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**DENSE-WAVELENGTH DEMULTIPLEXING AND MICROWAVE FREQUENCY
GENERATION USING NONLINEAR OPTICAL WAVEGUIDES**

by

EDWARD FRLAN, B. ENG., M. ENG.

**A thesis submitted to
the Faculty of Graduate Studies and Research
in partial fulfilment of
the requirements for the degree of**

Doctor of Philosophy

**Ottawa-Carleton Institute for Electrical Engineering
Department of Electronics
Faculty of Engineering
Ottawa, Canada**

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GENERATION USING NONLINEAR OPTICAL WAVEGUIDES"**

submitted by
EDWARD FRLAN, M. ENG.

in partial fulfillment of the requirements
for the degree of Doctor of Philosophy

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Abstract

Based on the concept of nonlinear mixing in optical, III-V semiconductor waveguides, two devices are developed for the microwave control of optical communications systems. The first demonstration and proof-of-concept of a dense-wavelength demultiplexing (DWDM) system utilizing an optical nonlinear waveguide spectrometer is undertaken, as well as the development of an optical downconverter module for implementation in an optically-fed phased array antenna.

Utilizing the sum-frequency nonlinear optical process, a nonlinear optical waveguide can be utilized as part of a DWDM system having microwave frequency resolution. Each of the optical channels can be spatially addressed by placing a detector array in the far-field of the optical nonlinear waveguide spectrometer. Possible use of the optical spectrometer as a deformation sensor having the ability to measure the induced strain in a semiconductor waveguide is also established.

The optical downconverter module is fabricated by utilizing the nonlinearity of the AlGaAs/GaAs optical waveguide to generate a difference-frequency. The generated difference-frequency is in the microwave frequency

range and is radiated from the waveguide into free space and collected by a waveguide horn antenna, whereupon the detected signal can be distributed to a microwave receive module.

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Table of Contents

	Page
Title Page	i
Acceptance Sheet	ii
Abstract	iii
Acknowledgements	v
Table of Contents	vii
List of Figures	x
1.0 Introduction	1
1.1 Motivation	1
1.2 Objectives	7
1.3 Thesis Organization	8
2.0 Overview of Relevant Nonlinear Optics and Optical Waveguides	10
2.1 Nonlinear Optics	10
2.2 Optical Waveguides	12
2.2.1 Molecular Beam Epitaxy	13
2.2.2 Theoretical Analysis of the Slab Waveguide	15
2.3 Sum-Frequency Generation	18

2.3.1 Enhanced GaAs Waveguide Structure for Efficient Second Harmonic Generation	22
2.4 Difference-Frequency Generation	34
2.4.1 Pulsed Microwave Generation	35
2.4.2 CW and Quasi-CW Microwave Generation	41
3.0 Sum-Frequency Generation for Application in a Wavelength Demultiplexing System with Microwave Resolution	44
3.1 Introduction	44
3.2 Theory of Surface-Emitting Nonlinear Waveguide Optical Spectrometer	45
3.3 Experimental Setup and Far-Field Diffraction Measurements	52
3.4 Summary	59
4.0 CW Microwave Generation by Optical Nonlinear Mixing	61
4.1 Introduction	61
4.2 Theory of CW Microwave Generation with Nonlinear Optical Waveguides	61
4.2.1 Modelling of a Hertzian Dipole on a Dielectric	62
4.2.2 Far-Field Radiation Pattern of a Hertzian Dipole Oriented Parallel and Perpendicular	66

to the Substrate Surface	
4.2.3 Finite Element Solution of Generated Electric Fields from a Current Source of Finite Length Oriented Perpendicular to the Substrate	74
4.3 Difference-Frequency Power Generated by the Nonlinear Waveguide	81
4.4 Experimental Setup	84
4.5 Measured Results	86
4.6 Discussion of Microwave Frequency Generation	90
5.0 Conclusions and Future Work	98
Appendix I Effect of Copropagating vs. Counterpropagating Signal Injection	100
Appendix II Derivation of Approximate Far-Field Microwave Power	102
Appendix III Derivation of Waveguide Wave Impedance, n_g	106
References	108

List of Figures

	Page
Figure 1.1 Optically fed phased array antenna	4
Figure 1.2 Optically fed subarray	6
Figure 2.1 Wave normals present in a slab waveguide	12
Figure 2.2 Schematic of an MBE growth system	14
Figure 2.3 Induced polarization vs applied electric field	19
(a) in a linear dielectric, and	
(b) in a crystal without inversion symmetry	
Figure 2.4 Second harmonic generation from two counter-propagating waves on a LiNbO ₃ substrate	23
Figure 2.5 Calculated A ⁿ vs slab thickness for bulk GaAs including effects of absorption	25
Figure 2.6 General waveguide geometry	27
Figure 2.7 Multilayer structure of GaAs SHG device	31
Figure 2.8 Nonlinear cross section for a 90%-70% AlGaAs multilayer waveguide	32
Figure 2.9 Observed bead pattern on (100) GaAs substrate	33
Figure 2.10 Radiated field from GaAs sample for various radiated photon energies with respect to the band edge	36
Figure 2.11 Experimental setup used for RF generation	37

Figure 2.12	Temporal response of 200 μm planar antenna	39
Figure 2.13	Diffraction limited microwave generation and detection	40
Figure 2.14	Experimental setup for noncollinear phase-matched FIR generation	42
Figure 3.1	Propagation of incident and sum-frequency wave vectors	46
Figure 3.2	Far-field diffraction measurement setup	48
Figure 3.3	(a) Far-field diffraction patterns for the distortionless and linear phase distortion cases	49
	(b) Far-field diffraction patterns for the distortionless and quadratic phase distortion cases	50
	(c) Far-field diffractoin patterns for the distortionless and cubic phase distortion cases	51
Figure 3.4	$\text{Al}_x\text{Ga}_{1-x}\text{As}/\text{GaAs}$ waveguide structure	52
Figure 3.5	Optical Waveguide DWDM Communications System	54
Figure 3.6	Variation of the offset frequency between the two ring lasers with laser cavity temperature	55
Figure 3.7	Sum-frequency emission angle vs input offset frequency for a 1-cm long waveguide	56
Figure 3.8	Far-field diffraction pattern for a 34-GHz offset between the carrier and reference signals	57

Figure 3.9	Far-field diffraction pattern for a dc offset between the carrier and reference signals	59
Figure 4.1	Transmitting dipole antenna with trapped rays as surface waves	63
Figure 4.2	(a) Slot on a dielectric (b) Dipole on a dielectric (c) E-Plane and H-Plane radiation patterns for slot and dipole antennas	64
Figure 4.3	Power distribution for an elementary dipole vs dielectric constant	65
Figure 4.4	Ray optics picture of surface-wave propagation	67
Figure 4.5	Propagating modes for a substrate with $\epsilon_r = 12.0$	67
Figure 4.6	Typical surface wave radiation patterns for TE and TM modes	68
Figure 4.7	E-plane pattern into dielectric for a Hertzian electric dipole	71
Figure 4.8	H-Plane pattern into dielectric for a Hertzian electric dipole	71
Figure 4.9	Perpendicular dipole situated in medium n_1, (ie. GaAs)	72
Figure 4.10	Finite Element Grid	77
Figure 4.11	Radial Poynting vector for dipole at 10 GHz	78

Figure 4.12	Radial Poynting vector for dipole at 1 GHz	79
Figure 4.13	Radial Poynting vector for dipole at 100 GHz	80
Figure 4.14	Nonlinear polarization induced in the optical waveguide	83
Figure 4.15	Experimental setup for CW difference-frequency generation	85
Figure 4.16	Measured output power vs input optical power	87
Figure 4.17	Frequency drift of one laser with respect to the other vs time (hours)	88
Figure 4.18	Detected microwave frequency vs tuned frequency offset	89
Figure 4.19	Detected microwave signal trace	90
Figure 4.19	Possible substrate lenses (a) hemispherical, and, (b) hyperhemispherical	92
Figure 4.20	Gold electrode pattern	93
Figure 4.21	Measurement test fixture	94
Figure 4.22	Generated power vs frequency for the nonlinear optical waveguide and a state-of-the-art MSM photodiode	96

Chapter 1.0 Introduction

1.1 Motivation

The efficient integration of RF technology with optical circuits has become a necessity due to the ever increasing demands placed on microwave communications systems. The distribution of analog and digital RF signals via fibre-optics is presently being seriously examined in several government and industry laboratories [1-3]. The key to the efficient solution of the above problem is the fibre-optic cable which is compact, lightweight, EMI-resistant, and has a large bandwidth capability. Although the bandwidth of the fibre-optic cable is very large, the frequency limits on the modulation rate of semiconductor lasers, which is presently at several tens of Gigahertz, severely restricts the amount of available optical bandwidth that can be utilized by any one laser. In an effort to utilize more of the available fibre-bandwidth, several lasers, each set to a specific frequency, can be combined and then transmitted over an optical cable in order to use more of the available fibre bandwidth. When the signals are demultiplexed at the receive end of the communications system, the signals can be downconverted by mixing the received signals with appropriate reference laser sources. However, an alternate method to wavelength-demultiplex a many-channel receive signal would be to use an optical spectrometer which transmits each channel to a separate portion of a

detector array. The advantage of spatially addressing different channels is that if the combined signal drifts, the output on the detector array only depends on the relative offset between each of the lasers and not on the actual absolute value of the optical carrier frequency.

In the area of phased array antennas, optical routing of receive and transmit signals using fibre-optics can lead to extremely high packing densities. Future military millimetre-wave phased arrays will typically require 1000 [4] elements and therefore, will pose a serious problem in terms of signal distribution if it were to be carried out exclusively at the microwave level. Typical microwave power divider and combiner networks are bulky, lossy and are bandwidth limited [5]. Therefore, the design constraints can be very severe for the case of a phased-array antenna feed network. Furthermore, since the microwave circuitry for these applications must have low power consumption and efficiency in terms of space utilization, most of the electronic processing circuitry (phase shifters, attenuators, digital switching circuitry) must be carried out using MMIC (monolithic microwave integrated circuit) technology. Typically, an MMIC function block will contain several active and passive circuits integrated onto a semi-insulating GaAs substrate, thereby providing high performance, small size, and low cost on a batch basis. GaAs MMIC's consist of MESFET's and Schottky-barrier diodes as the major active components, thin-film resistors, metal-insulator-metal overlay capacitors, and

spiral inductors as the passive component building blocks. Plated air bridges are used for circuit interconnections, and through-substrate via holes for ground interconnections [2]. Therefore it would be advantageous to ultimately combine the optical and microwave functions of phased array antennas on a single semiconductor chip, and therefore the motivation exists to fabricate efficient optical receive and transmit modules integratable with MMIC circuitry.

Much of the research and development taking place in photonics is being carried out on the mediums of gallium arsenide and indium phosphide, and, consequently, leads to the distinct possibility of combining micro/millimeter wave and photonic circuitry on one, or several GaAs or InP wafers. This could result in a reduction of the cost for systems that require a large number of identical circuits. At present, OEIC (optoelectronic integrated circuit) technology is attempting to provide the interface between the optical fiber and the GaAs MMIC's by combining GaAs lasers and photodiodes with MMIC MESFET-based mixer and amplifier circuitry. OEIC circuits have not matured yet, and are at least several years away from being used in commercial or military systems, largely due to the incompatibility of the fabrication techniques used for microwave and optical components.

A block diagram of a 4x4 optically fed phased array antenna is shown in Fig. 1.1 [6]. In this system, the intelligence at 1 GHz is impressed onto a 40

GHz carrier, which is in turn received and transmitted by each antenna array element via a fiber-optic feed.

Figure 1.1 Optically fed phased array antenna (from Herczfeld [6])

A general optically fed subarray is shown in Figure 1.2 [1]. It can be understood that for this proposed array there are four connections:

1. reference signal (to provide frequency and phase information)
2. information signal in

3. information signal out
4. control signals (to switch phase bits in and out).

The reference signal is used to provide the required frequency and phase information for the subarray. The information sent to the antennas must be processed through an optical receiver, and the information arriving from the antennas must be fed through an optical transmitter. Therefore, the optical downconverter developed in this thesis may become a key component for use in the optical receiver block. It will be shown that the nonlinear waveguide optical downconverter does not have the speed limitations of conventional optical receivers implemented using Schottky or PIN photodiodes. The control line is used to switch in and out different phase bits at each of the elements of the phased array. At the moment, this is typically being done by using MMIC FET/PHEMT (pseudomorphic high-electron mobility transistor) switches which are controlled by placing the appropriate voltage on the gate of the transistor. This could be undertaken by converting the optical control signal to an appropriate DC level using a photodiode [7], or having the optical signal directly control the semiconductor phase shifter or attenuator [8].

In addition to the advantages of large bandwidth and ease of routing a large number of signals, optically fed antennas also have the flexibility to allow optical processing to be undertaken. For example, they can generate true time

delays using optical delay lines in order to perform beam-forming [9]. This

Figure 1.2

Optically fed subarray (from Herczfeld [6])

would be particularly appropriate in wideband systems where a constant time delay rather than a phase shift would be required to steer the beam without dispersion. Although typically all of the signal processing (phase shifters, switching, etc.) is done individually at the antenna-radiating element, it may be more efficient to undertake some of the steps at the optical carrier level.

1.2 Objectives

The focus of this thesis will be to implement two kinds of optical demultiplexing systems based on the nonlinearity of optical waveguides. By injecting two optical signals so that they both propagate in the same direction (ie. copropagating), the first known optical downconverter at X-band, based on nonlinear waveguides, is used for demultiplexing signals, for use in phased array antenna receiving. Utilizing the same optical waveguide, but having the signals injected in the counterpropagating sense, an optical upconverter for DWDM applications with microwave frequency resolution will be demonstrated.

The downconverted frequency range that will be specifically investigated is the 10 to 30 GHz microwave range which will probably be the most common frequency allocation for future phased array antennas since MMIC technology is fairly mature at these frequencies. Also, a theoretical model will be developed for the radiation pattern of the Hertzian electric dipole induced in the optical waveguide utilizing the Lorentz reciprocity theorem, as well as modelling of the radiation pattern using a commercial antenna simulation program based on a finite elements algorithm to solve Maxwell's equation.

A similar waveguide can generate a sum frequency polarization which will be used to fabricate an optical spectrometer. The optical spectrometer is a potentially useful component in a DWDM (dense-wavelength demultiplexing) communications system, and can also be utilized as an optical deformation sensor based on the optical waveguide strain. Hence, demultiplexing can be achieved using both up- and down-conversion. In the case of down-conversion the demultiplexing is carried out directly at the microwave difference frequency, whereas for up-conversion, the demultiplexing is carried out utilizing a detector array which detects the different optical channels spatially.

1.3 Thesis Organization

Chapter 2 presents the theory of nonlinear optical waveguides which are investigated in this thesis. A discussion of sum-frequency and difference-frequency generation using these waveguides is then undertaken. Specific techniques previously used to generate difference frequencies at the far-infrared and microwave wavelengths are reviewed to provide a background for the present work. A theoretical calculation of the nonlinear cross-section of the optical waveguides for both sum- and difference-frequencies is shown, as well as the limiting factors in the overall efficiency of these frequency conversion processes. Chapter 3 discusses the nonlinear waveguide optical spectrometer

concept. The theoretical basis of the novel optical spectrometer is outlined, and its application in a DWDM optical communications system is examined. Measured results are compared with theory for the spectrometer and deformation experiments are verified. Chapter 4 compares the theoretical microwave output power of the downconverter to the actual measured power for the slab waveguide. Theoretical models for the radiation pattern of the induced polarization at the difference frequency are given, and a discussion of the implementation of the Electromagnetic Application Simulator (EMAS™) program, to model the radiation pattern of the induced dipole, is outlined as well. Chapter 5 draws conclusions based on this work and suggests possible future directions which could lead to greater efficiencies in the optical downconversion process.

Chapter 2.0 Overview of Relevant Nonlinear Optics and Optical Waveguides

In order to obtain optical to radio frequency conversion, the nonlinearity of some component, or set of components, must be optimized for an efficient mixing process to occur. Two possible methods are investigated. The first method considered is that of sum-frequency generation, which is to be discussed in the first part of this chapter. Following this, difference-frequency generation will be discussed as a means of generating a microwave frequency using two infrared beams offset from each other by a microwave frequency. These two cases are a subset of a large and diverse field known as nonlinear optics.

2.1 Nonlinear Optics

The relevant basics needed for the implementation of the various discussed devices is outlined in this section.

The induced polarization in a nonlinear material due to an applied electric field is given by

$$P = \chi^{(1)} E + \chi^{(2)} EE + \chi^{(3)} EEE + \dots \quad (2.1)$$

The $\chi^{(1)}$ term is the familiar linear susceptibility term, which relates the dielectric constant of the material to that of free space according to

$$\epsilon = 1 + \chi^{(1)} \quad (2.2)$$

The second-order nonlinear susceptibility term, $\chi^{(2)}$, leads to second harmonic generation, sum-frequency and difference-frequency mixing, as well as the electro-optic effect. The sum-frequency mixing is used to fabricate an optical spectrometer of microwave resolution by inputting two infrared wavelengths into a nonlinear waveguide and measuring the position of the sum-generated infrared beam. The difference-frequency mixing is used to generate microwave radiation at the offset frequency between two infrared laser beams. The polarization term due to $\chi^{(2)}$ is usually significantly less than the linear polarization term due to $\chi^{(1)}$, and therefore many of the effects attributed to the $\chi^{(2)}$ nonlinear susceptibility had not been observed until the advent of lasers which provided coherent radiation with a high electric field strength. The sum-frequency and difference-frequency cases will be elaborated upon later in the chapter, since a knowledge of these processes is fundamental to the various experiments described in this thesis. The third-order nonlinear susceptibility term, $\chi^{(3)}$, leads to the Kerr effect and the n_2 effect, which is an intensity-

dependent change in the refractive index of the nonlinear material. $\chi^{(3)}$ effects are not considered in the work of this thesis.

2.2 Optical Waveguides

The simplest optical waveguides used for the generation of the sum and difference frequency signals are planar and of the 'slab' variety, as shown in Fig. 2.1.

Figure. 2.1 Wave normals present in a slab waveguide (from Tamir [12])

The guiding takes place using three layers in which the index of refraction of the film, n_f , is greater than that of the substrate and cover, n_s and n_c respectively. The cover material is typically air having $n_c = 1.0$, and the film and semiconductor are made up of various semiconductor alloy systems such as AlGaAs/GaAs, InGaAs/AlGaAs or InAlAsP/InP. The film layer of the slab

waveguide can be grown using several different fabrication techniques. All of the AlGaAs/GaAs nonlinear optical waveguides utilized in this body of work were fabricated using molecular beam epitaxy (MBE) technology. It is also possible to fabricate slab waveguides using either polymers or glassware.

2.2.1 Molecular Beam Epitaxy

MBE technology can be utilized to grow various multilayer semiconductor structures such as optical waveguides, quantum well lasers, and MESFET-based transistors, among many other devices. MBE is an epitaxial process which involves the reaction of thermal beams of atoms or molecules with a crystalline substrate surface in an ultra-high vacuum chamber. A typical MBE growth can achieve an extremely high precision in the chemical profile and the doping content and thickness of the various layers.

Fig. 2.2 shows a schematic of a typical MBE growth system. The system shown is capable of growing either AlGaAs-based or InGaAsP-based semiconductor structures. Each of the elemental sources in the chamber is contained in a resistively heated crucible in order to produce a beam of thermally accelerated atoms. Each crucible contains a shutter in front of its opening in order to remotely turn on or off each of the beams as desired for each step of the growth cycle. The chamber containing the various beams is

kept under high vacuum conditions (approximately 10^{-10} Torr) in order that the atoms in the beams not be scattered by atmospheric molecules, and to also avoid contamination.

Figure 2.2 Schematic of an MBE growth system (from Hunsperger [10])

The extremely high purity of MBE systems is achieved by using three separate chambers: a loading chamber, a growth chamber, and an analysis chamber. This system of multiple chambers also increases throughput, since as one structure is being grown another can be prepared for growth in the preceding chamber. A well-designed MBE system should be able to achieve layer thickness of 1 % uniformity across a 2-inch wafer. As well layers less than 10 nm can be produced in order to fabricate quantum well devices such as semiconductor lasers [10,11].

2.2.2 Theoretical Analysis of the Slab Waveguide

Using ray optics, the propagation of light through a slab waveguide can be considered as the superposition of two uniform plane waves with wave normals which follow the zig-zag path shown in Fig. 2.1. Assuming that the waves are monochromatic with angular frequency, ω , they have a wave vector, kn , associated with them, where the absolute value of k is given by

$$k = \frac{2\pi}{\lambda} = \frac{\omega}{c} \quad (2.3)$$

with c the speed of light in a vacuum. The waves can be written as

$$Ae^{-jkn_x(\pm x \cos\theta + z \sin\theta)} \quad (2.4)$$

Only certain angles of θ are allowed in order for the propagating wave to be considered a guided wave (ie. nonradiating). This condition is given by the total phase shift for the wave in the transverse direction, ie. in the x -direction, being a multiple of 2π radians. This self consistency condition can be expressed mathematically, using the ray model, as

$$2kn_f h \cos\theta - 2\phi_s - 2\phi_c = 2\pi m \quad \text{where } m=0,1,2,\dots \quad (2.5)$$

where ϕ_s and ϕ_c are the phase shifts encountered at the reflection between the film and substrate and the film and cover respectively, and m refers to the

mode number. Equation (2.5) is the dispersion equation for a slab waveguide and yields the propagation constant, β , as a function of ω and film thickness, h , where $\beta = k n_f \sin \theta$.

Electromagnetic theory can be used to solve for the modes of a dielectric slab waveguide and to verify the intuitive ray optics picture just discussed. The Maxwell's equations to be solved for the sourceless and linear case are:

$$\nabla \times \vec{E} = -i\omega \mu \vec{H} \quad (2.6)$$

$$\nabla \times \vec{H} = i\omega \epsilon \vec{E} \quad (2.7)$$

The appropriate boundary conditions at the surfaces of materials where a change in material properties occurs, either in dielectric constant, ϵ , or magnetic permeability, μ , must be taken into account using:

$$\begin{aligned} \vec{e}_n \cdot (\vec{B}_1 - \vec{B}_2) &= 0 \\ \vec{e}_n \cdot (\vec{D}_1 - \vec{D}_2) &= 0 \\ \vec{e}_n \times (\vec{E}_1 - \vec{E}_2) &= 0 \\ \vec{e}_n \times (\vec{H}_1 - \vec{H}_2) &= 0 \end{aligned} \quad (2.8)$$

where $\vec{e}_n$ is a unit vector perpendicular to the surface boundary.

The solution for the slab waveguide can be found in any optical waveguide textbook such as those by Tamir and Yariv [12,13], and for the TE

modes is given by:

$$E_y = E_c \exp(-\gamma_c(x-h)), \quad \text{for } x > h \quad (2.9)$$

$$E_y = E_f \cos(\kappa_f x - \phi_s) \quad \text{for } 0 < x < h \quad (2.10)$$

$$E_y = E_s \exp(\gamma_s x) \quad \text{for } x < 0 \quad (2.11)$$

where,

$$\tan \phi_s = \frac{\gamma_s}{\kappa_f}, \quad \text{and} \quad \tan \phi_c = \frac{\gamma_c}{\kappa_f} \quad (2.12)$$

and equation (2.5) are found using the appropriate boundary conditions. As well, the following relation is obtained between the peak electric fields E_f , E_s and E_c where n_{eff} is the effective refractive index of the slab guide:

$$E_f^2(n_f^2 - n_{\text{eff}}^2) = E_s^2(n_f^2 - n_s^2) = E_c^2(n_f^2 - n_c^2) \quad (2.13)$$

where, $n_{\text{eff}} = \beta/k$

Most often optical waveguides are designed such that the zero-order TE mode is the dominant mode of the optical waveguide, and the higher order modes are designed to be below cutoff. The solution for the TM modes is solved by using a similar procedure as for the TE modes and can be found in [13].

The guides utilized for the sum-frequency generation are, however, multilayer waveguides and therefore have mode profiles which cannot be determined analytically. A computer program based on the numerical analysis of multilayer stack waveguides is therefore necessary and has been developed at the National Research Council for the determination of the various mode profiles for arbitrary multilayer waveguide structures. In certain cases, a three-layer model of the optical waveguide can lead to a good approximation of the mode profile, and can be used to estimate various properties such as the effective waveguide index of refraction.

2.3 Sum-Frequency Generation

Sum-frequency generation is induced in a material when an applied electromagnetic field is comparable in strength to the atomic fields in which the electrons sit, thereby exchanging energy between applied fields at different frequencies. The quadratic nonlinear optical susceptibility, $\chi^{(2)}$, which is directly responsible for sum-frequency and second harmonic generation, has been successfully modelled using the analogy of the anharmonic oscillator [14,15]. As well, various other optical phenomena such as difference-frequency mixing, optical rectification, the linear electro-optic effect, and parametric generation are due to the nonlinear $\chi^{(2)}$ effect. Only since the

advent of the laser have these effects been widely observed, due to the large power densities which are required.

Shown in Fig. 2.3 is the induced optical polarization dependence on the amplitude of the applied electric field in a nonlinear material. Hence, for large electric field strengths, $P^{(2\omega)}$ becomes comparable to $P^{(\omega)}$.

Figure 2.3 Induced polarization vs applied electric field (a) in a linear dielectric, and (b) in a crystal without inversion symmetry (from Yariv [13]).

The second order nonlinearity can generate a field at a frequency $2\omega_1$, where ω_1 is the frequency of the input field. As well, a sum frequency may also be created from two separate counter-propagating signals at frequencies ω_1 and ω_2 , yielding a sum signal, ω_3 , of frequency $\omega_1 + \omega_2$. The nonlinear susceptibility tensor, d_{ijk} , where $\chi^{(2)} = 2d_{ijk}^{(2)}$, is given by

$$P_i^{\omega_3} = 2d_{jk}^{\omega_3 - \omega_1 - \omega_2} E_j^{\omega_1} E_k^{\omega_2} \quad (2.14)$$

where P_i is the induced optical polarization. Similarly, the difference frequency tensor is given by

$$P_i^{\omega_3} = 2d_{jk}^{\omega_3 - \omega_1 - \omega_2} E_j^{\omega_1} E_k^{\omega_2} \quad (2.15)$$

Typically the nonlinear materials used are transparent over a large frequency range and hence the nonlinear susceptibility d_{jk} can be considered to be constant with frequency.

The Miller tensor can be written as a reduced two-dimensional 3 x 6 matrix

$$d_{jk} = \begin{bmatrix} d_{11} & d_{12} & d_{13} & d_{14} & d_{15} & d_{16} \\ d_{21} & d_{22} & d_{23} & d_{24} & d_{25} & d_{26} \\ d_{31} & d_{32} & d_{33} & d_{34} & d_{35} & d_{36} \end{bmatrix} \quad (2.16)$$

Several materials which have been used to generate harmonics are LiNbO_3 , various organic agents, and III-V compounds such as GaAs, AlGaAs, and InP. The first researchers to carry out experiments on GaAs slabs were Anderson and Boyd [16] in the early 1970s. For the case of GaAs, which is a tetragonal crystal, its nonlinear polarization can be calculated using the d_{jk} matrix as

$$\begin{bmatrix} P_x \\ P_y \\ P_z \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & d_{14} & 0 & 0 \\ 0 & 0 & 0 & 0 & d_{14} & 0 \\ 0 & 0 & 0 & 0 & 0 & d_{14} \end{bmatrix} \cdot \begin{bmatrix} E_x^2 \\ E_y^2 \\ E_z^2 \\ 2E_y E_z \\ 2E_x E_z \\ 2E_x E_y \end{bmatrix} \quad (2.17)$$

Equation (2.17) demonstrates that with a (100) GaAs crystal, a mixed TE-TM mode interaction is required to induce a nonlinear polarization, whereas, for a substrate oriented with the (111) plane, only a pure TE electric field would be required for the same effect. This is evident since the E_x^2 , E_y^2 , and E_z^2 terms of the resultant polarization are zero when the matrices are multiplied, whereas the cross terms $E_y E_z$, $E_x E_z$, and $E_x E_y$ have non-zero values.

Much work in nonlinear optics has been carried out using GaAs semiconductors since they have a relatively high value for their second order nonlinear optical susceptibility and a high transparency at the standard Nd:Yag and Ti:Sapphire wavelength ranges. Also, for second-harmonic and sum-frequency processes, the losses due to phonons are negligible when the input wavelength is in the infrared range. However, bulk GaAs material has a

Reststrahlen band centered at $37\text{ }\mu\text{m}$ due to transverse optical (TO) phonons, as well as multiphonon processes on either side of the band. Hence, generation of a far-infrared (FIR) frequency around 8.1 THz is highly inefficient due to high absorption at this frequency. Similarly, interband absorption at the sum-frequency can have a significant effect on the generated power since the energy of the sum-frequency photon will typically be greater than the bandgap of the AlGaAs nonlinear material. Finally, absorption due to free carrier effects at millimeter wave frequencies can have a like effect on the final power level of the generated microwave frequency. These influences will be elaborated upon when the results for the copropagating difference frequency generation in the nonlinear optical waveguide are discussed in Chapter 4.

2.3.1 Enhanced GaAs Waveguide Structure for Efficient Second-Harmonic Generation

Stegeman and Normandin first demonstrated nonlinear mixing of two counter-propagating waves in a Lithium Niobate waveguide [17]. The counter-propagating $1.06\text{-}\mu\text{m}$ Nd:YAG laser beams were coupled into the LiNbO_3 waveguide using rutile prisms. The radiated second-harmonic field at $0.53\text{ }\mu\text{m}$ was emitted perpendicular to the substrate, as was expected for the situation of counter-propagating waves of equal frequency in a waveguide geometry. A

liquid, with an arsenic-selenium base, placed on the surface of the waveguide, was used to extend the evanescent TE input electric field further out into the liquid, thereby generating a greater sum-frequency field by adding the generated polarizations constructively in the far-field of the waveguide [18].

The output signal power at $0.53 \mu\text{m}$ was enhanced by approximately 500 by placing the phase matching liquid on the surface of the

Figure 2.4 Second harmonic generation from two counter-propagating waves on a LiNbO_3 substrate (from Vella et al. [18])

waveguide. The above setup could easily be used in many types of important applications such as a picosecond processor [19], an optical transient digitizer [20], or an optical spectrometer [21].

An important parameter which is used to quantify the efficiency of various nonlinear materials for second-harmonic and sum-frequency generation is the nonlinear cross-section, A_{ω} . It is defined as

$$I_{\omega_1+\omega_2} = A_{\omega} I_{\omega_1} I_{\omega_2} \quad (2.18)$$

where the I 's represent power in W, and A_{ω} has units of W^{-1} , normalized to a waveguide having an interaction length of 1 cm, and a beam width of 1 mm. Equation (2.18) shows that the efficiency of the sum-frequency generating process goes as the square of the input power. Recently, it has been demonstrated that a limiting nonlinear cross-section of 10^{-6} W^{-1} [22], is feasible for a waveguide thickness approaching zero at a wavelength of $1.06 \mu\text{m}$, which is significantly higher than the value of $6 \times 10^{-13} \text{ W}^{-1}$ for the setup of Fig. 2.4, with the same parameters. Much of the effort in efficient second harmonic generation has concentrated on using GaAs and other III-V compounds since the technology required to grow the necessary layers for quasi-phase matching can be easily accomplished using MBE technology. Values for the nonlinear cross-section of bulk GaAs can be viewed in Fig. 2.5 [14] for different thicknesses of GaAs. The envelope of the A^{ω} curve of GaAs demonstrates that

as the slab thickness increases, the power of the emitted radiation decreases. With the counter-propagating waveguide structure being discussed here, there is no phase matching possible and, therefore, the resultant fields are much weaker than those established by traditional second harmonic generating devices.

Figure 2.5 Calculated A' vs slab thickness for bulk GaAs not including absorption (from Normandin et al. [22])

The second harmonic generation efficiency of a bulk GaAs crystal can be improved by properly constructing a multilayer $\text{Al}_x\text{Ga}_{1-x}\text{As}$ structure. By performing an analysis for a general waveguide structure, it can be optimized for the highest efficiency.

Assuming that the SHG nonlinear waveguide is being pumped from either side, as shown in Figure 2.4, and that each separate medium is isotropic, then the distribution of the input fields at ω can be obtained by solving Maxwell's equations, along with the interface boundary conditions. The appropriate starting equation is therefore

$$\nabla^2 \vec{E}(\vec{r}) + k^2 n^2(\vec{r}) \vec{E}(\vec{r}) = 0 \quad (2.19)$$

where n is the index of refraction in the medium of concern, and k , the vacuum wave vector. Once, the solution of the input guiding fields at ω is obtained, then the second-harmonic polarizations at 2ω can be used as the source term in the Maxwell's equation for the generated second harmonic field. Using this two-step analytical approach decouples the nonlinear generated fields from the two input fields resulting in one independent set of Maxwell's equations to be solved for the input fields and then solving a separate set of equations for the output field. The approach is valid for sum- and difference-frequency generation cases using the nonlinear AlGaAs waveguides since the input fields are never depleted enough to require the coupling of the Maxwell's equations.

Figure 2.6 General waveguide geometry (from Normandin et al. [22])

The solution, for the case of the TE_0 mode, is given by

$$E_z = E_{eff} e^{-\alpha z} \quad (2.20)$$

$$E_z = E_{guide} \left(\cos(k_1 y) - \frac{q}{k_1} \sin(k_1 y) \right) \quad (2.21)$$

$$E_z = E_{substrate} \left(\cos(k_1 h) + \frac{q}{k_1} \sin(k_1 h) \right) e^{p(y+h)} \quad (2.22)$$

where k_1 is the normal component of the electric field, E_z , and p and q are real numbers. The guide solution is therefore

$$E(\vec{r}, t) = E(\vec{y}) e^{i(\omega t - k_1 x)} \quad (2.23)$$

where $n_{eff} = k_1/k$. The derivation for the TM_0 mode follows the same procedure,

and is fully described by Yariv [13].

Utilizing equations (2.17) and (2.23), the nonlinear induced polarization can be obtained as

$$P_z^{nl}(y) = \epsilon_0 d_{14} E_1(y) E_2(y) (e^{2i\omega t} + e^{-2i\omega t}) \quad (2.24)$$

Assuming the nonlinear induced optical polarization drives the Maxwell's equations, one obtains,

$$\nabla \times \vec{H}(2\omega) - i\omega \epsilon \vec{E}(2\omega) = i\omega \vec{P}_N \quad (2.25)$$

$$\nabla \times \vec{E}(2\omega) + i\omega \mu_0 \vec{H}(2\omega) = 0 \quad (2.26)$$

The method used to solve the above problem is based on a surface Green's function, first developed by Sipe [23]. It should be mentioned that the above approach is based on classical electromagnetic theory for which the Maxwell's equations are not quantized. This method yields an answer which is accurate for high input field strengths which are typically utilized with second harmonic generation devices. However, the classical approach has limitations which must be properly understood in order that the obtained solutions do not become misleading. For instance, when either of the input field strengths or the sum-

frequency field strength have energies which are so small that classical theory cannot adequately model the interaction, the Maxwell's equations must first be quantized, and then consequently solved.

The approach taken here is to consider each polarization as a thin dipole sheet parallel to the waveguide surface, and then to integrate the effect of each contributing dipole sheet, according to the amplitude, position and phase of the induced nonlinear polarization, P_n . The solution for the electric field of the emitted radiation at the surface of the waveguide is [22],

$$\vec{E} = -i\omega \sqrt{\frac{\mu_0}{\epsilon_0}} \frac{n}{n+1} \int \frac{P_n(y')}{n(y')} e^{-ik_0 \int_0^{y'} n(y'') dy''} dy' \cdot \hat{k} \quad (2.27)$$

Using the Poynting vector in order to obtain the radiated field intensity,

$$\langle S \rangle = \frac{1}{2} \text{Re}(\vec{E} \times \vec{H}^*) = \frac{1}{2} \omega^2 \sqrt{\frac{\mu_0}{\epsilon_0}} \left(\frac{n}{n+1}\right)^2 S S^* \hat{y} \quad (2.28)$$

where,

$$S = \int \frac{P_n(y')}{n(y')} e^{-ik_0 \int_0^{y'} n(y'') dy''} dy' \quad (2.29)$$

It can be seen that equation (2.29), assuming a constant $n(y)$ term, is just the spatial Fourier transform of the nonlinear induced polarization.

Fig. 2.5 illustrates that unless the waveguide thickness is less than one half of the second harmonic wavelength, then the interference of the radiated second-harmonic field causes the far-field pattern to approach zero. One way to eliminate the undesirable interference effects of the uniform slab waveguide geometry is to have alternating linear and nonlinear layers, so that in the far field, the fields generated from each layer would sum together constructively. The problem with this technique is its complexity in terms of practical fabrication. A more practical method would be to change the index of refraction of the alternating layers in order to optimally modulate the phase of the induced polarization, instead of the nonlinear coefficient. The approach followed was to use a combination of different nonlinear susceptibility and indices of refraction for the alternating layers by selectively controlling the level of aluminum in the multilayer GaAs waveguide.

A structure with alternating layers of $Al_{0.9}Ga_{0.1}As$ and $Al_{0.7}Ga_{0.3}As$ of 1/2

wavelength thickness, in order to obtain quasi phase matching at the second harmonic wave, can be seen in Fig. 2.7. The change in aluminum concentration with each of the layers was taken into account since this affects the nonlinear susceptibility as well as the absorption of the generated signal. For wavelengths greater than $1.0 \mu\text{m}$, $\chi^{(2)}$, and absorption, α , are inversely proportional to aluminum concentration. Taking all of these various factors into account, it is found that this structure yields an improved nonlinear cross-section of 10^{-5} W^{-1} at a waveguide thickness of $1 \mu\text{m}$. As shown in Fig. 2.8, the overall cross-section of a multi-layered waveguide is almost independent of waveguide thickness. This design is found to be very tolerant to variations in layer thickness or index of refraction.

Figure 2.7 Multilayer structure of GaAs SHG device (from Normandin et al. [22])

Figure 2.8 Nonlinear cross section for a 90%-70% AlGaAs multilayer waveguide (from Normandin et al. [22])

From Fig. 2.8 it can be affirmed that the alternating structure theoretically provides a nonlinear cross section that is improved 3 million fold over bulk GaAs material [22]. In the experimental setup that was used for testing the various multilayer waveguides, the 1.06 μm Nd:YAG lasers were pulsed at 7.7 kHz, with a pulse width of approximately 200ns. 70 μW of average power was successfully coupled into the waveguide, and the emitted green light was easily visible across a room. Even when both lasers were running CW, the generated signal could be observed. Taking into account that

the TE and TM group velocities differ, a propagation interference effect was established; this effect can be likened to a set of beads, as depicted in Figure 2.9. The effect is predicted by observing the second term of equation (I.4) in Appendix I.

Figure 2.9 Observed bead pattern on (100) GaAs substrate

2.4

Difference Frequency Generation

In principle, difference frequency generation is related to second harmonic or sum frequency generation. It can be used to produce tunable infrared radiation from two optical sources, and, in the low frequency limit, it can provide the means for microwave signal generation.

Whereas second harmonic generation can theoretically be a 100% efficient conversion process, the same cannot be said for difference-frequency generation. Assuming that difference-frequency generation takes place at ω_3 using the input frequency ω_1 and ω_2 according to $\omega_3 = (\omega_1 - \omega_2)$, then the Manley-Rowe relations [24] for the power exchange between the different signals is given by:

$$\frac{\Delta P_{\omega_3}}{\omega_3} = \frac{\Delta P_{\omega_2}}{\omega_2} = -\frac{\Delta P_{\omega_1}}{\omega_1} \quad (2.30)$$

where ΔP refers to the change in power at the frequency indicated. Therefore a photon at frequency, ω_1 , is split into two photons at frequencies ω_2 and ω_3 . Therefore the process results in parametric amplification of the signal at ω_2 as well as the generation of the difference frequency at ω_3 . Hence, the maximum conversion efficiency from ω_1 to ω_3 is given by

$$\eta_{\text{eff}} = \frac{\omega_3}{\omega_1} \quad (2.31)$$

The above equation shows that the process of generating a 10-GHz difference frequency from an optical signal at $1.32 \mu\text{m}$ can have a maximum efficiency of 0.0044%, assuming 100% quantum conversion efficiency. Using standard $1.32 \mu\text{m}$ tuneable YAG lasers with an output power of 150 mW, the generated microwave signal can be no greater than $6.6 \mu\text{W}$! A conventional heterodyne detection scheme wherein two laser signals are used to obtain a microwave signal from a state-of-the-art Schottky photodiode results in several mW's of microwave power, which translates to an efficiency of several per cent.

2.4.1

Pulsed Microwave Generation

Radiation can be obtained in the microwave region by illuminating a bulk GaAs surface with a femtosecond optical pulse. Hu, Zhang and Auston [25] have demonstrated this principle by successfully detecting terahertz radiation via a photoconducting dipole. Through their research, it has been demonstrated that when a $\langle 100 \rangle$ undoped GaAs wafer is placed in a cryogenic cooler, conjointly with a dipole detector 2 cm from the sample, far-field radiation can be detected. The tunable laser source (815-870 nm), which irradiated the

surface of the sample, was pumped at 76 MHz, producing an average power of 80 mW, in addition to a 100 fs pulse width. The resultant measured radiation can be viewed in Fig. 2.10 below.

Figure 2.10 Radiated field from GaAs sample for various radiated photon energies with respect to the band edge (from Hu et al. [25])

For excitation which takes place below the band gap, only a reactive photocurrent is generated by the sample. The third-order nonlinear polarization of these virtual photocarriers is determined by the following expression,

$$P^M = \chi^{(3)}(0,0,-\omega_{op},\omega_{op}) E_{op} E_{op}^* E_{dc} \quad (2.32)$$

where $\chi^{(3)}(0,0,-\omega_{op},\omega_{op})$ is the nonlinear susceptibility due to the inverse Franz-Keldysh effect. The virtual photoconductivity is due to an optical field which changes the dc dielectric constant in a bulk semiconductor [26]. In the case of quantum wells, it is known as the ac-dc Stark effect. Hence, the above technique can be used to produce THz radiation from a narrow laser pulse, yet at the same time allows for the preservation of the phase information of the input signal, as well as its amplitude.

Figure 2.11 Experimental setup used for RF generation (from Smith et al. [27])

Another recently developed technique that has been used to generate microwave radiation was carried out by Smith, Auston, and Nuss [27] at AT&T Bell Labs. This technique was based on the illumination of a photoconducting antenna with femtosecond optical pulses. A diagram of the experimental setup can be viewed in Fig. 2.11. The planar dipole antennas were fabricated on silicon-on-sapphire substrates and were radiation-damaged in the area of the dipoles in order to produce a photoconducting effect. The transmitting substrate was illuminated by a 100 MHz pulsed laser, of 120 fs duration, and 100 pJ pulse energy. The photoconductor at the transmitter served as a current source to excite the dipole antenna. The received electromagnetic signal was then sampled by a delayed optical pulse via the photoconductor on the receiver substrate. In this way, the received pulse characteristics could be determined. A plot showing the temporal response of a 200 μm dipole is illustrated in Fig. 2.12. The peak average current induced in the antenna was 6.5 pA.

Fattinger and Grischkowsky [28] at the IBM Research Center have generated terahertz radiation in much the same manner as the previous group, by means of the transmitting and receiving photoconducting dipoles. They have modified the standard arrangement of Smith et al. by locating the dipoles at the focal point of a collimating lens, thereby producing diffraction-limited single cycle beams of 0.5 THz frequency. A diagram of the collimating and

focussing optics used, can be viewed in Fig. 2.13.

Figure 2.12 Temporal response of 200 μm planar antenna (from Smith et al. [27])

Fattinger obtained similar results to Smith in terms of the received signal and, therefore, will not be delved into any further.

The pulsed microwave generation experiments discussed here are useful for characterization of the physical properties of semiconductors, such as carrier lifetime and mobility, and also in the spectroscopic study of various gaseous molecules. However, due to the very short pulse widths, and consequently, large spectral range of the generated microwave signals, their use in communications and phased array antenna systems is quite impractical.

Quasi-CW and CW generated microwave radiation, discussed in the next section, is much more amenable to implementation in modern communications systems.

Figure 2.13 Diffraction limited microwave generation and detection (from Fattinger and Grischkowsky [28])

2.4.2 CW and Quasi-CW Microwave Generation

The first quasi-CW generation of far-infrared signals was undertaken by Zernike and Berman [29] in 1965. Radiation at $100\ \mu\text{m}$ (300 GHz) was obtained using quartz as the nonlinear medium and a pulsed Nd:glass laser as the source. Later, using two Q-switched CO_2 lasers at wavelengths of $9.6\ \mu\text{m}$ and $10.6\ \mu\text{m}$, Zernike [30] generated far-infrared (FIR) generation at wavelengths of $100\ \mu\text{m}$ (i.e. 3 THz) in a crystal of InSb. The first generation of infrared radiation using a GaAs crystal for the nonlinear medium was undertaken by Aggarwal, Lax, and Favrot [31,32] using TEA CO_2 lasers for the input sources. Recognizing that conservation of momentum must take place for maximum FIR radiation, they utilized a noncollinear phase matching technique. The index of refraction of the GaAs crystal was determined at the appropriate wavelengths, and the necessary external phase matching angle was consequently used in the experimental setup. Using the discrete lines of the CO_2 laser, a tunable difference frequency source from 150 GHz to 4.3 THz was obtained (2 mm to $70\ \mu\text{m}$). The mechanism used to obtain the mixing in the GaAs crystal was the nonlinear susceptibility, $\chi^{(2)}$. A schematic diagram of the setup can be viewed in Fig. 2.14. Later, a more sophisticated crystal arrangement, termed a 'folded' geometry, was utilized in order to achieve orders of magnitude increase in the output infrared power [33,34].

Figure 2.14 Experimental setup for noncollinear phase matched FIR generation (from Aggarwal et al. [33])

Another technique used to phase match the wave vectors of a difference-frequency generating process is to use a dielectric waveguide phase matching technique, as demonstrated by Thompson and Coleman [35]. Here, the input signals are both collinear, but phase matching is obtained by using the

dispersion of a planar GaAs waveguide to cancel the bulk dispersion of the GaAs crystal. Utilizing two Q-switched CO₂ lasers having 3 kW peak power, an output FIR power slightly less than a milliwatt was obtained at a wavelength of 240 μm (1.25 THz) yielding an efficiency of approximately 1.7×10^{-7} .

As well, narrowband microwave signals can be obtained by mixing two optical signals on a fast photodetector as demonstrated by Goldberg, et al. [36,37]. Direct current modulation of a laser, FM sideband optical injection locking, and offset phase locking can be used to generate the carrier signal, however, the downconversion of the optical signal to a microwave frequency is limited by the speed of present day fast photodetectors (presently several hundred GHz). By carrying out the optical downconversion directly in a nonlinear waveguide the inherent speed limitations of a photodiode are bypassed, potentially allowing operation at much higher frequencies.

Chapter 3. Sum-Frequency Generation for Application in a Wavelength Demultiplexing System with Microwave Resolution

3.1 Introduction

Two counterpropagating modes in a nonlinear waveguide can be used to generate sum-frequency light and thereby carry out certain signal processing functions such as dense wavelength demultiplexing (DWDM), optical autocorrelation, as well as time-division multiplexing. The DWDM application is based on the waveguide optical spectrometer, which, it has been found, can also be used as an optical deformation sensor for measuring strain in nonlinear waveguides.

This chapter of the thesis will demonstrate the construction of a 10-GHz resolution waveguide spectrometer using a 1-cm long GaAs-based optical waveguide. In seeking microwave resolution from the optical spectrometer, the effect of both extrinsic and intrinsic waveguide nonuniformities on the obtained resolution are shown via their effect on the far-field diffraction pattern of the radiating substrate aperture. It will be demonstrated how the far-field diffraction pattern can be used to monitor strain, as well as waveguide uniformity.

3.2 Theory of Surface-Emitting Nonlinear Waveguide Optical Spectrometer

In a nonlinear waveguide having two counterpropagating light beams, at frequencies ω_1 and ω_2 , the direction of the sum-frequency generated beam can be determined by utilizing conservation of wave vectors as given by

$$\vec{k}_{\omega_1} + \vec{k}_{\omega_2} = \vec{k}_{\omega_3} \quad (3.1)$$

where ω_3 is the generated sum frequency. For the geometry shown in Fig. 3.1, the angle of emission of the sum frequency beam relative to the surface of the waveguide, θ , is given by

$$\theta = \cos^{-1} \left(\frac{\beta_{\omega_1} - \beta_{\omega_2}}{k_{\omega_1 + \omega_2}} \right) \quad (3.2)$$

$$\beta_{\omega_1} = \frac{n_1 \omega_1}{c} ; \quad \beta_{\omega_2} = \frac{n_2 \omega_2}{c} ; \quad k_{\omega_1 + \omega_2} = \frac{\omega_1 + \omega_2}{c} \quad (3.2a)$$

where the β 's refer to the input wave vectors, n_1 and n_2 refer to the effective index of refraction at the input frequencies, and $k_{\omega_1 + \omega_2}$ refers to the wave vector of the sum-frequency generated light. Hence, the angle of emission is proportional to the difference in frequency between a reference and carrier signal. This can be measured by placing a CCD array in the far-field of the waveguide radiating aperture. The case shown in Fig. 3.1 is for emission into vacuum. The angle of emission into the substrate, θ_{sub} , will be slightly larger

due to the index of refraction of the substrate. Most of the substrate directed component is absorbed by the GaAs semiconductor material.

Figure 3.1 Propagation of incident and sum-frequency wave vectors

The resolving capability of an ideal spectrometer is given by the Rayleigh criterion [38],

$$\frac{\lambda}{\Delta\lambda} = \frac{nL}{\lambda}, \text{ or, } \Delta\nu = \frac{c}{nL} \quad (3.3)$$

where n is the effective index of refraction of the waveguide mode, L is the interaction length along the waveguide, and λ is the wavelength of the sum-frequency light. It is based on the diffraction of an optical beam from an aperture of length, L , and yields the minimum separation in frequency between

two optical beams in order to differentiate them in the far-field, so that the centre of the main diffraction lobe of one beam falls on the zero of the diffraction pattern of the other beam. The above equation shows that resolving capability increases with interaction length. For the case of a 1-cm long waveguide the resolution is 0.06 nm, or 10 GHz in frequency and this will be shown to agree with the measured result.

If there are any stress-induced variations in the waveguide surface due to extrinsic inhomogeneities, such as bending of the waveguide, then the aperture having the length of the waveguide and the width of the optical mode cannot be considered a line source with uniform phase. The aforementioned also applies for any intrinsic inhomogeneities, such as changes in thickness or material composition along the waveguide.

The modelling of the inhomogeneities can be undertaken by assuming that a nondistorted planar surface exists and adding on an appropriate phase variation, $\phi(x)$, to the sum-frequency polarization along the length of the waveguide. For the waveguide used in these experiments it was found that the Fresnel diffraction integral was sufficiently accurate (condition (4-8) of [38], ie. the Fresnel approximation) to model the sum-frequency far-field pattern, provided that the CCD array was far enough from the aperture [38]. Assuming that the focussing lens was placed in front of the GaAs waveguide and that the

paraxial approximation is valid, the Fresnel integral to be solved is given by,

$$U = \int_{-l}^l P \exp[i\phi(x)] \exp\left(-i \frac{2\pi}{\lambda f} x x_f\right) dx \quad (3.4)$$

where x is the displacement along the guide, P is the amplitude of the sum-frequency polarization, f is the focal length of the lens, and x_f is the displacement in the focal plane of the lens. Assuming that the phase distortion is relatively small, it can be adequately modelled by the Taylor series expansion

$$\phi(x) = K_0 + K_1 x + K_2 x^2 + K_3 x^3 \quad (3.5)$$

Consequently, the diffraction integral can be solved for several different phase variations.

Fig. 3.2 Far-field diffraction measurement setup

A linear phase distortion term across the guide, due to the $K_1 x$ term of equation (3.5), produces a sum-frequency output in the far-field that is identical to that of the distortionless case but whose diffraction pattern is translated in the x-direction of the focal plane by Δx_f , according to

$$\Delta x_f = \frac{\lambda f K_1 L}{2\pi} \quad (3.6)$$

This effect can be observed in Fig. 3.3(a) compared to the case where there is no phase distortion.

Figure 3.3(a) Far-field diffraction patterns for the distortionless and linear phase distortion cases

Figure 3.3(b) Far-field diffraction patterns for the distortionless and quadratic phase distortion cases

A quadratic phase variation across the waveguide, modelled using the $K_2 x^2$ phase term of equation (3.5) increases the intensity of the adjacent first-order sidelobes, and fills in the nulls between the various sidelobes. An inhomogeneity leading to cubic phase variation causes the far-field radiation

pattern to become asymmetrical, where sidelobes to one side of the main lobe decrease in amplitude, and those to the other side increase in amplitude, as shown in Fig. 3.3(c). Therefore, it can be concluded that the type of distortion observed in the far-field diffraction pattern can facilitate the identification of certain types of waveguide inhomogeneity.

Figure 3.3(c) Far-field diffraction patterns for the distortionless and cubic phase distortion case

3.3 Experimental Setup and Far-field Diffraction Measurements

The waveguide structure used in this experiment was designed to satisfy the quasi-phase-matched condition for second-harmonic generation of 1.32- μm light [39,40]. It consists of six layers of alternating 0.1- μm $\text{Al}_{0.8}\text{Ga}_{0.2}\text{As}$ and 0.081- μm GaAs, as well as AlGaAs cladding layers. A ridge was etched in the 1-cm-long waveguide in order to ensure that the coupled mode does not diverge laterally. Ridge widths were between 6 and 10 μm , and were etched down 10 nm. From the $\chi^{(2)}$ symmetry for the (100)-grown $\text{Al}_x\text{Ga}_{1-x}\text{As}$, both TE and TM modes are required for sum-frequency generation.

Figure 3.4 $\text{Al}_x\text{Ga}_{1-x}\text{As}/\text{GaAs}$ waveguide structure

The optical spectrometer setup, illustrated in Fig. 3.2 contains two

tunable, 1.32- μm , Nd:YAG ring laser sources, each with a 150-mW cw output power. Each of the sources has a 5-kHz linewidth and is independently tunable over a 35-GHz frequency range by a stable temperature-controlled cavity. The beams are coupled into fibers and then injected into either side of the nonlinear waveguide. In this experiment, the coupling parameter from the optical fiber to the waveguide is approximately 1 %, and the nonlinear cross-section, A_{nl} , is $3.9 \times 10^{-5} \text{ W}^{-1}\text{cm}^{-1}\text{mm}^{-1}$. The power of the reference and 'carrier' signals in the waveguide is 80 μW , and the generated sum-frequency signal is approximately 10 pW. The 'carrier' signal is offset from the reference signal by a microwave frequency of up to 30 GHz. The term carrier is used here because the spectrometer can also be utilized in an optical wavelength demultiplexing communications system. A DWDM system can be constructed using the nonlinear waveguide as demonstrated in Fig. 3.5. If two or more carrier signals, ω_1 , ω_2 , and ω_3 , at slightly different offset frequencies, are injected into one side of the waveguide, and the other has a reference frequency source, ω , then each of the sum-frequency beams from the different interactions is directed to a different area of the detector array. The minimum spacing of the elements in the array can be found utilizing equation (3.1), (3.2), and the focal distance to the detector array.

Figure 3.5 Optical Waveguide DWDM Communications System

In this experiment, the far-field diffraction pattern is imaged onto the CCD array by a lens with 494-mm focal length. The reference beam is held at a constant frequency, while the carrier is independently tuned over its 35-GHz frequency range with a temperature tuned cavity. In order to obtain a frequency offset curve between the two lasers, one of the lasers was held at a constant temperature while the other was tuned over its complete 30 degree Celsius range, and the signal was combined on a fast InGaAs photodiode and the difference frequency measured on an HP70000 spectrum analyzer.

Figure 3.6 Variation of the Offset Frequency Between the Two Ring Lasers with Laser Cavity Temperature

Fig. 3.6 shows the variation in the frequency offset between two Nd:YAG ring lasers, one of which is held at a constant cavity temperature of 20 °C, and the other, which is varied from 20 °C to 50 °C. The curve of Fig. 3.6 could be used to accurately 'dial in' a frequency offset between the two lasers for both the sum- and difference- frequency experiments to within a resolution of MHz. It can be noticed that there are certain cavity temperatures for which the lasers have two offset-frequencies at the same time. These are regions where the temperature-tuned laser hops from one of its cavity modes to a higher or lower

order cavity mode. The spacing between the cavity modes is approximately 10 GHz as can be verified from Fig. 3.6.

The curve of Fig. 3.7 shows the output angle of the sum-frequency generated beam in terms of the input offset frequency which was obtained from the curve of Fig. 3.6. The error bars signify the width of the sum-frequency generated beams in the far field in terms of μrad 's. Hence, if the input beam held at the constant frequency is considered to be the reference beam, and the input beam which is varied in frequency is considered to be the 'carrier', then it can be seen how the nonlinear waveguide can be used as an

Figure 3.7 Sum-Frequency Emission Angle vs Input Offset Frequency
for a 1-cm Long Waveguide

optical spectrometer. In this case the spectrometer has a variation of $6.86 \mu\text{rad}/\text{GHz}$. As well, there is the possibility of DWDM applications where each channel transmitted by a fiber would be routed to a different photodetector section on the CCD array.

Figure 3.8 Far-field diffraction pattern for a 34-GHz offset between the carrier and reference signals

Fig. 3.8 shows the far-field diffraction pattern for a 34-GHz offset. The peak at 0 GHz is due to the mixing of the reference beam with a second beam at the same frequency. Here the second beam is provided by facet reflection. The peak at 34 GHz represents the mixing term between the reference beam

and the carrier beam. The measured linewidth (obtained using the calibration curve of Fig. 3.8) is 10 GHz which is in close agreement to the diffraction-limited ideal case for a 1-cm long waveguide.

From Fig. 3.8 it can be seen that the measured linewidth of the different far-field mixing terms is approximately 10 GHz, as expected from equation (3.3) which gives the result for the linewidth resolution due to the ideal diffraction-limited case. The plot of far-field intensity for the case of a dc offset is shown in Fig. 3.9. It can be seen that there are lobes to the right of the main peak, and an absence to the left of the peak. This pattern demonstrates that there is an inhomogeneity in the waveguide which is causing the distortion in the far-field diffraction pattern. In this case, it was determined that there was a cubic phase distortion, along the waveguide of $K_3 = 2 \times 10^{-11} \mu\text{m}^{-3}$, by solving the diffraction integral of equation (3.4) using equation (3.5) [41]. This phase distortion along the waveguide corresponds to a displacement of 2 μm between the ends of the waveguide and was due to the warping of the sample due the adhesive used in the mounting procedure.

Figure 3.9 Far-field diffraction pattern for a dc offset between the carrier and reference signals

3.4 Summary

The present subnanometer resolution limit of the optical waveguide could be enhanced by simply using a waveguide with a longer overall length. The measured resolution limit of the optical spectrometer is approximately that of a conventional grating spectrometer having a 0.25-m focal length and a 1200-groove/mm grating. It should be mentioned that Guy et al. used 2-mm long waveguides and a numerical fitting procedure in order to measure frequencies

to an accuracy of 200 MHz, but for spectrometry and DWDM demultiplexing the channel spacing as dictated by the resolution is determined only by the diffraction-limited linewidth [42,43].

From the experimental measurements shown in chapter 3.3 it can be appreciated that small distortions due to bonding procedure can produce a stressed beam deformation in the waveguide that can have a significant effect on the generated diffraction pattern. A similar effect can also be obtained due to an intrinsic inhomogeneity, such as a varying effective index of refraction along the length of the waveguide. For instance, if the waveguide thickness were to vary quadratically along the guide by only 1 %, the same effect that was demonstrated by the stressed beam deformation would again be apparent.

Chapter 4.0 CW Microwave Generation by Optical Nonlinear Mixing

4.1 Introduction

This chapter of the thesis will investigate the difference-frequency generation obtained by injecting two copropagating waves into a several-millimetre length of the nonlinear waveguide structure that was utilized for the sum-frequency experiments with the optical spectrometer. A theoretical equation for the total generated microwave power will be developed as well as a qualitative and quantitative description of the far-field pattern produced by the induced difference-frequency polarization. Several measurements will be presented and compared to the theoretical calculation. A discussion of possible improvements to the generating system will be undertaken, primarily in terms of how to increase the radiated power to more practical levels.

4.2 Theory of CW Microwave Generation with Nonlinear Optical Waveguides

The theoretical models for a Hertzian dipole oriented parallel and perpendicular to the substrate surface are given since either polarizations are possible depending on the input polarizations of the two optical beams. As well, a model based on a finite element analysis is used to determine the

effects of the lateral dimensions of the substrate.

4.2.1 Modelling of a Hertzian Dipole on a Dielectric

The modelling of the induced difference-frequency polarization consists of an electric Hertzian dipole for two cases of possible polarization (i.e. perpendicular to and parallel to the substrate). As well as modelling the generated far-field radiation, it is important to take note of the percentage of generated microwave power which goes into surface waves, which can contribute to a transmitted power but with a multi-lobed radiation pattern. Using the ray optics picture of propagating waves, one can observe in Fig. 4.1 that for angles of propagation greater than the critical angle, θ_c , the wave is internally trapped by the dielectric. Although the radiation from a typical surface wave can be picked up by a receive antenna, it interferes with the directly generated radiation of the dipole. As well, if the dielectric has a finite thickness, or has more than one layer present, reflections from the generated electromagnetic waves must be taken into account in order to properly model the antenna radiation.

Figure 4.1 Transmitting dipole antenna with trapped rays as surface waves (from Rutledge and Muha [44])

Modelling of antennas on dielectrics is complicated by the fact that the antenna does not radiate equally above and below the plane of the dielectric. As is well-known from antenna theory, an antenna placed on a dielectric will radiate principally into the dielectric. This effect can be seen clearly in Fig. 4.2 where Fig. 4.2(c) which shows the E- and H-plane patterns of a slot and a dipole antenna on a dielectric. The E-plane is parallel to the dipole axis, and the H-plane is perpendicular to the dipole axis for the case of the dipole antenna, as shown in Fig. 4.2(b). Fig. 4.2(c) demonstrates that the dipole pattern, shown on the right half of the diagram, mostly radiates into the $\epsilon_r = 4.0$ dielectric. As well, it can be seen that the E-Plane pattern of the dipole is much more directive than that for the slot antenna which is a favourable characteristic of the dipole, especially when being implemented in an array,

where the mutual coupling between antennas must be reduced. For the case of the dipole antenna, the ratio of the power propagating into the dielectric vs. that into free space is approximately given by $\sqrt{\epsilon_r} : 1$ [45,46]. Radiation is favoured in the dielectric, since a dipole transmitting in a uniform dielectric of value ϵ_r radiates $\sqrt{\epsilon_r}$ more power than the equivalent antenna transmitting into

Figure 4.2 (a) Slot on a dielectric

(b) Dipole on a dielectric

(c) E-plane and H-Plane radiation patterns for slot and dipole antennas (from Button [45])

free space with a relative dielectric constant of 1.0. Fig. 4.3 demonstrates the amount of power that is transmitted into the air and dielectric regions around a dipole antenna vs. the relative permittivity of the dielectric material. It can be seen that the amount of power that is transmitted into the air drops off very quickly with relative dielectric constant.

Figure 4.3 Power distribution for an elementary dipole vs dielectric constant (from Engheta et al. [46])

Also, the waves at the air-dielectric interface propagate approximately at a velocity equal to that in a dielectric whose value is the mean value of the relative permittivity of the air and dielectric. Therefore, the wave generated at the interface propagates slower than it would in the air, and at the same time,

faster than it would if completely contained within the dielectric substrate. This is the same effect that takes place in a leaky-wave antenna where radiation primarily takes place into the dielectric [47].

4.2.2 Far-Field Radiation Pattern of Hertzian Dipole Oriented Parallel and Perpendicular to Substrate Surface

In all discussions of radiating elementary dipoles on thin substrates, one must be aware of the possibility of generating far-field surface waves. These are basically modes which propagate as trapped waves in the optical waveguide, but radiate once they reach either a facet or edge of the substrate. Fig. 4.4 shows a possible path of propagation for surface-wave rays along with the effective guide thickness, h_{eff} , which is made up of the actual substrate thickness, h , plus $2z_0$, which represents the apparent ray penetration depth on the top and bottom portions of the substrate, as required by the Goos-Hanschen shift, $2x_0$ [48]. Generally, surface-wave fields and powers are inversely proportional to h_{eff} , which means that well-confined modes have high surface-wave losses. Much of the work on the modelling of dipoles, on thick and thin substrates, including the effect of surface waves has been carried out by Alexopoulos et al [49-52].

Figure 4.4 Ray optics picture of surface-wave propagation (from Button [48])

The chart of Fig. 4.5 demonstrates at which guide thickness it is possible to propagate the different TE and TM surface waves. It is interesting to note that the grounded substrate configuration has a cutoff for the TE_0 mode, but there is no cutoff in an ungrounded substrate for this mode [45].

Figure 4.5 Propagating modes for a substrate with $\epsilon_r = 12.0$ (from Button [45])

The direction in which the generated far-field surface wave energy is directed is important as well. For the case of a dipole, TE power is directed normal to the dipole, and TM power is directed parallel to the dipole, as shown in Fig. 4.6 [45,53].

Figure 4.6 Typical surface wave radiation patterns for TE and TM modes (from Alexopoulos et al. [53])

The Lorentz reciprocity theorem [54] can be used to calculate the far-field radiation pattern of a dipole oriented parallel to the surface of the substrate. It has been determined that the E-plane pattern of such a dipole on an ungrounded substrate is given by [45]

$$G_s(\theta_E) = \frac{3}{2P_t} \frac{\cos^2\psi + \eta_{TM}^2 \sin^2\psi}{\cos^2\psi + \left[\frac{1}{2} \left(\eta_{TM} + \frac{1}{\eta_{TM}} \right) \right]^2 \sin^2\psi} \quad (4.1)$$

for radiation upwards into the air, and,

$$G_d(\theta_E) = \frac{3}{2P_t} \frac{\cos^2\theta_E}{\cos^2\psi + \left[\frac{1}{2} \left(\eta_{TM} + \frac{1}{\eta_{TM}} \right) \right]^2 \sin^2\psi} \quad (4.2)$$

for radiation downwards into the dielectric. As well, the H-plane pattern of the same dipole on the ungrounded substrate is given by

$$G_s(\theta_H) = \frac{3}{2P_t} \frac{\cos^2\psi + \eta_{TE}^2 \sin^2\psi}{\cos^2\psi + \left[\frac{1}{2} \left(\eta_{TE} + \frac{1}{\eta_{TE}} \right) \right]^2 \sin^2\psi} \quad (4.3)$$

for radiation into the air, and,

$$G_d(\theta_H) = \frac{3}{2P_t} \frac{1}{\cos^2\psi + \left[\frac{1}{2} \left(\eta_{TE} + \frac{1}{\eta_{TE}} \right) \right]^2 \sin^2\psi} \quad (4.4)$$

for radiation into the dielectric.

ψ , η_{TE} , and, η_{TM} , are respectively defined as:

$$\psi = k_d h \cos \theta_r \quad (4.5)$$

$$\eta_{TM} = \frac{\cos \theta_r}{n \cos \theta_E} \quad (4.6)$$

$$\eta_{TE} = \frac{n \cos \theta_H}{\cos \theta_r} \quad (4.7)$$

where k_d is the wave vector in the substrate dielectric, θ_r is the refracted angle in the dielectric, and, θ_H is the incident angle in air with respect to the surface normal [45].

A plot of the resultant far-field radiation patterns for the semiconductor substrate of 400- μm thickness, and with an infinitesimal dipole of 20 GHz frequency, parallel to the surface yields E-plane, and H-plane patterns of Figures 4.7 and 4.8.

Fig. 4.7 shows that maximum power in the E-plane will be generated normal to the substrate surface and it decreases to zero in the direction parallel to the substrate surface. This is expected, since for the parallel direction, the generated electric field is being completely reflected at the air-dielectric interface. The power transmitted into the air is negligible compared to that into the dielectric. Similarly, the H-Plane pattern generated fields approach zero

Figure 4.7 E-Plane Pattern into Dielectric for a Hertzian Electric Dipole

Figure 4.8 H-Plane Pattern into Dielectric for a Hertzian Electric Dipole

intensity for increasing radiation angles. However, for the H-plane case, the maximum does not occur normal to the substrate into the dielectric, but at 25° to the normal.

The case of a dipole which is oriented perpendicular to the substrate can be solved by utilizing the method of Lukosz and Kuntz [55-57]. The geometry of the dipole for this case can be viewed in Fig. 4.9. The angle, α , is the angle of observation with respect to the surface normal. The analysis assumes that the dielectric occupies an infinite half-plane, which is not the case encountered for the waveguide sample utilized in the experiment, but which can be assumed to give an approximate picture of the radiation pattern with certain restrictions.

Perpendicular dipole D located in medium 1 at distance z_0 from the interface. Unit vector $\hat{u}$ points into direction of observation.

Figure 4.9 Perpendicular Dipole Situated in Medium n_1 , ie. GaAs
(from Lukosz and Kunz [57])

The power generated by the dipole for the case of the electric dipole embedded in the GaAs waveguide at the surface of the substrate, and transmitted into the air ($0 \leq \alpha_2 \leq 90^\circ$), is given by [57]

$$G(\alpha_2) = \frac{3}{8\pi} \frac{n^5 \sin^2 2\alpha_2}{(n \cos \alpha_1 + \cos \alpha_2)^2} \quad (4.8)$$

where n is the index of refraction of the GaAs dielectric, α_1 is the angle of incidence in medium 1 and α_2 is the angle of refraction in medium 2 (ie. air). The case of the dipole transmitting into the dielectric is split up into the totally internally reflected case, ie. $\alpha_{1,c} < \alpha_1 \leq 90^\circ$, which is given by

$$G(\alpha_1) = \frac{3}{8\pi} n^2 (1 - n^2)^{-1} [(1 + n^{-2}) \sin^2 \alpha_1 - 1]^{-1} \sin^2 2\alpha_1 \quad (4.9)$$

and the case of $0 \leq \alpha_1 \leq \alpha_{1,c}$, which is given by

$$G(\alpha_1) = \frac{3}{8\pi} \frac{n^2 \sin^2 2\alpha_1}{(n \cos \alpha_1 + \cos \alpha_2)^2} \quad (4.10)$$

Investigating the above radiation pattern equations, for the two cases of the dipole oriented parallel and perpendicular to the substrate surface, it is noted that in both cases radiation into the dielectric is significantly greater than that into air when using GaAs substrates. However, in the case of the perpendicular dipole orientation, the radiation into the waveguide is significantly more directive than that for the parallel dipole orientation.

4.2.3 Finite-Element Solution of Generated Electric Fields from a Current Dipole of Finite Length Oriented Perpendicular to the Substrate

The generated radiation pattern from an induced difference frequency polarization, existing in the plane of the substrate, can easily be determined by adapting the general microwave case of a Hertzian microstrip dipole over a thick ungrounded dielectric slab, as outlined in the previous section using the theory of Lukosz and Kunz [57]. However, when the induced polarization is normal to the surface of the substrate, as it was expected to be for the experiment which will be outlined in section 4.3, then it is not possible to find an analytical closed-form expression for the generated dipole radiation. For this case a commercial finite element simulator (EMAS-Electromagnetic Application Simulator) [58,59] was used to solve the special case of an induced polarization perpendicular to the surface of the substrate. Since the polarization is induced in a dielectric that has finite lateral dimensions, the actual physical radiation pattern will depend not only on α , defined in the previous section, but on ϕ as well.

In order to determine the variation of the radiated power as a function of ϕ , the modelling of the dipole and its associated radiation pattern was undertaken using a two-dimensional grid, along with four-sided finite elements. Although the full problem can only be modelled utilizing all three spatial

dimensions, the two-dimensional picture will lead us to a general picture of the expected radiation pattern as a function of angle, ϕ . (It is known that a standard dipole, of infinitesimal length, in free-space, has zero Poynting vector in the direction of the dipole axis). The grid can be viewed in Fig. 4.10. Each square represents an area of $600\ \mu\text{m} \times 600\ \mu\text{m}$. The substrate is modelled using 5 squares by 30, yielding substrate dimensions of 3 mm x 18 mm, which corresponds to that of the standard sample utilized in the laboratory. The other group of squares, to either side of the substrate, are used to create a larger area which can then be extruded to a circular grid. The quadrilaterals around the area made up of squares lead to infinite boundary finite elements which must be modelled in the software in a spherical coordinate system. The radius from the centre of the plot is 20 mm, which places the entire plot in the near-field of the antenna. Although measurements were taken in the far-field of the dipole, useful information can still be extracted from the results obtained using this grid. With the present number of elements being almost 5000, it was decided not to use more, in order to guarantee the stability of the problem.

A point source of current was placed in the approximate centre of the substrate and its radiation pattern was examined for several cases. In order to verify that the element size was small enough to yield reliable results, the case of a dipole radiating in free space was first examined and passed all the required tests. As well, EMAS's built-in checker for finite element validity

verified that the distortion and skew of all the elements passed the acceptable limits. It was assumed that the finite elements were made up of ideal insulating GaAs/AlGaAs with no free carriers so that plasma effects on the radiation pattern were ignored.

It is expected that for the case of a 10-GHz dipole source perpendicular to the substrate, the relative permittivity of the GaAs material will have a significant effect on the radiation pattern of the dipole due to the fact that the dimensions of the substrate are comparable to the wavelength of the generated microwaves. The plot of the real part of the radial Poynting vector can be viewed in Fig. 4.11. It can be seen that the radiated power is greatest in the vertical directions and tapers off towards either side of the wafer. As well, plots for the same situation, but with source frequencies of 1 GHz, and 100 GHz can be viewed respectively in Fig. 4.12 and Fig. 4.13, respectively. It can be seen that for the case of the source radiating at a frequency of 1 GHz, the Poynting vector is approximately that of a dipole in a uniform dielectric. This is expected since the wavelength at 1 GHz in GaAs material is approximately 8.4 cm, which is much greater than either dimension of the sample. On the other hand, for the case of the infinitesimal dipole radiating at 100 GHz, it can be seen that maximum radiation occurs in the vertical directions with very little power emanating horizontally out from the sample.

- ⑥ dipole oriented perpendicular
to substrate surface

Figure 4.10 Finite Element Grid

Figure 4.11 Radial Poynting vector for dipole at 10 GHz

4.3 Difference-Frequency Power Generated by the Nonlinear Waveguide

From equation (I.1) of Appendix I, which describes the induced polarization for copropagating waves, it can be noted that the sum-frequency polarization component has a rapid sinusoidal variation along the waveguide axis, while the difference frequency has the slowly varying polarization term. For difference frequency generation, the above case is the desired situation, as opposed to injecting counterpropagating waves, which would lead to an induced polarization with a rapidly varying spatial dependence. Hence, the counterpropagating case would establish a dipole array in the waveguide which would cancel itself out in the far-field of the radiating structure. Actually, for the length involved and since identical polarizations are injected into the waveguide from the two lasers, the wave vectors differ only by a very small amount in the waveguide (the wavelength of the variation is cm's) and, hence, the induced polarization in the waveguide can be taken as being approximately constant along the several millimetre length of the waveguide.

The same multilayer $\text{Al}_x\text{Ga}_{1-x}\text{As}/\text{GaAs}$ optical waveguide that was used for the optical spectrometer and deformation sensor investigations was used for the microwave generation as well. For the case of microwave generation, the multilayer structure has no effect on the generated microwave radiation

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PM-1 3½"x4" PHOTOGRAPHIC MICROCOPY TARGET
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PRECISIONSM RESOLUTION TARGETS

pattern since the stack thickness is small compared to the wavelength of the microwave radiation. However, since the stack is made up of various layers with differing Al concentrations, the value of $\chi^{(2)}$ will in fact not be that of homogeneous GaAs. However, when a regular homogeneous GaAs slab waveguide was used there was no appreciable difference in the generated microwave power as expected.

Establishing a pure TE mode in the waveguide from each of the input beams results in a difference frequency polarization in the y-direction, as shown in Fig. 4.14. To a first order approximation, the induced difference frequency generation can be modelled using a small dipole approximation for the equivalent current from the induced polarization. For the case of a pure TM polarization, the microwave radiated power should be nil. Absence of microwave generation for this situation is expected since there are no cross-terms of electric field because of the (100) geometry of the semiconductor. When a combination of TE and TM modes is set up in the waveguide, a dipole in the y-direction, as well as an array of dipoles in the direction of propagation, are established in the semiconductor material. Since $k_{TE} \neq k_{TM}$, they would lead to destructive interference in the far-field of the antenna array due to the rapidly varying phase variation across the wafer. A full discussion of injecting TE and TM polarizations into the wafer from either side and observing the effect on the near-field sum-frequency-generated signal as a function of the

phase of the input signals can be found in [60], and [61].

Figure 4.14 Nonlinear polarization induced in the optical waveguide

In order to calculate the total radiated power generated from the nonlinear process, to a first-order approximation, it is necessary to first calculate the induced difference frequency polarization using the input electric field strengths from the two beams, and to then determine the equivalent current density of the dipole antenna. From that point an approximate equation can be used to find the total power radiated by an infinitesimal dipole (dipole length $\ll \lambda_0$) based on its induced current and electric length. The final result is given by [62,63]

$$P_{\text{microwave}} = K d_{14}^2 P_{\text{optical}}^2 (n^*)^2 t^2 (\omega_1 - \omega_2)^4 \quad (4.11)$$

where K is a constant which depends on the linear optical properties and geometry of the waveguide [22], d_{14} is the nonlinear susceptibility, P_{optical} is the total input optical power to the waveguide, n^* is the effective wave impedance in the waveguide material, and t is the waveguide thickness. A complete derivation of Equation (4.11) can be found in Appendix II.

4.4 Experimental Setup

The experimental setup can be viewed in Fig. 4.15. The waveguide is pumped by two $\lambda = 1.32 \mu\text{m}$ Nd-YAG ring lasers, just as it was for the optical spectrometer work, except that the beams are injected in the copropagating geometry. The frequency offset between the two lasers is tuned by varying the temperature of one of the laser cavities precisely as previously described for the case of the sum-frequency spectrometer application. The laser emissions are vertically polarized and, therefore, a $\lambda/2$ waveplate is used in order to rotate the polarization to any desired setting. The infrared beams are coupled into the optical waveguide using a 10X microscope objective. This reduces the waist of the light from several millimeters to approximately $10 \mu\text{m}$,

which is required in order to concentrate the light laterally in the waveguide for optimal coupling. The microwave radiation generated by the difference frequency mixing is collected by an X-Band waveguide horn and an HP70000 spectrum analyzer along with a pre-amplifier in order to reduce the noise floor to -15° dBm when sweeping across a 10 kHz-100 kHz frequency range centred at the microwave frequency offset between the two lasers. The microwave radiation was determined to be normal to the waveguide surface, since the receiving horn discriminates horizontal from vertical electric field polarization by at least 25 dB.

Figure 4.15 Experimental setup for CW difference frequency generation

At the same time that a difference frequency is being generated, a sum frequency is also produced and emitted approximately perpendicular to the

surface of the substrate. The presence of the red light emission at the second-harmonic wavelength ($0.66\ \mu\text{m}$) can be used to optimize the coupling of light into the waveguide which is one of the critical parameters in the setup. The counter-propagating geometry for the sum-frequency harmonic-generation is provided by reflection from the back facet. The sum-frequency power of the red light is on the order of several pW's and, therefore, the microwave power can be expected to be significantly less than this value due to the Manley-Rowe relation which was discussed in Chapter 2.

4.5 Measured Results

The input power to the nonlinear optical waveguide can be varied by rotating the $1.32\text{-}\mu\text{m}$ polarizing cube. Figure 4.16 shows the variation of output power against input power for a difference frequency of 10 GHz and with both of the beams having TE polarizations. The slope of the graph is approximately 2:1, confirming that the microwave generation is a $\chi^{(2)}(\omega)$ process. The peak power collected by the pyramidal horn was measured to be approximately -130 dBm.

Figure 4.16 Measured Output Power vs Input Optical Power

The 140 kHz linewidth of the received microwave signal was due to the frequency drift of the two YAG lasers. Since optical isolators were not used, there was some feedback from one laser into the other via reflections from various optical components, leading to a small amount of frequency instability in the lasers. A more significant problem in the experiment was determining the appropriate frequency range to scan. At the generated microwave levels of the experiment, a 1000 second scan was required along with a scan width of approximately 10 kHz in order to measure the generated signal which was 10 to 15 dB above the noise floor of the spectrum analyzer. Fig. 4.17 demonstrates that one laser drifts with respect to the other at the rate of 15 MHz per hour, and, therefore, it is very hard to 'catch' the microwave signal in the 10 kHz scan window. The long term offset frequency drift between the two lasers is cyclical where it first decreases at a uniform rate, then levels off

for a while and then increases at the same uniform rate. The nuisance of long term laser drift made it very difficult to gather data for the experiment and at present there is no immediate solution to the problem. A possible future approach would be to beat the two lasers on a photodiode and to have external electronic circuitry which would determine how far the offset frequency drifted and to precisely change the temperature of one of the cavities in order to keep a constant difference between the two lasers. In any practical system this would have to be undertaken, otherwise the problem of long-term drifting of each of the lasers would cause the loss of receive signal for long periods of time. Therefore the precision of this measurement technique far exceeds the available stability of the state-of-the-art laser sources at this time.

Figure 4.17 Frequency drift of one laser with respect to the other
vs time (hours)

The tuning curve of Fig. 4.18 shows the variation of the detected microwave frequency against the laser difference frequency. The laser difference frequency was obtained by combining the beams from the two laser sources onto a fast InGaAs photodiode and measuring the beat frequency on a spectrum analyzer. The input polarizations of the two beams were both TE and the polarization of the microwave generation was perpendicular to the surface of the substrate. The spectrum analyzer trace of Fig. 4.19 shows the detected microwave frequency along with the approximate 100 kHz bandwidth of the detected signal which was due to the jitter of the solid state laser sources.

Figure 4.18 Detected microwave frequency vs tuned frequency offset

Figure 4.19 Detected Microwave Signal Trace

4.6 Discussion of Microwave Frequency Generation

Equation (4.11) derived in section 4.3 shows that the most logical way to increase the generated microwave power would be to increase the frequency offset between the two lasers, but using the present experimental setup it is not possible to increase the offset much greater than 30 GHz. The derivation of the free carrier screening effect (Appendix III) demonstrates that it does play a certain role in reducing the generated microwave power and, hence, its

reduction in a future system would be advantageous. This could be undertaken by determining exactly what the excess doping is in the nonlinear waveguide and to compensate for the excess p-type carriers by properly n-type doping the waveguide. However, radiation could be augmented by growing a ridge waveguide instead of the standard slab waveguide structure. Comparing a guide of several mm length with and without ridges has yielded an increase in sum-frequency power by a factor of ten for the ridge waveguide [64]. In this way, the optical input beams would be confined in both the lateral and vertical dimensions of the guide.

Placing a lens on the back side of the substrate with the same dielectric constant as the GaAs material would be the most efficient way to decrease the amount of power that goes into surface waves, since, for this case, the generated waves do not undergo complete internal reflection [65,66]. Hence, more of the generated energy would go into the desired far-field radiation of the antenna. Two possible lens systems are shown in Fig. 4.20.

As well, including a multilayer stack of asymmetric quantum wells in the waveguide could significantly augment the d_{14} value for input optical wavelengths near the semiconductor bandgap, and thereby have an impact on the output power [67-69].

Figure 4.20 Possible substrate lenses: (a) hemispherical, (b) hyperhemispherical (from Neikirk et al. [66])

It should be mentioned at this point that there exists another method to collect the generated power from the induced difference frequency polarization. Evaporating electrodes directly onto the surface of the semiconductor substrate at the appropriate spacings and the appropriate distance between electrodes would also yield a combined voltage, which could then be measured using a high-speed microwave jig (as shown in Figures 4.21 and 4.22). The length of the electrodes would be dictated by the difference in the propagation constants of the fundamental TE and TM modes of the nonlinear waveguide, and the spacing of the electrodes would be given by the waist of the propagating

beams inside the nonlinear waveguide (approximately $10\text{ }\mu\text{m}$ when focussed down by a 10X microscope objective). Since the sign of the induced polarization flips as one goes along the waveguide, the electrodes would have to be appropriately designed as shown in Fig. 4.21. For experimental purposes, electrodes were designed where there were not a series of them but only one in case the spacing of the electrodes was actually slightly off from the previously measured value.

The above method should be a much more efficient technique to gather the generated microwave energy. Unlike the previous method where a part of the total transmitted energy must be collected by a microwave horn, in this case almost all of the induced voltage should be 'trapped' by the electrodes. The expected voltages on the electrodes would be in the μV range.

Figure 4.21 Gold electrode pattern

Figure 4.22 Measurement test fixture

Once the electrodes were tested, using a setup very similar to that of Fig. 4.15, it was found indeed that microwave generation was taking place but not from the expected source. Rotating the polarization of one of the input beams did not change the output from the electrodes at all. As well, when the generated pulse was measured its time constant was found to be much larger than expected. Although this was partially due to parasitics from the external circuit, the main effect was due to the semiconductor-metal interface acting effectively as a Schottky diode. Hence, the receive signal contained a very large component due to the Schottky photodiode effect with the swamping of the nonlinear difference-frequency generation process.

Although the electrode technique for measuring the induced microwave

difference frequency is not presently feasible, it may be possible in the future to engineer an interface between the semiconductor and the substrate which would have an insignificant Schottky barrier height and which could therefore be considered as an ohmic contact. This could be done by highly doping the entire substrate in order to create an ohmic contact between the semiconductor and metal, and then selectively etching off the high conductivity layer to the semi-insulating waveguiding portion of the multilayer structure.

Observing equation (4.11) it is evident that the generated microwave power increases by 40 db/decade with increase in frequency. Although the power is extremely small for difference frequencies in the K-band range, for frequencies in the THz range the efficiency can easily exceed that of a conventional Schottky barrier photodiode whose efficiency decreases with frequency. Hence, there is a certain frequency at which the difference-frequency optical waveguide will generate more power than a metal-semiconductor-metal photodiode, and can, therefore, be considered one of the main advantages of the nonlinear waveguide for signal generation in the far infrared wavelength spectrum.

Using Equation (4.11), a graph of radiated power vs frequency can be shown for the nonlinear optical waveguide as displayed in Fig. 4.23. As well, one can also view the detected power vs frequency for a 25 nm finger-spacing

GaAs MSM photodiode (curve taken from Fig. 6 of [70] with the assumption that 1 mW of microwave power is generated from the input of 100 mW of optical power which is approximately the case for these types of photodiodes). It can be seen that the curves intersect at 16 THz and that, hence, difference frequencies greater than this value will be more efficiently generated utilizing a nonlinear waveguide rather than a present day MSM photodetector assuming that the various assumptions that went into formulating Equation (4.17) are valid.

Figure 4.23 Generated Power vs. Frequency for the Nonlinear Optical Waveguide and a State-of-the-Art MSM Photodiode (photodiode response taken from Chou et al. [70])

State-of-the-art metal-semiconductor-metal (MSM) photodiodes are known to be limited by transit time as well as intrinsic and mounting parasitics [70,71]. The only way to increase the 0.5 THz bandwidth [70] of present-day MSM photodiodes is to decrease the finger spacing which at present is limited to the nanoscale range due to fabrication technology. As well, the intrinsic parasitics of the detector's electrodes must be decreased by decreasing the electrode width as well as the ratio of the finger width to finger spacing, which again is limited due to fabrication technology. Hence, it can be seen that since the various Schottky and p-i-n photodiodes [72] use carriers to generate a microwave current, these types of photodetectors will always be limited by transit time effects as well as recombination lifetime, whereas a device that uses the inherent nonlinearity of a semiconductor such as GaAs is not limited by these factors in the near- to far-infrared wavelength spectrum. Therefore the future development of Terahertz nonlinear optical waveguide photodetectors should be a very productive research area as the limits of present day photodetectors are rapidly being reached.

Chapter 5.0 Conclusions and Future Work

The nonlinear waveguide optical spectrometer concept has been demonstrated for possible use as a demultiplexer in a DWDM optical communications system, which does not have the main disadvantage of the conventional optical demultiplexing scheme wherein laser drift can potentially destroy the system. Also, the optical spectrometer work leads to a practical optical deformation sensor based on the change in far-field radiation pattern of the nonlinear waveguide depending on the type of inhomogeneity present in the structure. A comparison between the theoretical and measured results for the deformation experiment were in very good agreement.

The difference-frequency generation was carried out on the AlGaAs/GaAs nonlinear optical waveguide and yielded powers which were expected from the theoretical calculations. Although the generated power was not large there are definitely several techniques which could be used in order to augment the power. Future work could include increasing the difference frequency by a factor of ten, in order to augment the power by a factor of four, leading to generated powers in the pW range. As well, although the electrode measurement approach was not successful, by fabricating high quality ohmic contacts on the GaAs material, more of the generated power will be able to be collected, since the power will be collected by the electrodes and not an

external antenna, which can only gather a certain amount of the radiated power density. The optical downconverter has the greatest potential to become a practical device in the THz frequency range where its efficiency is a lot higher and where the photodiode response is severely limited by finite transit time through the electrodes on the semiconductor.

Appendix I Effect of Copropagating vs Counterpropagating Signal Injection

When utilizing the planar slab waveguides for sum- or difference-frequency generation, it is important to determine whether it be more useful to inject the two input beams in a copropagating sense or a counterpropagating sense. Assuming that two signals E_1 and E_2 are injected in a copropagating sense then they can be expressed as:

$$\begin{aligned} E_1 &= e^{j(k_1 z - \omega_1 t)} \\ E_2 &= e^{j(k_2 z - \omega_2 t)} \end{aligned} \quad (I.1)$$

and utilizing equation (2.12) the induced polarization is determined to be

$$P_2 = \frac{1}{2} \chi^{(2)} \left[e^{j[(\omega_1 + \omega_2)t + (k_1 + k_2)z]} + e^{j[(\omega_1 - \omega_2)t + (k_2 - k_1)z]} \right] \quad (I.2)$$

It can be seen that for this case the sum-frequency polarization is associated with the rapidly varying $(k_1 + k_2)$ spatial term, whereas the difference frequency is associated with the slowly varying $(k_2 - k_1)$ term. Hence, it would not be advantageous to inject copropagating beams into the waveguide since the induced polarization would create a multi-lobed far-field pattern. The above case is more suitable for difference-frequency generating processes where the spatial term is slowly varying and would, therefore, generate more of an ideal dipole far-field radiation pattern.

Now assuming that the two signals E_1 and E_2 are injected in a counterpropagating sense they can be expressed as:

$$\begin{aligned} E_1 &= e^{j(k_1 z - \omega_1 t)} \\ E_2 &= e^{j(-k_2 z - \omega_2 t)} \end{aligned} \quad (1.3)$$

and once again, utilizing equation (2.12), the resultant induced polarization is determined to be

$$P_2 = \frac{1}{2} \chi^{(2)} \left[e^{j[(\omega_2 - \omega_1)t + (k_1 + k_2)z]} + e^{j[(\omega_1 + \omega_2)t + (k_2 - k_1)z]} \right] \quad (1.4)$$

For this case, the sum-frequency term has the slowly varying spatial term and the difference-frequency term has the rapidly varying spatial term. Hence, counterpropagating signals are more useful for sum-frequency generation (for use in optical upconversion applications) and not as amenable to difference-frequency generation.

Appendix II**Derivation of Approximate Far-Field Microwave Power**

The following derivation assumes that the case of TE polarized input beams are used for the generation of the microwave power. This first-order derivation assumes that the optical energy is concentrated completely within the $1\text{-}\mu\text{m}$ thickness of the waveguide. The fact that this is not entirely the case will not significantly affect the final calculated microwave received power since there are other factors in the final equation whose values are known to a far lesser accuracy. The following formula which relates electric field strength in the y-direction, E_z , to the power in the guide is used to calculate E_z ,

$$E_z = \sqrt{\frac{2P_g \eta_g}{A_{\text{interaction}}}} \quad (\text{II.1})$$

where P_g is the power in the guide and is found using the coupling factor K' according to

$$P_g = K' P_{\text{optical}} \quad (\text{II.2})$$

and η_g is the wave impedance of the waveguide and $A_{\text{interaction}}$ is the interaction area of the input optical beam and is given by (WT) where W is the lateral dimension of the input beam, and T is the effective waveguide thickness. Substituting equation (II.2) into (II.1) we obtain

$$E_z = \sqrt{\frac{2K' P_{\text{optical}} \eta_g}{WT}} \quad (\text{II.3})$$

Using equation (2.11) and substituting the electric field strength calculated from equation (II.3), the induced polarization P_{induced} is given as

$$P_{\text{induced}} = \frac{4d_{14}K'P_{\text{optical}}n_g}{WT} \hat{y} \quad (\text{II.4})$$

In the above equation, P_{induced} has no spatial dependence due to the fact that the propagation constants for the TE modes set up from the two input beams are only slightly different. Although there is a spatial variation for the induced polarization, it can be considered to be negligible over the waveguide length of several millimeters.

The equivalent induced current present in the waveguide can be found using the following equation

$$I_o = |\vec{P}_{\text{induced}}| \omega Wl \quad (\text{II.5})$$

and yields the result

$$I_o = \frac{4d_{14}K'P_{\text{optical}}n_g l}{T} \quad (\text{II.6})$$

Since the free space wavelength of the generated microwave radiation is much greater than the length of the induced current dipole, then the well-known result for the total radiated power of an infinitesimal dipole antenna can be

used. Also, (kr) [38], (ie. the free space wave vector of the microwave difference frequency multiplied by the distance from the antenna at the observation point) which is used to determine in which region the receiving horn is positioned, is found to be much greater than one and, therefore, it can be confirmed that indeed the measurements were all taken in the far-field of the infinitesimal dipole antenna. The total radiated power from an infinitesimal dipole is known to be,

$$P_{rad} = \eta_o \left(\frac{\pi}{3} \right) \left(\frac{I_o l_{ant}}{\lambda_o} \right)^2 \quad (II.7)$$

Substituting equation (II.6) and the variable T , for l_{ant} , into equation (II.7), and using the fact that $\lambda_o = 2\pi c_o / \omega$, the total radiated power is given by

$$P_{rad} = \frac{4}{3} \frac{\eta_o d_{14}^2 K^{1/2} P_{optical}^2 n_g^2 l^2 \omega^4}{c_o^2 \pi} \quad (II.8)$$

Combining some of the terms we arrive at

$$P_{rad} = K d_{14}^2 P_{optical}^2 l^2 \omega^4 \quad (II-9)$$

where,

$$K = \left(\frac{4}{3} \right) \frac{\eta_o K^{1/2} n_g^2}{c_o^2 \pi} \quad (III.10)$$

Due to the finite aperture of the receive antenna, the fraction of the total

radiated power which impinges on the antenna could be found. This can be done by finding the ratio of the power to the input horn antenna in terms of the total radiated power which is given by

$$\frac{P_{rec}}{P_{rad}} = \frac{3\pi}{8} \int_0^\pi \int_0^{2\pi} \sin^3\theta \, d\theta \, d\phi \quad (11.11)$$

using the equations in [63].

In order to calculate the value for P_{rad} using equation (11.9), the following values are substituted for the input variables:

$$d_{14} = (72/9) \times 10^{-22} \text{ C/V}^2$$

$$K' = 0.1$$

$$P_{optical} = 75 \text{ mW}$$

$$n_g = 105 \, \Omega \text{ (derived in Appendix II)}$$

$$l = 3 \text{ mm}$$

$$\omega = 1.26 \times 10^{11} \text{ rad/sec}$$

Therefore, the total radiated power is expected to be 0.05 fW.

Appendix III Derivation of Waveguide Wave Impedance, n_g

Any semiconductor material has different regions of operation in terms of its effective index of refraction for the propagating signal. It may be either transparent, absorptive, or reflective, depending on the frequency of operation and the free carrier concentration of the crystal as outlined in the Drude-Zener model [73,74]. The formula for the effective dielectric constant of a semiconductor is given by

$$\epsilon^*(\omega) = \epsilon_o \left(1 + \frac{4\pi N e^2}{\epsilon_o \omega} \left[\frac{(m_e^*)^{-1}}{\omega + \frac{j}{\tau_e}} + \frac{(m_h^*)^{-1}}{\omega + \frac{j}{\tau_h}} \right] \right) \quad (\text{III.1})$$

where ϵ_o is the high frequency dielectric constant, N is the carrier density, e is the electronic charge, m_e^* is the effective mass of electrons, m_h^* is the effective mass of holes, τ_e is the scattering time for electrons and τ_h is the scattering time for holes. From the above equation the plasma frequency of a semiconductor system can be calculated as

$$\omega_p = \sqrt{\frac{4\pi N e^2}{\epsilon_o m^*}} \quad (\text{II.2})$$

and represents the frequency at which conductive effects can predominate and the semiconductor acts as an optical absorber, and will basically screen out the induced electromagnetic radiation.

For the case of MBE grown nonlinear waveguide material we know that the material is p-type and that its properties are:

$$N = 10^{15} \text{ cm}^{-3}$$

$$\tau_h = 56 \text{ fs}$$

$$m_h = 0.34 m_0$$

and upon substituting these values into equation (III.1) the $\text{Im} [\epsilon(\omega)]$ is 33 at a frequency of 20 GHz which translates to an effective wave impedance of 137 Ω . Hence, since P_{out} is proportional to n_p^2 , the output power is decreased by approximately 10 dB due to free-carrier absorption effects.

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