.

2.2 Slow Light in Photonic Crystals & Tight Binding Mechanism

Slow light based applications have attracted numerous researcher because of the difficulties in the control of the optical signals in time domain [16]. The velocity of the light in vacuum is approximately $c = 3 \times 10^8 m/s$ which corresponds to a movement of 30 centimeters in $1ns$. This ultrahigh speed is desired for efficient and fast data transmission in optical circuit. However, some additional mechanisms are required to slow the speed of light to handle the technical difficulties [16–18]. Next generation optical network devices require solutions that perform the task with low power consumption, high data rate and high throughput. In addition, all optical processing is needed to prevent the losses caused by the conversion of the signal from optical to electronic or vice versa. Optical switches and optical buffers can handle the problems of the transmission of the data optically by storing and arranging the timing of the optical data. In the current technology, the temporary storing and the delaying of the information flow is performed by mechanical delay lines. However, if the velocity of the light can be controlled, the response of the systems can be further accelerated. Moreover, decreasing the velocity of light can also provide the advantage of easier control of the optical phase in interferometric devices. Further, light-matter interactions can be enhanced by increasing the storing of the optical data for the purpose of improving the nonlinear effects, absorption and gain performances of the optical systems [19, 20].

A significant amount of effort has been shown for the purpose of employing photonic crystal (PC) based structures in slow light applications [21–23]. As mentioned previously, PC structures offer intriguing dispersion properties such as photonic band gap (PBG) region. Moreover, cavities, waveguides and optical resonators with high confinement properties can be designed by using PBG effect in PC structures. For PC waveguide structure, it has been shown that the velocity of the guided light is also lower than the speed of the light in regular media for some specified frequencies. The waveguide band lies inside the PBG region which also means that the field distributions of the guided modes experience strong confinement along transverse direction to the propagation of the wave.

In addition, the zero group velocity near the edges of the waveguide band can be deduced by investigating the dispersion diagrams. However, the modes at the band edges are strongly dispersion because the relation between the frequency and the wave vectors are nonlinear which causes huge distortion in the propagation of the guided waves. Therefore, numerous studies are performed to decrease the group velocity dispersion (GVD) [24, 25].

The most useful and meaningful definition of the velocity of light for slow light applications is the group velocity v_g , which corresponds to the speed of the pulse packet during the propagation. The group velocity is given by the inverse of the first-order dispersion [17]

$$v_g = \left(\frac{\partial k}{\partial \omega}\right)^{-1} \quad (2.2)$$

where ω and k are the angular frequency and the wavenumber, respectively. The effective refractive index that the slow light wave experiences is usually expressed with the group index

$$n_g = \frac{c}{v_g} = c\left(\frac{\partial k}{\partial \omega}\right) \quad (2.3)$$

can also regarded as the slow-down factor of the speed of the light from the velocity in vacuum. In addition, the GVD can be calculated with the dispersion diagram and the group velocity diagram given as $\partial^2 k / \partial \omega^2$.

A detailed investigation of this slow light formation has been made in a previous study [26]. In this study, we focus on the theoretical aspect of the ultra slow modes in a PC waveguide constructed by using weakly coupled cavities. The behaviour of the ultra slow modes is explained by tight-binding (TB) formalism, which is used to express the nature of the waveguide modes in PCs [27]. TB description in PCs is an analog of classical wave TB explanation, which corresponds to the eigen-mode splitting of the evanescent cavity modes. The impurity states of the semiconductors are an equivalent mechanism to the defect modes of the PC cavity structures. The plane wave expansion (PWE) and the finite difference time domain (FDTD) methods are used to calculate the group index values and eigen-mode splitting in the cavity modes. These calculations will be

used in extracting TB parameters and in the comparison of TB formalism and numerical results.

PC waveguides are constructed by introducing a line defect inside the PC structure to control the propagation of the incident EM field with high localization. The frequencies of the allowed modes are in the vicinity of the forbidden band of the PC, which is called photonic band gap. Furthermore, the modes are bounded by the waveguide mode that is created by the formation of linear defect. It can be deduced that the extra bound is a sign of high confinement and eventually the appearance of slow light modes. The confinement can be increased by introducing dielectric rods of different radii inside the waveguide and coupling cavities. Thus, the bound of the allowed modes is squeezed by the coupled cavity formation, which results in ultra low group velocity. The formation of the PBG inside a classical two dimensional PC waveguide and the alternative coupled cavity structure that is used to investigate the TB formalism in this study is shown in 2.1.

The coupled cavities work as an impurity chain in the PC structure. Every cavity that is placed inside the structure creates a new resonance. The resonances eventually converge to a continuous band, creating a waveguide from the cavity chain. Strong coupling between the cavities causes a combined effect, which is crucial to realize the analogy of the TB mechanism.

Each resonance occurring due to the waveguide can be represented by the eigenmodes of each cavity. The relation between the resonances can be deduced by introducing different number of coupled cavities into the system, which results with an eigenmode splitting scheme. The eigenmode splitting was expressed as a set of formulation [27]. The eigenmode splitting can be demonstrated by only two different configurations, which are the single cavity and double cavity structures. There is a single resonance peak for the single cavity case centered at ω_c . For the double cavity, the two resonance peaks split symmetrically obeying the following relation [27];

$$\omega_{1,2} = \omega_c \frac{1 \pm \beta}{1 \pm \alpha + \Delta\alpha} \quad (2.4)$$

where β , α and $\Delta\alpha$ are the TB parameters. $\Delta\alpha$ is very small compared to α

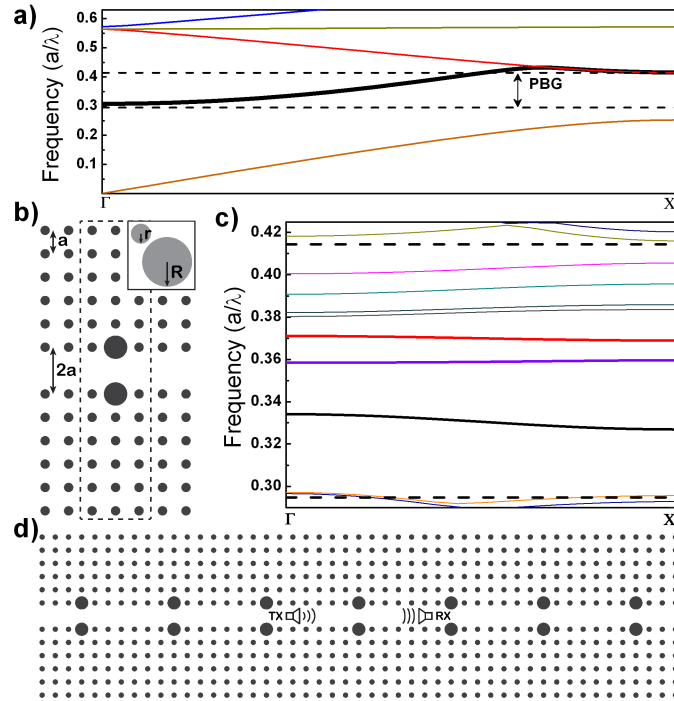


Figure 2.1: (a) The dispersion relation under transverse magnetic (TM) field excitation of the two dimensional PC waveguide structure; the black curve is the waveguide mode. (b) The super-cell of the PC coupled cavity configuration. The dashed lines enclose a coupled cavity waveguide, which has smaller coupled cavities (c) The dispersion relation of the PC coupled cavity structure, the studied waveguide bands are given bold, which are observed to be approximately flat bands. (d) The complete PC coupled cavity configuration.

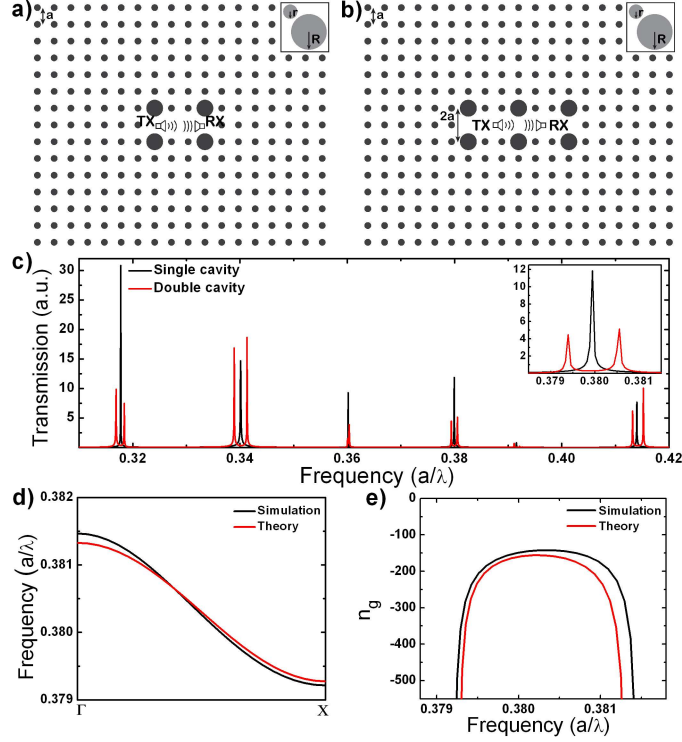


Figure 2.2: (a) Single PC cavity formation. (b) Double PC cavity formation to observe eigen-mode splitting. (c) The transmission spectra for the single and double cavity structures. (d) The comparison of dispersion relations of a selected waveguide band of the PC coupled cavity structure, which is constructed by the demonstrated cavities. (e) The comparison of the numerical and theoretical group index values.

and β thus it can be ignored. The splitting mechanism is demonstrated using the transmission data of the two configurations as given in Figs. 2.2(a) and 2.2(b).

The splitting separation decreases by introducing more cavities. The resonance peaks will emerge to each other and form a waveguide band, eventually. The dispersion relation of the occurring waveguide is given as [27];

$$\omega^2(k) = \omega_c^2 \frac{1 + 2\beta \cos(kb)}{1 + 2\alpha \cos(kb)} \quad (2.5)$$

where b is the intercavity distance. The parameters, α and β are ignorable compared to unity in flat waveguide modes as in the ultra slow modes. Thus, the

dispersion relation can be simplified as [27];

$$\omega(k) = \omega_c(1 + \kappa \cos(kb)) \quad (2.6)$$

where κ is equal to $\beta - \alpha$. The group index of a waveguide band can be calculated by combining Eq. 2.2, Eq. 2.3 and Eq. 2.6;

$$n_g = -\frac{c}{\omega_c \kappa b \sin(kb)} \quad (2.7)$$

The comparison of the tight-binding aspect and the numerical results of a slow light mode are shown in Fig. 2.2.

Furthermore, the relation between the group index and frequency can be made independent of the constant, κ , which results in an efficient method to calculate the group index by the knowledge of the transmission spectrum. The relation between κ and frequency is given as;

$$|\kappa| = \frac{\Delta\omega}{2\omega_c} \quad (2.8)$$

where $\Delta\omega$ is the frequency bandwidth of the waveguide. Combining Eq. 2.7 and Eq. 2.8, the final expression for group index is given as;

$$n_g = -\frac{2c}{\Delta\omega b \sin(kb)} \quad (2.9)$$

It is apparent in Eq. 2.9 that group index is inversely related to bandwidth and inter-cavity distance. To sum up, tight-binding mechanism is applied to examine the characteristics of the slow light modes of a coupled cavity structure. This formulation provides flexibility in determining the group velocity of an incident field inside the PC by the knowledge of the transmission spectrum, if the waveguide band supports TB mechanism. Furthermore, the characteristics of ultra slow light modes can easily be determined without phase measurements.

2.3 Optical De-multiplexer Design by Utilizing Slow Light in Photonic Crystals

As mentioned in the previous sections, PCs are strongly sensitive to the wavelength of the input source. This sensitivity also provides significant changes in the output transmission even for the slight geometrical modifications in the PC structures. The wavelength selectivity may be utilized for constructing optical devices that are commonly used in optical communication systems such as optical filters which have the most usage in wavelength division multiplexing (WDM) systems [28–30]. Several de-multiplexer (DEMUX) methodologies, i.e. arrayed waveguide gratings (AWG), Bragg grating and thin film filters, have been proposed for achieving WDM operations that can be feasible for planar light wave circuits [31–33]. Some technical properties are desired such as compactness, cost-effectiveness, polarization independence and insensitivity, designs with low cross-talk and employing high spectral resolution for DEMUXs to achieve high quality optical communication [34, 35]. The conventional designs that are mentioned above mostly meet the demands regarding the resolution and cross-talk properties whereas the compact and cost effective designs have not reached state-of-art and feasible solutions [33]. PCs can be good candidates for new generation DEMUXs that are used in planar light wave circuits due to intriguing dispersion characteristics provided in smaller geometries.

Wavelength selectivity properties of PCs have been investigated to realize compact DEMUX designs by various methodologies such as using directional couplers, frequency selective micro PC cavities, multimode waveguides and employing self imaging and super prism properties. In the work using PC directional couplers, demultiplexing operation is achieved by creating different resonance frequencies with different perturbations that are also responsible for the coupling of the operation frequencies into the coupled parallel waveguides which are employed as output channels [32, 33, 36]. Similar mechanism for creation of different resonance frequencies is also achieved by designing PC cavities with slightly different geometrical properties in Ref [36]. In the studies that multimode waveguides

are employed, the operation mechanism can be explained such that the waveguide modes in the photonic band gap (PBG) region also creates a mini stop band (MSB) where the modes experiences different group velocities [37–40]. The higher order waveguide modes have larger field penetration depth because of experiencing lower group velocities in which can be extracted from the output channels if the channel is thinned. The demultiplexing mechanism is finally achieved by constructing different output channel thicknesses according to the group velocities of each waveguide mode. Moreover, demultiplexing mechanism is also investigated in the structures that consist of bulk PCs where the self-imaging and super prism properties are employed [41–43]. Further, a recent study proposes a general methodology for demultiplexing mechanism by employing coupling between different PC waveguides [44]. In all of these PC based DEMUX studies, it can be said that the large area and complex operation mechanisms can be eliminated by using PCs which provide strong wavelength selectivity due to rich dispersion characteristics that are occurred for the wavelengths that are comparable to the period of the crystals.

2.3.1 Optical De-multiplexing Methodology

Optical de-multiplexing is mainly based on both creating different paths for different operation wavelengths and provide an effective extraction with minimum cross-talk between the output channels. The first step is designing a cascaded structure which has subsections that are sensitive to different wavelength intervals. In our studies, we constructed PC waveguides which has different geometrical periodicities which causes slight shifts in the allowed wavelengths to achieve different operation frequencies for each subsection. Secondly, we designed output channels such that each channel is only sensitive to the operation frequencies of the corresponding waveguide. This spatial sensitivity is achieved by matching the effective impedance of the output channel and the interested wavelength in the corresponding waveguide. In other words, the wavelengths that exhibit lower group velocities are extracted from the output channels. One should also note that the design of the output channels are arranged such that the slow light modes

with wider spatial distributions are easily penetrated into the output channels. By combining these two spatial and spectral selectivity properties, an efficient de-multiplexing operation is realized.

For the purpose of realizing spectral selectivity, we proposed two different mechanism which are using PC waveguide subsections with different periodicities and PC waveguides with different widths. In the first design, we aimed to achieve de-multiplexing of single wavelengths from each output channels. In the second design, we tried to achieve a DEMUX structure that the output channels collect an interval of wavelengths which is also called as frequency division DEMUX.

2.3.2 Single Wavelength Extraction

In single wavelength extraction operation, we report a DEMUX design using slow light phenomena in PC waveguides. We investigate the transmission characteristic of each output channel for each targeted single wavelength. The main target of this study is based on employing a DEMUX operation in a rather compact and small system. Thus, fulfilling aforementioned desired technical properties of DEMUX designs are not in the scope of this work.

The methodology of the wavelength selectivity mechanism is based on the modulation of the PBG region by altering the unit cell periodicity and electric-field distributions in the slow light regime. The dielectric filling factor of each PC section is altered that in turn modulates the bounds of the PBG region. The allowed waveguide modes lie inside the PBG region because there is not any other mechanism than PBG effect that supports guidance to propagating fields inside the waveguide. Therefore, other propagating modes leak out of the structure in the absence of guiding mechanism. As discussed in 2.2, the oscillation period along the longitudinal direction increases for the input wave if the wave experiences higher effective dielectric permittivity in the PC waveguide. Consequently, the increase in the longitudinal direction leads to penetration deeper toward the transverse plane to the propagation direction. Further, if the width of the PC is decreased on one side and coupled to a new PC waveguide channel which acts as

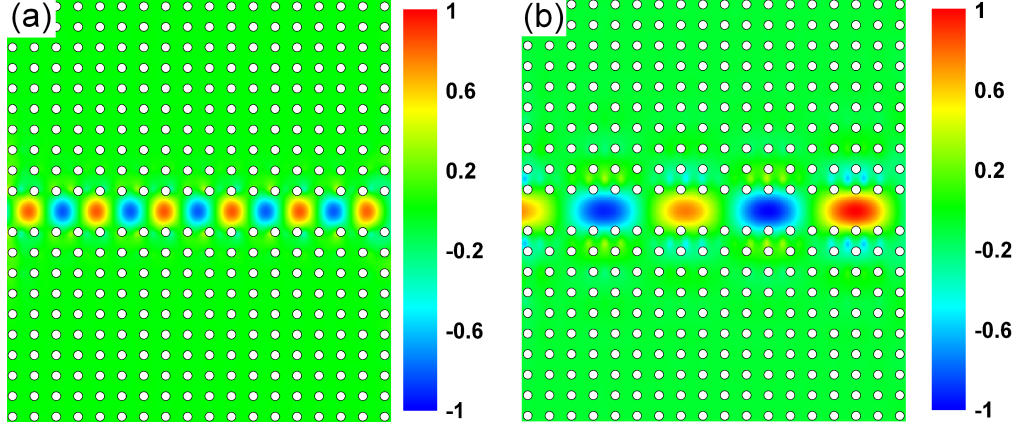


Figure 2.3: (a) The steady-state electric field distribution when the wavelength of incident light corresponds to a propagating mode within the PC waveguide. (b) The field distribution for a different frequency, which is in the vicinity of the slow light regime. The same PCW is used in both cases.

an output channel, the wavelengths that have broader spatial distribution in the transverse plane start to leak out at the thinned place of the PC waveguide into the output channel. If this mechanism is provided for PC sections with different periodicities, the extraction of different wavelengths, slow light regions, can be achieved. The difference in the spatial distributions of the guided regular and slow light modes can be seen in Fig. 2.3.

The geometry of the DEMUX design consists of three cascaded PC waveguide sections which have different unit cell periods, as shown in Fig. 2.4. The solid lines indicates the interfaces of each PC waveguide section. The main waveguide is constructed by the removal centerline row along the Γ X symmetry direction. The output channels are created by removing 2 rows of rods in the upper side of the main waveguide. The distance between two consecutive PC sections are kept as the periodicity of the section placed at the right part of the structure.

The numerical investigations of the PC structures are performed by utilizing finite-difference time-domain (FDTD) [45] which is a time-domain based numerical analysis technique generally used for electrodynamic modeling. The simulation region is surrounded by perfectly matched layer (PML) to apply fully absorbing boundary condition [46]. The unit-cell computational lattice a is divided into 32 grid points to sense the effect of incremental change in the periodicities

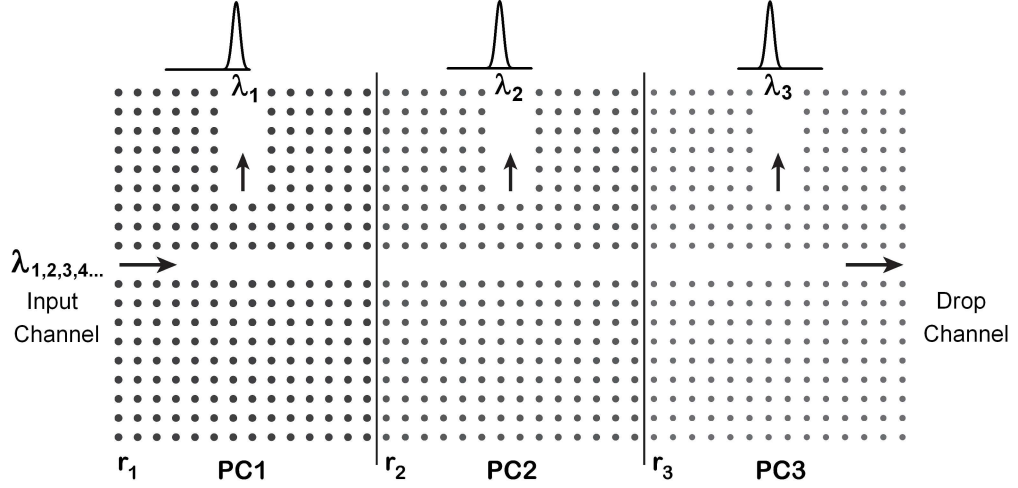


Figure 2.4: The schematic of the device that is designed for the wavelength de-multiplexing. The different wavelengths are spatially separated at different locations along the x-direction. The complete structure is composed of three PCWs of different dielectric filling factors.

for the purpose of investigating the wavelength shift in each output channel. The incoming wave is chosen as a Gaussian wave whose peak frequency is changed to excite the different output channels. The gaussian wave is intentionally selected to resemble the horn antennas which will be used in the experimental efforts. The input wave is set to TM polarization (i.e., the electric field component is perpendicular to the propagation plane) to create a PBG region and a highly confined waveguide mode for the proposed PC type.

The dielectric permittivity of the rods is selected as 9.61 in the numerical investigations for the purpose of achieving a relatively high contrast between the air and the PC structures. The material is chosen as Alumina (Al_2O_3). In the case of TM polarized light illumination, the spatial localizations of the electric field is mainly concentrated on the rods. Therefore, the high dielectric permittivity contrast strengthens the interaction between the waves and the rods. Furthermore, if materials with higher dielectric permittivities are used, the separation between the operation wavelengths are expected to become larger.

As mentioned above, the DEMUX structure consists of three PC sections where the r/a ratios are slightly altered to achieve spectral selectivity. In other

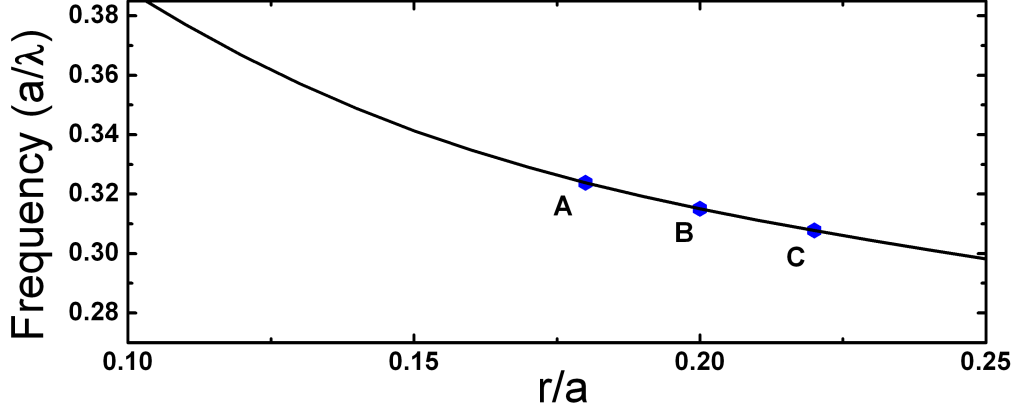


Figure 2.5: The relation between the waveguide band cut-off frequency and the dielectric filling factor. The highlighted frequencies correspond to the selected PCW sections.

words, each PC section performs a testing by checking the wavelength of the input wave. The geometrical parameters of each PC section are selected by numerically investigating the cut-off frequencies of the waveguide bands of the consecutive PC waveguides. In Fig. 2.5, the selection of the r/a ratios of the consecutive PC waveguides is shown.

The first PC waveguide section has the largest rods, in other words highest dielectric filling factor (f), where f equals to $\pi r^2/a^2$. The first PC waveguide section has $f = 15.2\%$ and the radius of the rods is $r_1 = 0.22a$ (which is shown as point C in Fig. 2.5). The filling factors of the second and the third section are decreased to 12.57% and 10.18% , respectively, by setting $r_2 = 0.20a$ and $r_3 = 0.18a$ (points B and A, respectively). As a result of decreasing the filling factors, the cut-off frequency of the PC waveguide increases as shown in Fig. 2.5.

The spectral properties and the evolution of the waveguide modes with the change in the r/a ratios of the PC structures can be seen in Fig. 2.6. The selection of the operating wavelengths in Fig. 2.6(b) is based on the group velocities of those. As the group velocity of the wave is given with $\partial w/\partial k$, the slow light regime of the waveguide modes in Fig. 2.6 is in the vicinity of the band-edge. Although the minimum group velocity occurs at the band-edge ($k=0$) the targeted wavelengths are chosen as slightly lower because the highest output transmission

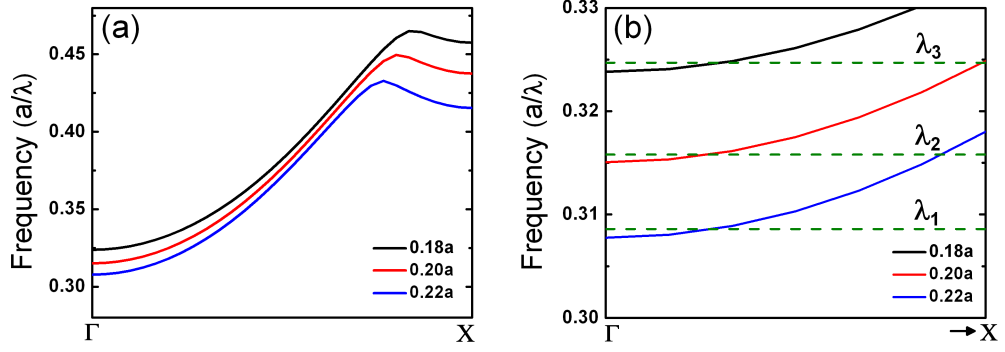


Figure 2.6: (a) The evolution of the waveguide bands due to the different dielectric filling factors, (b) the enlarged view of the vicinity of the slow light region, where the dashed lines correspond to the operating frequencies.

does not occur at the band-edge. There are two main reasons that the output transmission does not reach its maximum value along the waveguide band: Firstly, the waveguide modes near $k=0$ have higher impedances resulting in a higher impedance mismatch between the air and the main waveguide channel. Secondly, the maximum of the transmitted intensity does not occur at a mode that has a higher group velocity that is in the linear region of the waveguide mode as a result of the spatial distribution of the modes. Furthermore, the spatial distribution of the targeted mode is not only related with its frequency. The geometrical parameters of the PC waveguide is also effective on the spatial distribution. In other words, wider spatially distributed slow-light mode can be achieved by increasing the length of the waveguide. Thus, the increase in the length of the PC waveguide results in a wider spatial distribution which is desired for the proposed DEMUX methodology. However, it can be deduced from the dispersion diagrams that slower modes occur at the band-edge which also means that slower modes have the lowest frequency along the allowed waveguide modes in the PC waveguide. Consequently, the frequency shift of the peak transmission from the band-edge and the length of the PC waveguide are inversely proportional. The targeted operating frequencies are chosen by investigating the transmission properties of each PC waveguide section selected in Fig. 2.5 by considering the aforementioned explanations.

The transmission spectra of each PC waveguide section is given in Fig. 2.7

which shows both the numerical and experimental characteristics of output transmissions. It can be seen that the transmission of the slow light modes are bounded from two sides which creates resonance like transmission characteristics. The lower frequency bound is caused by the PBG effect. The higher frequency bound is caused by the impedance mismatch and the spatial distributions of the modes. As it can be seen from the dispersion diagrams, the group velocity of the waveguide modes increases with the increase in the frequency. Therefore, the spatial distributions of the modes become less wide which means that the coupling between the main waveguide channel and the output channel decreases. Consequently, each PC section behaves as a frequency filter and allows the propagation of only a portion of the modes that are already propagating in the main waveguide channel. This narrow band transmission can be seen as a coupling mechanism that has a quality factor of nearly 1000.

As mentioned previously, the design of the main waveguide and the output channels are optimized to maximize the output transmission at the operating wavelengths. The length of the PC waveguide and the geometrical parameters of the output channels are optimized to achieve the maximum output efficiency for this specific DEMUX device. The output channel is also designed as a wider waveguide than the waveguide to create a waveguide band which provides a lower effective dielectric permittivity for the targeted wavelength to achieve gradual permittivity transition from the main waveguide. Gradual transition in the permittivity also provides a more efficient match between the impedances of the main waveguide and the output channel which in turn results stronger coupling and higher output transmission. The design of the output channel can be achieved by constructing single- or multi-mode PC waveguide with a waveguide band which constitutes lower effective dielectric permittivity for the corresponding wavelengths. For an easier and more feasible design, the output channels are selected as multi-mode PC waveguides which has a width of $3a$.

The selected wavelengths for the DEMUX operation are given in Fig. 2.7. As provided in Fig. 2.5, operating normalized frequencies for each PC waveguide section are selected as $a_1/\lambda_1 = 0.3086$, $a_2/\lambda_2 = 0.3158$, and $a_3/\lambda_3 = 0.3247$. For the purpose of achieving the targeted filling factors and the normalized frequencies,

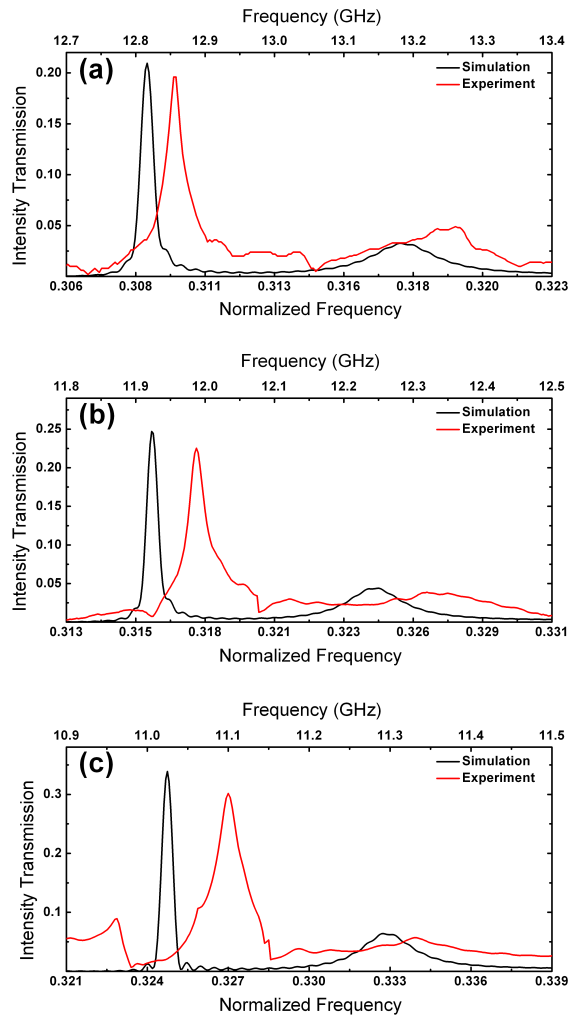


Figure 2.7: The output frequency spectra of the channels of each PCW section that has a r/a ratio of (a) 0.22, (b) 0.2, (c) 0.18, respectively.

the unit cell of each PC waveguide section is changed and the radii of rods are kept constant which can also be seen that each normalized frequency for each PC waveguide section is unique. Therefore, each targeted frequency is extracted from the corresponding output channel. In Fig. 2.7, the experimental output transmission spectra for each PC waveguide section are also shown for the purpose of comparison with the numerical results. The radii and the length of the alumina rods that are used in the experiments are selected as 0.158 cm and 15.32 cm, respectively. Consequently, the actual unit cell periods of each PC section can be calculated as 0.878 cm, 0.79 cm, 0.718 cm for $r = 0.18a$, $r = 0.20a$ and $r = 0.22a$, respectively. The proposed PC waveguide sections is installed between two flat platforms that have comparable dielectric constants to that of free space. Two standard horn antennas are used to conduct the experiments. The antennas are both placed 10 cm away from the main and the output channels. For the purpose of measuring the transmission coefficients, HP-8510C network analyzer (Agilent Technologies, USA) is used. The demultiplexed frequencies and the similar output transmission spectra are observed in the experiments. However, because of the fabrication inconsistencies and impurities in the used materials, a slight frequency shift occurred between the numerical and experimental results. Usage of alumina rods with finite length also cause power leakage in the z -direction which also can be seen in the experimental results. However, the leakage is minimized by means of using relative long rods ($\sim 100r$) compared to the operating wavelengths. Moreover, the numerical and experimental results are in good agreement which can also be seen in Fig. 2.7. This good agreement also proves that the leakage in the z -direction is significantly prevented. In Fig. 2.7, the spectra of each section are shown separately because changing the unit cell period of the PC sections instead of radii of the rods to achieve the targeted normalized frequencies creates different normalized frequency axes for each PC waveguide section.

The demultiplexing mechanism occurs from the largest wavelength to the smallest one in the proposed PC structure. Each PC waveguide section behaves as a band pass filter and extracts the targeted wavelength into the output channel. Furthermore, assuming that the light has the appropriate wavelength to propagate through the first PCW, the second PCW checks the wavelength and allows

the light to pass if the wavelength of light corresponds to a propagating mode, in other words, whether it is distant from the slow light region. Otherwise, the wave will be extracted into the output channel due to its spatial distribution and the high coupling between the main waveguide and the output channel. Briefly, each PC section performs a similar task for different wavelength intervals. Each section either extracts the wave into the corresponding output channel or couples it to the next PC section. If the input wavelength is not extracted from one of the output channels, then the input will propagate through the main waveguide without any widening in the transverse direction and reaches the end of the main waveguide. One should also note that the injection ratio of the incoming wave between the main waveguide and the air is lower than that of a wave having a comparable group velocity with air since modes in the slow light regime have higher impedances. Nonetheless, the injection ratio for the slow light modes can also be maximized by adding an additional input waveguide to achieve a more gradual transition in the effective permittivities which results in a better impedance match. Similar approach can be applied to the output channels to increase the overall extraction efficiency. Actually, this approach is already used for the purpose of designing output channels. The selection of a multi-mode waveguide with a width of $3a$ also provides a smoother impedance transition which in turn results a better coupling efficiency. This approach can also be thought as an incentive idea to design more efficient input and output couplers for PC based slow light devices. The output power at the output channels are smaller than the power at the output of the main waveguide when a single PC section is considered as the DEMUX device. However, the composite structure consisting multiple PC waveguide sections is also designed to inhibit the propagation of the wavelengths that are already demultiplexed in the previous PC sections. Then, the inhibited wavelengths reflect because they coincides with the PBG region of the next PC waveguide section. Therefore, the extracted power at output channels become larger than the power at the end of the main waveguide as a result of the reflection of the previously de-multiplexed wavelengths at the surface of the next PC waveguide section.

Different than the transmission spectra, the numerical results of the combined

PC device are also simulated by keeping the unit cell period of each PC waveguide section and changing the radii of the rods to achieve the same filling factors. The steady-state electric field distributions of the targeted demultiplexed wavelengths are shown in Fig. 2.8. The electric field distribution for the normalized frequency $a/\lambda = 0.3086$ is shown in Fig. 2.8 (a). As it can be seen in the dispersion diagrams and in the transmission spectra, this input wavelength coincides with the PBG region of the first PC waveguide section and below that of the second PC waveguide section. Therefore, the extraction of this wavelength is expected to be from the first output channel. The field distribution depicted in Fig. 2.8 (a) confirms that the extraction mechanism works properly. Likewise, the input frequency is set to $a/\lambda = 0.3158$ and the field distribution is shown in Fig. 2.8 (b). It is expected to observe that this wavelength propagates through the first PC waveguide section without any spatial widening and dropped to the output channel of the second PC waveguide section. Finally, for the wavelength at $a/\lambda = 0.3247$, the extraction from the third channel is shown in 2.8 (c). Lastly, a wavelength that does not coincide with any of the slow light regimes of all PC waveguide section but also allowed to propagate through those, is selected and the propagation through the main waveguide can be seen in 2.8 (d). As a result, it is observed using these figures that each PCW section filters a different wavelength and directs the selected wavelength toward the appropriate drop channel.

The spatial distributions of the demultiplexed wavelengths at the output channels along the x -direction are also calculated and given in Fig. 2.9. The spectral content of the targeted wavelengths are determined by applying Fourier transform on the time domain data. The operating wavelengths are chosen such that, at these wavelengths, the output transmission is maximized whereas the cross-talk between the channels is minimized. As it can be observed in Fig. 2.9, the maximum level of the cross-talk for the first channel is calculated as -13.7 dB. The cross-talk levels for the second and the third channels are obtained as -15.6 dB and -28.7 dB, respectively. The proposed demultiplexing methodology also provides low cross-talk ratios at the output channels because of the inhibition of the wavelengths that are already demultiplexed in the previous output channels by utilizing the PBG regions of each PC waveguide section. The cross-talk ratios

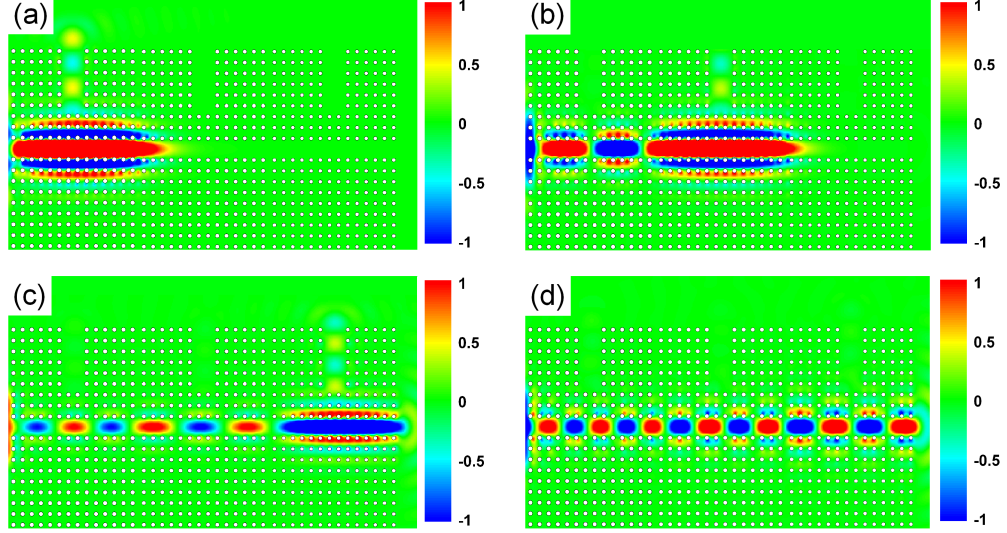


Figure 2.8: The steady-state electric field profiles of the three de-multiplexed frequencies for each waveguide channel where the input beam is placed $5a$ away from the PC structure, (a) $a/\lambda = 0.3086$, (b) $a/\lambda = 0.3158$, (c) $a/\lambda = 0.3247$, (d) the steady-state electric field profile of an unselected frequency, which has a normalized frequency of $a/\lambda = 0.34$.

can further be decreased by a more effective selection of targeted wavelengths from the transmission spectra of the output channels. Moreover, cross-talk ratios can also be reduced by using rods that have higher dielectric permittivities which causes stronger separation of the selected wavelengths as a result of deeper interaction between the PC structure and the waves that are in the slow light regime. In Fig. 2.9, it can be seen that the output distributions have sinc-like profile in x -direction which indicates that the coupling mechanism between the main waveguide and the output channels properly works. Moreover, sinc-like distribution profiles also show that the dispersion characteristics of the demultiplexed wavelengths in the main waveguide transform into the dispersion characteristics of the output channels.

The proposed methodology shows that the spectral and the spatial selectivity of the targeted wavelengths can be obtained. The performance of the proposed structure can be increased by geometrical modifications such as increasing the length of the PC waveguide sections which results in a peak transmission wavelength with a smaller group velocity. Furthermore, the spectral selectivity can be

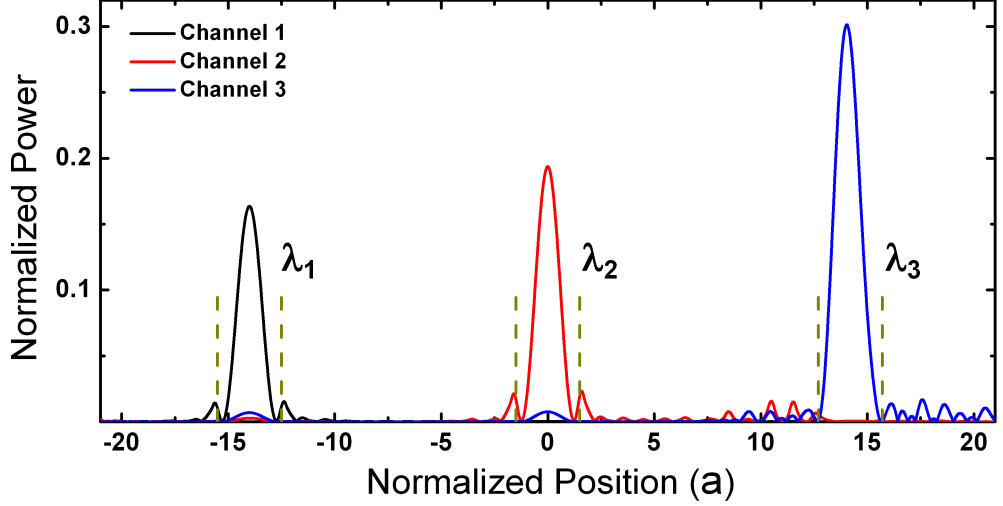


Figure 2.9: The spatial distributions of the operating frequencies at three output channels in x-direction. The output levels are calculated nearly 5a away from the output channels. The green dashed lines show the positions of the output channels in x-direction.

increased by keeping the difference of the filling factors between the PC waveguide sections smaller. In other words, to demultiplex wavelengths with smaller shifts can be achieved by cascading PC waveguide sections with small changes in the filling factors. The number of the PC waveguide sections can be increased in order to demultiplex more than three wavelengths. Additional PC waveguide sections should also be carefully designed for the purpose of achieving low cross-talk ratio between the other channels by considering the overlap of the new targeted wavelength with the waveguide bands of the previous channels to inhibit the transmission of the previously demultiplexed wavelengths. Lastly, the usage of PC waveguide sections with equal lengths leads to a compact design, although the management of cross-talk between channels becomes more difficult as a consequence of the absence of the minimization of back reflections from each section.

2.3.3 Frequency Division Demultiplexing

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In the single wavelength extraction section, we demonstrated the implementation of a different approach that benefits from the spatial distribution characteristics of the waveguide modes for the purpose of obtaining the desired frequencies from the corresponding cascaded photonic crystal waveguide (PCW) blocks [48]. In the present study, we studied the potential of combining the ultra compact single frequency de-multiplexing operation proposed in Ref. [48] and the modulation of the waveguide bands by modifying the waveguide widths, which is proposed in Refs. [35] and [49]. It should be noted that the dielectric constants and the periodicity of the consecutive PC sections are not varied for the purpose of achieving frequency division de-multiplexing. In addition, the possibility of obtaining a better low cross-talk ratio compared to the study given in Ref. [50] is investigated and this aim is achieved by adjusting the cut-off frequencies of each waveguide band with the careful optimization of the parameters.

The de-multiplexing mechanism in this study is designed and optimized for achieving frequency division operation in a specific frequency interval by the involvement of the slow light regime of cascaded PCWs. The main purpose in this study is to create an ultra compact design with a very small footprint that can be achieved by the utilization of PC based structures. Moreover, the entire de-multiplexing scheme is bounded such that the frequency bands of interest lie inside the stop band region of the PC and that the optimization results in a minimized cross-talk ratio between the output channels. Initially, a classical two dimensional PC is designed to create a photonic band gap (PBG) region. Afterwards, one entire row of dielectric rods is removed along the ΓX direction from the PC to create a PCW along the propagation direction. As it can be deduced, a modulation of the waveguide bands in the frequency axis is required to achieve frequency selectivity between the cascaded PCWs. We propose that a modulation can be introduced by altering the waveguide widths. Thus, a combination of two PCWs with different waveguide widths can be used for suppressing specific frequencies that are not desired in the next PCW. After the construction of a frequency selective cascaded waveguide scheme, the introduction of output channels is necessary for the extraction of the de-multiplexed frequencies. At this point, we implement a novel mechanism for the extraction of the desired frequencies

that includes the utilization of the spatial distribution characteristics of the slow light modes inside the PCW that correspond to the close vicinity of the $k=0$ band edge of the Brillouin zone of the corresponding PCW. The field distributions of the slow light modes along the direction that is perpendicular to the propagation direction are spatially more broadened as a consequence of the increase in the oscillations in that direction. It is intuitive to predict that opening a new waveguide along the perpendicular direction gives rise to the leakage of the modes that have broadened field distribution along the transverse plane by evanescent coupling to the new waveguide. Therefore, the output channels are formed by structuring a new PCW that is created by removing extra rods along the perpendicular direction. The output channels are designed to have a waveguide width of $3a$, where a is the periodicity of the PC structure, that corresponds to the removal of two entire rows of dielectric rods. Although the output waveguide exhibits multi-mode propagation, it is necessary that it has a width that is larger than the main waveguide width in order to realize a gradual impedance transition for the de-multiplexed frequencies that results in an optimal coupling between the waveguides. To sum up, the proposed design is finalized after forming the output channels that provide for the extraction of the desired frequencies.

The geometry of the proposed DEMUX design is depicted in Fig. 2.10. Two PCWs that contain rods of the same radii and have the same lattice periodicity but unequal waveguide widths are cascaded. The unequal waveguide widths generate a modulation in the waveguide bands that makes each PCW sensitive to a different band of frequencies and different regions of the slow light modes. The frequency sensitivity also comes from the nature of the PC structure that prevents the propagation of the modes that lie inside the PBG region.

The proposed structure can be investigated by starting with a classical two dimensional PC structure. The periodicity of the lattices and the dielectric constant of the dielectric rods are selected as $r/a = 0.24$ and $\epsilon = 12$, respectively. The dispersion diagram of the PC structure with the given lattice periodicity and the dielectric constant value is shown in Fig. 2.11(a). The dispersion characteristics of the PC are determined by using the Plane-Wave Expansion Method (PWM). The structure has a PBG between the normalized frequencies of 0.25

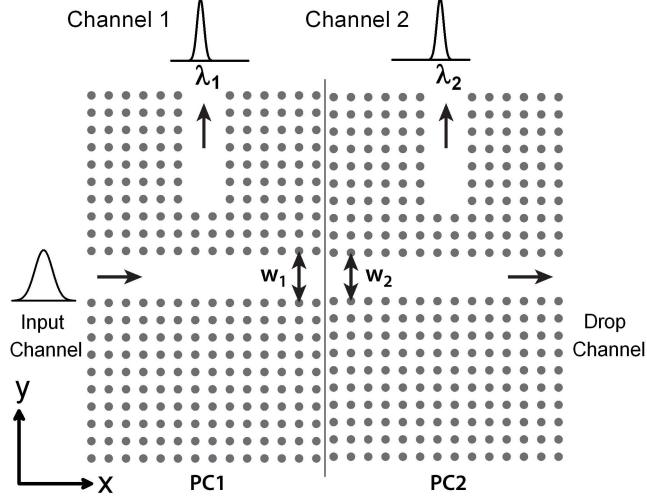


Figure 2.10: The geometry of the proposed DEMUX structure. The design consists of two cascaded PCWs with unequal waveguide widths w_1 and w_2 , respectively. The radii of the dielectric rods and the lattice periodicities are identical.

and 0.38. The removal of an entire row of rods in the x -direction results in a set of allowed modes in the vicinity of the normalized frequencies of 0.285 and 0.39 that form a waveguide band. The cut-off frequency of the waveguide band can be modulated by the alteration of the waveguide width. In the numerical calculations, the width of the waveguide is changed between $1.6a$ and $2.4a$. The PBG region remains approximately constant since the total filling factor of the PC structure does not change while altering the channel width. However, the position of the waveguide band on the frequency axis changes as a result of the modification of the waveguide width. The evolution of the waveguide band with respect to the waveguide width is shown in Fig. 2.11(b). For all the waveguide widths, the cut-off frequencies of the waveguide bands lie inside the PBG region and that provides a large parametric interval for optimizing the design such that it exhibits the desired DEMUX behavior. The expected de-multiplexed frequency intervals are the slow light modes on the waveguide band that corresponds to the flat regions of the waveguide band. Therefore, the widths of the cascaded PCWs should be selected to be separated enough such that the slow light regions do not coincide. Accordingly, the waveguide widths of the PCWs are chosen as $w_1 = 2a$ and $w_2 = 1.8a$. This cascaded waveguide structure provides two separated slow light regions for each one of the cascaded PCW sections. Finally, an extraction

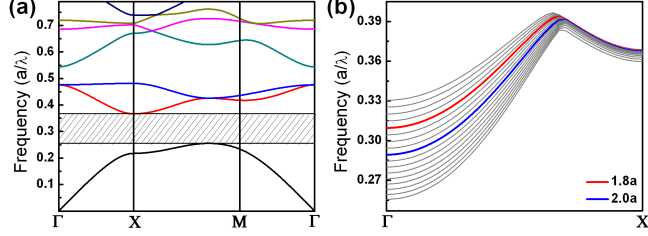


Figure 2.11: (a) The dispersion diagram of a two-dimensional classical PC with a r/a ratio of 0.24 and a dielectric constant of 12. (b) The evolution of the waveguide band with respect to the waveguide width. The blue and red lines correspond to the widths for the DEMUX design. The PCW widths are altered linearly from $1.6a$ to $2.4a$ with an increment of $0.5a$

mechanism is required in order to direct these slow light modes into different output channels.

A multimode PCW is designed as an output channel for each PCW section. Indeed, a single mode classical PCW can serve the same mechanism to the system, but the underlying reason for this choice can be explained by inspecting the dispersion characteristics of the designed output channel. Firstly, the output channels are constructed by removing two rods along the y -direction to drop the slow light modes using their broadened spatial distribution along the y -direction. Secondly, it is known that the group velocities of the targeted frequency interval are lower than the other modes that are allowed to propagate in the PCW. Thus, the impedance mismatch between the air and the structure for the targeted modes is larger than the other allowed frequencies that in turn results in a significant amount of reflection in the coupling process between the main PCW and the output channels. A gradual transition in the impedances is needed to couple the targeted slow light modes into the free space efficiently. Moreover, it is known that a broader PCW exhibits a set of allowed modes that have higher group velocities than a classical PCW with a waveguide width of $w = 2a$. Hence, such a broad PCW can be a good candidate for a more optimized power extraction through the output channel. Finally, a PCW is designed with a waveguide width of $3a$ in order to propose a simpler design and to create a gradual impedance transition for the slow light modes between the main channel and the air. The DEMUX design is finalized by the formation of the output channels.

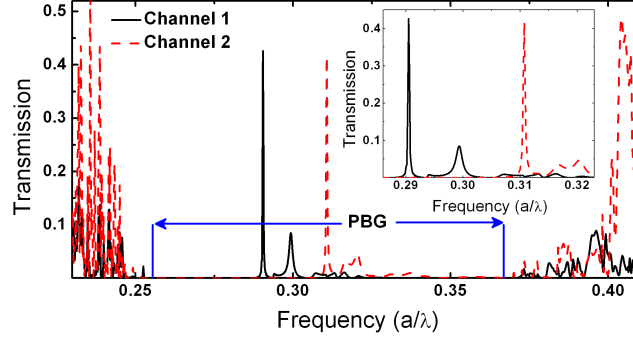


Figure 2.12: The transmission spectra of the output channels. The inset shows the enlarged view of the de-multiplexed frequency intervals.

The transmission spectra of the output channels are given in Fig. 2.12. The numerical analysis of the proposed design is performed by using FDTD SOLUTIONS (version 6.0) from Lumerical Solutions, Inc. (Vancouver, Canada), which is a commercial software program based on the finite difference time domain method (FDTD). The geometrical parameters and the dielectric constant of the design are given in the previous section. The input source is chosen as a Gaussian wave that has a wide range of frequency components in order to cover the entire PBG region of the classical PC. The input source is positioned far enough from the DEMUX structure in order to illuminate the structure with almost flat phase fronts for achieving an approximate plane wave illumination. The electric field intensity is monitored at the output of each output channel. The intensity values are normalized with respect to the intensity of the input source. Fig. 2.12 clearly shows that, the extraction of the slow light modes inside each PCW section can be achieved with a very high quality factor.

The transmission through the output channels is restricted by the PBG region and the slow light region as explained in detail herein above. However, there is a second transmission peak with a much lower intensity for each output channel that corresponds to faster modes compared to the narrow transmission peaks. The occurrence of the second peaks is related to the back reflections at the output of the each PCW section. This transmission which is due to the internal reflections inside the main channel can be decreased by creating longer PCW sections that can be the subject of a further optimization. It is observed that the

extraction of modes other than the slow light modes in channel 2 is much less than channel 1. This observation shows that the reflection in the second PCW section is less crucial than the reflections from the interface between the PCW sections. Moreover, the de-multiplexed frequency intervals at each output channel do not overlap and the center frequencies are similar to the cut-off frequencies of the waveguide bands shown in Fig. 2.11(b) and that shows that the results derived from the PWM and FDTD methods are consistent. One can see that the center frequencies of the extracted slow light modes are not exactly the same as the cut-off frequencies of the corresponding waveguide widths but are slightly higher. The reason behind that can be explained by the nature of the modes at the $k=0$ band edge. The modes at the $k=0$ band edge have very large group indices that correspond to a high effective dielectric constant that can result in an approximately unity reflection at the PCW input. Therefore, the peak of the transmission at the output channel slightly shifts to higher frequencies. The transmission is also restricted by the waveguide modes with higher frequencies because their group indices decrease with the increase in the frequency and their field distributions in the transverse direction become narrower and which gives rise to a decreased coupling between the main channel and the output channels. Furthermore, these two limitations on the de-multiplexed frequency intervals result in a narrowband transmission in frequency. The quality factors of the de-multiplexed frequency intervals are calculated approximately as 1930 and 1550 for channel 1 and 2, respectively.

The spatial distributions of the de-multiplexed frequencies are shown in Fig. 2.13. The behavior of the slow light modes in the main channel and the coupling between the main channel and the output channel can be clearly observed. The investigated frequencies are selected such that they are the frequencies of peak transmission of the de-multiplexed frequency bands for each output channel. The separation between the cut-off frequencies of the waveguide modes of the chosen PCW sections can be seen in Fig. 2.13(a) as zero transmission through the boundary of PC1 and PC2. Moreover, the broad spatial distributions of the slow light modes for both channels are clearly demonstrated. In Fig. 2.13(b), the propagation of the frequencies to be de-multiplexed in the channel 2 along the

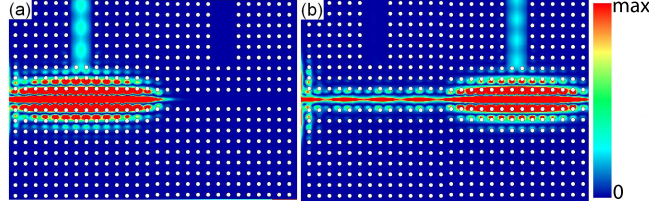


Figure 2.13: The spatial distributions of the pulses with normalized frequencies of (a) $a/\lambda = 0.2905$ and (b) $a/\lambda = 0.3105$ that correspond to the centers of the de-multiplexed frequency bands at PC1 and PC2 sections, respectively.

cascaded structure is shown. It is an allowed waveguide mode with a moderate group index whereas it becomes a slow light mode after entering the waveguide section of PC2.

We also examined the spatial distributions of the center frequencies of the de-multiplexed frequency bands along the x -direction at the output of the output channels and these results are shown in Fig. 2.14. The degree of the cross-talk between output channels can be calculated by the usage of the electric field intensities along the x -direction. The maximum cross-talk values for channel 1 and channel 2 are calculated as -20dB and -40dB, respectively.

The restrictions on the de-multiplexed frequencies result in a very low cross-talk between the channels. Actually, the underlying reason for having obtained such small values for the cross-talk is the modulation of the waveguide bands that is performed by changing the waveguide widths. The slight difference in the cut-off frequencies causes the de-multiplexed frequencies to be prohibited for the next PCW section. In the proposed design, the de-multiplexed frequencies in channel 1 are already in the stop band of channel 2 which yields a very low cross talk for channel 1. The reason for the low cross-talk for channel 2 is the physical mechanism of the de-multiplexing operation. The de-multiplexed frequencies in channel 2 propagate with moderate group indices along the waveguide of the PC1 section. Therefore, they cannot be dropped along channel 1 as a consequence of their narrow spatial distribution along the y -direction.

The de-multiplexing operation is investigated and achieved by utilizing the slow light regime of PCWs. Further optimizations can be performed in order to

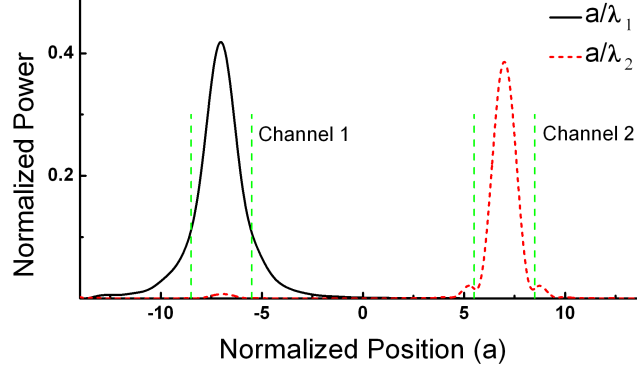


Figure 2.14: The spatial distributions of the center frequencies of the de-multiplexed frequency bands at the output channels along the x -direction. The vertical green dashed lines correspond to the positions of the output channels.

create a design with narrower de-multiplexed frequency bands, a larger power extracted power and a lower cross-talk ratio between the channels. The length of the PCWs can be modified in order to obtain different center frequencies. The waveguide widths of the cascaded PCW sections can be adjusted in order to modulate the de-multiplexed frequencies inside the PBG region. The number of the PCW sections can be increased for the purpose of de-multiplexing more than two sets of frequency bands. The back reflections inside the waveguides can be minimized by a further careful optimization of the geometrical properties of the PCWs.

In this study, the utilization of a high permittivity dielectric has two significant advantages. Firstly, introducing a higher dielectric contrast between the periodic structure and the background medium results in a larger PBG and enables the easier control of modulation by the modification of the PC waveguide width. Secondly, the waveguide bands have smaller slopes at the $k=0$ band-edge due to the usage of high permittivity dielectrics and that result in slower modes. However, a similar mechanism can be obtained by using materials with smaller dielectric constants, which are more appropriate concerning the optical frequencies, with some drawbacks such as a narrower PBG region, which corresponds to a less flexible design, and de-multiplexed frequency intervals with lower quality factors as an implication of the lower group indices at the operating frequencies.

2.4 Optical Diode-like Devices based on Graded Photonic Crystals

Compact optical diode-like devices are one of the most demanded devices for on chip photonic circuits [51–61]. Due to reciprocity theorem for linear and pure dielectric media, implementing diode-like electromagnetic wave propagation brings several technical difficulties whereas the confinement and routing of the light in photonic circuits are rather easier by means of index and band gap guiding. One-way light propagation has significant importance because its electronic analogue is widely used in on chip electronic circuits named as electronic diode. In electronic diodes, the inhibition of the electrical signal propagation is provided by changing the polarity of the voltage difference of the signal from positive to negative. In the proposed design, the direction of the illumination determines the propagation of the optical pass through the device. For the ideal proposal, one-way propagation should work as to allow input signal to travel through the device for forward direction illumination whereas the transmission should be inhibited for the backward direction illumination. Early optical diode proposals have utilized different mechanism to obtain the desired effect such as breaking the time-reversal symmetry. Nonetheless, it is needed to introduce material nonlinearity or metallic and magnetic materials in the early approaches [51, 53, 58, 61]. For designs including material nonlinearity, high-power input is generally required because majority of the nonlinear materials constitutes significantly low nonlinear susceptibility coefficients. In addition, for the designs including metallic or magnetic inclusions, high optical losses are generally observed. Despite the fact that the most of the early designs includes nonlinear, metallic or magnetic media, some different approaches may be realized for asymmetric light propagation by only using linear and pure dielectric inclusions. For instance, asymmetric gratings can be used to manipulate the direction of the light direction by creating higher diffraction orders and suppressing the *zeroth* order for one of the illumination directions. Likewise, higher order diffractions can be eliminated and the *zeroth* order diffraction dominates the power flow of the input signal to achieve asymmetric light propagation [54–57]. This similar methodology is also used for devices that

can be employed for acoustic waves by using sonic crystals [62, 63]. Additionally, photonic crystal based approaches are also presented for optical diode-like applications such as using hetero junction structures that utilizes directional band gap difference effect [64].

2.4.1 Asymmetric Light Propagation Methodology

The electronic analogous of the asymmetric light propagation is the electronic diode mechanism. In electronic diodes, the flow of the electrons is inhibited under reverse biasing whereas the electrons flow without any prevention mechanism under forward bias. Therefore, electronic diode can be classified as a nonlinear circuit element. Similarly, this concept can be carried to the photonics circuits because of the significant importance of asymmetric information processing. Optical diode mechanism can be fully realized by proposing designs that break reciprocity. However, we propose a partial solution for asymmetric power collection at the detectors placed at the ends of the structure with keeping the media reciprocal.

The main aim of this design is concentrating the input wave into an area for one of the illumination directions while deflecting the same frequency under opposite illumination direction. A composite structure that has focusing abilities can further increase the difference between the concentration and the deflection ratios of the operating frequencies under forward and backward illumination cases. A classical PC waveguide allows a bundle of frequencies to propagate through which is also named as waveguide band. Moreover, the dispersion characteristics of the waveguide band can be significantly changed by modifying the geometric properties. For instance, the cut-off frequencies of the allowed modes can be controllably altered by changing the width of the waveguide. Likewise, the unit cell period of the PC waveguide can also be modified to achieve new dispersion characteristics. For the purpose of introducing an asymmetric wave deflection to the system, we preferred to change the interdistances between the neighbouring columns of the rods with increasing increments. One should notice that the linear change in the interdistances can result in a simultaneous modification of

the dispersion characteristics of the waveguide band. In other words, the change in the interval of the allowed modes for each neighbouring column of rods can lead to deflect some of the modes at the spatial position that the modes are not hosted in the guiding mechanism any more. Similarly, some of the modes can be guided after a propagation inside the PC section without any mechanism due to the simultaneous change in the dispersion characteristics.

The following path of the light is shown in Fig. 2.15. The major component of the path of the waves are highlighted with red arrows. As explained in the previous paragraph, under forward illumination, the input wave propagates through the center of the structure at first. After a few lattice distance, electric fields starts to leak towards the y -direction. This leakage is due to the change in the unit cell period of the rods along x -direction. The input wave exits the structure at the upper sides of the structure. The gradual change in the unit cell periods along x -direction results in loss of guiding mechanism in the structure. As shown in Fig. 2.15(b), the main component of the input wave exits the structure at the end of the waveguide for backward illumination. Therefore, it can be seen that the proposed structure provides a contrast ratio for opposite directions of illuminations at the ends of the waveguides in Fig. 2.15. In addition, the electric field distributions of the targeted frequency for the proposed linearly graded PC structure are shown in Fig. 2.16 which have the general characteristics of the proposed light path methodology.

The lack of symmetry along the x -direction in the proposed gradual PC waveguide structure weakens the guiding mechanism for forward illumination. However, this lack of symmetry also strengthens the guiding mechanism for the input wave that propagates inside the PC structure for backward illumination. As the separation between the columns of rods increases, the waveguide mode shifts outside of the PBG region which means that the guiding mechanism disappears. Then the waveguide modes start to leak out from the central part of the waveguide into the bulk PC part of the structure. The targeted frequencies exit the structure in the form of two main side lobes. This similar path is followed by the targeted frequencies for the backward illumination. However, this similar path corresponds to the entering of the targeted frequencies with the form of two main side lobes

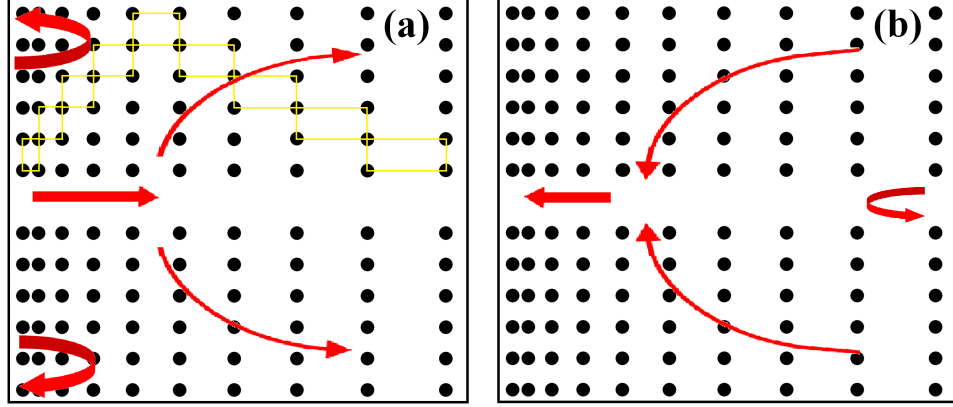


Figure 2.15: Schematic representation of a chirped photonic crystal waveguide that enables lights asymmetric propagation is provided. (a) and (b) correspond to left-to-right and right-to-left propagation, respectively. Yellow boxes in (a) show unit cell variations along the propagation direction.

into the PC structure and then concentrating into the center of the waveguide at the exit.

2.4.2 Numerical Results of the Proposed Device & Experimental Verification

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As mentioned previously, time-reversal symmetry is given as $\omega(k) = \omega(-k)$ where k represent the wave vector. In other words, if a structure provides optical diode-like devices, it also means that the time-reversal symmetry of the system has already been broken. In this study, we proposed a graded photonic crystal (PC) waveguide structure consisting only linear and pure dielectric materials for the purpose of achieving asymmetric light deflection operation. Different than the classical PC waveguides, we break the periodicity of the PC structure along the direction that is parallel to the waveguide to create a media constituting gradual effective dielectric permittivity.

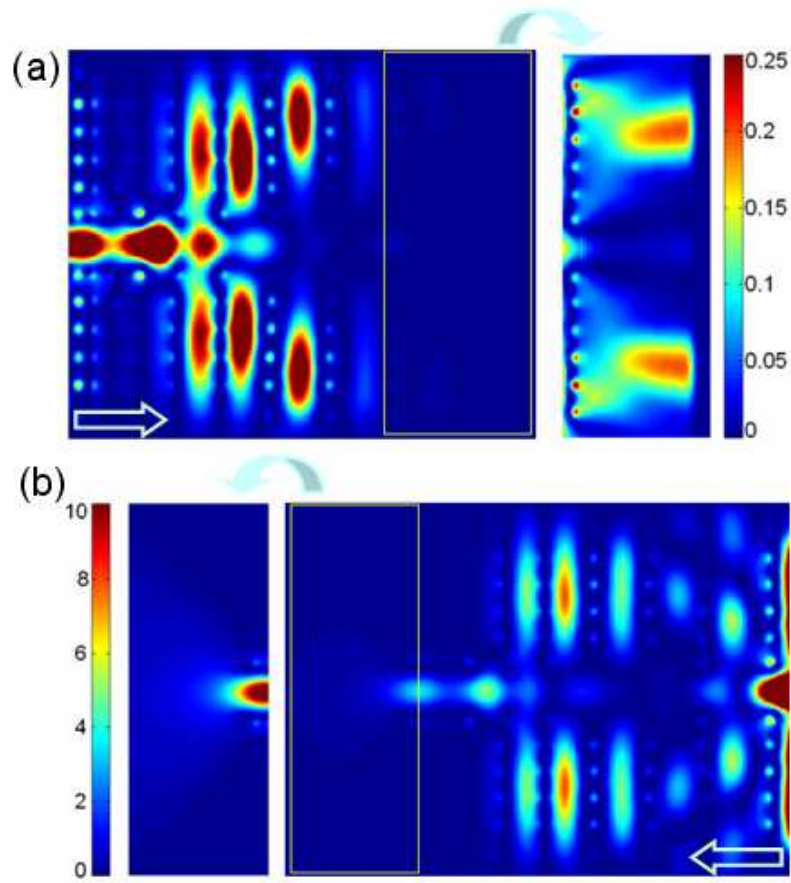


Figure 2.16: The detailed presentations of the intensity distributions in linearly chirped waveguide configuration for the forward and backward light propagations are shown in (a) and (b), respectively.

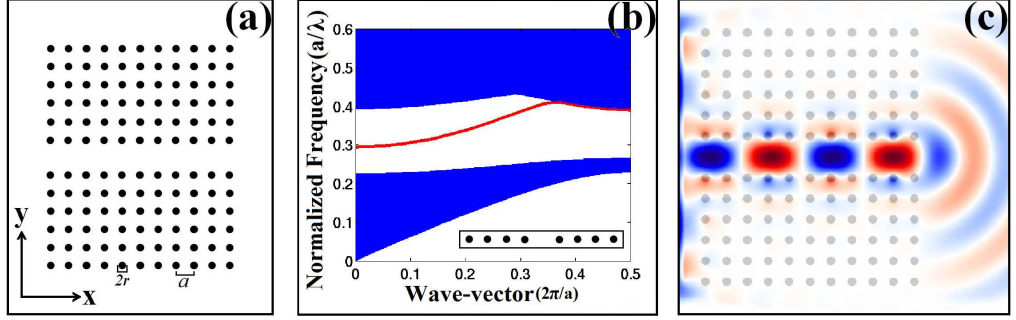


Figure 2.17: (a) Schematic of a regular photonic crystal waveguide. (b) The dispersion diagram of regular photonic crystal waveguide. (c) Incident light propagation from left-to-right in the waveguide structure.

First of all, the field profiles of the waveguide modes in the regular PC waveguide are calculated. Then, the spatial distributions of the modes that are allowed in a designed graded PC waveguide are compared with the regular PC waveguide. It is observed that the behavior of the waves propagating through the waveguide with opposite illumination directions have some differences. By changing the geometrical parameters of the graded waveguide, the asymmetric wave propagation properties of the waveguide are optimized. For comparison purposes, the designed square-lattice PC waveguide is shown in Fig. 2.17(a). The radii of the rods are chosen as $0.22a$ where the unit cell period of the rods is denoted as a . The dielectric permittivity of the rods are selected as 12. The PC waveguide is constructed by introducing a line defect along ΓX direction. The spatial distributions of the modes that lie inside the waveguide band for forward and backward illumination cases are calculated under inputs that are TM polarized where electric field components of the waves are parallel to the plane of propagation. The spatial distributions are numerically calculated by utilizing finite-difference time-domain method (FDTD). The dispersion diagram of the regular PC structure is given in Fig. 2.17(b) by employing plane wave expansion method (PWM) [66]. The electric field distribution of one of waveguide modes is shown in Fig. 2.17(c) where the input wave is illuminated from left to right. The field distribution under backward illumination case is also calculated and the results yield same pattern through the system which means that the system does not support any asymmetric propagation mechanism.

The distance between each neighbouring column of the dielectric rods in the regular PC waveguide is intentionally altered incrementally to break the structural symmetry of the system which is shown in Fig. 2.18(a). The unit cell period of the rods are kept constant along y -direction ($\Delta y = a$). The alteration of the interdistance between the rods along x -direction is denoted as $a(x)$. The total length of the graded PC waveguide is around $16a$ long and the function of $a(x)$ is altered between $0.50a$ and $2.50a$. The spectra of the contrast ratio between the output transmission under forward and backward illumination is provided in Fig. 2.18(b). The contrast ratio is defined as

$$T = \frac{T_{-\hat{x}} - T_{+\hat{x}}}{T_{-\hat{x}} + T_{+\hat{x}}} \quad (2.10)$$

where $T_{+\hat{x}}$ and $T_{-\hat{x}}$ denote the transmission intensity under $+x$ and $-x$ directions, respectively. The non-zero regions in the contrast ratio spectra mean that asymmetric power collection or optical diode-like behaviour can be provided for the presented structure. The peak contrast ratio occurs at the normalized frequency of $a/\lambda = 0.3288$. In Fig. 2.18(c) and 2.18(f), the field distributions of the targeted frequency is given for forward and backward illumination, respectively. The electric field distributions along the y -direction at the output of the waveguide can be seen in Figs. 2.18(d) 2.18(e). Considering the results given in Fig. 2.18, the waveguiding mechanism does not hold for the targeted frequency when the input light is illuminated from the right side of the structure where as the main output transmission is concentrated just at the end of the waveguide which corresponds to a strong guiding mechanism. One should note that the following path of the wave does not change by changing the direction of illumination under the assumption that the input wave is not a point source. For a input source that has a comparable width to the width of the PC structure, the concentrating of the majority of the power at the ends can differ which means that the wave can be deflected unidirectionally.

The similar approach is also investigated for using parabolic chirping parameter instead of the previous linear increment between each neighboring column of rods. The two dimensional cross section of the structure is shown in Fig. 2.19(a). Due to usage of a different chirping parameter, the total length of the graded PC

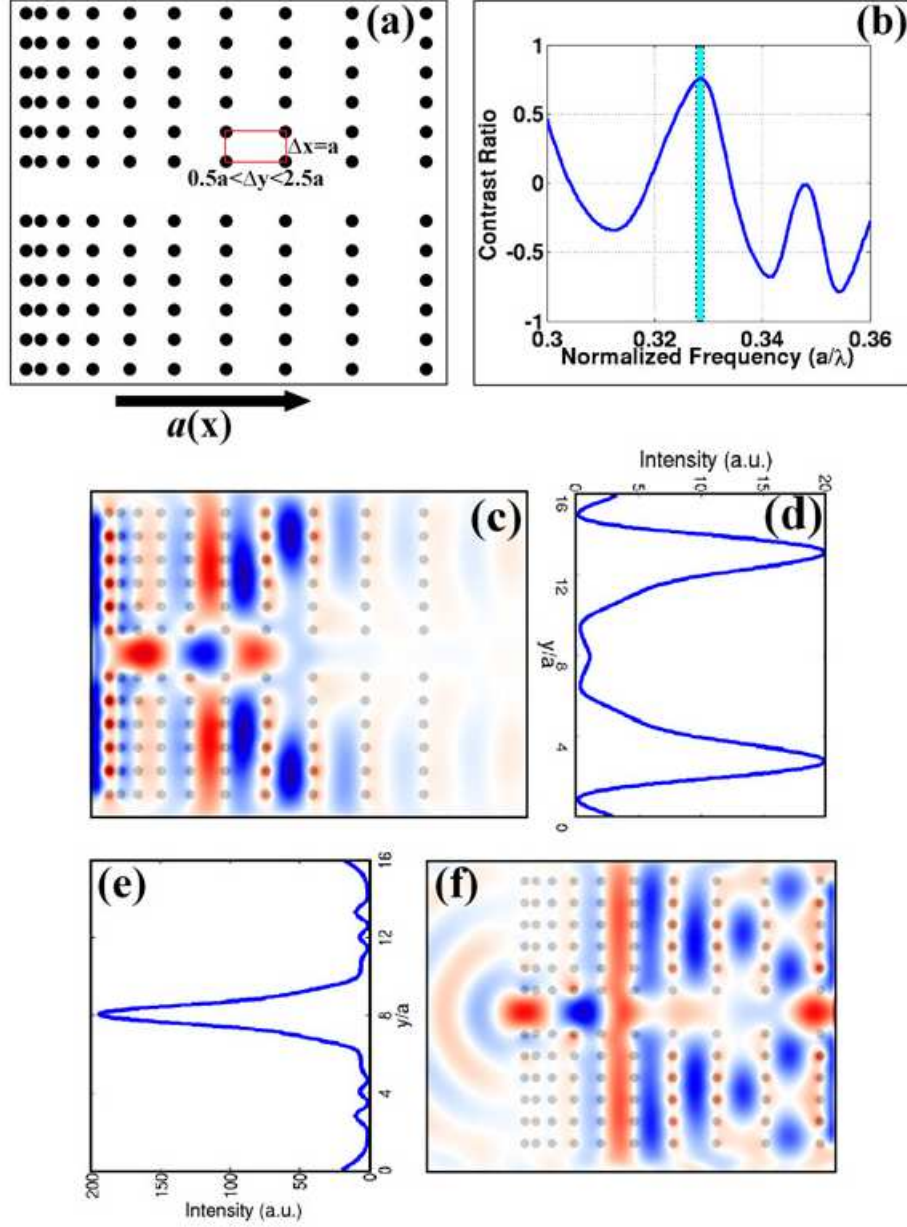


Figure 2.18: (a) A photonic crystal waveguide that has linearly increased the distance separation between each neighboring column of rods. (b) The spectral content of the contrast ratio. The blue rectangle designates the selected frequency value. (c) Time-domain snapshot of propagation of light from left to right. (d) The transverse field profile across the end face of the structure. (e) The transverse field profile across the front side of the structure. (f) Time-domain snapshot of light propagation from right to left.

waveguide is increased to $16.89a$. The reason behind this new chirping parameter is to achieve a higher contrast ratio between the forward and backward transmission. In Fig. 2.19(b), the contrast ratio spectrum of the structure reveals that the operating frequency of $a/\lambda = 0.2992$ experiences the maximum difference for opposite illumination directions. The electric field distributions of the targeted frequency for forward and backward illumination is illustrated in Fig. 2.19(c) and 2.19(f), respectively. The electric field distributions at the ends of the waveguide along the y axis is also shown in Fig. 2.19(d) and 2.19(e). As it is evident in Fig. 2.19, the results also reveals that this PC waveguide structure also shows optical diode-like mechanism. Furthermore, it can be seen that the contrast ratio of the graded PC waveguide structure with parabolic chirping coefficient is higher than the linear increment case when the ratio of the intensities for opposite directions at the centerline of the waveguide is calculated.

After the numerical investigations, the experimentals are performed for the purpose of confirming the numerical results for the graded PC waveguides. The experiments are conducted in the microwave regime using an Anritsu 37369A network analyzer. The material of the rods in the experiments are selected as alumina. The radii and the dielectric permittivity of the rods are given as $r = 0.159$ cm and $\epsilon_{Al_2O_3} = 9.61$, respectively. Different than the proposed design, two additional rows of dielectric rods are included in the upper and lower side of the structure. The unit cell period of the PC waveguide structure along the y -direction is kept similar with the numerical investigations. The corresponding unit cell period is calculated as $a_y = 0.7227$ cm. The height of the rods are selected as $h = 15.32$ cm ($\sim 100r$) for the purpose of resembling the design with the two dimensional proposal as much as possible. This selection of long rods also prevents the leakage of the wave through the waveguide along the z -direction. Naturally, the transmission properties of the PC waveguide structures have been modified because of the significant change in the dielectric permittivities of the rods that are used in the numerical and the experimental studies. As well as the contrast ratio is changed due to material changes, it is needed to re-define the targeted frequencies where the maximum contrast ratio occur. Thus, the transmission characteristics are recalculated numerically for the purpose of comparison with

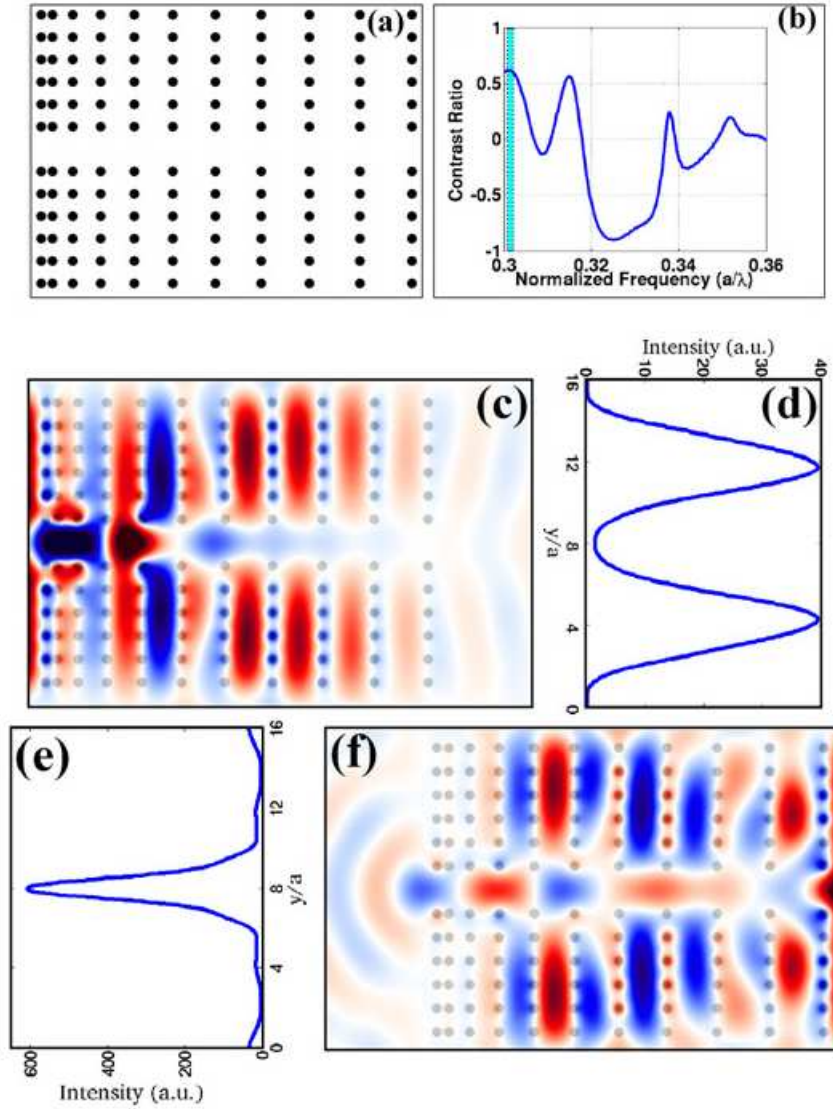


Figure 2.19: (a) Another chirped photonic crystal waveguide is shown. The distance between each neighbouring column of rods increased in a parabolic pattern. (b) The spectral content of the contrast ratio. The blue rectangle designates the selected frequency value. (c) Time-domain snapshot of propagation of light from left to right. (d) The transverse field profile across the end face of the structure. (e) The transverse field profile across the front side of the structure. (f) Time-domain snapshot of propagation of light from right to left.

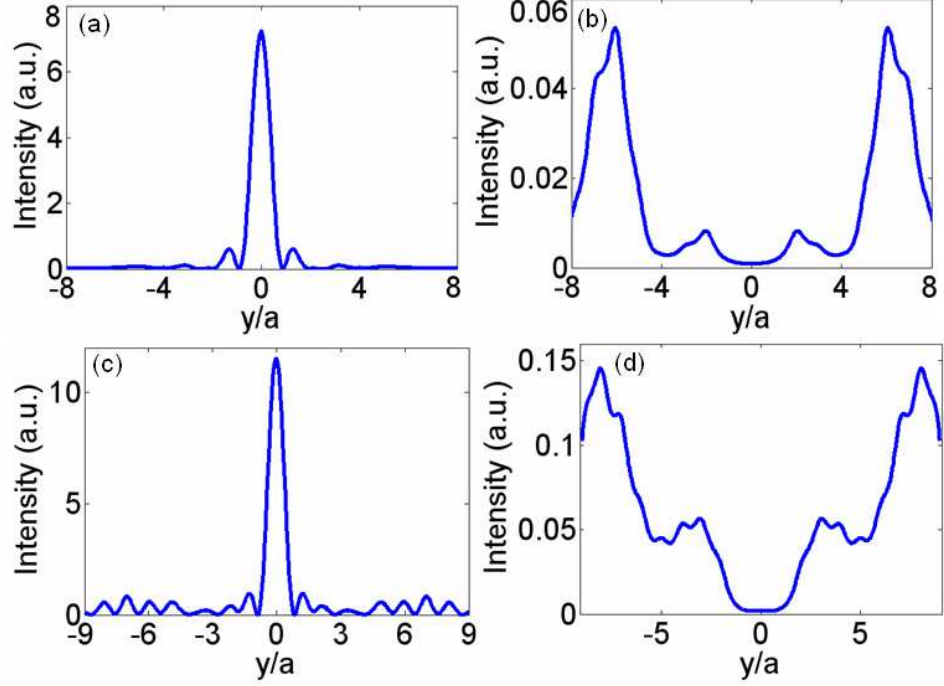


Figure 2.20: The simulated intensity distributions are provided at the outside of the structure for microwave region. (a) The transverse field profile across the end face of the linearly chirped waveguide. (b) The transverse field profile across the front face of the same structure. (c) The transverse field profile across the end face of the parabolically chirped waveguide. (d) The transverse field profile across the front face of the same structure.

the experimental efforts. Only for the targeted frequencies, the electric field profiles along the y -direction at the ends of the waveguides are shown in Fig. 2.20. The targeted frequencies for the linear and parabolic chirping are 12.28 GHz and 12.77 GHz, respectively. The transverse electric field distributions for the linear chirping case for forward and forward illumination is given in Fig. 2.20(a) and Fig. 2.20(b), respectively. The similar plots are provided for the parabolic chirping case in Fig. 2.20(c) and Fig. 2.20(d). As it can be seen in Fig. 2.20, the wave deflection is also achieved for one of the illumination directions even in the usage of materials with lower dielectric permittivities.

For the practical purposes, the PC waveguides are covered with a fully absorbing layer which is made of carbon sheets with only opening a window at the input and the output of the structures. The reason behind the placing absorbing

layer can be explained as to reach a meaningful calibration of the horn antennas. The transmission coefficients are calculated by the capture of the entire incident wave from the transmitter antenna. Receiving antennas are placed just behind the end of the structure to measure the transmission coefficients. The transmitter antennas are placed far enough (the distance between the antennas is adjusted to 50cm) for the purpose of illuminating the whole PC interface with a plane-wave.

The transmission coefficients of the linearly and parabolically chirped PC waveguides are measured in the experiments for the targeted frequencies of 12.36 GHz and 12.83 GHz, respectively. One should note that the experimental errors and the impurities in the materials that are employed in the structures are the main reason of the slight frequency shifts observed in the experiments than the numerical results. Nonetheless, the experimental transmission characteristics are in agreement with the numerical ones. In addition, it can be seen in Fig. 2.21 that the experimental results reveal the asymmetric light propagation properties of both the graded PC waveguide structures. In the experiments, the transmission characteristics are measured for the frequencies that the maximum contrast ratio occurs between the forward and backward illuminations. The placement of the receiver antenna is changed with increments of $0.5a$ for the purpose of completing the electric field distributions of the targeted frequencies at the ends of the structure along y -direction. The experimental transmission characteristics of the targeted frequency for the linearly chirped PC waveguide structure is given in Fig. 2.21(a) and Fig. 2.21 (b) for the forward and backward illumination case, respectively. One should compare the numerical and the experimental results by checking the Figs. 2.20(a) and 2.20 (b) with Figs. 2.21(a) and 2.21 (b). For the parabolically chirped PC waveguide structure, the numerical and the experimental results can be compared by looking at the (c) and (d) subplots of the Figs. 2.20 and 2.21. By inspecting the Figs. 2.20 and 2.21, it can be said that the experimental results are in agreement with the numerical ones. However, the limited spatial scan of the detector antennas results in non-smooth lines which can be refined by reducing the step size of the measuring along the y -direction.

Despite the fact that the numerical and the experimental results demonstrate the similar transmission characteristics for the targeted operation frequencies, one

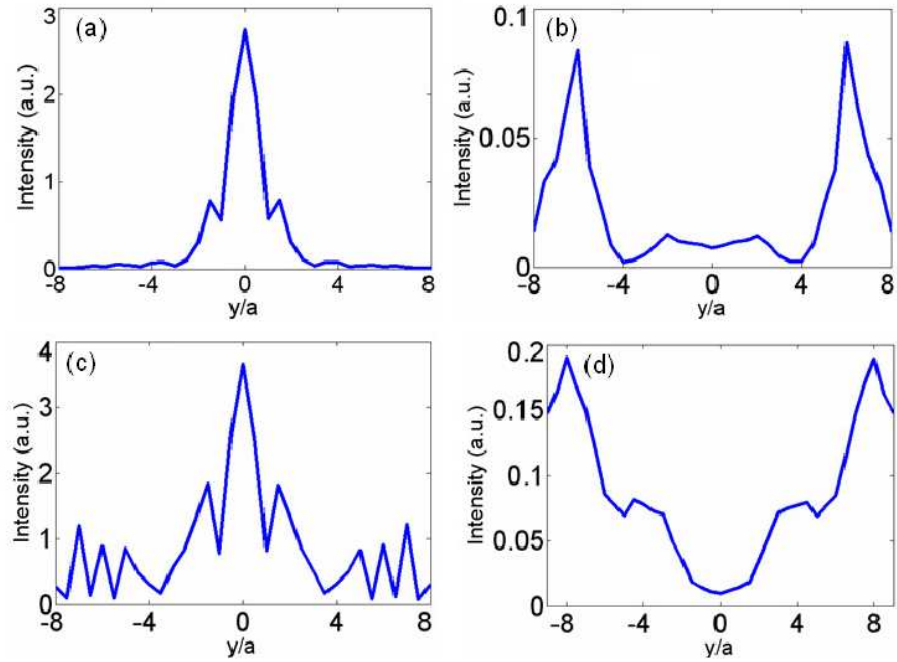


Figure 2.21: The measured intensity distributions are provided at the outside of the structure. (a) The transverse field profile across the end face of the linearly chirped waveguide. (b) The transverse field profile across the front face of the same structure. (c) The transverse field profile across the end face of the parabolically chirped waveguide. (d) The transverse field profile across the front face of the same structure.

should note that the transmission ratios are observed lower in the experiments. The inconsistencies in the calibrations of the antennas and the power leakage in the z -direction are the main reason for the lower experimental transmission ratios. One can also calculate the contrast ratios of the graded PC waveguide structures, as well. The contrast ratios of the linearly and parabolically chirped designs are calculated as 24.8 dB and 30.8 dB, respectively.

2.5 Compact Photonic Crystal based Fabry-Perot Cavity Design for Electrooptic Modulation

Photonic crystal (PC) based cavities have attracted considerable interest because of their having rich dispersion characteristics and enabling the enhanced interaction of light with matter in a small modal volume [3, 67, 68]. By using different geometrical and structural approaches, various studies have been performed for the purpose of designing a PC cavity with an ultra-high quality-factor (Q) and an ultra-small modal volume [69–71]. In spite of the fact that the achievement of a high Q and a small modal volume by utilizing a PC cavity implies the reduction of the transmitted power, such cavity designs has the potential to find applications in several fields including low-threshold lasers [72], coherent optical devices [73], optical communication and quantum information processing systems [74, 75], optical filters [76], and nonlinear photonic elements for photonic chips [77]. Particularly, the possibility of creating a strong light-matter interaction in the cavity region as a result of the highly localized fields can lead to the enhancement of the nonlinear and electro-optic effects. For instance, due to the enhancement of the nonlinear effects, a PC cavity can lead to a significant modification in the refractive index of the material that constitutes the cavity region and, in turn, a significant modulation of the operation wavelength can be realized. Furthermore, PC cavities can be highly sensitive to changes in the refractive index of the cavity region inasmuch as the optical path of the traveling wave inside the cavity region can be much larger than the physical length of the cavity.

Considering the aforementioned features of PC cavities, it is possible to realize high- Q , small modal-volume, compact, and high throughput devices that can be utilized as efficient optical modulators by exploiting the possibility of achieving enhanced nonlinear and/or electro-optic effects inside the cavity region [78–80].

PC cavities can be constructed in several ways, including point and line defect cavities [81], which are conventional PC cavities and obtained by the removal of either a single rod/hole or multiple rods/holes in the cavity region, heterostructure cavities [82] that consist of cavity regions having different periodicities and/or lattice constants compared to those of the surrounding media, air-bridge cavities [83] that consist of input/output coupling lines and a free-standing PC slab stage that is utilized for the purpose of minimizing the output plane losses by benefiting from total internal reflection, and, finally, nanobeam cavities [84] that are composed of a single line of rods/holes and exhibit high- Q s in ultra-small volumes.

In this study, we demonstrate a new type of PC cavity that is two-dimensional and based on Fabry-Perot type resonances. The design target is the simultaneous achievement of nearly unity power transmission and high field localization inside the cavity region. For the satisfaction of these design goals, we utilize an air-hole in slab configuration and the proposed design is shown in Fig 2.22. The reasoning behind the use of an air-hole configuration is the enabling of the investigation of the effect of changing the refractive index inside the cavity region on the transmission spectrum. For the purposes of this investigation, we select Lithium Niobate (LiNbO_3) as the host material. We numerically demonstrate that the high sensitivity of the effective optical path to the refractive index of the cavity region can be exploited for enabling the strong modulation of the resonance/operation frequency. Accordingly, we state that such a cavity design can be employed for enhancing the electro-optic effect of LiNbO_3 inside the cavity region due to strong field localization and enhanced light-matter interaction. In addition, we show that it is possible to alter the operation frequency and Q of the cavity by changing the cavity length, L_c , and the number of the hole columns on either side of the cavity, N , respectively. Finally, we investigate the possibility of achieving a dual-mode operation regime for a particular polarization state and a

dual-polarization regime, where resonant transmission peaks for both transverse magnetic (TM) and transverse electric (TE) polarization states are observed inside the photonic band-gap (PBG) region, by altering L_c and the r/a ratio.

2.5.1 Proposed Cavity and Sensitivity to Changes in Dielectric Permittivity

As shown in Fig. 2.22, a Fabry-Perot type cavity, which is formed between two identical PC sections that act as side mirrors, is formed and the length of the cavity is given by L_c . The host material is selected as LiNbO₃, which has a trigonal molecular crystal system and exhibits Pockels effect and nonlinear optical polarizability [85]. Owing to its trigonal crystal distribution, which is known to lack inversion symmetry, LiNbO₃ displays electro-optic effect with the application of a static or time-dependent electric field. For the numerical analysis, the incident wave is defined as a plane wave that propagates in the $+z$ direction. In the practical realization stage, the plane wave field distribution can be approximated by coupling the incident light to a wide waveguide with the aid of a fiber. It is assumed that the structure is infinitely long and periodic in the $\pm x$ and $\pm y$ directions, respectively. It is noteworthy that holes with a length that is much larger than the operation wavelength and a large number of holes in the y -axis are required in the practical realization stage. Moreover, open boundary condition is applied for the $\pm z$ directions. The radius and the periodicity of the holes are denoted by r and a , respectively. The main polarization state for operation is the TM polarization, for which the electric field is along the y -axis as shown in Fig. 2.22.

In the first step of the design stage, we construct a bulk PC for the purpose of forming a broad PBG region. As a design goal, we desire the PBG region to include $\lambda_0 = 1.55 \mu\text{m}$, which is the projected operation frequency. As a result of birefringence, LiNbO₃ has two different refractive indices: n_o , for ordinary rays, and n_e , for extraordinary rays. Here, in the numerical analysis, we prefer to use n_e , which is approximately given by 2.14 in the wavelength interval of interest.

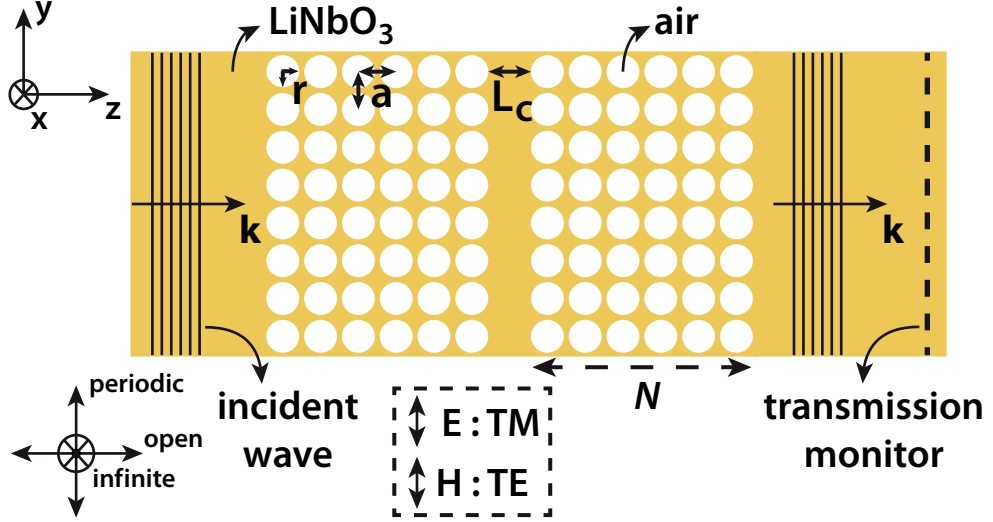


Figure 2.22: The cross-sectional view of the proposed Fabry-Perot cavity geometry. The PC structure is assumed to be periodic and infinitely long along the $\pm y$ and $\pm x$ directions, respectively. The incident wave is assumed to be a plane wave propagating in the $+z$ direction. The transmission spectrum is obtained with the aid of the transmission monitor.

Accordingly, we set the geometrical parameters for the bulk PC as $a = 510$ nm and $r/a = 0.4$. For confirming the existence of a PBG in the desired wavelength range for TM polarization, in addition to the calculation of the dispersion diagram, we also investigate the transmission spectrum. The calculation of the dispersion diagram is performed by utilizing the plane wave expansion (PWE) method [66]. The transmission spectrum is obtained by running finite-difference time-domain (FDTD) simulations (FDTD Solutions, Lumerical Inc.), in which one period of the corresponding structure is divided into 128 mesh cells for ensuring the obtaining of accurate data.

Secondly, we design the Fabry-Perot type cavity, see Fig. 2.22, for the purpose of obtaining a cavity resonance at $\lambda_0 = 1.55$ μm . One reason for working in the TM polarization is the possibility of the existence of a wider PBG region compared to the TE case. In this design, r/a is kept at 0.4 and, basically, the cavity is formed by the removal of one column of holes in the middle of the bulk PC. For $a = 510$ nm and $r = 205$ nm, we calculate $L_c = a - 2r = 100$ nm. It should be noted here that the bulk PC sections act as highly reflective blocks and

their performance is strongly dependent on N . As it will be evident in Fig. 2.25, the selection of N affects Q and the peak power transmission, e.g., increasing N implies increasing Q and decreasing peak power transmission. By noting the strong field localization inside the cavity, it can be predicted that the sensitivity of the proposed structure to changes in the refractive index of the host medium is expected to be higher than that of conventional optical devices as a result of the lengthening of the effective optical path due to multiple reflections from the cavity mirrors. Here, the utilization of LiNbO₃ can enable the modulation of the refractive index as a consequence of the application of a static electric field. It has recently been shown that the effective modulation of the refractive index, Δn , in structures having strongly localized static and optical fields can be expressed as follows [86]:

$$\Delta n = -\frac{1}{2}n_e r_{33} f_{\text{opt}}^2 f_{\text{el}} \frac{V}{L}, \quad (2.11)$$

where n_e and r_{33} denote the refractive index of LiNbO₃ for extraordinary rays and the electro-optic coefficient of LiNbO₃, respectively. In addition, f_{opt} and f_{el} are used for the optical and the static field localization factors, respectively. Finally, V and L denote the applied voltage and the distance between the electrodes, which supply the external static voltage to the system, respectively. Qualitative interpretation of Eq. 2.11 shows that the refractive index of the cavity region can be significantly modulated in geometries displaying the strong localization of the static and optical fields, i.e., high- Q cavities.

In the numerical consideration of the proposed geometry, we imitate the electro-optic effect by changing the refractive index of the host material *solely* in the cavity region. Note that it is expected that the effective refractive index modulation is much stronger in the cavity region compared to the rest of the structure. Therefore, we assume that the existence of the electro-optic effect can be neglected outside the cavity region. Accordingly, the effect of changing n_e inside the cavity region on the transmission spectrum is investigated.

Following the investigation of the effect of refractive index modulation on the transmission spectrum, different possible functionalities of the proposed geometry, such as dual-mode and dual-polarization operation regimes, are reported. For the

enabling of dual-mode operation, $L_c = 610$ nm, which corresponds to the removal of an additional column of holes, is employed and, therefore, the fitting of the two transmission resonances inside the PBG region is made possible. In the analysis of the dual-mode regime, we do not modify the geometrical parameters besides L_c . The dispersion diagram and the transmission spectrum for TM waves for the dual-mode case are also provided. Moreover, we demonstrate that the simultaneous achievement of a high (nearly unity) power transmission and a large Q is possible for both of the resonances inside the PBG for the dual-mode case.

Finally, we analyze the possibility of the achievement of a high transmission and a large Q for both TM and TE polarization states concurrently. For the geometrical parameters that are used in the single-mode case, the resultant PBG regions for TM and TE waves do not overlap and, as a consequence, the resonance frequencies for TE and TM polarizations have a significant spectral difference. Therefore, for enabling the achievement of two resonances that are fairly close to each other in terms of wavelength, we modify the aforementioned geometrical parameters and the parameters used for the analysis of the dual-polarization regime are given by $a = 540$ nm and $r/a = 0.46$. The obtained transmission spectra for TM and TE waves reveal that it is possible to utilize the proposed geometry for both polarization states since both resonances exhibit a high transmission and a high Q .

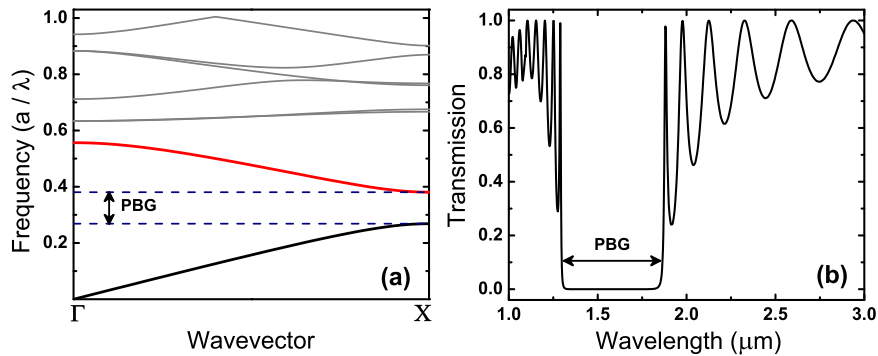


Figure 2.23: (a) The dispersion diagram and (b) power transmission spectrum of the bulk PC structure. The red and the black curves represent the upper and the lower band of the PBG region, respectively.

As it has been mentioned above that the dispersion diagram and the transmission spectrum of the bulk PC with $a = 510$ nm and $r/a = 0.4$ for TM waves are calculated using the PWE and FDTD methods, respectively. The corresponding results are shown in Fig. 2.23. According to Fig. 2.23(a), the bulk PC structure displays a PBG region between the normalized frequencies of 0.27 and 0.39 for the ΓX direction of wavevectors. The transmission spectrum shown in Fig. 2.23(b) confirms the existence of the PBG region and demonstrates that transmission is suppressed between approximately $\lambda_0 = 1.2$ μm and $\lambda_0 = 1.7$ μm .

Next, we show the dispersion diagram and the transmission spectrum for the PC cavity geometry depicted in Fig. 2.22 where $N = 6$. The corresponding dispersion diagram is shown in Fig. 2.24(a) and the mode arising due to the existence of the cavity is denoted by “Band of interest”. Figure 2.24(b) shows the transmission spectrum of the PC cavity between approximately $\lambda_0 = 1.1$ μm and $\lambda_0 = 2.3$ μm and Fig. 2.24(c) is simply the enlarged view of the region enclosed by the dashed rectangle in Fig. 2.24(b).

Figure 2.24(c) reveals that the normalized peak power transmission is approximately given by 0.85 and, under the assumption that it is reasonably small, Q can be approximated as follows [87]:

$$Q \approx \frac{w}{\Delta w} = -\frac{2\pi c_0}{\lambda_0} \frac{\lambda_0^2}{2\pi c_0} \frac{1}{\Delta \lambda_0} = -\frac{\lambda_0}{\Delta \lambda_0}, \quad (2.12)$$

where Δw and $\Delta \lambda_0$ are the full-width half-maximum linewidths evaluated when the horizontal axis denotes the angular frequency and the wavelength, respectively. Finally, c_0 is the speed of light in vacuum. The minus sign in Eq. 2.12 shows that λ decreases when w increases and vice versa. For our purposes, this minus sign can simply be neglected. By using Eq. 2.12, we calculate that $Q \approx 1275$ for the spectrum given in Fig. 2.24(c). In addition, note that the transmission resonance occurs in the close vicinity of $\lambda_0 = 1.55$ μm .

It has been stated that Q and the transmission spectrum are strongly dependent on the reflectivity of the PC sections surrounding the cavity region and, therefore, N . In order to reveal this strong relation, N is incremented from 5 to 7.

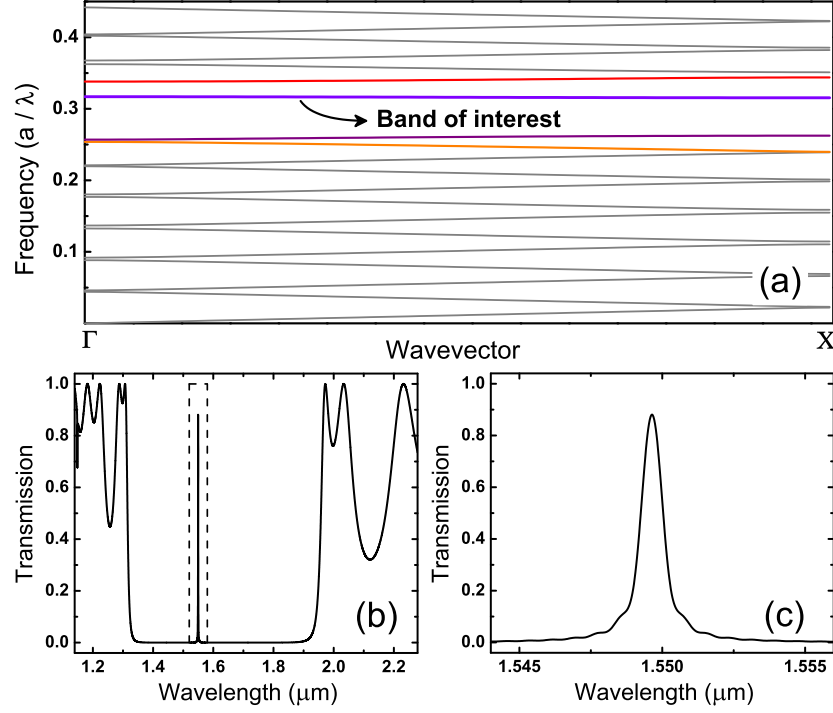


Figure 2.24: (a) Dispersion diagram and (b) power transmission spectrum of the proposed Fabry-Perot type cavity. (c) The enlarged view of the cavity resonance that is enclosed by the dashed rectangle.

Figure 2.25 shows the evolution of the transmission spectrum of the PC cavity for the above-mentioned N values. On the one hand, it is observed that Q increases with increasing N . On the other hand, however, the peak power transmission at the cavity resonance decreases with increasing N , since the reflectivity of the cavity mirrors increase with increasing N . Meanwhile, the spectral position of the cavity resonance is not modified because of the fact that L_c is kept constant. Using Eq. 2.12, Q values of the cavity resonances for $N = 5$ and $N = 7$ cases are evaluated as 455 and 4340, respectively.

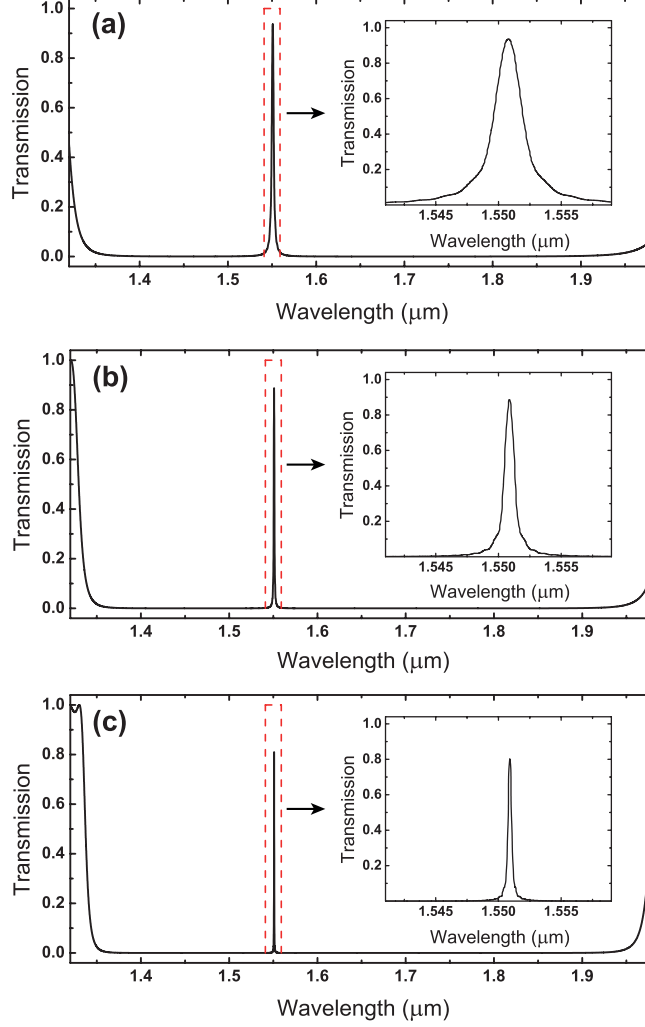


Figure 2.25: Power transmission spectra of the Fabry-Perot cavity for the cases of (a) $N=5$, (b) $N=6$, and (c) $N=7$. The insets show the enlarged view of the regions enclosed by the dashed rectangles.

Afterwards, we analyze the field distributions for the $N = 5$, $N = 6$, and $N = 7$ cases at the wavelengths where the transmission of the cavity mode is maximum and these results are shown in Fig. 2.26. As expected, the optical field is highly localized inside and in the vicinity of the cavity region. As a result of the increase of the reflectivity of the cavity mirrors, increasing N conduces to an increase in the effective optical path of light and therefore, the optical localization factor, f_{opt} . Inside the cavity region, it is estimated that the average f_{opt} is approximately given by 13, 20 and 35 for $N = 5$, $N = 6$, and $N = 7$, respectively. Finally, as can be seen in Fig. 2.26, the localized fields inside

the cavity region have an approximately uniform distribution similar to what is observed in a traditional Fabry-Perot resonator.

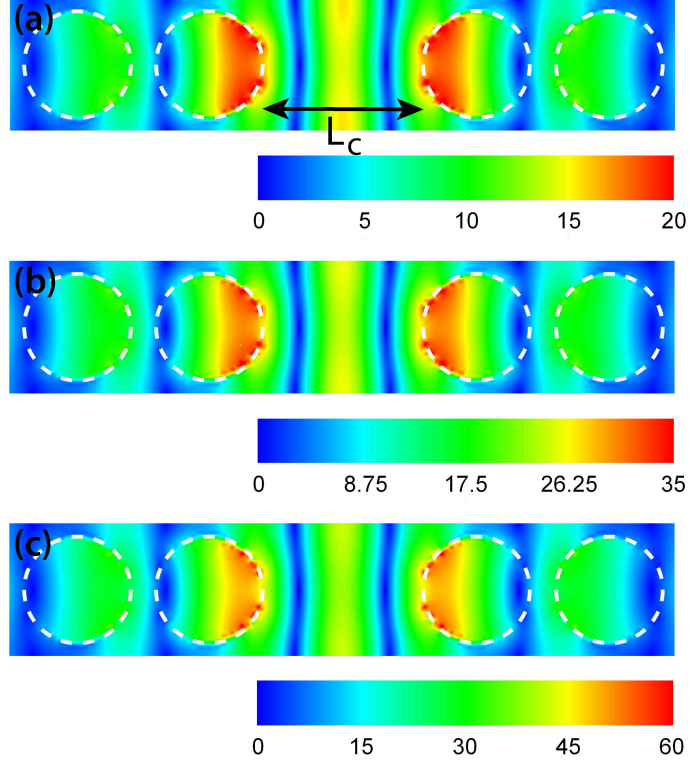


Figure 2.26: Electrical field distributions of the Fabry-Perot cavities for the cases of (a) $N=5$, (b) $N=6$ and (c) $N=7$. The white dashed circles denote the positions of the holes.

The evolution of the transmission spectrum with respect to changes in the refractive index of the cavity region, Δn , is demonstrated in Fig. 2.27. The practice of changing the refractive index of the cavity region is well-justified since Fig. 2.26 shows that the field distribution is approximately uniform in the cavity region. In the practical realization stage, the modification of n corresponds to the application of an external static field. In other words, in this context, we model the electro-optic effect by modifying the refractive index. The refractive index of the cavity region is altered in linear increments of 0.05. The relationship between Δn and the peak transmission wavelength of the cavity mode reveals that the designed Fabry-Perot cavity has a high refractive index sensitivity. The amount of average redshift in the peak transmission wavelength is evaluated as

approximately 15 nm per 0.05 increase in the refractive index. Finally, Fig. 2.27 shows that the alteration of the refractive index affects the value of peak power transmission negligibly and, therefore, the proposed structure can be a good candidate for practical applications that demand the enhancement of the electro-optic effect.

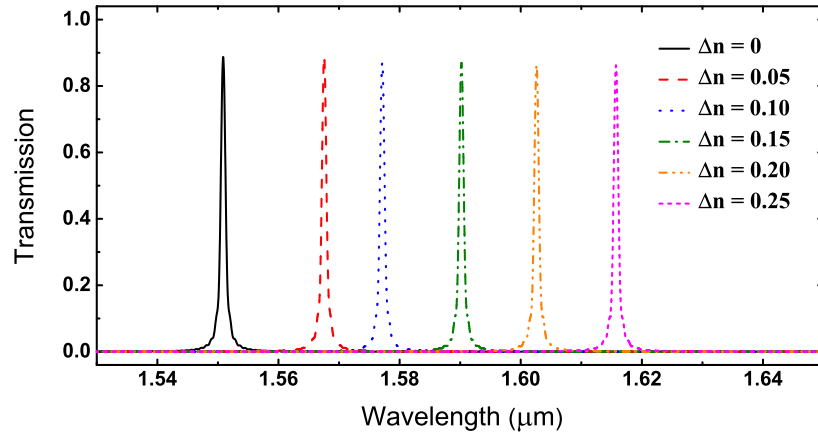


Figure 2.27: Evolution of the spectral position of the cavity resonance with respect to the change in the refractive index of the cavity region, Δn .

Next, we investigate the dual-polarization regime by modifying the geometrical parameters of the proposed geometry as stated previously, namely $a = 520$ nm and $r/a = 0.46$. For these parameters, we calculate $L_c = a - 2r = 41.6$ nm. The reasoning behind this modification is the enabling of the PBG regions of TM and TE waves to coincide. Correspondingly, Fig. 2.28 shows the transmission spectrum and the cavity resonances for TM and TE polarization states. From the viewpoint of practical applicability, it is advantageous that both of the cavity resonances yield similar peak power transmissions and Q s. The distinction between the spectral positions of the cavity resonance for TM and TE cases can be qualitatively explained by noting that different polarization states can encounter different cavity mirror reflectivity and, in addition, the propagation constants at the cavity resonance wavelength can differ for TM and TE waves, so that the phase gained for TM and TE waves in a single round-trip can be different.

It is known that L_c determines the spectral position and the inter-distance

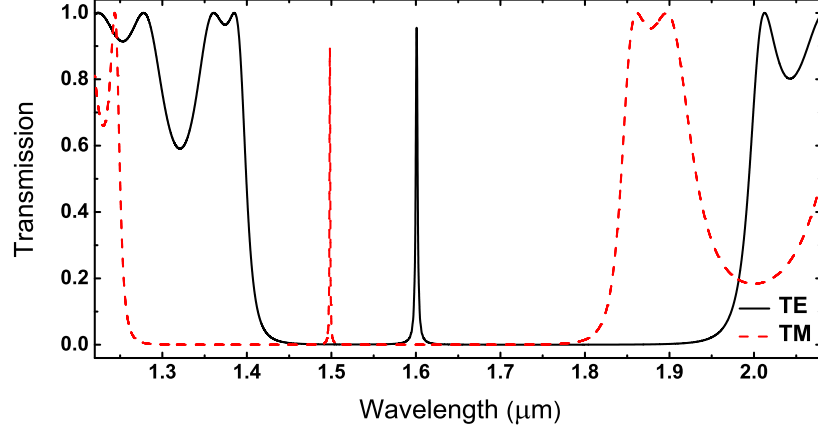


Figure 2.28: Power transmission spectrum of the proposed Fabry-Perot type cavity for the dual-polarization operation regime.

between the consecutive allowed modes of a Fabry-Perot cavity. For this particular Fabry-Perot cavity geometry, it is required for the observed cavity resonances to lie inside the PBG region of the corresponding PC structure. In the initial design, for which $a = 510$ nm, $r/a = 0.4$, and $L_c = 100$ nm, L_c has been carefully selected with the purpose of supporting only one cavity mode inside the PBG region. Nonetheless, by increasing L_c , we can expect the number of the supported cavity modes inside the PBG region to increase as a result of decreasing inter-distance between consecutive modes. For this reason, we set $L_c = 610$ nm by removing an additional column of holes neighbouring the initial cavity region. In this case, as predicted, the PBG region supports two distinct cavity modes at different wavelengths. The dispersion diagram and the transmission spectrum of the dual-mode design are shown in Fig. 2.29. In particular, Fig. 2.29(a) demonstrates the occurrence of the two cavity modes that are denoted by “Bands of interest”, compare to Fig. 2.24(a) where only one cavity mode is allowed inside the PBG region. The transmission spectrum given in Fig. 2.29(b) reveals the occurrence and the excitation of the two high transmission and high Q cavity modes. The normalized peak power transmission values are approximately given by 0.95 for the mode at $\lambda_0 = 1.36$ μm and 0.85 for the mode at $\lambda_0 = 1.65$ μm . Finally, we calculate the Q s by using Eq. 2.12 as 680 for the mode at $\lambda_0 = 1.36$ μm and 2870 for the mode at $\lambda_0 = 1.65$ μm .

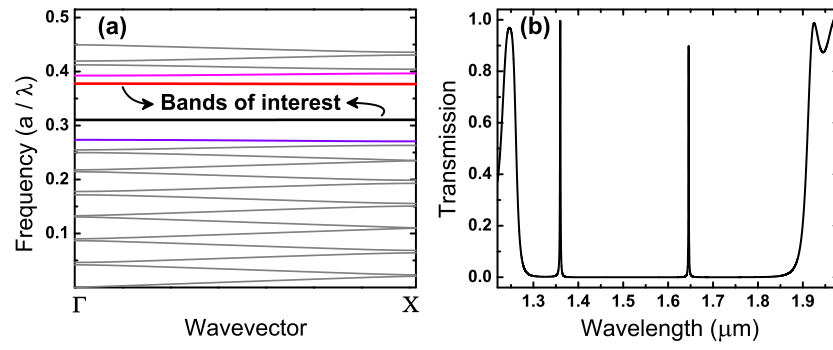


Figure 2.29: (a) Dispersion diagram and (b) power transmission spectrum of the dual-mode operation regime of the proposed Fabry-Perot type cavity.

Chapter 3

Circular Polarization Generation and Asymmetric Transmission Utilizing Chiral Metamaterials

3.1 Introduction

Due to the possibility of achieving negative refraction using a chiral route in metamaterial based applications, the chiral metamaterials (CMMs) has attracted significant interest [4]. CMMs are not identical to its mirror image, i.e., it cannot be brought into congruence with its mirror image unless it is lifted off the substrate [88]. The most significant electromagnetic property of CMMs can be explained as the existence of the cross coupling between the electric and magnetic fields of the wave. This cross coupling also leads to different transmission coefficients under right hand circularly polarized (RCP) and left hand circularly polarized (LCP) wave illumination [5]. This difference in the transmission coefficients also be utilized for the purpose achieving rotation in the polarization of incoming wave [89]. Besides negative refraction, CMMs can be utilized in several optical devices because of other intriguing properties such as giant optical activity and circular dichroism [90–95]. The strength of the cross coupling between the magnetic and electric fields are modelled with the chirality parameter, κ , for CMMs. In addition, the constitutive relations between the fields in a chiral

medium are given with [96]

$$\begin{pmatrix} \mathbf{D} \\ \mathbf{B} \end{pmatrix} = \begin{pmatrix} \epsilon_0 \epsilon & i\kappa/c_0 \\ -i\kappa/c_0 & \mu_0 \mu \end{pmatrix} \begin{pmatrix} \mathbf{E} \\ \mathbf{H} \end{pmatrix}. \quad (3.1)$$

For time harmonic waves in bianisotropic media, the Maxwell equations can be written as

$$\nabla \times \mathbf{E} = -jw\bar{v} \cdot \mathbf{E} - jw\bar{\mu} \cdot \mathbf{H}, \quad (3.2a)$$

$$\nabla \times \mathbf{H} = jw\bar{\epsilon} \cdot \mathbf{E} + jw\bar{\varsigma} \cdot \mathbf{H} + \mathbf{J}. \quad (3.2b)$$

where $\bar{v}$ and $\bar{\varsigma}$ are the coefficients for cross coupling between the electric and magnetic fields.

Lorentz reciprocity theorem claims that

$$\langle a, b \rangle = \langle b, a \rangle \quad (3.3)$$

which is valid for the bianisotropic media under special conditions and the isotropic media. In Eq. 3.3, a and b symbolizes the points that the transmitting and receiving electromagnetic fields are measured or calculated. $\langle a, b \rangle$ relates the reaction of the field $\mathbf{E}_a$, which is produced by the source $\mathbf{J}_a$, to the source $\mathbf{J}_b$, which produces the field $\mathbf{E}_b$. Likewise, $\langle b, a \rangle$ relates the reaction of the field $\mathbf{E}_b$, which is produced by the source $\mathbf{J}_b$, to the source $\mathbf{J}_a$, which produces the field $\mathbf{E}_a$. Therefore the $\langle a, b \rangle$ and $\langle b, a \rangle$ can be written as [97]

$$\langle a, b \rangle = \int_V \mathbf{E}_a \cdot \mathbf{J}_b dV, \quad (3.4a)$$

$$\langle b, a \rangle = \int_V \mathbf{E}_b \cdot \mathbf{J}_a dV. \quad (3.4b)$$

under the assumption that the magnetic current density is zero. As it is deduced from Eq. 3.4, the lorentz reciprocity theorem does hold for the bianisotropic media if the $\bar{\epsilon}$ and $\bar{\mu}$ tensors are diagonally symmetrical [97].

However, for bianisotropic media that has not symmetrical $\bar{\epsilon}$ and $\bar{\mu}$ tensors, the coefficients in the constituent equations should be changed to create a complementary medium for the purpose of achieving reciprocity equations. The changed coefficients can be written as $\bar{\epsilon} \rightarrow \bar{\epsilon}^T$, $\bar{\varsigma} \rightarrow -\bar{v}^T$, $\bar{v} \rightarrow -\bar{\varsigma}^T$, and $\bar{\mu} \rightarrow \bar{\mu}^T$. The reciprocity theorem can be rewritten with complementary coefficients and complementary fields for bianisotropic media.

If the linear transmission matrix for forward propagation is written as

$$T_l^f = \begin{pmatrix} T_{xx} & T_{xy} \\ T_{yx} & T_{yy} \end{pmatrix}, \quad (3.5)$$

the same matrix for backward propagation can be expressed as

$$T_l^b = \begin{pmatrix} T_{xx} & -T_{yx} \\ -T_{xy} & T_{yy} \end{pmatrix}. \quad (3.6)$$

Chiral metamaterials which consist of linear constituents hold this relation between T^f and T^b [98]. Similarly, the transmission matrices for forward and backward circular wave illuminations can be written as

$$T_c^f = \begin{pmatrix} T_{++} & T_{+-} \\ T_{-+} & T_{--} \end{pmatrix}, \quad (3.7)$$

and

$$T_c^b = \begin{pmatrix} T_{++} & T_{-+} \\ T_{+-} & T_{--} \end{pmatrix}. \quad (3.8)$$

The cross coupling terms in the linear transmission matrices for forward and backward illumination are interchanged. Thus, the asymmetric transmission is achieved by creating the difference between T_l^f and T_l^b . However, one should also consider the fact that the polarization states of the backward and forward waves should be similar.

For a x -polarized incoming wave, the difference between the transmitted powers between forward and backward illumination can be written as

$$\Delta_l = P^f - P^b = |T_{yx}|^2 - |T_{xy}|^2. \quad (3.9)$$

Likewise, the difference for the right hand circularly polarized wave case can be expressed as

$$\Delta_c = P^f - P^b = |T_{-+}|^2 - |T_{+-}|^2. \quad (3.10)$$

$\Delta_c \neq \Delta_l$ is the general condition for the asymmetric transmission. However, some attention is needed to the points that the degree of the asymmetric transmission may differ from one incoming wave polarization state to another. In addition, it is noteworthy to point that this asymmetric transmission effect is a reciprocal effect and achieved in a reciprocal medium.

3.2 Circular Polarizer based on Chiral Metamaterials

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In this section, we study the potential of an electrically thin, asymmetric chiral circular polarizer structure, which is composed of isotropic materials, in obtaining LCP and RCP waves using an x -polarized incident wave. Circular polarization is an important property for antenna applications [100–102], laser applications [103, 104], remote sensors, and liquid crystal displays [105, 106]. In the design process, we benefit from the metamaterial structure composed of two resonators, which are rotated by 90° [6] with respect to one another. The design proposed here is similar to those in Refs. [5, 107]; however, it is distinguished in that asymmetry is introduced by reducing the sizes of the electrically excited split ring resonators (SRRs) in order to modify the transmission at the x -polarization. By using this size reduction, the magnitudes of the transmitted electric fields in

the x and y directions are equalized at the resonance frequencies, with a 90° phase difference. The structure works as a left-hand and right-hand circular polarizer around the lower and higher resonance frequencies, respectively.

Circular polarization can be obtained from a linearly polarized incident wave by using a CMM, due to the different circular polarization transmission coefficients for LCP and RCP waves at the resonances. For instance, for the CMM structures given in Refs. [5, 90, 108], the transmission differences between RCP and LCP waves are 8 dB, 1518 dB, and 10 dB, respectively. The structure proposed in this Letter aims to introduce a larger magnitude difference between LCP and RCP waves at the output interface to obtain circular polarization.

Figure 3.1 shows the geometry of the CMM unit cell used in the simulations and experiments. The unit cell consists of four double-layered U-shaped SRRs placed on both sides of an FR-4 board with a relative permittivity of 4 and a dielectric loss tangent of 0.025. Copper that is $30\ \mu\text{m}$ thick is used for the metallic parts. SRRs are positioned so that they are rotated by 90° with respect to their neighbors.

We started the analysis with numerical simulations of the CMM structure using CST Microwave Studio (Computer Simulation Technology AG, Germany), which is a commercial software that is based on the finite integration method. The boundaries are selected to be periodic in the x and y directions and open in the z direction. The material is excited by a plane wave propagating in the direction of $\mathbf{k}(-z)$, as shown in Fig. 3.1, with the electric field in the x direction. The following geometric parameters are used in the simulations and experiments: $a_x = a_y = 15\ \text{mm}$, $s_1 = 6\ \text{mm}$, $s_2 = 4.2\ \text{mm}$, $w_1 = 0.7\ \text{mm}$, $w_2 = 0.5\ \text{mm}$, $d = 2.6\ \text{mm}$, and $t = 1.5\ \text{mm}$. The structure is electrically thin since t/λ is 0.024 and 0.03 for 5.1 GHz and 6.4 GHz, respectively. In addition, the periods in the x and y directions are also electrically small, since ax/λ is 0.255 at 5.1 GHz and 0.32 at 6.4 GHz.

For the experiments, we fabricated the structure with the dimension of 15 by 15 unit cells. The experiment is conducted using two standard horn antennas facing each other at a 50 cm distance. The structure is placed in the middle

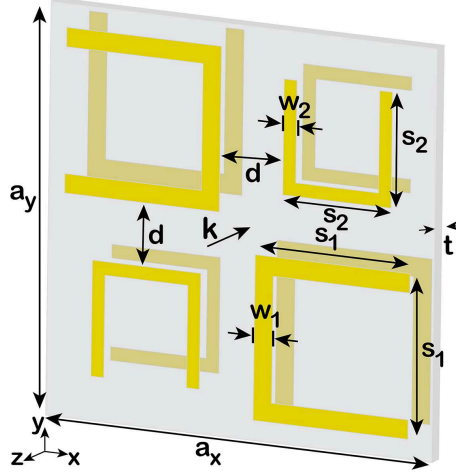


Figure 3.1: Geometry of the simulated and fabricated chiral metamaterial unit cell.

between the antennas. The transmission coefficient measurements are performed using an HP-8510C network analyzer (Agilent Technologies, USA). In order to obtain the transmission characteristics of the structure, the x and y components of the transmitted field are studied in the simulations and experiments in terms of the transmission coefficients T_{xx} and T_{yx} .

The ratio $|T_{yx}|/|T_{xx}|$ and the phase difference $\Phi(T_{yx}) - \Phi(T_{xx})$ are presented in Figs. 3.2(a) and 3.2(b). The resonance peaks in Fig. 3.2(a) correspond to the frequencies where the electromagnetic coupling between the top and bottom layers is strong, so that T_{yx} is large. According to the numerical results in Fig. 3.2(b), phase differences between T_{yx} and T_{xx} are (A) 89.8° and (B) -89.2° at 5.1 GHz and 6.4 GHz, respectively. In addition, at these frequencies, the ratios of the magnitudes of T_{yx} and T_{xx} are 1.03 and 0.994, respectively. These results prove that the transmitted waves are LCP and RCP waves at 5.1 GHz and 6.4 GHz, respectively. The experimental data shown in Fig. 3.2 is in good agreement with the numerical results.

Circular polarization transmission coefficients can be calculated from the linear transmission coefficients T_{xx} , T_{yx} , T_{xy} , and T_{yy} [11]. Since the metamaterial structure lacks C_4 symmetry because SRRs of different sizes are used, the four

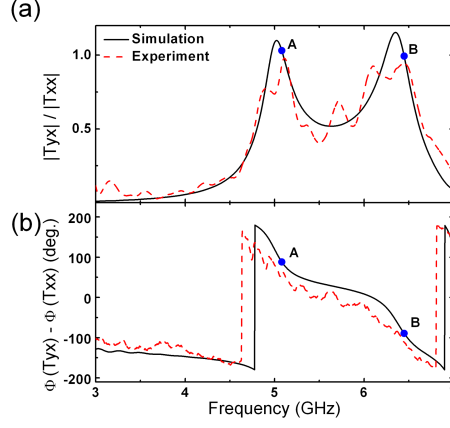


Figure 3.2: (a) Ratio of magnitudes of T_{yx} and T_{xx} , and (b) phase difference between T_{yx} and T_{xx} .

circular transmission coefficients, T_{++} , T_{+-} , T_{-+} , T_{--} should be calculated in order to completely characterize the response of the CMM. However, the structure provides circular polarization only if the incident wave is x -polarized. Therefore, E_y is assumed to be zero. Under this assumption, we define the circular transformation coefficients C^+ (RCP) and C^- (LCP). The conversion from the linear transmission coefficients to circular transformation coefficients becomes, simply, $C^\pm = T_{xx} \pm iT_{yx}$ [3]. It is noteworthy that the calculated transformation coefficients are only valid for an x -polarized incident field, since the structure lacks C_4 symmetry.

Numerical and experimental circular transformation coefficients for the LCP and RCP components are presented in Fig. 3.3. In this context, these parameters are defined as the magnitude of the RCP and LCP waves at the output interface due to an x -polarized incident field. In Fig. 3.3, 0 dB corresponds to the magnitude of the incident x -polarized electric field. The minimum contribution of the RCP component is observed at 5.1 GHz as 37 dB in Fig. 3.3(a) and, similarly, the minimum for the LCP component is obtained at 6.4 GHz as 43 dB. The experimental results provided in Fig. 3.3(b) are in good agreement with the numerical results.

It follows from the presented results that the circular polarizer feature of the studied CMM originates from the different circular polarization transformation

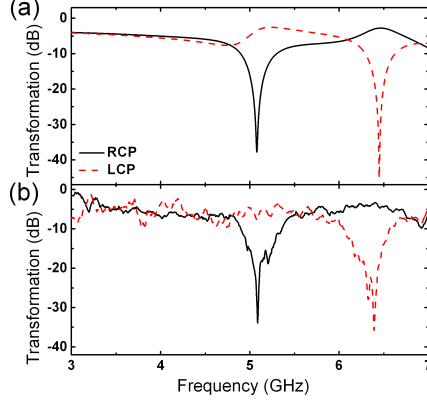


Figure 3.3: (a) Numerical and (b) experimental results for the circular transformation coefficients.

coefficients for the RCP and LCP components, due to an x -polarized incident wave. One of the circular polarizations is eliminated at the resonance frequencies, whereas the other polarization is transmitted with a small loss. As a result, at 5.1 GHz, the transmitted wave is LCP. Similarly, at 6.4 GHz, the RCP wave is transmitted.

As mentioned above, the proposed structure is similar to the structure studied in Ref. [3]. In this work, the sizes of all the SRRs are equal (geometric parameters given as s_1 and w_1), which means that at the frequencies of minimum RCP and LCP wave transmission, all the SRRs are in the vicinity of resonance. As a result, T_{xx} is around 20 dB near the resonance frequencies, while T_{yx} is obtained to be 6 dB at those frequencies. Hence, since T_{xx} and T_{yx} are significantly different, the resulting polarization is elliptical.

In the proposed design, the sizes of the SRR pairs, which are shown in Fig. 3.1, are reduced in order to increase T_{xx} . Numerical results reveal that, in this case, T_{xx} is increased (to approximately 6.5 dB around the resonance frequencies) by using the effect exerted by the smaller SRR pairs being out of resonance. In order to understand the nature of the resonances occurring at 5.1 GHz and 6.4 GHz, the surface current distribution is studied numerically. The directions of the induced surface currents at the two resonances are shown in Fig. 3.4. Smaller arrows are used for the smaller SRR pairs in order to indicate that the

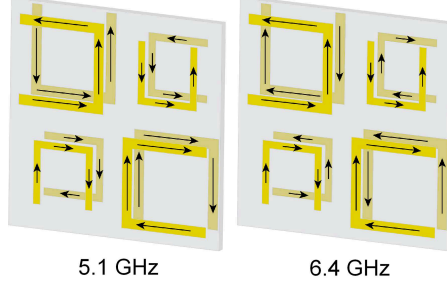


Figure 3.4: Directions of the surface currents on the SRRs due to the incident electric field polarized in the x -direction, at 5.1 GHz and 6.4 GHz

magnitude of the currents on those SRRs are smaller when compared to the large SRR pairs as an effect of being out of resonance. As a result, transmission of the x -polarization, T_{xx} , is increased, and the geometrical parameters are carefully optimized such that at the frequencies where $|\Phi(T_{yx}) - \Phi(T_{xx})| = 90^\circ$, $|T_{yx}|/|T_{xx}|$ is approximately equal to unity. According to the simulations performed for a y -polarized incident field, the transmitted wave is elliptical, with $|T_{xy}|/|T_{yy}| = 0.44$ and $|\Phi(T_{yx}) - \Phi(T_{xx})| = 105^\circ$ at 5.1 GHz. At 6.4 GHz, $|T_{xy}|/|T_{yy}| = 0.39$ and $|\Phi(T_{yx}) - \Phi(T_{xx})| = 5^\circ$, and these values do not correspond to circular polarization.

For the mutually 90° rotated SRR pairs, the resonance levels are determined in line with the longitudinal magnetic dipole to magnetic dipole coupling [9]. Numerical study shows that at the lower resonance frequency, the surface currents on the twisted SRR pairs are in the same direction, leading to parallel magnetic dipole moments for the twisted pairs. In contrast, at the higher resonance frequency, the surface currents on the twisted SRR pairs are antiparallel, which results in antiparallel magnetic dipole moments.

The major advantage of the structure is being electrically very thin. Another advantage is that the structure is planar and easy to fabricate. On the other hand, narrowband operating frequency can be a drawback, depending on the application. However, broadband response may be achieved by extending the design to three dimensions.

3.3 Polarization Angle Dependent Rotation

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It was recently shown that a carefully designed metamaterial reflector can be utilized in order to rotate the polarization of a wave [110]. However, due to the reflection based operation, the reflected wave and the incoming wave interfere, which causes the design to be inconvenient for practical applications. Consequently, transmission based polarization rotators have been proposed [111–113]. Those structures exhibit remarkable polarization rotation efficiencies in the vicinity of their operation frequencies. However, for rotationally asymmetric structures operating as polarization rotators, transmitted waves are elliptically polarized for certain polarization angles [113]. On the other hand, for rotationally symmetric polarization rotators, the amount of rotation is independent of the polarization angle of the incident wave [111]. In order to overcome these limitations, in the present section, we study the potential of an electrically thin polarization rotator, whose rotatory power is polarization dependent, i.e., the linear transmission coefficients are dependent on the polarization angle of the incoming wave. As a consequence, the proposed chiral metamaterial (CMM) also exhibits an asymmetric transmission of linearly polarized waves at 6.2 GHz. Asymmetric transmission of electromagnetic waves is an optical phenomenon that has been studied for different structures in several studies [114–117]. Using the theoretical calculations, we obtained four polarization angles for which the transmission of the structure is symmetric. In order to have linearly polarized eigenwaves for the CMM, we optimized the phases of the transmitted waves, so that any linearly polarized wave can be transmitted as a linearly polarized wave.

Such a design can be utilized in order to dynamically control the polarization-mode dispersion (PMD) in optical communication systems [118], modify the polarization of a laser output arbitrarily and dynamically, obtain an arbitrary linear polarization from a steady antenna and scan certain polarization directions in order to characterize the response of a certain material to different incident polarizations. The proposed CMM can be integrated into a rotating stage to easily

obtain the desired polarization state by simple rotation of the stage with respect to the analytical description provided in the subsequent sections. In addition, the proposed CMM can be stacked while avoiding the coupling effects [119] in order to obtain a larger rotation than one layer can provide.

3.3.1 Proposed Geometry

In order to obtain asymmetric transmission and incident wave polarization angle dependent polarization rotation, the CMM structure that is composed of four U-shaped split ring resonator (SRR) pairs is a good candidate [5, 107, 120]. It is possible to introduce asymmetry by altering the dimensions of certain SRR pairs. As a result, illuminating the structure by an incident wave polarized in a certain direction is not equivalent to illuminating the structure from the opposite direction by a wave with the same polarization state. The introduction of such an asymmetry also breaks the fourfold rotational C_4 symmetry. Thus, circularly polarized waves are not eigenwaves of the proposed design. This fact can be used as an advantage for the purpose of creating polarization angle dependent rotation. As a consequence of the broken C_4 symmetry, orthogonal electric field components (x and y components in this context) of an incident wave encounter different transmission coefficients, both in terms of magnitude and phase. This discrepancy between the transmission coefficients can be optimized to yield a linearly polarized transmitted wave whose polarization angle is a function of the polarization angle of the linearly polarized incident wave. In this study, we demonstrate the results of this optimization numerically and experimentally. Afterwards, a closed form relationship is derived that relates the polarization rotation introduced by the CMM to the polarization angle of the incident wave.

The unit cell of the proposed CMM structure is depicted in Fig. 3.5. It is an asymmetric version of the structure that has been studied in Refs. [5, 107, 120]. In a previous study, it has been demonstrated that a similar variant of this design operates as a circular polarizer for incident waves that are linearly polarized in the x -direction [99]. For the circular polarizer, the coupling between the meta-atoms has been optimized in order to provide maximum wave ellipticity in the vicinity

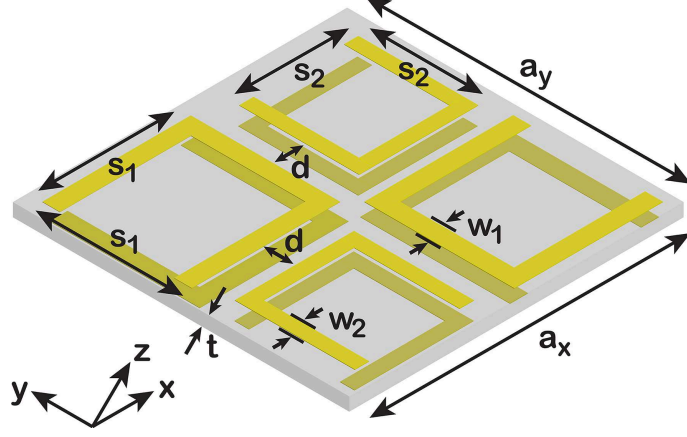


Figure 3.5: Geometry of the simulated and fabricated chiral metamaterial unit cell.

of the resonance frequencies as to ensure that the transmitted wave is circularly polarized. Conversely, the design goal in this study is minimizing the transmitted wave ellipticity as to rotate a linearly polarized incident wave without distorting the linearity of the polarization.

The following geometrical parameters for the unit cell are used in the simulations and experiments: $a_x = a_y = 13.6$ mm, $s_1 = 6$ mm, $s_2 = 4.8$ mm, $w_1 = 0.7$ mm, $w_2 = 0.55$ mm, $d = 1.4$ mm, and $t = 1.5$ mm. A FR-4 board with a relative permittivity of 4 and a dielectric loss tangent of 0.025 is utilized as the substrate. For the metallic parts, copper that is $30\text{ }\mu\text{m}$ thick is used. As it will be presented subsequently, the operating frequency of the CMM is 6.2 GHz. At this frequency, the structure is electrically thin with $t/\lambda \approx 0.031$. In addition, the periodicity in the transverse plane is electrically small at 6.2 GHz, since $a_x = a_y$ corresponds to 0.281λ .

3.3.2 Numerical Results

We initiated the analysis with numerical simulations of the proposed CMM unit cell using CST Microwave Studio (Computer Simulation Technology AG, Germany), which is a commercially available software that utilizes the finite integration method. During the simulations, boundary conditions along the x and

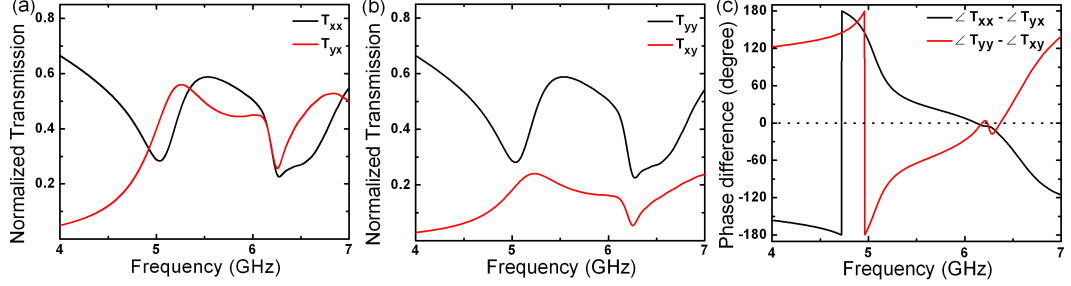


Figure 3.6: Magnitudes of the linear transmission coefficients when the CMM is illuminated by (a) x -polarized and (b) y -polarized incident waves. (c) The mutual phase differences between the transmission coefficients.

y directions are adjusted to be periodic in order to obtain periodicity in the transverse plane. The boundary condition along the z direction is selected to be absorbing. In order to characterize the response of the CMM, the structure is illuminated by plane waves propagating in the z direction. Assuming the linearity of the CMM, linear transmission coefficients T_{xx} and T_{yx} are obtained from the simulations when the incident wave is x -polarized. Similarly, for a y -polarized incident wave, linear transmission coefficients T_{xy} and T_{yy} are obtained. The magnitudes of the four linear transmission coefficients are shown in Figs. 3.6(a) and 3.6(b). The mutual phase differences between T_{xx} and T_{yx} , and T_{yy} and T_{xy} are shown in Fig. 3.6(c).

Using Fig. 3.6, it is observed that the CMM creates an electric field that is orthogonal to the incident field, since the cross-coupling terms T_{xy} and T_{yx} are non-zero. In addition, the simulation results reveal that T_{xx} and T_{yy} are equal in terms of magnitude and phase (phase not shown here). In other words, the transmitted x -polarization due to an x -polarized incidence is equal in terms of magnitude and phase to the transmitted y -polarization due to a y -polarized incidence. Conversely, it is noticed that the transmitted y -polarization due to an x -polarized incidence is strongly different than the transmission of the x -polarization due to a y -polarized incidence. Using these observations, it can be deduced that this CMM configuration creates polarization angle dependent chirality.

The asymmetric transmission of the structure is also explicable using the

linear transmission coefficients. Due to the geometry, illuminating the structure by a plane wave polarized in the y -direction and propagating in the z direction is equivalent to illumination by a plane wave polarized in the x -direction and propagating in the $+z$ direction. The two waves encounter the same transmission coefficients with respect to their polarization direction, e.g., T_{xx} becomes T_{yy} and T_{yx} becomes T_{xy} . Thus, the transmission of the CMM is asymmetric.

Figure 3.6(c) shows the mutual phase differences between T_{xx} and T_{yx} , and T_{yy} and T_{xy} . At 6.2 GHz, which is the operating frequency of the device, the mutual phase differences are approximately equal to 0° , in turn demonstrating that at this frequency optical activity is observed. Although the mutual phase differences are 0° , the phases of all the elements must be equal in order to avoid the transmission of elliptically polarized wave. We investigated the numerical results and observed that at 6.2 GHz, the phases of all linear transmission coefficients are equalized. As a result, combining the two orthogonal cases, linearly polarized waves are eigenwaves of the CMM at 6.2 GHz and are transmitted with a polarization rotation.

Thereafter, circular transformation coefficients [99] are calculated using

$$C_x^\pm = T_{xx} \mp iT_{yx}, \quad C_y^\pm = T_{yy} \pm iT_{xy} \quad (3.11)$$

for the two separate incident polarizations. In order to characterize the polarization of the transmitted wave, the polarization azimuth rotation angle θ is calculated using the formula

$$\theta_{x,y} = \arg(C_{x,y}^+) - \arg(C_{x,y}^-) \quad (3.12)$$

and the ellipticity of the transmitted wave, which is defined as

$$\eta_{x,y} = \tan^{-1} \left(\frac{|C_{x,y}^+| - |C_{x,y}^-|}{|C_{x,y}^+| + |C_{x,y}^-|} \right) \quad (3.13)$$

is determined. The results retrieved from the simulations for the polarization rotations and the transmitted wave ellipticities are shown in Fig. 3.7. In Fig. 3.7(a), it is observed that the ellipticities of the transmitted waves for x -polarized

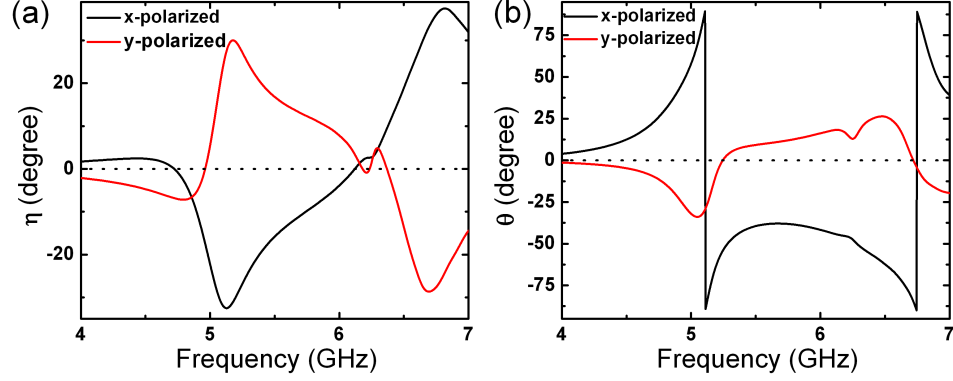


Figure 3.7: (a) Ellipticities and (b) polarization azimuth rotation angles of the transmitted waves for x -polarized and y -polarized illumination.

and y -polarized incident fields are approximately equal to zero at 6.2 GHz, which corresponds to pure optical activity at this frequency. The corresponding azimuth polarization rotation angles for x -polarized and y -polarized incident fields are shown in Fig. 3.7(b).

According to the results given in Fig. 3.7(b), an incident wave that is linearly polarized in the x direction is rotated by 46° at 6.2 GHz, whereas a y -polarized incident wave is rotated by 15° at the same frequency. Due to different rotations for x and y polarizations, each incident polarization angle encounters a different rotation. The relationship between the incident polarization angle and the resulting polarization rotation can be derived by performing simple geometrical calculations based on the rotation values provided above. However, for simplicity, we will employ the transfer matrix formulation subsequently for determining this relation.

3.3.3 Experimental Results

In order to characterize the behavior of the CMM and examine the validity of the simulation results, we performed experiments. We fabricated the structure with a dimension of 16 by 16 unit cells. The experiment is conducted using two standard horn antennas facing each other at a 60 cm distance. The structure is placed in the middle between the antennas. The transmission coefficients are

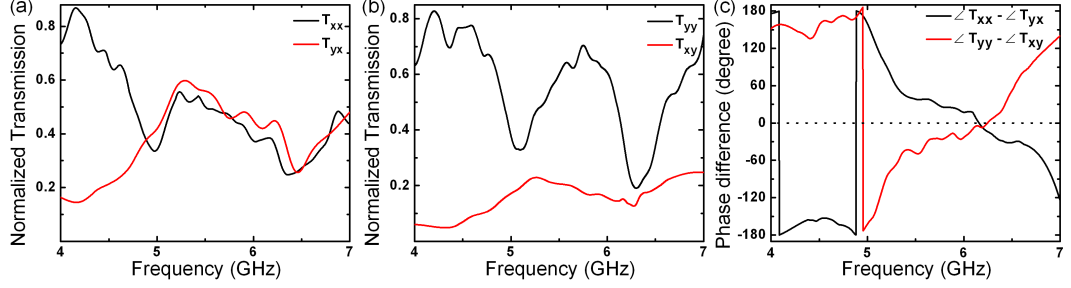


Figure 3.8: (a) Experimental magnitudes of T_{xx} and T_{yx} , and (b) T_{xy} and T_{yy} . (c) Mutual phase differences obtained from the experiments.

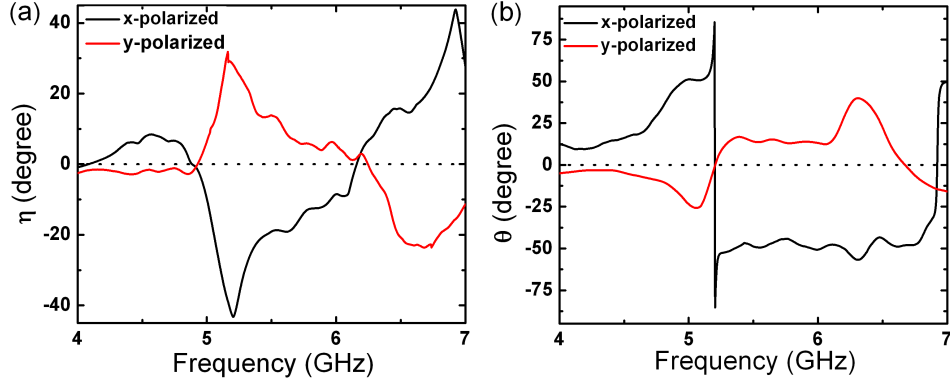


Figure 3.9: (a) Experimental ellipticities and (b) polarization rotations of the transmitted waves due to x - and y -polarized incident waves.

measured using an HP-8510C network analyzer (Agilent Technologies, USA).

The x and y components of the transmitted fields due to x - and y -polarized incident waves are measured for characterization. Phase information is also obtained from the network analyzer in order to calculate the rotation and ellipticity for the transmitted waves. The experimental magnitudes of T_{xx} , T_{xy} , T_{yx} , and T_{yy} are shown in Fig. 3.8, as well as the phase differences between T_{xx} and T_{yx} , and T_{yy} and T_{xy} . In Fig. 3.6(c), the phase differences are approximately equal to 0° at 6.2 GHz. For the experiment results, at 6.2 GHz we obtain -6° for the phase difference between T_{xx} and T_{yx} and -7° for the T_{yy} and T_{xy} case. Throughout the scanned frequency range, the experiment results agree closely with the simulations.

In addition to the transmission results, we extracted the ellipticity and polarization rotation information from the measurement data. These results are shown in Fig. 3.9, which is the experimental analog of Fig. 3.7. At 6.2 GHz, the ellipticity of a transmitted wave due to an x -polarized incidence is 4.2° . Similarly, for a y -polarized incident wave, the ellipticity of the transmitted wave is 2.9° . These results are very close to the numerical results, where both ellipticities are approximately equal to 0° . Subsequently, we examine the experimental results for polarization rotation. Experiment results indicate that the amount of rotation is 49° for an x -polarized incident wave at 6.2 GHz. This value differs by 3° from the simulation results, which indicates a good agreement. However, we obtain a 26° rotation for a y -polarized wave at the same frequency. Numerical results provided 15° rotation for a y -polarized field, which differs by 11° from the experiment result. These discrepancies can be attributed to the inaccuracies in the fabrication stage, multi-reflections in the experiment setup, diffraction effects from the sharp edges of the CMM structure, probable misalignments, and non-zero cross-polarization response of the antennas. In addition, the resonance frequencies slightly shift to higher frequencies in the experiments. The possible reasons are the variance of the dielectric permittivity of the FR-4, experiment inaccuracies, and simulation inaccuracies, i.e., mesh size affects the resonance frequencies. Overall, we conclude that the agreement between the numerical and experiment results is good and the operation of the CMM is verified experimentally.

3.3.4 Formulation

Assuming that the CMM structure is a two-input and two-output system, where the fields associated with the x and y directions represent the two inputs and the two outputs of the system, a transmission matrix T , with the elements T_{xx} , T_{xy} , T_{yx} , and T_{yy} can be defined so that the following relation holds

$$\begin{pmatrix} E_{xd} \\ E_{yd} \end{pmatrix} = \begin{pmatrix} T_{xx} & T_{xy} \\ T_{yx} & T_{yy} \end{pmatrix} \begin{pmatrix} E_{x0} \\ E_{y0} \end{pmatrix}, \quad (3.14)$$

where E_{xd} and E_{yd} are the transmitted fields along the x and y directions, respectively. Similarly, E_{x0} and E_{y0} represent the electric field components of the incident field along the x and y directions. In linear equation form, Eq. 3.14 can be rewritten as the follows:

$$E_{xd} = T_{xx}E_{x0} + T_{xy}E_{y0}, \quad (3.15a)$$

$$E_{yd} = T_{yx}E_{x0} + T_{yy}E_{y0}. \quad (3.15b)$$

Subsequent to the calculation of transmitted x and y components for an arbitrary input to the system, the polarization angle of the transmitted wave is calculated as

$$\phi = \tan^{-1} \left(\frac{T_{yx}E_{x0} + T_{yy}E_{y0}}{T_{xx}E_{x0} + T_{xy}E_{y0}} \right). \quad (3.16)$$

In order to relate the transmitted polarization angle to the incident polarization angle, without loss of generality, we assume that the magnitude of the incoming wave is unity in all cases. Under this assumption, Eq. 3.16 is modified as

$$\phi = \tan^{-1} \left(\frac{T_{yx} \cos \varphi + T_{yy} \sin \varphi}{T_{xx} \cos \varphi + T_{xy} \sin \varphi} \right), \quad (3.17)$$

where φ denotes the polarization angle of the incident wave. It is noteworthy that the inverse tangent is a multi-valued function that requires special attention. The quadrant where ϕ lies depends on the signs of the numerator and denominator of Eq. 3.17. Finally, the corresponding polarization rotation is defined as

$$\theta = \varphi - \phi. \quad (3.18)$$

Then, to calculate the polarization rotation introduced by the CMM at 6.2 GHz, we constructed the transmission matrices using the simulation results for waves propagating in the z and $+z$ directions. Thereby, the asymmetric transmission of the design would be demonstrated simultaneously with the incident polarization angle dependent polarization rotation. For an incident wave propagating in the z direction, the elements of the transmission matrix are given as $T_{xx} = 0.3568$, $T_{xy} = 0.1104$, $T_{yx} = 0.3599$, and $T_{yy} = 0.3568$. In general, these elements are complex quantities carrying phase information. However, in this case we omit

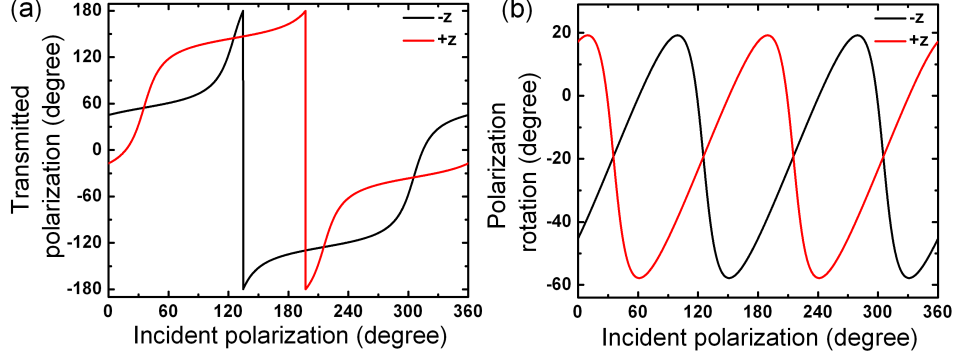


Figure 3.10: (a) Polarization angles of the transmitted linearly polarized waves with respect to the polarization angle of the incident wave, at 6.2 GHz, for the $-z$ and $+z$ propagating waves. (b) Introduced polarization rotation to the $-z$ and $+z$ propagating waves with respect to the incident polarization angle, at 6.2 GHz.

the phases of the elements, since the simulation and experiment results prove that all the elements are in-phase. Thus, using only the magnitude information is sufficient. Similarly, for a $+z$ propagating wave, the transmission matrix elements are given as $T_{xx} = 0.3568$, $T_{xy} = 0.3599$, $T_{yx} = 0.1104$, and $T_{yy} = 0.3568$. The two transmission matrices demonstrate the asymmetric transmission of the structure, since T_{xy} and T_{yx} values are not equal for the $+z$ and $-z$ cases. Figure 3.10(a) shows the calculated polarization angle of the transmitted wave, using Eq. 3.17, due to incident waves linearly polarized from 0° to 360° and propagating in the $-z$ and $+z$ directions. The corresponding polarization rotation calculated using Eq. 3.18 is presented in Fig. 3.10(b).

As it is apparent in Fig. 3.10(a), at several angles, $-z$ and $+z$ propagating waves are transmitted symmetrically, with the same polarization. Equating ϕ given by Eq. 3.17 for the above-mentioned transmission matrices, we obtain the following transcendental equation to calculate these angles:

$$\tan^{-1} \left(\frac{0.3599 \cos \varphi + 0.3568 \sin \varphi}{0.3568 \cos \varphi + 0.1104 \sin \varphi} \right) = \tan^{-1} \left(\frac{-0.1104 \cos \varphi + 0.3568 \sin \varphi}{0.3568 \cos \varphi - 0.3599 \sin \varphi} \right). \quad (3.19)$$

Numerical solution for Eq. 3.19 yields $\varphi = 35.3^\circ, 125.3^\circ, 215.3^\circ$, and 305.3° . Figure 3.10(b) presents θ as a function of ϕ , which is calculated using Eqs. 3.17 and 3.18.

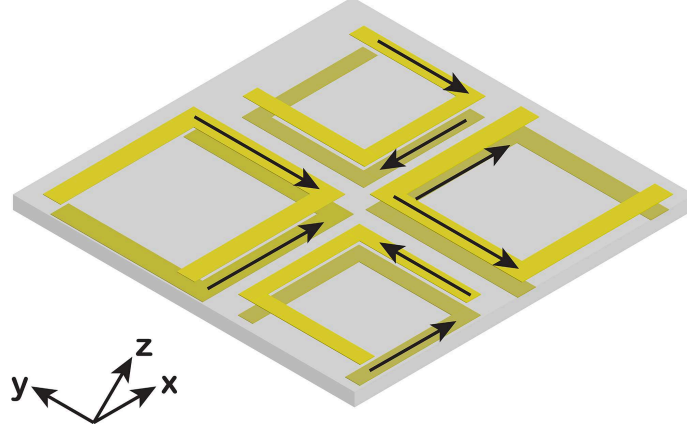


Figure 3.11: Directions of the induced surface currents due to x -polarized plane waves propagating in the $-z$ and $+z$ directions at 6.2 GHz.

It is seen that the proposed CMM is reciprocal, i.e., it shows the same θ for the waves propagating in the $-z$ and $+z$ directions, which are linearly polarized at the angles those are being equal to the solutions of Eq. 3.19. In turn, for polarization angles, which do not satisfy Eq. 3.19, the transmission is asymmetric, i.e., different for waves propagating in the $-z$ and $+z$ directions. In both cases, linearly polarized waves are eigenwaves for both directions.

3.3.5 Surface Currents

In order to explain the asymmetric transmission, we investigated the induced surface currents at 6.2 GHz, when the CMM is excited by x -polarized waves propagating in the $-z$ and $+z$ directions. The simulation results indicate that the directions of the induced surface currents are identical for both excitations. The directions of the surface currents are shown in Fig. 3.11.

As the next step, we have simulated a single SRR pair, where one SRR is rotated by 90° with respect to the other. We observed that coupling from the x polarization to the y polarization decreases when the electric field vector of the incident wave is parallel to the slit of the SRR that is closer to the source. However, rotating the pair by 90° , while keeping the electric field vector direction constant, does not change the transmission of the x polarization. In the case

where each SRR pair has the same dimensions, rotation does not affect the transmission coefficients. In the asymmetric case, rotation decreases coupling from x polarization to y -polarization, whereas the transmission of the x -polarization is not changed. When the structure is rotated, smaller SRR pairs are not at the resonance since their resonance frequencies are larger. In addition, according to the simulation results, larger SRR pairs begin to produce less y polarization compared to the previous case. Thus, illuminating the structure by an x -polarized wave is not equivalent to illumination by a y -polarized wave. It should also be denoted that due to the geometry of the CMM, illuminating the structure by a y -polarized wave propagating in the $-z$ direction is equivalent to illuminating it by an x -polarized wave propagating in the $+z$ direction. Hence, as a result of different transmission coefficients for the x and y polarized waves, the transmission of the structure is asymmetric.

3.4 Diodelike Asymmetric Transmission

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Simultaneous breaking of time reversal and spatial inversion symmetries has been considered as necessary for nonreciprocal transmission in volumetric structures, enabling the obtaining of $w(\mathbf{k}) \neq w(-\mathbf{k})$ while involving anisotropic constituents [122, 123]. The possibility of nonreciprocal light propagation owing to the symmetry of the parity-time operator has also been studied, for example, in two-channel structures with a properly chosen symmetry of the complex refractive index [124]. At the same time, there were attempts to achieve asymmetric but still reciprocal transmission while using conventional isotropic, linear and low-loss/lossless materials. Among them, the asymmetric transmission of linearly polarized waves in diffractive nonsymmetrical volumetric gratings based on photonic [54] and sonic [62] crystals, or made of ultralow-index materials [125], and in nonsymmetrical metallic gratings supporting surface plasmons [113, 126] should be mentioned. The asymmetric transmission of circularly polarized waves

has been demonstrated at normal incidence in thin structures composed of meta-atoms [115–117,127]. A thin structure composed of three-dimensional meta-atoms without any rotational symmetry has been suggested that allows asymmetric transmission for an arbitrary, also linear, polarization of the incident electromagnetic wave [114].

In this section, we propose a new, ultrathin, three-layer structure made of isotropic and linear materials, which enables a diode-like asymmetric transmission of certain linearly polarized, normally incident waves. Nearly total transmission is demonstrated in cross-polarization regime in one direction, while the transmission in the opposite direction at the same polarization of the incident wave is blocked. Possible applications include microwave and optical isolation, integrated photonic circuits, ultrafast information processing, and optical interconnects [128]. To our knowledge, diodelike asymmetric transmission of linearly polarized waves at normal incidence has not been observed before. In fact, it requires the diagonal components and one of the off-diagonal components of the Jones matrix to be zero, while the only non-zero element must be approximately unity for maximum efficiency. These requirements are not easily achievable, also in a chiral structure, without the support of additional physical mechanisms, in our case the utilization of elliptical eigenstates and the tunneling of electromagnetic waves.

3.4.1 General Idea

The exploited physical mechanism is based on the combination of the coupling of electric and magnetic fields in mutually rotated split ring resonator (SRR) pairs [129] and electromagnetic wave tunneling through a metallic subwavelength mesh [130]. Tunneling is an exciting phenomenon that utilizes the potential of nonpropagating waves in transmission. Unlike the previous experience of utilizing the tunneling effect [130], we perform an ABC stacking type instead of an ABA type. Layers A and C are, in fact, the same layers which are rotated by $\pm\pi/2$ with respect to each other. Layer B is a subwavelength mesh that acts as a negative permittivity medium throughout the interested frequency range independent of the polarization. Because of the inclusion of layer B, the phase difference between

the transmission coefficients of the eigenwaves of the ABC stack may be properly optimized, yielding a strong modification in the transmission, i.e. enhancement of one of the two off-diagonal (cross-polarized) components of the Jones matrix (T_{xy}) and suppression of all the others (T_{xx} , T_{xy} , T_{yy}). As a result, strongly asymmetric, diode-like transmission takes place for the incident waves that are linearly polarized in the x or y directions. Similar to Refs. [114–117, 127], asymmetric transmission in the suggested structure is a completely reciprocal phenomenon.

One of the main ideas behind the proposed design is choosing two layers, which do not exhibit C_4 symmetry, as Layer A and C. In addition, these layers must be chosen such that they exhibit strong optical activity when brought together. Accordingly, we choose the separate layers of the metamaterial structure given in Ref. [109] as Layer A and C.

Adopting the Jones matrix formalism, we relate the electric field amplitudes of incident (I_x , I_y) and transmitted (T_x , T_y) fields [114], for the $+z$ (forward) propagation as follows:

$$\begin{pmatrix} T_x^f \\ T_y^f \end{pmatrix} = \begin{pmatrix} T_{xx}^f & T_{xy}^f \\ T_{yx}^f & T_{yy}^f \end{pmatrix} \begin{pmatrix} I_x^f \\ I_y^f \end{pmatrix} = \begin{pmatrix} \delta & \gamma \\ \beta & \delta \end{pmatrix} \begin{pmatrix} I_x^f \\ I_y^f \end{pmatrix}. \quad (3.20)$$

Then, for the $-z$ (backward) propagation, we have

$$\begin{pmatrix} T_x^b \\ T_y^b \end{pmatrix} = \begin{pmatrix} \delta & -\beta \\ -\gamma & \delta \end{pmatrix} \begin{pmatrix} I_x^b \\ I_y^b \end{pmatrix}. \quad (3.21)$$

According to Eqs. 3.20 and 3.21, if $I_y^f = I_y^b = 0$ (x -polarized incident wave) and $\delta = 0$, all transmitted energy corresponds to the nondiagonal (cross-polarized) elements, leading to a strongly asymmetric transmission if these elements are significantly different in magnitude. Diodelike asymmetric transmission also occurs when $I_x^f = I_x^b = 0$. However, in this case, the direction of the strong transmission should be reversed. Next, we will demonstrate that almost all the incident wave energy can be converted to that of the cross-polarized transmitted wave for one of the incidence directions, while it is largely reflected for the opposite direction, so that a diodelike behavior occurs. In the ideal case, one must obtain $\gamma = 1$ and

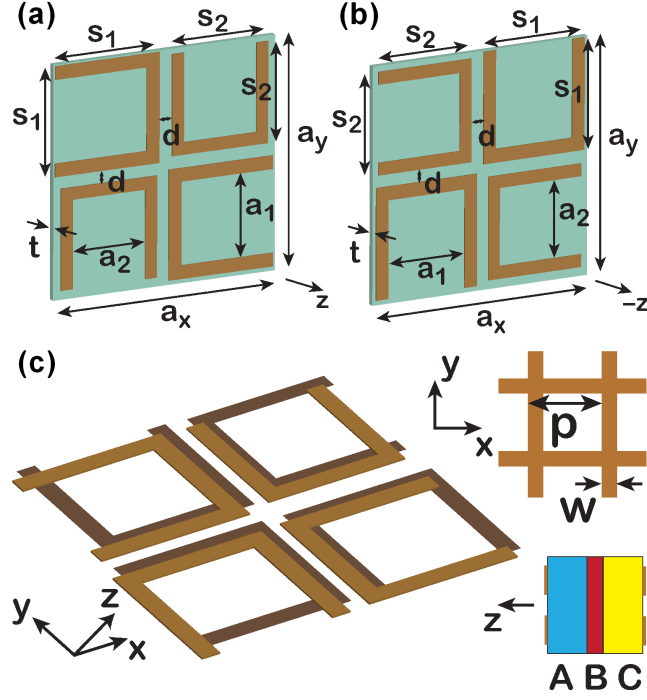


Figure 3.12: Geometries of (a) layer A and (b) layer C. (c) Arrangement of A and C with respect to each other along with the mesh geometry and the stacking scheme.

$\beta = 0$. For the purposes of this study, we seek for the regimes with $\gamma \approx 1$ and $\beta \ll \gamma$.

3.4.2 Method

The unit cell of the proposed structure that is assumed to be periodic in the transverse plane is depicted in Fig. 3.12. It is composed of three distinct layers that are referred to as Layer A, B and C. Fig. 3.12(a) and 3.12(b) shows Layer A and Layer C, respectively. They both contain four U-shaped SRRs and show the lack of C_4 symmetry due to the involvement of the SRRs of different dimensions. The arrangement of layers A and C, and the scheme of stacking are presented in Fig. 3.12(c) together with the geometry of Layer B. In the arrangement of A and C, the SRRs of the same dimensions are paired together. Layer B is positioned symmetrically between A and C. The geometric parameters for Layer B are given

by $w = 0.5$ mm and $p = 2.56$ mm. The geometric parameters of both layers A and C are given by $s_1 = 6$ mm, $a_1 = 4.6$ mm, $s_2 = 5.52$ mm, $a_2 = 4.23$ mm, $d = 0.64$ mm, $t = 1$ mm and $a_x = a_y = 12.8$ mm. Teflon dielectric substrates with a relative dielectric permittivity of 2.1 and a loss tangent of 0.0002 are utilized for the layers A and C. Layer B is considered to be a stand-alone metallic structure with a thickness of 20 μ m. The metal utilized for the structure of the SRRs and the mesh is copper with a 20 μ m thickness. The total thickness of the structure is 2.06 mm, which corresponds to $\lambda/21$ at 7 GHz while $a_x/\lambda \approx 0.3$.

We started the analysis with numerical simulations using CST Microwave Studio (Computer Simulation Technology, AG, Germany), by using periodic boundary conditions along the x and y directions and an open boundary condition for the z direction, while the excitation source is an x -polarized plane wave. Then, in order to experimentally verify the diodelike transmission, we fabricated the structure with the dimension of 23 by 23 unit cells. The experiment was conducted using two standard horn antennas facing each other at a 50 cm distance. The sample is positioned in the middle between the antennas. The transmission coefficient measurements are performed using an HP-8510C network analyzer (Agilent Technologies, USA). For the experimental characterization of the structure, the transmitted fields for the forward and backward propagating incident waves are studied in terms of the transmission coefficients T_{xx} and T_{yx} .

3.4.3 Results and Discussion

As the first step, we studied the responses of the single layers, A, B, and C. Figures 3.13(a) and 3.13(b) show $|T_{xx}|$ and $|T_{yx}|$, respectively. We see in Fig. 3.13(a) that Layer A is resonant at 7.7 GHz whereas the resonance frequency is 7 GHz for Layer C. Field distributions (not shown) reveal that at the resonance frequency, the responses of the layers are governed by the SRRs that have slits parallel to the direction of the electric field. In the case of Layer C, such SRRs are larger which implies that their resonance frequency is lower. Figure 3.13(b) shows that the $|T_{yx}|$ values of layers A and C show maxima at around 8.5 GHz. However, $|T_{yx}|$ does not exceed 0.06, indicating that the coupling between electric

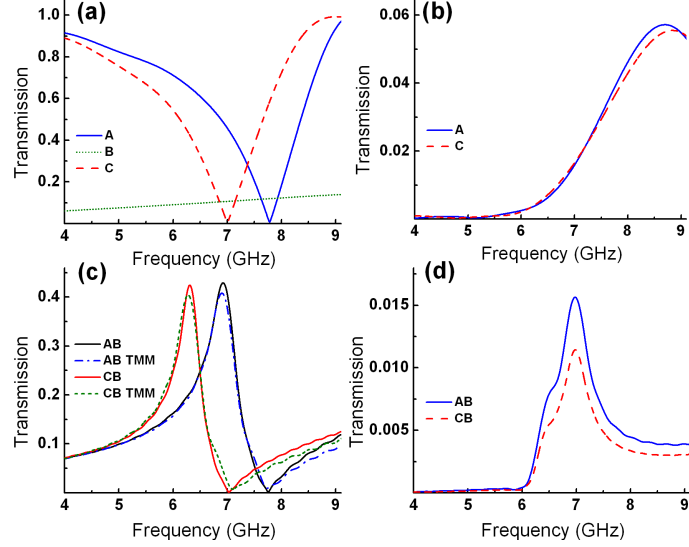


Figure 3.13: $|T_{xx}|$ for (a) separate and (c) double layers, and $|T_{yx}|$ for the same (b) separate and (d) double layers.

and magnetic fields in the transverse plane is relatively weak. The small difference in $|T_{yx}|$ values for layers A and C is caused by the weak symmetry breaking of the substrate [131].

The numerical results of the AB and CB cases are presented in Fig. 3.13(c) for $|T_{xx}|$ and in Fig. 3.13(d) for $|T_{yx}|$. Using Fig. 3.13(c), we observe transmission bands at 6.2 GHz and 7 GHz for the AB and CB cases, respectively. Moreover, the frequencies of zero transmissions are not modified by adding Layer B. Noting that $|T_{yx}|$ is negligible compared to $|T_{xx}|$ in these cases, we can approximately model the transmission of AB and CB utilizing the transfer matrix method (TMM), while assuming the eigenwaves to be linearly polarized, i.e., similarly to Ref. [130]. Using the effective medium theory (EMT), for x -polarized incident waves, we characterize the separate layers as homogenous dielectric slabs with $\epsilon_A = 5.67 + 252/(7.78^2 - f^2)$, $\epsilon_B = 3.65 - 55.7^2/f^2$, and $\epsilon_C = 6.33 + 286/(7^2 - f^2)$ where f is the frequency in GHz. The effective thickness required to reproduce the simulated magnitude and phase values is 1.22 mm for layers A and C, and 1.5 mm for layer B. The results obtained from the TMM calculations are shown in Fig. 3.13(c) and are in excellent agreement with the simulations. At the same time, we observe a peak at 7 GHz for $|T_{yx}|$; see Fig. 3.13(d). In order

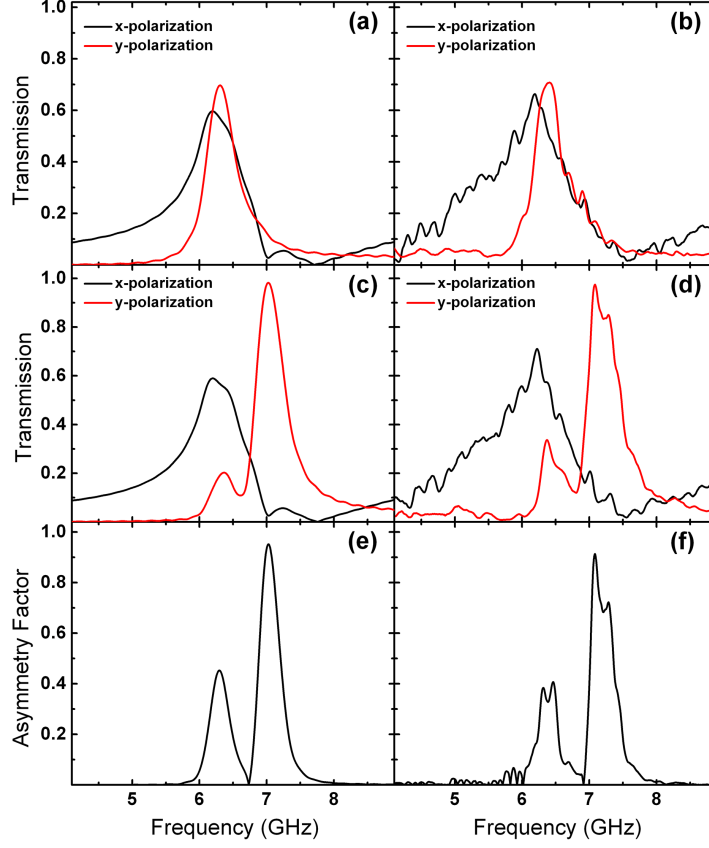


Figure 3.14: Numerical and experimental transmission spectra for the ABS stack, for x -polarized (a), (b) forward and (c), (d) backward propagating waves; (e), (f) numerical and experimental asymmetry factor.

to theoretically describe $|T_{yx}|$ using EMT, one should determine the elliptically polarized eigenwaves for layers A and C and retrieve the effective permittivity for each eigenwave. Afterwards, TMM calculations should again be performed.

Thereafter, we simulated the composite structure with ABC stacking-type. The numerical results reveal that increasing the distance between the layers decreases $|T_{xy}|$ as a result of the decreased evanescent-wave coupling. Thus, one should not leave any air gaps. The numerical and experimental results that demonstrate the diodelike asymmetric transmission phenomenon are shown in Fig. 3.14. According to these results, the frequency of the diodelike transmission is 7 GHz. In Fig. 3.14(c), it is observed that $|T_{yx}|$ approaches unity for the backward propagating waves, whereas $|T_{xx}|$ is close to zero for both propagation directions. In turn, Fig. 3.14(a) shows that only 10% of the incident field is

transmitted when the structure is excited by a forward propagating wave. To be exact, $|T_{xx}|$ is approximately 0.03, and $|T_{yx}|$ is 0.1 and 0.98 for the forward and backward propagating waves, respectively. These parameters provide 1.09% intensity transmission for the forward propagation and 96.1% intensity transmission for the backward propagation. These values correspond to the contrast of 88 (19.5 dB), which is obtained as the ratio of the two transmitted intensities. For the backward propagation, the magnetic field distribution inside the composite structure (not shown) that is similar to the one given in Ref. [130] reveals the occurrence of the electromagnetic tunneling due to the high magnetic fields (see Fig. 2 in Ref. [130]). The transmission of the composite ABC structure can also be modeled using EMT as we have done for the AB and CB cases. However, in the composite case, the eigenwaves are elliptically polarized, so that significant effort is needed to accurately apply the EMT framework. Instead, subsequently, we will describe the behavior of the structure in terms of the eigenstate interpretation.

According to Fig. 3.14, the experiment results are in good agreement with the numerical results, except for a slight shift of the frequency of the maximal asymmetry (7.1 GHz). The possible reasons for this shift can include (i) the difference of the effective permittivity of the Teflon substrate from the simulations, (ii) the small differences in substrate and metal thicknesses, and (iii) the finite size effects. Experimental data show that $|T_{xx}| = 0.141$ and $|T_{yx}| = 0.171$ for a forward propagating wave. These transmission coefficients yield 4.9% intensity transmission at 7.1 GHz. On the other hand, for the backward propagation, we obtain that $|T_{xx}| = 0.112$, whereas $|T_{yx}| = 0.974$, which corresponds to 96% intensity transmission. Hence, the experimental contrast is 19.6 (13 dB).

In this context, the *asymmetry factor* is defined as $|T_{xx}^b|^2 + |T_{yx}^b|^2 - |T_{xx}^f|^2 - |T_{yx}^f|^2$. We demonstrate the asymmetry factor of the structure for an x -polarized incident wave numerically and experimentally in Figs. 3.14(e) and 3.14(f), respectively. In particular, one can see two peaks of the asymmetry factor that are positioned at 6.3 GHz and 7 GHz. Such behavior is quite expectable while these two frequencies correspond to the transmission peaks of $|T_{yx}|$, where the transmitted intensity difference is maximized as a result of increasing difference between the forward and backward $|T_{yx}|$ values. The simulation results state a

0.45 and 0.95 asymmetry factor at 6.3 GHz and 7 GHz, respectively. Experimental results show great agreement in terms of the asymmetry factor, which is now of 0.38 at 6.3 GHz and 0.91 at 7.1 GHz. The direction of diodelike transmission is reversed for a y -polarized wave as is obvious from Eqs. 3.20 and 3.21. The asymmetry factor is strongly dependent on the incident polarization and has its maximum for the x - and y -polarized waves. On the other hand, the asymmetry factor is zero for an incident wave that is linearly polarized at 45° , meaning complete symmetry in transmission.

3.4.4 Eigenstate Interpretation

For the achievement of diode-like asymmetric transmission, the Jones matrix descriptions in Eqs. 3.20 and 3.21 state that $T_{xx} = T_{xy} = T_{yy} = 0$ and $T_{yx} = -1$ must be achieved for backward propagation. However, using the present ABC-type stacking approach, the transmission matrix for backward propagation is achieved as

$$T = \begin{pmatrix} 0 & -\alpha e^{j\varphi} \\ -1 & 0 \end{pmatrix}, \quad (3.22)$$

where $\alpha = 0.1$. As it will be demonstrated subsequently, φ does not affect the magnitudes of the transmission coefficients and therefore, its value is not explicitly given. Afterwards, the eigenwaves of the ABC structure are found as [98]

$$\mathbf{i}_1 = \begin{pmatrix} 1 \\ \alpha^{-1/2} e^{-j\varphi/2} \end{pmatrix}, \quad \mathbf{i}_2 = \begin{pmatrix} 1 \\ -\alpha^{-1/2} e^{-j\varphi/2} \end{pmatrix}. \quad (3.23)$$

The eigenstates are orthogonal and counter-rotating elliptically polarized waves. We can define the transmitted waves as $T_1 \mathbf{i}_1$ and $T_2 \mathbf{i}_2$ when $\mathbf{i}_1$ and $\mathbf{i}_2$ are incident on the structure, respectively. Here, T_1 and T_2 are complex transmission coefficients for the eigenwaves. Next, by decomposing an x -polarized incident field in terms of the eigenwaves of the system, we obtain

$$T \begin{pmatrix} 1 \\ 0 \end{pmatrix} = \frac{1}{2}(T_1 \mathbf{i}_1 + T_2 \mathbf{i}_2) = \begin{pmatrix} (1/2)(T_1 + T_2) \\ (\alpha^{-1/2}/2)e^{-j\varphi/2}(T_1 - T_2) \end{pmatrix}. \quad (3.24)$$

Thus, for the purpose of achieving $|T_{xx}| = 0$, we need $T_2 = -T_1$. In other words, the two counter-rotating eigenwaves must be transmitted with a phase difference of π . At this point, the phase difference is a critical parameter. The subwavelength mesh in the structure acts as a necessary degree of freedom that enables the obtaining of the desired phase difference between the transmitted eigenwaves. Then, $|T_{yx}|$ is found as $\alpha^{-1/2}|T_1|$ and equal to *unity* at 7 GHz and, therefore, for this frequency, $|T_1| = |T_2| = 0.316$. Electromagnetic tunneling becomes evident as a result of the constructive interference of the transmitted eigenwaves in the y direction. Next, we decompose a y -polarized incident wave in terms of the eigenwaves as follows:

$$T \begin{pmatrix} 0 \\ 1 \end{pmatrix} = \frac{e^{j\varphi/2}}{2\alpha^{-1/2}}(T_1\mathbf{i}_1 - T_2\mathbf{i}_2) = \begin{pmatrix} (\sqrt{a}e^{j\varphi/2}/2)(T_1 - T_2) \\ (1/2)(T_1 + T_2) \end{pmatrix}. \quad (3.25)$$

In that case, $|T_{yy}|$ is equal to zero at 7 GHz. On the other hand, $|T_{xy}|$ is given by $\alpha^{1/2}|T_1|$ and equal to 0.1. Moreover, the ratio between $|T_{yx}|$ and $|T_{xy}|$ is found as α^{-1} , as was expected from Eq. 3.22. Note that, due to Lorentz reciprocity $T_{xy}^b = -T_{yx}^f$. Therefore, we conclude that the presented diodelike asymmetric transmission originates from the common effect of the two counter-rotating elliptically polarized eigenwaves that are transmitted through the structure with a phase difference of π . The phase difference is of crucial importance since it determines the magnitudes of the co- and cross-polarized transmission. One should choose the eigenwaves such that the magnitude of their component along the y -direction ($\alpha^{-1/2}$) should be as large as possible for maximizing the transmission asymmetry. The calculations based on the Jones matrix formalism reveal that, if $\alpha \ll 1$ is achieved, any structure exhibiting the transmission of the elliptically polarized eigenstates with a phase difference of π would exhibit diodelike asymmetric transmission. It is noteworthy that the modification of the mesh thickness can decrease the asymmetry factor significantly as a consequence of modifying the phase difference between the transmitted eigenstates.

Chapter 4

Circular Polarizer Design based on High-Contrast Gratings

4.1 Introduction

Due to exhibiting broadband and diffraction-free high reflectivity regimes, high-contrast gratings (HCGs) have attracted numerous researchers [6,7]. This advantageous high-reflectivity property can be utilized in flat reflection lenses [132,133], vertical cavity surface emitting lasers [134], broadband polarization independent reflectors [135], polarizing beam splitters [136], ultra low loss hollow core waveguides [137], beam steering devices [138] and optical diode-like devices [57]. In addition, by means of the ease in the fabrication steps and the geometrical design flexibility, HCGs can be seen as good candidates for constructing compact optical devices. Furthermore, the theoretical analysis of the binary HCGs are rather easier than the complex periodic electromagnetic structures like photonic crystals and metamaterials because of the simpler periodic constituents of the HCG structures.

In this chapter, we propose a novel broadband circular polarizer design based on binary HCGs. The numerical and theoretical analysis of the proposed device is investigated for a stand-alone structure. Additionally, a more realistic design on a substrate is proposed and the experimental verification of the proposed

circular polarization generation operation is performed around the wavelength of $\lambda = 1.55\mu m$.

The circular polarization generation mechanism can be realized by optimizing the complex transmission coefficients under the transverse magnetic (TM) and transverse electric (TE) polarized wave illumination. For the theoretical investigation purposes, the conditions for the unit transmittances in both polarizations are analytically calculated and the phase components for both polarization are compared. The unity transmission for both polarizations and the phase difference of 90° between the polarizations are required for the generation of circular polarization. Further, as it mentioned in the previous section, the direction of the linearly polarized wave determines the handedness type of the generated circular polarization.

4.2 Theoretical and Numerical Investigations of the Proposed Structure

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As shown in Fig. 4.1, the proposed design consists of periodic two dimensional binary HCGs which are assumed as infinitely long in the y -direction. Except the ridges in Region II are filled with high index material (Si), the materials for other regions and the grooves in Region II are selected as SiO_2 . Region III can be seen as the substrate and the reason behind the usage of the same material with other regions except the high index region can be explained as the purpose of achieving unity transmission by eliminating any impedance matching concerns. Further, for the purpose of achieving the infinite medium effects for practical applications, regions I and III can be constructed by antireflective coatings and sufficiently thick SiO_2 substrate, respectively.

In adopting the theoretical analysis given in Ref. [140], the condition for the

unit transmittance of a TM polarized incident wave can be written as follows:

$$|\sum_m (a_m - a_m^\rho) \Lambda^{-1} \int_0^\Lambda h_{y,m}^{in}(x) dx| = 1, \quad (4.1)$$

where a_m and a_m^ρ are the coefficients of the $+z$ and $-z$ propagating field components of the m^{th} waveguide mode, respectively, and $h_{y,m}^{in}$ represents the lateral magnetic field distribution inside region II. The condition given in Eq. (4.1) implies that the magnitude of the sum of the transmission coefficients for each mode should add up to unity. It is noteworthy that the obtaining of identical phases for the transmission coefficients of the individual modes is not a requirement. Benefiting from the duality, the unit transmittance condition for TE waves can be written by replacing $h_{y,m}^{in}$ in Eq. (4.1) with $h_{x,m}^{in}$.

In order to enable the obtaining of a structure that acts as a circular polarizer for linearly polarized incident plane waves with a polarization angle of $\pi/4$, Eq. (4.1) must be simultaneously satisfied for both TM and TE waves. Furthermore, the phase difference between the TM and TE transmission coefficients must be $\pm\pi/2$. This phase condition implies the following equation:

$$\angle \left(\sum_m \mathbf{H}_{0,m}^{\text{TM}} A^{\text{TM}} \right) - \angle \left(\sum_m \mathbf{H}_{0,m}^{\text{TE}} A^{\text{TE}} \right) = \pm\pi/2, \quad (4.2)$$

where $A = (a_m - a_m^\rho)$ and $\mathbf{H}_{0,m}^{\text{TM}}$ is given as follows:

$$\mathbf{H}_{0,m}^{\text{TM}} = \Lambda^{-1} \int_0^\Lambda h_{y,m}^{in}(x) dx. \quad (4.3)$$

In the calculation of $\mathbf{H}_{0,m}^{\text{TE}}$, $h_{y,m}^{in}$ in Eq. (4.3) should be replaced by $h_{x,m}^{in}$. Furthermore, one can show that the longitudinal wavenumber inside the grating region, β_m , is given by

$$\beta_m^2 = (2\pi n_g / \lambda_0)^2 - k_{g,m}^2 = (2\pi n_{bar} / \lambda_0)^2 - k_{r,m}^2, \quad (4.4)$$

where λ_0 is the free-space wavelength and $k_{g,m}$ and $k_{r,m}$ are the lateral wavenumbers of the m^{th} waveguide mode inside the grooves and the ridges, respectively. The dispersion relation between k_g and k_r for TM waves is given as follows:

$$n_{bar}^{-2} k_{r,m} \tan(k_{r,m} r/2) = -n_g^{-2} k_{g,m} \tan(k_{g,m} g/2). \quad (4.5)$$

For TE waves, the dispersion relation is given by

$$k_{r,m} \tan(k_{r,m} r/2) = -k_{g,m} \tan(k_{g,m} g/2). \quad (4.6)$$

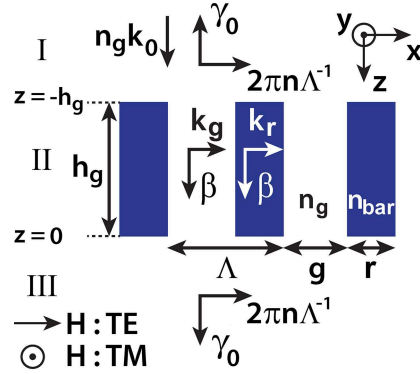


Figure 4.1: Geometrical description of the HCG circular polarizer. The blue regions indicate the presence of Si, whereas the white regions are SiO₂.

As a result of the different dispersion relations, the longitudinal and lateral wavenumbers of the TM and TE modes can be different. At this point, we invoke a two-mode approximation such that the allowed values of m are 0 and 1. The corresponding k_g , k_r , and β values for a certain h_g , r , g , and Λ geometrical parameter set can be determined by the simultaneous solution of Eqs. (4.4) and (4.5) for TM waves and Eqs. (4.4) and (4.6) for TE waves. The wavelength of interest (λ_0) is selected as 1.55 μm . n_{bar} and n_g are set to 3.48 and 1.47, respectively, which correspond to the refractive indices of Si and SiO₂ at this frequency. Thereafter, with the usage of a custom parametric optimization code, the geometrical parameters that satisfy Eqs. (4.1) (for TM and TE waves simultaneously) and (4.2) are obtained as $r = 160$ nm, $g = 220$ nm, $\Lambda = 380$ nm, and $h_g = 550$ nm.

According to the given geometrical parameters, for TM waves, the lateral

wavenumbers inside region II are calculated as $k_{g,0}^2 = -4.2/\Lambda^2$, $k_{r,0}^2 = 19.43/\Lambda^2$ and $k_{g,1}^2 = 31.3/\Lambda^2$, $k_{r,1}^2 = 54.9/\Lambda^2$. In a waveguide, the mode cut-off corresponds to $\beta_m = 0$ and accordingly, in this system, the mode cut-off condition, which is derived from Eq. (4.4), is given by $k_r^2 = (n_{bar}/n_g)^2 k_g^2$. For larger k_r , the resulting β_m values are negative, which implies that the corresponding waveguide mode is exponentially decaying in the propagation direction ($\pm z$). Thus, the first TM mode is a propagating waveguide mode, whereas the second one corresponds to an exponentially decaying mode.

By performing a similar analysis for TE waves, the lateral wavenumbers inside the grating region are obtained as $k_{g,0}^2 = -13.1/\Lambda^2$, $k_{r,0}^2 = 10.48/\Lambda^2$ and $k_{g,1}^2 = 30.2/\Lambda^2$, $k_{r,1}^2 = 53.8/\Lambda^2$. The lateral wavenumbers in the TE case again imply that the first mode and the second mode are propagating and evanescent modes, respectively.

The longitudinal wavenumber of the n^{th} order diffracted wave in transmission and reflection is denoted by γ_n and defined as $\gamma_n^2 = (2\pi n_g)^2 (\lambda_0^{-2} - n^2 \Lambda^{-2})$. Since $\lambda_0/\Lambda = 4.1$ in the proposed design, only the 0^{th} diffraction order is propagating, whereas all of the higher orders are evanescent, which results in the fact that the reflected and transmitted waves are plane waves propagating in the $-z$ and $+z$ directions, respectively.

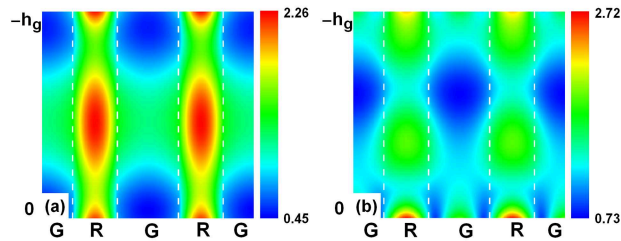


Figure 4.2: The normalized magnetic field distributions at $\lambda_0 = 1.55 \mu\text{m}$ inside region II, (a) $|H_y|/|H_0|$ and (b) $|H_x|/|H_0|$ for TM and TE waves, respectively. The distributions are obtained under the two-mode approximation. Grooves and ridges are separated by the white dashed lines and denoted by G and R, respectively.

The magnitudes of the corresponding field transmission coefficients are calculated for the TM and TE cases as 0.988 and 0.936, respectively. The phase

difference between these two coefficients is defined as $\angle T_{TM} - \angle T_{TE}$ and its value is obtained as 92° , which is desired for the obtaining of a circularly polarized wave at the output interface. Accordingly, it can be deduced that a circularly polarized wave should be transmitted if the structure is illuminated by a plane wave that is linearly polarized at an angle of $\pi/4$, meaning that the electric field components that correspond to the TE and TM polarizations have equal magnitudes and are also in-phase. The magnetic field distributions inside the region II obtained under the two-mode approximation for TM and TE waves are shown in Fig. 4.2. The unequal lateral and longitudinal wavenumbers for the TE and TM modes, which also lead to unequal a_m and a_m^ρ values along with unequal $h_{y,m}^{in}$ and $h_{x,m}^{in}$ distributions, result in significantly different field distributions inside the periodic waveguide. For the further details of the calculation of the transmission coefficients and the field distributions, one can refer to Ref. [140], which provides a detailed theoretical analysis for vertical binary gratings.

In order to verify the validity of the results obtained via the theoretical model, we employ the Rigorous Coupled Wave Analysis (RCWA) [141]. Using this method, between the free space wavelengths of $1 \mu\text{m}$ and $3.5 \mu\text{m}$, the magnitudes of the TM and TE transmission coefficients and their phase difference are determined and these results are shown in Fig. 4.3.

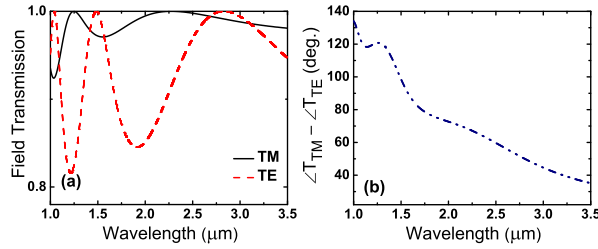


Figure 4.3: (a) Field transmission coefficients for TM and TE waves and (b) the phase difference between the transmission coefficients.

Thereafter, the circular conversion coefficients, which indicate the amplitudes of the right-hand circularly polarized (RCP, +) and left-hand circularly polarized (LCP, -) waves at the output interface, for an incident wave that is linearly polarized at an angle of $\pi/4$ in the xy -plane can be calculated with a simple transformation from the linear base to the circular base as $C_{\pm} = 0.5 (T_{TM} \mp iT_{TE})$

[142]. At this point, we define a conversion efficiency in order to characterize the performance of the proposed structure, which is denoted by C_{eff} and can be given as follows:

$$C_{eff} = \frac{|C_+|^2 - |C_-|^2}{|C_+|^2 + |C_-|^2}. \quad (4.7)$$

In order to incorporate the effects arising due to the material dispersion of Si and SiO₂, we perform Finite-Difference Time-Domain (FDTD) simulations using FDTD Solutions (Lumerical, Inc.), where the refractive index data given in Ref. [143] is utilized. Figure 4.4 shows the circular conversion coefficients obtained from the RCWA along with the conversion efficiencies calculated using the RCWA and FDTD methods.

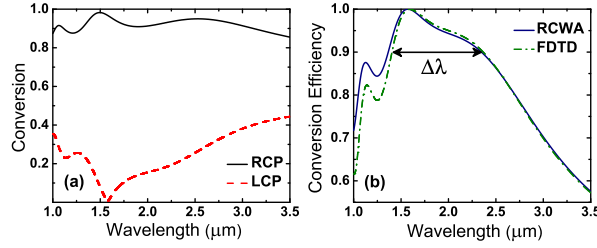


Figure 4.4: (a) Circular conversion coefficients obtained using the RCWA and (b) the conversion efficiencies obtained from the RCWA and FDTD. The wavelength range for the FDTD result satisfying the 0.9 efficiency threshold is denoted by $\Delta\lambda$.

The RCWA method suggests that a conversion efficiency that is larger than 0.9 is achieved between the wavelengths of 1.36 μm and 2.36 μm. According to the FDTD results, where the material dispersion is taken into account properly, the 0.9 threshold is achieved between 1.4 μm and 2.36 μm. As a consequence of the strongly pronounced material dispersion effects for Si for wavelengths smaller than 1.5 μm, a slightly narrower bandwidth is achieved. If the percent bandwidth is defined as follows [144]:

$$BW\% = 200\% \frac{\lambda_H/\lambda_L - 1}{\lambda_H/\lambda_L + 1}, \quad (4.8)$$

where λ_H and λ_L are the higher and lower corner wavelengths, respectively, a BW% of 54% and 51% is achieved using the RCWA and FDTD methods, respectively. Despite causing an observable modification in the conversion efficiency

spectrum, the involvement of the material dispersion effects does not significantly modify the percent bandwidth, which shows that the proposed structure in the present study can be practically realizable.

4.3 Experimental Verification

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In a recent study, we have shown that a two-dimensional HCG structure with optimized geometrical parameters can be utilized for implementing a broadband circular polarizer [139]. The optimized geometrical parameters have been suggested by benefiting from the periodic dielectric slab waveguide interpretation, which is studied rigorously in [140]. The spectral transmission results obtained from the rigorous coupled-wave analysis [141] and finite-difference time-domain (FDTD) simulations (FDTD Solutions, Lumerical Inc.) have suggested that an operation percent bandwidth of 54% and 51% can be achieved, respectively, under the assumption that the proposed structure is said to be within the operation regime if the conversion efficiency exceeds 0.9. For the sake of maximizing the operation bandwidth in this theoretical study, the material that is employed outside the grating region as well as in the grooves between the silicon gratings has been assumed to be silicon dioxide.

4.3.1 Proposed Geometry

In this paper, we study the potential of the experimental realization and characterization of an HCG based broadband quarter-wave plate that can be designed by following the methodology given in [139]. By considering the potential challenges in the fabrication stage and evaluating the limitations of the devices used in the nanofabrication processes, without sacrificing significantly from the bandwidth, we decided to design an HCG structure with sapphire substrate and silicon

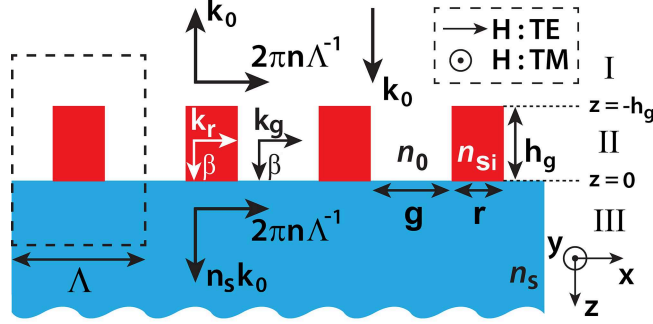


Figure 4.5: Illustration of the proposed quarter-wave plate geometry. The dashed square box on the left denotes one period. The geometrical parameters are given by $r = 220$ nm, $g = 350$ nm, $h_g = 320$ nm, and $\Lambda = 570$ nm. In the theoretical and numerical consideration, for the sake of simplicity, regions I and III are assumed to be infinite in the $-z$ and $+z$ directions, respectively. The materials constituting region III and the ridges in region II are sapphire and silicon, respectively, whereas region I is free-space. n_0 , n_{si} , and n_s represent the refractive indices of free-space, silicon, and sapphire, respectively. *Grating direction* is defined such that it corresponds to the y direction.

gratings, where the grooves between the gratings are free-space as well as the region that the wave is incident onto the structure from. The proposed geometry is depicted in Fig. 4.5. The polarization states are defined such that a plane wave is transverse electric (TE) and transverse magnetic (TM) polarized if the magnetic field is in the x and y directions, respectively, where the incident wave propagates in the $+z$ direction.

In the present study, a theoretical solution of the specific HCG design problem is not provided due to the fact that the solution is fundamentally the same as the ones given in [139, 140]. However, we make a remark on the difference of this problem compared to the previously mentioned studies. As a result of the selection of different materials for filling regions I and III, i.e., free-space and sapphire, respectively, the wavenumbers of the zeroth reflection and transmission orders are expected to be different. The wavenumber of the zeroth reflection order is equal to k_0 and given by $\gamma_{0,\text{I}} = 2\pi\lambda_0^{-1}$, whereas the wavenumber of the zeroth transmission order is equal to $n_s k_0$ and, therefore, given by $\gamma_{0,\text{III}} = 2\pi n_s \lambda_0^{-1}$. This discrepancy results in unequal lateral magnetic field distributions in regions I and

III [denoted by $H^I(x, z \leq -t_g)$ and $H^{III}(x, z \geq 0)$ in [140], respectively]. Therefore, ensuring the utilization of the correct lateral distributions while matching the boundary conditions at $z = -h_g$ and $z = 0$ is of crucial importance and requires special attention.

The optimized HCG design is achieved with the aid of the theoretical calculations and the rigorous coupled-wave analysis for vertical binary gratings [141]. The utilized geometrical parameters are provided in the caption of Fig. 4.5. Subsequently, FDTD simulations (FDTD Solutions, Lumerical Inc.) are run in order to numerically characterize the transmission of the proposed HCG structure for TM and TE waves, while taking the effects arising because of the material dispersion into account, i.e., dependence of n_{si} and n_{s} on the wavelength of the incident light. In the simulations, the refractive index data given in [143] is employed for n_{si} and n_{s} . The two-dimensional simulations are constructed such that the periodic boundary condition is employed along the x direction, whereas the perfectly matched layer (PML) boundary condition is adopted along the z direction. The sapphire substrate is assumed to be extending to infinity in the $+z$ direction for the purpose of avoiding the oscillations in the transmission spectrum that originate because of the Fabry-Perot modes created within a substrate of a finite thickness. Moreover, as a result of the typical short coherence length of broadband light sources, such an effect is not expected to be visible in the experimental results, unless a substrate with a thickness that is comparable to the operation wavelength is employed.

4.3.2 Numerical Results

The numerically obtained normalized transmitted intensities for the HCG for normally incident TM and TE waves ($|T_{\text{TM}}|^2$ and $|T_{\text{TE}}|^2$, respectively) are shown in Fig. 4.6(a). The proposed HCG geometry is optimized such that, in the close neighborhood of $\lambda_0 = 1.55 \mu\text{m}$, the conditions $|T_{\text{TM}}| \simeq |T_{\text{TE}}|$ and $\angle(T_{\text{TM}}) - \angle(T_{\text{TE}}) \simeq \pi/2$ are satisfied simultaneously, which result in the transmission of a right-hand circularly polarized (RCP, +) and left-hand circularly

polarized (LCP, $-$) wave assuming that the structure is illuminated by a normally incident plane wave that is linearly polarized with a polarization plane angle of 45° and -45° with respect to the x -axis on the xy -plane, respectively. The circular conversion coefficients for 45° polarization angle are subsequently calculated by $C_{\pm} = 0.5 (T_{\text{TM}} \mp iT_{\text{TE}})$ [142] and shown in Fig. 4.6(b). Obviously, T_{TM} and T_{TE} are complex numbers, which carry the phase information besides the magnitude. However, in the present experimental study, the phases

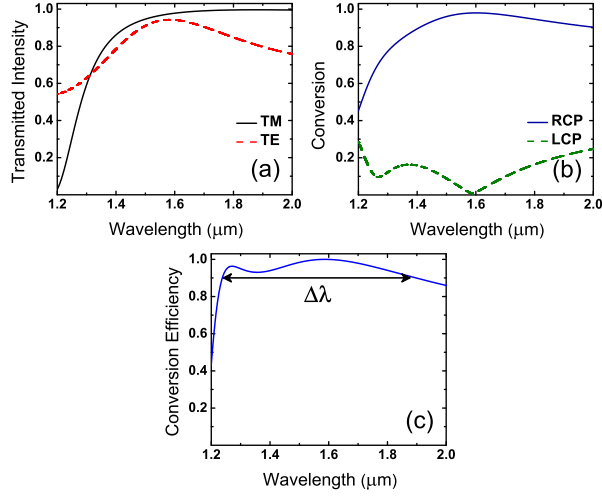


Figure 4.6: Numerical (a) TM and TE transmitted intensities, (b) circular conversion coefficients, and (c) conversion efficiency spectrum obtained via FDTD simulations. The wavelength interval of operation is denoted by $\Delta\lambda$. The numerical conversion efficiency spectrum yields a percent bandwidth of 42%.

of the numerical transmission coefficients are not explicitly given, since the phase information is not required for the obtaining of the experimental conversion coefficients. Instead, as it will be evident subsequently, these coefficients can directly be measured by the utilization of a linear polarizer and a quarter-wave plate. Finally, the conversion efficiency, which is shown in Fig. 4.6(c), is calculated as follows [139]:

$$C_{\text{eff}} = \frac{|C_+|^2 - |C_-|^2}{|C_+|^2 + |C_-|^2}. \quad (4.9)$$

At this point, we impose a condition on C_{eff} such that the proposed structure is said to be in the operation regime if $C_{\text{eff}} \geq 0.9$ is satisfied. This condition implies that, at the output interface, the intensity of the RCP wave should be, at least, 12.8 dB larger than that of the LCP wave. The numerical results suggest

that the C_{eff} condition is satisfied between $\lambda_0 = 1.24 \mu\text{m}$ and $1.90 \mu\text{m}$, which corresponds to a percent bandwidth of 42%, if the percent bandwidth is defined as follows [144]:

$$\text{BW}\% = 200\% \frac{\lambda_H/\lambda_L - 1}{\lambda_H/\lambda_L + 1}, \quad (4.10)$$

where λ_H and λ_L represent the higher and lower corners of the operation band, respectively. It is noteworthy that in the case of illumination by a plane wave with a polarization plane angle of -45° , C_+ and C_- would simply be swapped.

In addition to the conversion efficiency interpretation, one can define the conversion characteristics of the proposed design by calculating the ellipticity spectrum for transmitted waves. Under the assumption that the HCG is illuminated by a normally incident plane wave with a polarization plane angle of $\pm 45^\circ$, the ellipticity, η , can be calculated as follows [5]:

$$\eta = \arctan \left(\frac{|C_+| - |C_-|}{|C_+| + |C_-|} \right). \quad (4.11)$$

By combining (4.9) and (4.11), η can simply be calculated from the C_{eff} spectrum as follows:

$$\eta = \arctan \left(\frac{\sqrt{(1 + C_{\text{eff}})/(1 - C_{\text{eff}})} - 1}{\sqrt{(1 + C_{\text{eff}})/(1 - C_{\text{eff}})} + 1} \right). \quad (4.12)$$

Using (4.12), it can be shown that a conversion efficiency of 0.9 corresponds to $\eta = 32^\circ$. Finally, due to the one-to-one correspondence between η and C_{eff} , only

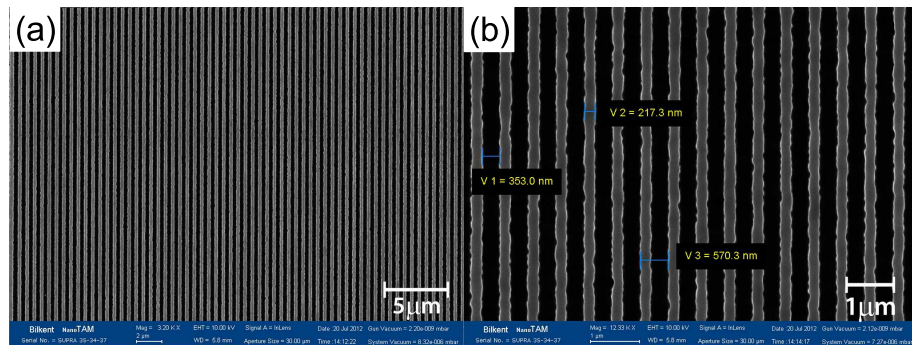


Figure 4.7: (a) Zoomed out and (b) zoomed in top view SEM micrographs of the fabricated HCG structure. In (b), the legends V1, V2, and V3 denote the geometrical parameters g , r , and Δ , respectively.

the C_{eff} spectrum is provided in the present study.

4.3.3 Fabrication

The grating structure is fabricated onto a silicon-on-sapphire (SOS) wafer with a 100 mm diameter, which is purchased from Valley Design Corporation. The wafer consists of a 600 ± 60 nm thick $\langle 100 \rangle$ silicon layer that is grown epitaxially on a 0.53 ± 0.05 mm thick R-plane sapphire. Firstly, the SOS wafer is diced into square shaped chips with the dimension of 6 mm by 6 mm. The fabrication process starts with the reduction of the thickness of the silicon layer to 320 nm by means of the reactive ion etching (RIE) technique, while using sulfur hexafluoride (SF_6) as the etchant gas. Afterwards, polymethyl methacrylate (PMMA), which is an electron beam sensitive photoresist, is spun onto the sample. Subsequently, the desired grating structure is patterned onto the SOS sample using e-beam lithography. Following the development operation of the PMMA, the final step is accomplished by etching the unexposed silicon regions using the RIE system. The scanning electron microscope (SEM) micrographs of the fabricated HCG sample are shown in Fig. 4.7.

4.3.4 Experimental Setup

The outlines of the two experimental setups, which are constructed for the measurement of the linear transmission coefficients and the circular conversion coefficients of the HCG sample, are illustrated in Fig. 4.8 in detail. In the experiments, the sample is illuminated by a broadband near-infrared (NIR) light source (Spectral Products, ASBN-W100-F-L) and, with the aid of standard commercial NIR objectives and an adjustable aperture (not shown in Fig. 4.8), the spot size of the incident light is adjusted to be $50 \mu\text{m}$. The normalization of the transmitted fields is performed with respect to a solely sapphire region that is obtained by etching the silicon layer on the SOS chip.

In the first measurement, which is illustrated in Fig. 4.8(a), the linear intensity

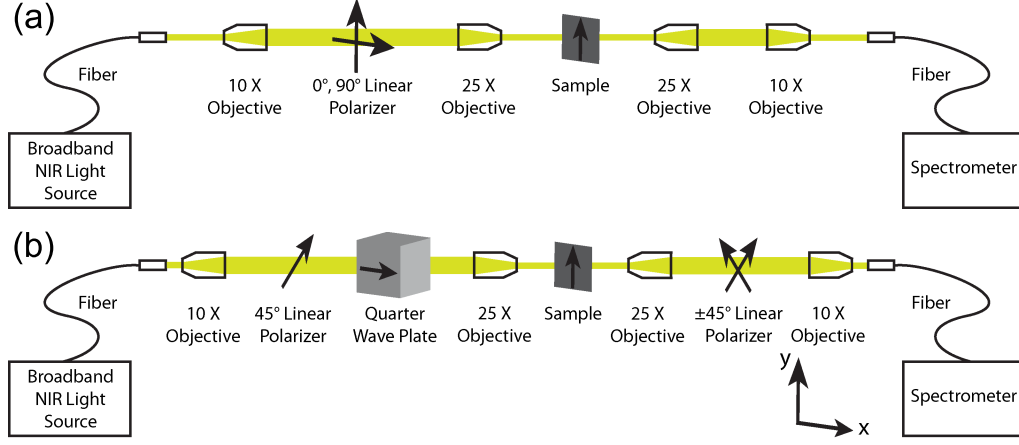


Figure 4.8: Visual illustrations of the experimental setups that are utilized for the measurement of (a) linear transmission coefficients and (b) circular conversion coefficients. The arrows, which lie inside the HCG samples and point upwards, denote the grating direction that is defined in the caption of Fig. 4.5.

transmission coefficients ($|T_{TM}|^2$ and $|T_{TE}|^2$) are obtained by the utilization of an adjustable linear polarizer (Thorlabs, DGL10) that is used for enabling the illumination of the grating region with TM and TE polarized light. In the second measurement, which is illustrated in Fig. 4.8(b) and conducted with the purpose of measuring the circular conversion coefficients, the grating region is illuminated by an RCP wave as a result of the employment of an adjustable linear polarizer (with its transmission axis making a 45° angle with the x – axis) followed by a quarter-wave plate (Thorlabs, WPQ05M-1550) that is oriented such that its fast axis is along the x direction. At the farfield, the transmitted wave through the grating region is passed through an adjustable linear polarizer that is oriented such that its transmission axis makes angles of $+45^\circ$ and -45° with respect to the x – axis. Afterwards, the intensities of the waves that are transmitted through the linear polarizer are measured using the spectrometer (Ocean Optics, NIR256-2.5) in order to find the square of the magnitudes of the RCP and LCP conversion coefficients ($|C_+|^2$ and $|C_-|^2$), respectively. By the utilization of the Jones matrix formalism, it can easily be shown that by adjusting the linear polarizer to $+45^\circ$ and -45° , $|C_+|^2$ and $|C_-|^2$ can be measured directly, respectively. As a result of the period of the proposed HCG structure being equal to 570 nm, FDTD simulations and experiments are conducted between the free-space wavelengths

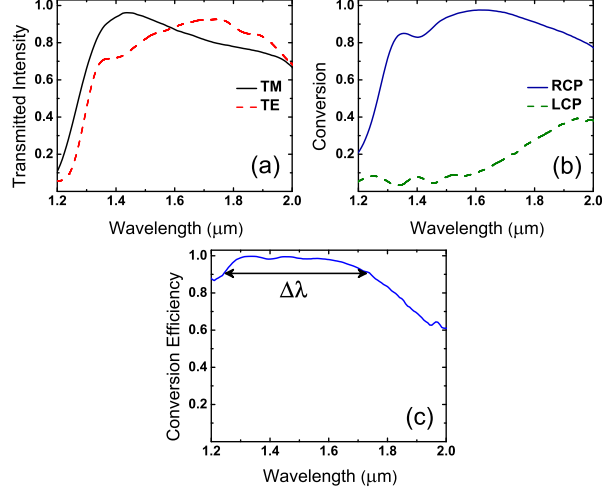


Figure 4.9: Experimentally obtained (a) TM and TE transmitted intensities, (b) circular conversion coefficients, and (c) conversion efficiency spectrum. The wavelength interval of operation is denoted by $\Delta\lambda$. The experimental conversion efficiency spectrum yields a percent bandwidth of 33%.

of 1.2 μm and 2 μm with the purpose of avoiding the wavelengths for which the HCG is diffractive, i.e., $\lambda_0 < n_s\Lambda$.

The proposed HCG based quarter-wave plate can be combined with a HCG based [146], a wire grid [147], a thin-film grating [148], or a plasmonic [149] linear polarizer for the purpose of implementing a compact circular polarizer.

4.3.5 Experimental Results

The linear intensity transmission and circular conversion coefficients (for an incident wave with a polarization plane angle of 45°), and the conversion efficiency spectrum obtained from the experiments are shown in Fig. 4.9. Conversion efficiency data given in Fig. 4.9(c) suggests that the experimental lower and higher corner wavelengths are obtained as $\lambda_L = 1.25 \mu\text{m}$ and $\lambda_H = 1.75 \mu\text{m}$. The usage of Eq. 4.10 results in an experimental percent bandwidth of 33%. In spite of the achievement of a smaller percent bandwidth compared to the numerical predictions, the experimental results exhibit a good agreement with the numerical results shown in Fig. 4.6.

We conjecture that the most significant factors responsible from the discrepancies between the numerical and experimental results are finite-size effect and finite substrate thickness. Finite-size effect originates from the usage of finite number of periods and finite grating length in the experiments (in contradiction with the two-dimensional periodic FDTD simulations, where the number of periods and the grating length are infinite) and possibly modifies the lateral and longitudinal wavenumbers of the excited modes inside region II and, additionally, this effect can result in diffraction from the edges of the sample. The utilization of a substrate with a finite thickness can result in multiple back-reflections inside the substrate, modifying the overall behavior of the corrugated surface and, in addition, Fresnel reflection from the back-side of the substrate can also modify the transmission characteristics of the sample.

In addition to the aforementioned effects, the trapezoidal grating profile achieved after the final etch step and possible deviations of the geometric parameters of the fabricated sample from the design parameters can conduce to the discrepancies between the numerical and experimental results. Furthermore, less significant effects such as the employment of non-ideal optical elements, i.e., linear polarizer and quarter-wave plate, probable small deviation of the actual refractive indices from the ones adopted in the simulations, and human related errors while setting the polarizer transmission axis angle and the angle of incidence might as well contribute to the deviations in the experimental results from the numerical ones.

Chapter 5

Conclusions

The aim of this thesis was to propose compact and highly efficient optical devices utilizing periodic meta-structures such as photonic crystals, metamaterials and high contrast gratings.

In Chapter 2, firstly, we propose a wavelength de-multiplexer design based on concatenated photonic crystal waveguides, for which dielectric filling factors are varied in order to target the slow light region. The frequency selectivity of the device originates from the light behavior in the vicinity of the slow light regime due to the high leakage as a result of the wider spatial distribution of the electromagnetic waves inside the main waveguide. The spatial selection of different wavelengths occurs within consecutive photonic crystal waveguide sections and we numerically and experimentally demonstrate the successful de-multiplexing of three wavelengths in a compact manner.

The preliminary results of the demultiplexer (DEMUX) design employing the slow light phenomena are encouraging. However, the DEMUX design can be further studied in order to obtain higher output power levels at each output channel and a linear spacing in de-multiplexed frequencies. Moreover, using the proposed de-multiplexer design idea, slab dielectric PC structures with air holes in triangular lattice form can be a good candidate to create similar devices that work at optical frequencies avoiding the power leakage in the z-direction. In addition, the length of each PCW can be optimized to create a more compact

design.

Secondly, a frequency division de-multiplexer, which benefits from the slow light regime of the photonic crystal waveguides and is formed by cascading waveguides of unequal widths, is proposed. The de-multiplexing mechanism originates from the spatial distribution characteristics of the slow light modes along the photonic crystal waveguides. The wavelength selectivity of the cascaded photonic crystal sections is obtained by changing the waveguide width of each section. This waveguide band modulation also yields the spatial selectivity of the de-multiplexed frequency bands. The spatial and spectral properties of the de-multiplexed frequency bands are numerically investigated and demonstrated. One strict limitation for the usage of slow light modes is the relatively low extracted power from output channels as a consequence of the high effective dielectric permittivity encountered by the slow light modes compared to the previous de-multiplexing efforts. In spite of the fact that the proposed design is optimized to achieve maximum coupling by introducing a gradual transition in the effective dielectric permittivities between the main channel and the output channels, input and output couplers can be appended to the current geometry for the purpose of smoothing the transition of effective permittivities between the air and the PC design for enabling the obtaining of increased extracted power. The utilization of the slow light regime for de-multiplexing purposes has several advantages at first glance. Firstly, the de-multiplexers that are based on PC structures generally require the occupation of large areas for the achievement of spatial selectivity; however, the proposed structure has a very small footprint compared to the other designs. Secondly, the de-multiplexed set of frequencies can have a very narrow bandwidth and therefore, high quality factors can be obtained as a result of the limited slow light regions in the photonic crystal waveguide bands. Finally, a very low cross-talk can be achieved by the utilization of the slow light regime with a proper modulation of the geometries of the consecutive photonic crystal waveguide sections.

Thereafter, the aim of next study was to obtain an optical diode-like acting structure by utilizing chirped photonic crystals. For this goal, the distances between each neighboring column of rods of square lattice photonic crystal are

modified to obtain a chirped device. The standard waveguide shows no diode like behavior due to the symmetry of the structure. Optical pulse propagates identically in either direction in a standard waveguide. Nonetheless, in the proposed waveguide designs, the propagation and transmission of light collected at the center of the structure are not the same in the two opposite directions. We witness the transmission asymmetry in the light propagation through the chirped PCW. The experimental results obtained at the microwave frequencies confirm the numerically observed asymmetric light propagation. The presented design based on the linear optic concept is a practical choice and may help the advancement of research activities utilizing the asymmetric light propagation of light. The ultimate engineering of chirping parameters not only in terms of lattice spacing but also using other parameters such as rods radii and refractive index parameters may give rise to the superior performance of the optical diode-like photonic structures.

Lastly for the photonic crystal based applications, we propose a Fabry-Perot type cavity with side mirrors that are formed by an air-hole photonic crystal structure for the purpose of achieving a high transmission and a high quality-factor cavity resonance. It is stated that the proposed structure can be employed in electro-optic effect based applications in order to enhance this effect. It is demonstrated that the spectral position and quality factor of the cavity resonance can be modified by changing cavity length, L_c , and number of rows, N , respectively. In addition, we provide the electrical field distributions inside the cavity region at the resonance wavelengths and show that we obtain, on the average, $f_{opt} = 13, 20, 35$ for $N = 5, 6, 7$, respectively. With the aid of these results, we state the possibility of utilizing the proposed design for the enhancement of the electro-optic effect. Then, we show that, as a result of the achievement of a large f_{opt} , the spectral position of the cavity resonance becomes highly sensitive to the refractive index of the material constituting the cavity region. Finally, we study the possibility of realizing other functionalities by using the proposed cavity geometry, i.e., dual-mode and dual-polarization operation regimes. In terms of practical applicability, the proposed structure can be employed in low-threshold lasers, coherent optical devices, optical communication and quantum information processing systems, and optical filters.

In Chapter 3, first of all proposed chiral metamaterial(CMM) based applications, we have designed an asymmetric CMM structure based on four double-layered U-shaped split ring resonators (SRRs). Left hand circularly polarized (LCP) and right hand circularly polarizer (RCP) waves are obtained in the vicinity of 5.1 GHz and 6.4 GHz, respectively, when the structure is illuminated by an x -polarized wave propagating in the z direction. The design can be utilized as a circular polarizer for microwave applications.

Secondly, we have designed an electrically thin CMM structure that allows obtaining polarization rotation, which is dependent on the polarization angle of the incident linearly polarized plane wave at 6.2 GHz. We derived a closed form expression, which relate the degree of polarization rotation to the polarization angle of the incident wave. In addition, it has been demonstrated that the transmission matrix, T , depends on the propagation direction of the incoming wave. Thus, the transmission through the structure is asymmetric for linearly polarized waves. On the other hand, four angles have been found for which the transmission is symmetric. Finally, surface current distributions at 6.2 GHz are studied in order to explain the underlying mechanism behind the asymmetric transmission. The CMM can be utilized in microwave applications as a configurable polarization rotator.

The last effect we investigated related to chiral metamaterials is the unidirectional transmission. we demonstrated a diodelike asymmetric transmission for normally incident x - and y -polarized waves in an electrically thin reciprocal structure. In addition to the planar ultrathin design, the usage of isotropic and linear materials only makes this and similar structures the promising candidates for practical applications. The backward-to-forward transmission contrast of the structure at the operating frequency 7 GHz is 19.5 and 13 dB according to the simulation and experiment results, respectively. The asymmetry factor at this frequency is 0.95 for the simulation results and 0.91 for the experiments.

In Chapter 4, we studied both theoretically and experimentally on a circular polarizer design based on binary high contrast gratings (HCGs). Initially, we have numerically and theoretically designed an optimized two dimensional HCG such

that the transmitted wave is circularly polarized if the structure is illuminated by a plane wave with a polarization plane angle of $\pi/4$. It has numerically been shown that a conversion efficiency that is larger than 0.9 can be achieved within a 51% bandwidth.

After the numerical and theoretical study, we have shown the fabrication and the experimental characterization of a broadband quarter-wave plate that is based on two-dimensional silicon HCGs, where sapphire is employed as the substrate. It has been shown that a conversion efficiency that is higher than 0.9 can be achieved in a percent bandwidth of 42% and 33% numerically and experimentally, respectively. RCP and LCP waves can be obtained by adjusting the transmission axis of the linear polarizer to 45° and -45° with respect to the x -axis, respectively. In terms of practical applicability, the proposed structure can be employed in laser and nanoantenna applications, liquid crystal displays, and remote sensors.

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Appendix A

Publication List

1. M. Mutlu, A. E. Akosman, A. E. Serebryannikov, and E. Ozbay, “Asymmetric chiral metamaterial circular polarizer based on four U-shaped split ring resonators,” *Opt. Lett.*, vol. 36, pp. 1653-1655, May 2011.
2. M. Mutlu, A. E. Akosman, A. E. Serebryannikov, and E. Ozbay, “Asymmetric transmission of linearly polarized waves and polarization angle dependent wave rotation using a chiral metamaterial,” *Opt. Express*, vol. 19, no. 15, pp. 14290-14299, 2011.
3. A. E. Akosman, M. Mutlu, H. Kurt, and E. Ozbay, “Compact wavelength de-multiplexer design using slow light regime of photonic crystal waveguides,” *Opt. Express*, vol. 19, pp. 24129-24138, Nov 2011.
4. M. Mutlu, A. E. Akosman, A. E. Serebryannikov, and E. Ozbay, “Diodelike asymmetric transmission of linearly polarized waves using magnetoelectric coupling and electromagnetic wave tunneling,” *Phys. Rev. Lett.*, vol. 108, p. 213905, 2012.
5. H. Kurt, D. Yilmaz, A. E. Akosman, and E. Ozbay, “Asymmetric light propagation in chirped photonic crystal waveguides,” *Opt. Express*, vol. 20, pp. 20635-20646, Aug 2012.
6. A. E. Akosman, M. Mutlu, H. Kurt, and E. Ozbay, “Dual-frequency division

de-multiplexer based on cascaded photonic crystal waveguides,” *Physica B: Condensed Matter*, vol. 407, no. 20, pp. 4043–4047, 2012. Proceedings of the conference - Wave Propagation: From Electrons to Photonic Crystals and Metamaterials in honor of Prof. C. M. Soukoulis 60th birthday.

7. M. Mutlu, A. E. Akosman, and E. Ozbay, “Broadband circular polarizer based on high-contrast gratings,” *Opt. Lett.*, vol. 37, pp. 2094-2096, Jun 2012.
8. M. Mutlu, A. E. Akosman, G. Kurt, M. Gokkavas, and E. Ozbay, “Experimental realization of a high-contrast grating based broadband quarter-wave plate,” *Opt. Express*, vol. 20, pp. 27966-27973, Dec 2012.