Ultra compact ans sensitive Terahertz Heterodyne receiver based on quantum cascade laser and hot electron bolometer
François Joint

To cite this version:
Ultra Compact and Sensitive Terahertz Heterodyne Receiver based on Quantum Cascade Laser and Hot Electron Bolometer
Abstract

The understanding of the stars life cycle in the interstellar medium and its role in the galaxy evolution has been possible with the continuous increase in sensitivity and spectral resolution of the far-infrared (FIR) and sub-millimetre heterodyne receivers in observatories such as the Hershel Space Observatory, the Stratospheric Observatory for Infrared Astronomy (SOFIA) or the Atacama Large Millimetre/submillimetre Large Array (ALMA). Further studies in the 1.2 $\rightarrow$ 5 Terahertz (THz) frequency range ($\lambda = 250 - 60 \mu m$) are however required, especially for the molecular spectroscopy of the rotational transition lines HD, $NH_3$ and $H_2$. The first key element of a heterodyne receiver is its non-linear mixer (detector) that down converts the THz signal of interest into a much lower frequency range. Superconducting hot electron bolometers (HEB) are good candidates for detecting weak signals in the THz frequency range. Such heterodyne detectors require radiant power from a local oscillator (LO)- the second key element - to achieve the mixing process in the frequency range of observations. The THz LO signals are overlaid with the signal of interest (RF) and coupled to the mixer using a diplexer or a beam-splitter. THz quantum cascade lasers (QCL) are promising semiconductor laser sources operating above 1 THz with mW of output power available. The combination of a QCL as LO and a HEB as mixer is an ideal THz heterodyne detection system that is currently employed by on-board missions for astrophysical observations. Further developments are however necessary to improve the integrations of QCL-based LOs with HEBs for future space missions, in terms of compactness, DC power dissipation of the QCL and quasi-optical coupling of the LO signal with the HEB.

This work has put the emphasis on the development of a suitable LO for a compact and ultra sensitive heterodyne receiver in the THz frequency range. To meet the required specification of THz heterodyne applications, we have successfully developed a 3$^{rd}$-order DFB THz QCL with single mode operation at 2.7 THz with less than 200 mW of power dissipation. The 3$^{rd}$-order DFB approach has proven to be extremely competitive as devices can be made very small, and hence require low power, without compromising the output beam quality. The beam pattern divergence was measured with a FWHM of $10^\circ \times 10^\circ$. The output power ($\sim 1$ mW) was sufficient to fully pump the HEB into its
normal resistive state. These characteristics made the fabricated devices suitable for a compact integration with the mixer in a heterodyne receiver.

A difficulty, however, remains in obtaining a precisely pre-defined emission frequency and an optimal output beam pattern simultaneously, as both are determined by the same parameter: the grating duty cycle. The optimal far-field beam profile is obtained when the effective index of the lasing mode is exactly $n_{\text{eff}}=3$. If one departs from this phase matching condition, for which there is perfect constructive interference in the far-field, the output beam profile will be degraded. Therefore, we have proposed a practical technique to re-shape the far-field emission pattern of THz frequency 3rd-order DFB lasers into almost perfect Gaussian beams independently of the quality of the laser output beam profile. The technique is based on the use of dielectric hollow waveguides (HWs). This technique reliably increases the laser beam directivity up to 55 dBi. The 3rd-order QCL coupled to a HW enables almost perfect Gaussian far-field emission profiles to be created with full-width-at-half-maximum values of $5^\circ \times 5^\circ$.

Upon a sufficient LO output power provided by the 3rd-order DFB QCL and the high directivity provided by the HW, a novel compact heterodyne receiver has been proposed where the LO and RF signals are coupled onto the HEB mixer without the use of diplexer. The laser was closely integrated with the HEB into the same dewar. A heterodyne characterization setup has been developed to enable repeatable and reliable measurements of the receiver sensitivity. A measured receiver noise temperature at 2.7 THz was recorded at 881 K which was close to the state of the art with a similar mixer in the same frequency range. Based on the experiences with the new quasi-optical coupling scheme, we have shown the possibility to integrate a QCL as LO into a heterodyne receiver and the novel coupling technique didn’t degrade the receiver sensitivity. In the contrary, it reduces the optical losses due to the absence of the beam splitter and the shorter length in the optical path. These results open a new horizon for THz coherent imagery where compactness and ultra high sensitivity are required.
Acknowledgements

First and foremost, I would like to acknowledge my supervisors Yan Delorme and Raffaele Colombelli for their continuous guidance and support during my Ph.D. Their patience and encouragement gave me the freedom which allowed me to test the boundaries of my capabilities, as well as being there to help me in moments where I got stuck. Their enthusiasm and humility were a remarkable lesson that I will keep as a guidance all along my future work. I am indefinitely grateful to them.

This research would not have been possible without the support of CNES and Christophe Goldstein, who also kindly agreed to serve on my Ph.D committee. The CNES and Christophe always kept a kind eye over my work and provided me with many opportunities to confront my ideas with other young researchers, from the organisation of the JC2 rally to the meeting at the Paris Air Show.

L’Observatoire de Paris and the Center for Nanoscience and Nanotechnology Orsay have been smashing places to work, most notably for the brilliant researchers who worked alongside me. I am especially indebted to Etienne Herth who assisted me many times in the clean room. I will miss his sense of humour and his patience, especially when the SEM pictures of the samples were keeping him waiting. The experimental work wouldn’t have been possible without Gregory Gay from LERMA and Stefano Pirotta at C2N. They get all the credits for keeping the labs up and running, while bringing new brilliant ideas for the experimental setups. I couldn’t possibly enumerate the ways in which they both helped me out through this Ph.D. My gratitude also extends to my laboratory colleagues from the Schottky group, Sylvain Caroopen, Alexandre Féret, Frederic Dauplay, Jeanne Treuttel, Sabrina Mignonni and Alain Maestrini who’s good wit and engaging conversations kept me sane during long lab hours. I would like to thanks Jean-Michel Krieg for his support. I have truly enjoyed working with Jean-Michel Manceau, Adel Bousseksou, Thibault Laurent, Pierre Laffaille and Fabien Defrance who were always prompt to help the Ph.D students in the clean room or cheering them up with a good joke. I would like to thank Maryvonne Gerin and Ouali Acef who made sure that the thesis work was going well. Martina Wiedner has been a very valuable source of advises on presenting my work effectively and supports. Thibaut Vacelet has been a great technical help in almost all
the laboratory fields, from sample mounting and wire-bonding to quasi-optical coupling, giving me clever tricks to ease my work. Most importantly, he always encouraged me and I am grateful to him for that. I had also the pleasure to collaborate with Haotian Zhu, who brought interesting concepts and ideas that were a great source of excitement during my Ph.D. I would like to thank Pierre Bonnay who kindly provided me with the glass for my experiments.

I am grateful for the engineers in the fabrication lab at C2N Orsay Marie-Paul Plante, Xavier Le Roux, David Bouville, Nathalie Isaac, Fabien Bayle, Cédric Villebasse, Jean-René Coudevyile and Samson Edmond for spending time to teach me how to use the tools and the instruments. They have also been tremendously helpful when the clean room went down to restart as quickly as possible the equipment and keep up running the lab. I would like to thank Dominique Decanini who welcomed me at LPN and showed me the 2-photons lithography system. Dominique made the project of the dielectric diagonal horn antenna a reality.

I have also benefited from many great discussions on Gaussian optics and RF electronics with Maurice Gheudin. His invaluable knowledge and kindness as well as his dedication to science were a source of inspiration and admiration.

I would like to thank Pascale Snini, Valérie Audon, Géraldine Gaillant, Carole Bonnot, Bernadette Laborde and Olga Julien for their constant help regarding the administrative work, planning the trips to the conferences and forgiving me for being late filling up the paper work.

The past three years wouldn’t have been as much fun without my fellow mates student and post-doc from the Ultra-QCL group: Pierre-Baptiste Vigneron, Claire Abadie, Laurent Boulley, Arnaud Jollivet, Ngoc Linh Tran, Paul Goulain, Mario Malerba. We have built altogether an agreeable working space with a challenging atmosphere, a constant problem solving mindset and friendships.

Several individuals outside my research group have also helped to make my work in the labs easier and funnier: at C2N, Guillaume Marcaud and its incredible energy, Aurore Ecarnot, Loïc Guillemot, Joan-Manel Ramirez, Mathias Berciano and at LERMA, Duccio Delfini and Anastasia Pienkina. I have enjoyed many good times around the coffee machine with the GaN people: Valerio Piazza, Nuno Amador, Nan Guan, Lu Lu, Omar Saket and Farsane Tabataba-Vakili.

I am also grateful to Meriem El Yajouri who organised and lead the Space Bus Morocco, some of the best memories I have from these past three years. We criss-crossed Morocco with Meriem and the other volunteers and showed the stars and the planets to the Moroccans. It has been a great human experience.

A special thank goes to Michèle Ba Trung who helped me not only on the technical
parts but also cheered me up with her considerate words. Working in the Lallemand building with Michèle was delightful and I will miss her way to see the lighter side of life. We did also share the same appetite for bike accidents.

I would like to thank Marielle and Guy who helped me carrying on during the difficult parts of my Ph.D. I am infinitely grateful to Sarah who stood patiently by my side since the beginning. Finally I would like to thank my parents, Vincent and Dominique Joint who provided me with an enormous amount of support and encouragement throughout. I hope they are proud.
Contents

Abstract

Acknowledgements

1 Introduction

1.1 Direct vs Heterodyne detection
1.2 Mixers
1.3 Local Oscillator
1.4 QCL used as LO in THz Heterodyne Receiver
1.5 QCL Requirements
1.6 Thesis Overview

2 Basic Operating Principles of THz QCLs

2.1 Intersubband Transitions & THz Quantum Cascade Lasers
2.2 Light Matter Interaction
2.3 Miniband electronic transport
2.4 Rate equations
2.5 LI395 laser design
2.6 Waveguides for the THz QCLs
2.7 Intrinsic linewidth of QCLs

3 Third Order Distributed Feedback Lasers

3.1 Distributed Feedback Lasers
3.2 Overview of the coupled modes analysis
3.3 Calculation of the coupling coefficient for lateral DFB gratings
3.4 Engineering the Distributed Feedback Structure
3.5 Emission
3.5.1 Antenna model
3.5.2 Finite Element Method Simulations
4 Third Order distributed feedback: results

4.1 Results on similar devices in the literature 47
4.2 Experimental measurements 48
4.3 Single mode emission at the target frequency 53
4.4 Far-field emission of 3rd-order DFB lasers 53
4.5 3rd-order DFB on L1390 57
4.6 Broad Band Extractor 60

5 Dielectric hollow core waveguides 67

5.1 Ray optics model for Hollow Waveguide 68
5.2 Launch condition of the lowest order hybrid mode 73
5.3 Total loss in hollow core waveguide 75
5.4 Coupling of a 3rd order DFB QCL to a HW 76
5.5 Further Developments: Diagonal Dielectric Horn Antenna 82

6 Superconduction Hot Electron Bolometer Mixer 85

6.1 Introduction 85
6.2 Bolometer effects 85
6.3 Superconducting Hot Electron Bolometer 86
  6.3.1 Heterodyne mixing with Hot Electron Bolometer 88
  6.3.2 Current-Voltage Characteristics of the Superconducting HEBs 89
  6.3.3 Cooling mechanism in HEBs 91
  6.3.4 Hot Spot Model 92
  6.3.5 Noise in HEBs 96
6.4 Requirements for the HEB 98
6.5 LO requirements 99
6.6 Coupling to the HEB 100
  6.6.1 Planar Antennas on dielectric substrate 101
  6.6.2 Quasi-optical Coupling 102
6.7 HEB developed at LERMA 104

7 Heterodyne measurements 107

7.1 Sensitivity of a heterodyne receiver system 107
7.2 Direct detection with the HEB 112
7.3 Linewidth characteristics of THz QCL 112
7.4 Noise temperature measurements 117
7.5 Heterodyne measurements with a novel quasi-optical coupling system 120
  7.5.1 Quasi-Optical Coupling with the Planar Antenna 120
8 Conclusion and Future Work

8.1 Conclusion .................................... 129
8.2 Future Work ................................... 130

A Metal-Metal Waveguides Fabrication Process 133

Bibliography 137

Financements 153

Publications 155
List of Figures

1.1 Calculated atmospheric transmission from 1 THz to 10 THz. The calculation was made using the model proposed in [1] for 30 cm long optical path, 25% of humidity at 300 K. ......................................................... 2

1.2 Heterodyne experimental schematic. ........................................... 4

1.3 Sensitivity of the available THz heterodyne mixers compared with their operating frequency. Figure adapted from [2]. ........................................ 5

1.4 SEM picture of an anti-parallel pair of Schottky diodes from Maestrini et al. [3] ................................................................. 6

2.1 Conduction band of heterostructure active region with a LO-phonon depopulation scheme with a bias of 64mV/module from Williams et al. [4]. The radiative transition takes place between the level 5 to 4 with a $E_{54}$=14 meV (3.4 THz). ................................................................. 18

2.2 Conduction band diagram of heterostructure active region L1395 with squared moduli wavefunctions for a bias of 57 mV/module. The heterostructure consists of a chirped super-lattice assisted with LO-phonon depopulation scheme. The radiative transition takes place between the level 2 or level 1, or both at the same time to the underneath miniband, while the phonon transition occurs from the miniband to the level 3 or level 2. ................................................................. 20

2.3 Lasing spectra of a 4 mm long and 150 $\mu$m wide laser ridge of L1395 just after threshold and maximum output power at 10 K in pulsed operation. ......................................................... 21

2.4 Cross-section of a metal-metal waveguide with intensity distribution of the fundamental laser mode. The typical dimensions are 15 $\mu$m for the height of the active region, 20–200 $\mu$m for the width of the waveguide. ..................................... 22

2.5 Calculated a) relative permittivity and b) conductivity of Au versus the frequency. ................................................................. 24
2.6 Calculated a) modal losses, b) confinement factor and c) effective index for $TM_{00}$ and $TM_{01}$ modes at 2.7 THz in function of the waveguide width for a metal-metal waveguide with a thickness of 10 $\mu$m.

3.1 a) Reflection on Bragg gratings. b) Waveguide with embedded Bragg gratings. For an embedded grating onto a laser, $\theta_i = \pi/2$ and for the diffracted wave $\theta_d = \pi/2$. For $p = 2$, $\theta_d = 0$ and the diffracted waves couple out from the guide.

3.2 Third-order with lateral corrugation grating coupling coefficient values calculated with the standard formula and with the adjusted formula.

3.3 a) Schematic of the DFB structure with lateral gratings simulated. Simulations of (b) the damping of the two band edge modes of the photonic band structure and (c) the coupling coefficient $\kappa$ of the lowest transverse magnetic mode $TM_{00}$ calculated with the eingen-frequency formula. $\kappa$ is calculated for laterally corrugated waveguides with $W_{\text{narrow}} = 5\mu m$, $W_{\text{wide}} = 15\mu m$, $h = 14\mu m$, grating periodicity $\Lambda = 55 \mu m$ and the grating duty cycle $DC$ being varied.

3.4 Eigen-frequency simulated of the symmetrical band edge modes of a laterally corrugated 3rd-order DFB laser varying with the grating duty cycle $DC$. The grating periodicity is $\Lambda = 55 \mu m$.

3.5 Simulated band structure of infinite 3rd-order DFB cavities using a finite element method where in a) the width of the waveguide is 15$\mu$m at the wide section and 5$\mu$m at the narrow section and in b) respectively 20$\mu$m and 5$\mu$m for the wide/narrow section. In c) is presented the $E_z$ profile of both higher and lower modes $\omega_+$ and $\omega_-$ for the second design. The mode profiles of the band edge modes for the first design is similarly distributed in the waveguide.

3.6 Schematic of a rectangular antenna aperture on an infinite electric ground plane with the defined referential for the far-field calculation.

3.7 (a) Schematic of the aperture antenna array and simulated beam patterns of different third-order DFB lasers using the dipole antenna array modeling. (b) in perfect phase-matching condition $n_{\text{eff}} = 3$, (c) $n_{\text{eff}} = 3.2$, (d) $n_{\text{eff}} = 3.4$ and (e) $n_{\text{eff}} = 3.6$. The grating period is $\Lambda = 55 \mu m$.

3.8 Simulated beam patterns of different third-order DFB lasers using the dipole antenna array modeling in perfect phase-matching condition ($n_{\text{eff}} = 3$). (a) 15 periods, (b) 50 periods, (c) 150 periods.
3.9 Simulated beam patterns of different third-order DFB lasers using the dipole antenna array modeling with $n_{eff} \neq 3$. (a) 15 periods, (b) 50 periods, (c) 150 periods ..................................................... 44

3.10 Simulated beam emission pattern using FEM simulations and Fourier transform model at 2.7 THz. The simulated waveguide consists of 15 periods with 15 $\mu m$ width for the wide section and 5 $\mu m$ for the narrow section. a) Schematic of the simulated 3rd-order DFB with deep lateral corrugations and showing the coordinates system used for the farfield calculation. The grating duty cycle was fixed to be 29% and the grating periodicity was for (b) $\Lambda = 55 \mu m$, (c) $\Lambda = 55.5 \mu m$ and (d) $\Lambda = 56 \mu m$ ............... 45

4.1 SEM pictures of second generations devices: the contact pad is optically decoupled from the waveguide to avoid any perturbations for the light extraction ..................................................... 49

4.2 L-I-V characteristic in pulsed operation of a 150 $\mu m$ x 4 mm Fabry-Pérot SP QCL. .......................................................... 50

4.3 Spectra of 3rd-order DFB from (a) first design and (b) second design. The grating periodicity is changed between each lasers while the grating DC is kept at 29% for (a) and 12% for (b). The lasers were driven in pulsed operation at 77K. The insets shows the mode wavelength in function of the grating periodicity. ............................................... 52

4.4 Typical electro-optical characterisation of the 3rd order DFB: the ratio of the wide-narrow section of the waveguide is 20$\mu m$-5$\mu m$ and the waveguide is composed of 15 periods with $\Lambda = 55\mu m$. (a) Light-current-voltage characterisation of the 3rd order DFB. The threshold current is as low as 25mA and the overall DC power dissipation is below 200mW for the whole current-voltage range. (b) Spectrum of the same device as above which shows a single mode operation at 90cm$^{-1}$ at 4K in pulsed operation (c) Emission spectra of 3rd-order DFB QCLs of the first design at 77K in continuous wave operation compared with emission from a similar laser heterostructure L1395 in single-plasmon waveguide (4 mm x 150 $\mu m$) ... 54
4.5 a) Schematic of the coordinate system used for the farfield experimental measurements b) Measured spectrum for a 3rd-order DFB QCL with the first design with $\Lambda = 54.6\mu m$ and DC = 28.8% with a single mode emission on the lower-band edge mode $\omega_- = 90.73 cm^{-1} \pm 0.25 cm^{-1}$ c) Measured far field corresponding to (a). d) Measured spectrum for 3rd-order DFB QCL with the second design with $\Lambda = 54.6 \mu m$ and DC = 12% with two lasing modes corresponding to the upper (main mode) $\omega_+ = 89.99 cm^{-1}$ and lower band edge mode and (e) Measured far field corresponding to (d).

4.6 Spectra and corresponding beam emission for devices from the first design with different value of $n_{eff}$. Figures a) and b) correspond to a $n_{eff} = 2.96$, figures c) and d) $n_{eff} = 2.94$ and figures e) et f) $n_{eff} = 2.92$

4.7 Spectra and corresponding beam emission for devices from the second design with different value of $n_{eff}$. Figures a) and b) correspond to a $n_{eff} = 3.05$ and figures c) and d) $n_{eff} = 3.09$

4.8 (a) Measured spectrum for a Metal-Metal Fabry-Pérot waveguide of the L1390 heterostructure and (b) corresponding Light-Current-Voltage characteristic in pulsed mode at 4K for the same laser. (c) Measured spectrum for the 3rd-order DFB waveguide with $\Lambda = 55 \mu m$ and $DC = 29\%$, the ridge section ratio between the narrow and wide part of the waveguide is 15/5$\mu m$ and (d) corresponding L-I-V.

4.9 Measured farfield emission pattern of the 3rd order DFB on L1390. The laser was operated in pulsed mode at 4 K.

4.10 Calculated far-field pattern of the extractor for different frequencies.

4.11 SEM pictures of the fabricated devices showing on a) an array of extractor devices, b) and c) showing details on the waveguide section with the 1st-order DFB with deep-etched lateral corrugations and the second section for the light extraction mechanism with 3rd-order DFB lateral corrugations.

4.12 a) Measured spectra and b) farfield beam emission of Fabry-Pérot waveguide coupled to broad extractor with squared teeth. The measurements were recorded at 10 K in pulsed operation. The spectra were measured just after lasing threshold (black curve) and at maximum output power (red curve) and the far field was measured at the maximum output power.

4.13 a) and c) Measured spectra for broad extractors on the L1395 heterostructure with different 1st-order DFB grating periodicity $\Lambda$. b) and d) corresponding beam diagram in pulsed operation at 77 K for the same lasers.
5.1 Modal intensity distributions of the first transverse electric and hybrid modes of a hollow core dielectric waveguide with a bore diameter of 4 mm and calculated for a wavelength of $\lambda = 110\mu m$ ................ 69

5.2 Cross-section of the hollow core waveguide, where the core is filled with air and the claddings is glass .................................. 70

5.3 $HE_{11}$, $HE_{12}$ and $HE_{13}$ mode angle $\theta_N$ in function of the bore diameter for $\lambda_0 = 110\mu m$ for a Pyrex HW. ............................................. 72

5.4 Calculated attenuation coefficients for the first four hybrid modes in a Pyrex HW. ..................................................... 73

5.5 Calculated coupling coefficients for a HW with a bore diameter of 4 mm and $\lambda_0 = 110\mu m$ ................................................. 74

5.6 Transmission of a HW in Pyrex with different core diameter for $\lambda_0 = 110\mu m$ and a waveguide length of $L = 10$ cm in function of the waist size ......... 75

5.7 Schematic diagram showing the copper block mount enabling the QCL to be coupled to the HW. b) CW light-current (L-I) characteristics of the 2.7 THz $3^{rd}$ order DFB QCL with dielectric HW length of 20 mm and 30 mm, and in the absence of the HW. ................................................. 77

5.8 a) Typical farfield beam pattern obtained from a perfectly phase matched 2.7 THz $3^{rd}$ order DFB QCL measured from 11 cm away from the centre of the laser (cf chapter 4) and b) the same measurements but for an oblong and diverging $3^{rd}$ order DFB QCL. c) and d) the corresponding far-field beam patterns for the quasi-Gaussian and oblong $3^{rd}$ order DFB QCLs measured with the HW in place, respectively. .......................... 78

5.9 Physical optics simulations of dielectric hollow waveguides: a) The coupling loss as a function of the laser diode divergence angle for different HW radii. b) Beam pattern of a 4-mm-bore-diameter HW obtained 110mm away from the waveguide aperture. .............................................. 79

5.10 Optimal system for launching a beam into a hollow waveguide. $w_s$ is the minimum waist .................................................. 80

5.11 a), b) and c) Mode profiles of the $3^{rd}$-order DFB QCL coupled to the HW with two parabolic mirrors at distances of 120 mm, 160 mm and 200 mm from the exit of the tube measured with a $R - \theta - \varphi$ far field setup. d) Light-current-voltage of the $3^{rd}$-order DFB coupled with a HW of 4 mm bore diameter in the configuration depicted in 5.10. The measurements were recorded at 10 K in continuous wave operation and the output power of the QCL was measured with a Thomas Keating absolute power meter (courtesy of M. Douared). ........................................ 81
5.12 a) Schematic for the coupling scheme of the QCL with the Dielectric Diagonal Horn Antenna and b) Initial design of the Dielectric Diagonal THz Horn Antenna in Si. c) Calculated Farfield pattern of the horn antenna coupled to a M-M waveguide.

5.13 SEM picture of the SU-8 model of the horn antenna processed with a 2-photons lithography from Nanoscribe (courtesy of Dominique Decanini, C2N).

6.1 Schematic of a bolometer.

6.2 Thermalisation schematic in hot electron bolometer adapted from [5].

6.3 Current-versus-voltage curve of a diffusion cooled HEB with NbN superconducting material grown on HR-Si and measured in a liquid helium dewar at 4.2 K. For the blue curve, the LO is not radiating on the HEB while for the pink, red and black curves, the LO optical power is increased until the negative differential resistance has vanished.

6.4 a) Schematic of the superconducting HEB showing the different diffusion mechanisms: electron-electron scattering, electron-phonon scattering and phonon-phonon scattering. The Andreev reflections marked by the curved arrow on the schematic limit the heat diffusion to the gold pads (normal metal heat sink) and form the hot-spot delimitations. The figure is adapted from [6]. b) Schematic of a 1-D electron temperature in the HEB as a function of the bridge length.

6.5 HEB equivalent circuit.

6.6 Relative mixer gain versus IF frequency calculated for a 4 nm thick NbN films and bath temperature of 6 K with the calculated noise temperature contribution from the Johnson noise and the thermal fluctuation noise, and the receiver noise temperature versus IF frequency.

6.7 Simulated beam patterns at 2.7 THz of the log-spiral antenna in the plane $\phi = 0^\circ$, $\phi = 45^\circ$, $\phi = 90^\circ$ when a) the antenna is surrounded by free space, b) for the case where the antenna is at the interface of the free space and an infinite dielectric half space with the effective refractive index $n = 2$ and c) for the case where the antenna is lying on silicon with $n = 3.42$.

6.8 Physical Optic simulations on Zemax of the extended hemispherical lens with the HEB and its planar antenna placed at the focal point of the lens.
6.9 Scanning electron microscope pictures of a niobium nitride (NbN) HEB on silicon substrate. Picture a) shows the double helix log spiral antenna and picture b) shows the micro-bridge with a passivation layer on the top. (photos from Roland Lefevre) .................................................. 105

7.1 Noise temperature as a function of the physical temperature for a black body radiator at 2.7 THz, according to the Planck, Rayleigh- Jeans and Callen-Welton formulas. ............................................................... 110

7.2 Block diagram of the heterodyne receiver showing the various noise contributions and gain of the optics guiding the RF signal, the mixer which down-converts the RF signal to the intermediate frequency and the IF chain. 110

7.3 Calculated a) reflection and b) transmission coefficient for an incident plane wave at 45° with polarisation in the plane of incidence ($E_P$) and perpendicular to the plane of incidence ($E_S$) for mylar thickness of 50µm. c) and d) Reflection and transmission coefficient respectively for an incident polarized $E_P$ and $E_S$ plane wave at 45° for mylar thickness of 13µm. ... 113

7.4 I-V curves of a pumped HEB at 2.7 THz with a QCL as LO and with hot and cold black body load: the HEB is more pumped with the hot black body compared to the situation with the cold black body. ................. 114

7.5 Schematic of the experimental set-up showing the THz 3rd-order DFB QCL and the solid state source with a frequency emission close to the of the QCL. 115

7.6 I-V curves of the superconducting HEB mixer at 4 K for different radiant power of the local oscillator and the QCL. .......................... 116

7.7 Heterodyne beat note of the QCL with the solid state AMC for a resolution bandwidth of 10 kHz and video bandwidth of 10 The sweep time for this scan is 3 ms. ................................................................. 116

7.8 a) Spectrum and b) farfield measurement of the 2nd-order DFB QCL used as LO for the heterodyne measurements (courtesy of G. Xu for the measurements of the farfield). .................................................. 118

7.9 Experimental set-up for heterodyne sensitivity measurements with HEB and QCL ................................................................. 121

7.10 Heterodyne measurements with 2nd-order QCL and HEB a)I-V curves of the HEB for different level of $P_{LO}$ coupled to the HEB and b) output power of the intermediate frequency measured with a power-meter for two different black body temperatures and noise temperature of the overall heterodyne detector calculated with the Y factor technique. .................. 122

xvii
7.11 Schematic picture of the measurement setup for the ultra compact heterodyne receiver with QCL as LO and HEB as mixer. Notice the absence of beam diplexer for the overlaying of the RF and LO signals. The incoming signals are symmetrically coupled on the HEB bridge and the QCL can be compactly integrated into the cryostat with the mixer block. 124

7.12 HEB current versus voltage characteristics with different level of LO pumping. The HEB’s temperature was 6.1 K without QCL and was 6.49 K when the QCL was switched on. 125

7.13 HEB mixer for two different pumping levels: under pumped and over pumped. In a) IF output power in $\mu$W as a function of the bias Voltage, the red curves correspond to the measurements of the hot and the blue curves of the cold blackbody. b) the DSB receiver noise temperature as a function of bias Voltage. 126

7.14 a) Measured DC-IV curve for the NbN HEB with an optimal LO power for a room temperature and 77 K radiating black bodies. b) Measured DSB receiver noise temperature as a function of the voltage bias of the HEB versus bias voltage with an IF bandwidth of 100 MHz centred at 1.5 GHz. 128
List of Tables

1.1 Few molecular and atomic fine structure emission lines in the THz ........ 2
1.2 Summary of the LO most important required parameters ............... 9

2.1 Calculated parameters of the structure L1395 where $\hbar\omega$ is the photon energy, $z$ the dipole matrix element, $f$ the oscillator strength and $f'$ the normalized oscillator strength, $\Delta$ the miniband width, $F$ the electrical field strength at the anti-crossing and $\tau$ the upper lasing state lifetime. ........ 19
2.2 List of the M-M waveguide material parameters used for the simulation. The Drude relaxation times are found in [7] and [8]. ......................... 23

3.1 Feedback and radiation losses mechanism for different gratings order $M$ and diffracting order $p$ ................................................................. 29

7.1 Optical components with their respective calculated losses. ............. 118
Chapter 1

Introduction

The interstellar medium (ISM), the space which lays between the stars in a galaxy, plays a major role in the galaxy evolution. The ISM is not void but is constituted with dust and gas gathered in large molecular clouds. During their life cycle, the stars reject materials into the ISM through stellar winds, supernova explosions and new stars are being formed by gravitational collapse of the clouds. The galaxy evolution is then characterised by this constant recycling and associated material enrichment between the stars and the ISM. To understand the star formations and the star life-cycle and further the galaxy evolution, the physical processes of the ISM must be studied. Due to the enhancement of the sensitivity and the resolution of the far-infrared (FIR) and submillimeter observatories such as the Herschel Space Observatory \[9\], Stratospheric Observatory for Infrared Astronomy (SOFIA) \[10\] and Millimetre/submillimetre Large Array (ALMA), high spectral and spatial resolution spectroscopic observations of interstellar molecular clouds have been possible. Thanks to these observations, it is now possible to study the physical chemistry that determines the formation and destruction of interstellar molecules, the star formation and their interaction with their surrounding ISM. More than 190 interstellar molecules have been discovered over the past 80 years in the ISM and in circumstellar sources \[11\]. Although, this list is not complete yet. This progress has been possible thanks to the continuous development of highly sensitive spectrometers in the field of laboratory spectroscopy as the detection of new molecules in the ISM starts, most of the time, from getting high-spectral resolution laboratory spectra of molecules. Finding the molecules spectral signatures help to their precise identification in the molecular clouds \[12\]. The majority of the molecules in the ISM contains carbon and organic substances, for which their detections are made in the microwave and submillimeter regime via their characteristic rotational transition frequencies. This latter point represents a real challenge as this frequency band is mostly absorbed by the Earth atmosphere due to the presence of water vapor. The atmosphere has a number of “windows,” in which the absorption is neverthe-
<table>
<thead>
<tr>
<th>Species</th>
<th>$\nu$ (THz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>N$^+$</td>
<td>1.461</td>
</tr>
<tr>
<td>C$^+$</td>
<td>1.901</td>
</tr>
<tr>
<td>N$^+$</td>
<td>2.495</td>
</tr>
<tr>
<td>OH</td>
<td>2.514</td>
</tr>
<tr>
<td>HD</td>
<td>2.674</td>
</tr>
<tr>
<td>NH$_3$</td>
<td>2.948</td>
</tr>
<tr>
<td>OH</td>
<td>3.545</td>
</tr>
<tr>
<td>O</td>
<td>4.746</td>
</tr>
</tbody>
</table>

Table 1.1 – Few molecular and atomic fine structure emission lines in the THz

less smaller (cf figure 1.1). Terrestrial-based FIR observatories must work within these windows. Otherwise, observations have to be done from very high altitudes or in locations where the atmosphere is very dry as for ALMA in the Atacama desert in Chili, from the stratosphere as for SOFIA or from space observatory with Herschel space telescope. In table 1.1 are listed some of the emission-line and atomic fine structures which are under study in the ISM and fall in the THz frequency regime [13].

1.1 Direct vs Heterodyne detection

There are two main competitive techniques for the sub-mm and mm detection. The first of them is the direct detection (incoherent) which allows the measurement of the amplitude of the signal of interest. The second one is the heterodyne detection (coherent) which allows to detect at the same time the amplitude and the phase of the signal of interest.

![Figure 1.1](image-url) – Calculated atmospheric transmission from 1 THz to 10 THz. The calculation was made using the model proposed in [1] for 30 cm long optical path, 25% of humidity at 300 K.
1.1. Direct vs Heterodyne detection

The direct detectors are widely used for spectroscopy measurements and imaging in the infrared, far-infrared, visible and ultraviolet frequency regime. In the sub-mm and mm frequency regime for instance, Golay cells and pyroelectronic detectors are employed for direct detection at room temperature for reasonable sensitivity and time response (usually \( \tau \simeq 10^{-2} - 10^{-3}s \)). A figure of merit for the sensitivity of sub-mm and mm detectors is the noise equivalent power (NEP) which expresses the minimum detectable power per square root bandwidth which can be written, for direct detectors, \( \text{NEP} = \sqrt{\frac{2h\nu}{\eta W_B}} \) (in \( W/\sqrt{Hz} \)) where \( W_B \) represents the background radiation power and \( \eta \) the detector quantum efficiency. Lower is the NEP and more the detector is sensitive. The noise equivalent power for room temperature detectors is typically from \( 10^{-10} - 10^{-9} W/\sqrt{Hz} \). Bolometer and micro-bolometer cooled down to cryogenic temperatures (liquid nitrogen or liquid helium) allows much better sensitivity. Niobium (Nb) bolometer cooled down to 4K can achieve \( \text{NEP} \leq 10^{-11} W/\sqrt{Hz} \). A review of the different incoherent sub-mm and mm detectors can be found in [14] and [15]. Lenses and mirrors are used to focus the incoming radiations on the detector and most of the time, a filter is employed to reduce the background radiations other than the signal of interest. Lock-in amplifier can be used to have a higher signal to noise ratio.

On the other hand, coherent detectors need a local oscillator (LO) to be phase sensitive and operate a mixing: the THz or RF signal at \( \nu_{RF} \) is down-converted to an intermediate frequency (IF) \( \nu_{IF} = |\nu_{RF} - \nu_{LO}| \). The IF signal conserves the phase and amplitude information of the original signal around \( \nu_{LO} \) and is then further amplified and analysed on a spectrometer. The instantaneous spectral range of \( \nu_{RF} \) around \( \nu_{LO} \) is limited and depends on the type of mixer used (cf section 1.2): it is generally referred as the IF bandwidth. Figure 1.2 schematically illustrates the heterodyne down-conversion system. This kind of detection is especially employed when high spectral resolution (\( \frac{\nu}{\Delta \nu} \simeq 10^6 \)) is required. The key element in heterodyne receiver is the detector which performs the non-linear mixing. The RF and LO signal are usually overlaid with diplexers or beam splitter to operate the beating of the two signals. The main advantage of the heterodyne detection is that the frequency and phase information of the RF signal \( \nu_{RF} \) in the THz frequency range are down converted at much lower frequency \( \nu_{IF} \), for which amplifiers, spectrum analysers are available.

Common mixers for heterodyne are devices with a strong quadratic nonlinearity: schottky diodes, superconductor–insulator–superconductor (SIS), superconducting hot electron bolometer (HEB). The sensitivity of a heterodyne detector is usually expressed in terms of single side band (SSB) or double side band (DSB) noise temperature (DSB noise temperatures being two times smaller than the SSB noise) and can be correlated
Chapter 1. Introduction

Figure 1.2 – Heterodyne experimental schematic.

with the mixer noise equivalent power:

\[ NEP_{\text{mix}} \propto k_B T_{\text{mix}} \]  (1.1)

The fundamental limit of sensitivity of heterodyne detectors is the quantum limit expressed as the equivalent quantum noise temperature \( \frac{hf}{2k_B} \) (for DSB receivers) where \( h \) is the Planck constant, \( f \) the frequency and \( k_B \) the Boltzmann constant [16]. This limit doesn’t exist for direct detectors, which means that these detectors can detect low level signals but background noise and saturation power level become problematic. To sum up the needs for detectors in astronomical observations, when a high spectral resolution is required \( \left( \frac{\nu}{\Delta \nu} \approx 10^6 \right) \) the heterodyne detection scheme is preferred but it requires a LO source close in frequency from the absorption or emission line of interest which can be difficult above 2 THz. On the other hand, direct detectors can operate in a wider spectral range when the noise background is low, but have a moderate spectral resolution \( \frac{\nu}{\Delta \nu} \approx 10^3 \) as for [17]. The direct detection is also more convenient for imaging where the sensitivity is more important than the spectral resolution.

1.2 Mixers

In the 1.2 \( \leftrightarrow 5 \) Terahertz (THz) frequency range (\( \lambda = 250 - 60 \mu m \)) the detectors employed in the radio-telescopes are usually composed of a mixer for heterodyne detection. The heterodyne detection is the technique which provides with the highest spectral resolution. There are 3 main mixer technologies in the THz frequency range: GaAs Schottky barrier diodes, superconductor- insulator-superconductor (SIS) tunnel junctions and hot electron bolometers (HEB), the latest stand out to be the most sensitive for frequencies
1.2. Mixers

Figure 1.3 – Sensitivity of the available THz heterodyne mixers compared with their operating frequency. Figure adapted from [2].

above 1.5 THz. Figure 1.3 compares the different available mixers in the THz frequency range with their respective sensitivity.

1 Schottky Mixers

Schottky mixers have a broad operating frequency range up to several THz. They can be operated at room temperature which makes them extremely attractive for spaceborne observatory where cryogenic liquid reservoirs used to cool down usual superconducting detectors are limited. However they need high LO power to reach their highly sensitive state ($\geq 3-8$ mW [18]) and their performances are degraded for frequencies above 1 THz: A double-side-band (DSB) noise temperature $T_{\text{rec}}$ of $\sim 8000$ K at 2.5 THz was measured in [19]. A DSB $T_{\text{rec}}$ was measured at 1500 K for frequencies ranging from 520 to 622 GHz [20] [21] at room temperature. Their sensitivity also increases at lower temperature. Nevertheless, the bandwidth of a Schottky mixer can be large, up to 50 GHz. A SEM picture of Schottky diodes is shown in figure 1.4.

2 SIS Junctions

Below 1 THz the SIS junctions are extremely sensitive mixers (near the quantum noise limit) with recorded equivalent noise temperature of 30 K at 100 GHz [22] and 85 K at 500 GHz [23]. Delorme et al. [24] have demonstrated SIS mixers for Herschel- HIFI band 1, covering the frequency range 480 to 640 GHz with a noise temperature of $\sim 50$ K,
Chapter 1. Introduction

Figure 1.4 – SEM picture of an anti-parallel pair of Schottky diodes from Maestrini et al. [3]

(few times of quantum limit). The maximum operating frequency of the SIS mixers is determined by the frequency gap of the superconducting material: for niobium nitride mixers, this frequency is about 1.4 THz. Their main advantage is that they require a relatively low LO power on the order of a few \(\mu\)W. SIS mixer with 11 GHz of bandwidth has been reported in [25].

3 Hot Electron Bolometer

So far, HEB mixers are the most sensitive mixers for frequencies above 2 THz. They rely on a bolometric effect and contrary to the SIS junctions, the HEBs have not a upper frequency limit. Their superconductor bridge needs to be cooled down at temperatures lower than critical temperature of the superconducting bridge (\(T_C\) around 9 K for NbN HEB) and their IF bandwidth can reach 3 to 4 GHz but they typically require a few hundred nanowatt of LO power to operate heterodyne measurements. HEBs are the preferred mixer for frequencies above 1.5 THz and have already been used up to 1.9 THz in Herschel observatory with the Heterodyne Instrument for the Far-Infrared (HIFI) [26] and at 2.5 THz on SOFIA GREAT [27] and upGREAT [28]. Below 2 THz, waveguides HEB and quasi-optical HEB have relatively similar performances. Currently, the majority of mixers above 2 THz are made with quasi-optical technology since the manufacturing of the waveguides becomes increasingly difficult at very high frequencies. Nobium Nitride (NbN) HEB fabricated at LERMA and C2N have been demonstrated extremely high sensitivity with a recorded DSB noise temperature of \(T_{rec} = 790\) K at 2.5 THz [29]. NbN HEB is the first building block of our THz heterodyne receiver and this technology will be further investigated in the chapter 6. Therefore, the next section will present few LO available in the THz frequency range.
1.3 Local Oscillator

Heterodyne mixers down-convert the frequency of the RF signal by superimposing it with a frequency of reference, the LO signal. The LO has to generate a single and stable frequency with enough output power to pump the mixer into its sensitive state. The emitted frequency needs to be close to the frequency of interest.

**Backward wave oscillators** The Backward wave oscillators (BWO) are vacuum tubes used to generate microwaves and terahertz signal. BWOs are composed of a gun which forms an electron beam that traverses an RF circuit. The slow wave structure imposes modulation on the beam that creates an RF wave that travels from the collector end of the circuit back toward the electron gun. The RF power is coupled out through a waveguide at the gun end of the circuit. The main features of a BWO include the high polarization and monochromatic radiation degree, and the tunability of the output frequency ($\geq 100$ GHz), which depends on the high-applied voltage. BWO can operate from few dozens of GHz up to THz range with mWs of output power. Nevertheless, those THz sources are relatively heavy, consume a lot of power ($> 100$ W) and require water cooling.

**Optically pumped lasers** Optically Pumped Terahertz Lasers (OPTL) use a $CO_2$ laser to excite the roto-vibrational levels of gas molecules at pressures in the sub-mbar range. The most widely used gas is methanol, which has an emission line at 118 µm. OPTL can provide up to several hundreds of milliwatts output power, working in CW, and covering a large range of frequencies. More recently, it has been shown that an OPTL with a $NH_3$ gas medium, can be pumped by a mid-infrared Quantum Cascade Laser (QCL) which operates in CW at room temperature around 1 THz with dozens of µWs of output power [30].

**Amplifier Multiplier Chain** Above 1 THz, only a few sources are available. The most widely used LO is the Amplifier Multiplier Chain (AMC) which emits several dozens of µW at 1 THz but its output power falls down to only few µW above 2 THz. The AMC multiplies an input signal in the tens of GHz frequency range and then amplifies to reach the THz frequency regime. Crowe et al. [31] have measured peak power in the range of 3 µW at 2.7THz. AMC output frequency can be tuned by 10% to 20%.

**Quantum Cascade Laser** The quantum cascade lasers (QCL) are promising alternative THz sources which can provide several mW of output power for frequencies above 2 THz. The QCL used as LO is the second building block of our THz heterodyne receiver and will be discussed in the chapter 2. The state of the art performances of the THz
Chapter 1. Introduction

### Operating Frequency

- 0.8 - 5 THz

### Output Power

- • 24 µm active region embedded into a surface-plasmon
  - 2.4 W at 4.4 THz in pulsed operation at 10K
  - 1.8 W at 77 K [144]

### Output Power in single mode emission

- • 170 mW in single-mode at 3.4 THz
  (pulsed operation 62 K) [145]
- • 6 mW in CW [146], [34]

### Maximal lasing temperature

- • 199.8 K [55]

### Beam Divergence

- • 4° × 4° with 1.35 W output power (6 K) at 3.4 THz [147]

### Wavelength Tuning

- • 1.6–3.5 THz in CW mode using difference-frequency generation (DFG-QCLs) [148]
  Conversion efficiency of 0.6 mW/W²
- • Peak power of nearly 1mW at room temperature,
- • 240 GHz around 3.8 THz of single-mode continuous electronic tuning using MEMS at 4K [149]

Table 1.2 – THz QCL state of the art.

QCL are reported in table 1.2. Although QCLs have large output power compared to the frequency multiplier chains, pumping of a HEB mixer is not a straightforward task since the beam of the QCL has to be coupled with the one of the HEB. Single mode operation and output beam quality can be addressed by a variety of approaches including through the use of photonic crystals [32], surface-emitting distributed feedback lasers [33] and 3rd-order DFB QCL [34]. This work has focused on this latter solution as it has been proven to have a fairly good beam quality and could be made very small, then requiring less electrical power. The theory, design and simulations of this structure can be found in chapter 3. The experimental measurements of our 3rd-order DFB QCL is presented in chapter 4.

1.4 QCL used as LO in THz Heterodyne Receiver

Previous groups have reported successful use of the THz QCL as LO in a heterodyne receiver. Gao et al. have reported in [35] double sideband receiver noise temperature of 1400 K at 2.8 THz and 4.2 K using a HEB and free running QCL. Later, Richter integrated the QCL with the HEB into the same pulse-tube cooler which leaded to a noise temperature of 2000 K at 2.5 THz [36]. In this experiment, they extracted the THz QCL light from a first dewar window and re-injected it through a second window using lenses for the optical coupling with the HEB, as the QCL had a high thermal dissipation. It was a step-forward into the development of a compact THz heterodyne receiver as only one dewar was used on the experimental set-up. More recently, the H-channel on the GREAT (German REceiver for Astronomy at Terahertz frequencies) heterodyne
1.5. QCL Requirements

<table>
<thead>
<tr>
<th>LO Requirements</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output Power</td>
</tr>
<tr>
<td>DC Power Dissipation</td>
</tr>
<tr>
<td>Central Frequency</td>
</tr>
<tr>
<td>Tunability</td>
</tr>
<tr>
<td>Linewidth</td>
</tr>
<tr>
<td>Beam Divergence</td>
</tr>
</tbody>
</table>

Table 1.3 – Summary of the LO most important required parameters

The aim of this thesis is to develop a compact and ultra sensitive heterodyne receiver at 2.7 THz with a QCL as local oscillator and a NbN HEB as mixer. The QCL LO needs to fulfil a number of requirements which are listed in table 1.2. The first of them is the available output power. As it will be discussed in the chapter 6, the HEB needs to be pumped to its sensitive state. The power required for our HEB is about 270 nW and is mostly depending on the dimensions of the HEB micro-bridge. If we roughly estimate the power losses due to absorption, Mylar beam splitter efficiency, coupling mismatch between the QCL beam and the HEB’s antenna at -20 dB (99% of losses, including a safety margin), this translates into a required QCL output power of 100 µW. The QCL power instability is an important issue for heterodyne receiver as it might reduce the sensitivity but it can be settled by using a proportional-integral-derivative (PID) feedback loop to stabilize the QCL output power as in [38]. For a compact integration of the QCL with the HEB into the heterodyne receiver, a relatively low DC power dissipation of the QCL operating after threshold is required. For a compact integration of the QCL, the laser chip will need to
be mounted on the same cold stage plate as the HEB which needs to stay at temperatures below the superconducting material critical temperature all the time of the measurements. Our pulsed tube generator used in chapter 6 can dissipate \( \geq 450 \text{ mW} \) at 4 K. While the cryo amplifier for the IF signal dissipate only few milliwatt, the 300 K radiation coming through the dewar window and the electrical wiring decrease the available cooling power of the cryo-pulser. The QCL power dissipation needs to be much below 500 mW for optimal operation of the heterodyne receiver. Another important consideration is the frequency emission of the QCL. We are interested into the investigation of the strong ground-state line of deuterated molecular hydrogen, HD at 112 \( \mu \text{m} \) (2.677THz). The IF -3 dB bandwidth of our HEB mixer is approximately 3 GHz. The cryo amplifier of the IF channel has a cut-off frequency at 4 GHz with a low-frequency cut-on of 500 MHz. The frequency of the LO should be within 4 GHz frequency offset from the line of interest. The tunability of the QCL is mostly required for astronomical observations of sources in our galaxy with a Doppler shift of about 100 km/s, which corresponds to 1.0 GHz at 2.7 THz \([39]\). The intrinsic QCL linewidth has also to be considered. Without phase lock loop system, the LO linewidth should be at least ten times less than the linewidth of the line of the observed source. A LO linewidth in the order of 1 MHz would allow the observation of several molecular lines. The frequency stability of course has to be on the order of 1 MHz. For comparison, the HIFI instrument on Herschel Observatory had a spectral resolution of 1.14 MHz which corresponds to a velocity resolution of 0.27 km/s. Then a LO linewidth of about 1 MHz would provide a similar velocity resolution at 2.7 THz. Finally, the beam profile of our HEB mixer is determined by the planar antenna coupled to the silicon lens and has a HPBW (half power beam width) of the main lobe around 2.4° \([40]\). The profile at 2.7 THz is Gaussian shaped. The QCL output beam has to match the antenna beam in order to have an optimal quasi-optical coupling with the HEB.

1.6 Thesis Overview

This thesis is focused on the development of a compact and ultra sensitive terahertz heterodyne receiver based on quantum cascade lasers and hot electron bolometers. This work puts the emphasis onto the compact integration of the QCL used as LO. The characteristic required for the LO has been addressed in the introduction. Chapter 1 introduces the basic operation of a THz QCL. The structure that has been used all along this work is also presented. Chapter 2 presents the distributed feedback (DFB) mechanism. The calculation of the coupling coefficient for laterally corrugated waveguide is shown. The simulations and the designs of two DFB structures are presented. Chapter 3 gives the experimental measurements on the THz QCL with the designs previously simulated.
Chapter 4 describes the use of dielectric hollow core waveguides to shape the beam of THz QCL into a nearly Gaussian beam with high directivity. Chapter 5 introduces the operation principles of hot electrons bolometers. Chapter 6 gives the heterodyne experimental measurements and presents the novel concept used for our compact and ultra sensitive heterodyne receiver.
Chapter 2

Basic Operating Principles of THz QCLs

Most of the THz QCLs, and as it will be shown later in this manuscript including the heterostructures used in this work, are composed of a succession of wells and barriers of planar $GaAs/Al_xGa_{1-x}As$ grown with Molecular Beam Epitaxy (MBE) on a GaAs substrate. The heterostructure gives rise to a two dimensional confinement of the carriers and electronic levels quantization in the growth direction of the energy level. The artificially engineered subbands within the conduction band allow a controlled emission wavelength of the photonic transition in the well with the width and the height of the wells and the barriers. The following overview of the inter-subband and QCL theory has been inspired by refs [41] [42] and [32]. The heterostructures used for this work and presented below have been previously designed and measured by Wienold et al. [43] for frequencies from 2.91 to 3.21 THz. A modified version has been grown for a frequency range centred at 2.7 THz. The MBE process has been done in Leeds University by L. H. Li, A.G. Davies and E. H. Linfield from the School of Electronic and Electrical Engineering at University of Leeds.

2.1 Intersubband Transitions & THz Quantum Cascade Lasers

In quantum mechanics, the particle is described by the wavefunction $\Psi(r, t)$. The Schroedinger equation to determine the wavefunction $\Psi(r, t)$ is written:

$$-\frac{\hbar^2}{2m} \nabla^2 \Psi(r,t) + V\Psi(r,t) = i\hbar \frac{\partial \Psi(r,t)}{\partial t}$$

(2.1)

where $\hbar$ is the Planck constant $h$ divided by $2\pi$ and $m$ is the mass of the particle. $V$ is the potential energy function. After applying the separation of variables $\Psi = \Psi(r)\phi(t)$, the time-independent Schroedinger equation is given by:

$$-\frac{\hbar^2}{2m} \nabla^2 \Psi(r) + V\Psi(r) = E\Psi(r)$$

(2.2)
We consider a heterostructure composed of GaAs wells and $Al_xGa_{1-x}As$ barriers, where $x$ is the percentage of the aluminium content. As layers of semiconductor material with different band gap are deposited with a very thin thickness in the order of a few nm, the electrons are confined in the growth direction. When alternating material layers with different bandgap, the conduction band splits into a series of subbands. In the envelope function approximation, the wavefunction is a result of the slowly varying envelope function, $\psi(r)$, modulating the Bloch wavefunction $u(r)$: $\Psi(r) = \psi(r)u(r)$. As the electrons are confined in the growth plane ($xy$ plane), a separation of variables on the envelope function can be applied $\psi(r) = \psi_\parallel \psi_z$ with the component of the wavefunction along the growth plane being $\psi_\parallel = \frac{1}{\sqrt{S_{xy}}} e^{ik_\parallel \cdot r_\parallel}$ where $k_\parallel$ is the in-plane wavevector and $S_{xy}$ is the cross-sectional surface of the quantum-well layers. One takes into account for the interaction of the electron with the periodic potential of the GaAs crystal by multiplying the electron mass by a coefficient equal to 0.067 for GaAs which corresponds to the effective mass of the electron $m_e^* = 0.067m_0$. The effective mass is used to describe the conduction band curvature. To take into account the band discontinuities at the hetero-junction, a spatially varying effective mass $m_e^*(z)$ is used. $\psi_z$ has to satisfy the following Schroedinger equation:

$$\frac{-\hbar^2}{2m_e^*(z)} \frac{\partial^2 \psi_z}{\partial z^2} + V_c(z)\psi_z + V_{el}(z)\psi_z = E\psi_z$$

(2.3)

where $V_c(z)$ represents the conduction band offset energy profile of the hetero-junction and $V_{el}(z)$ the electrical potential due to an external applied electrical field. The engineered levels are controlled by the Al content in the barriers and the well/barrier thickness.

### 2.2 Light Matter Interaction

This section deals with the generation of photons in the heterostructure from the transition of an electron between two sub-bands. This will help to determine the optical gain of the QCL. The rate of radiative transitions, its lifetime, can be evaluated by the Fermi’s Golden Rule. Considering only the stimulated emission, the time dependant harmonic perturbation $H'$ is added to the time-independent Hamiltonian in [44].

$$H = H_0 + H'$$

(2.4)

where

$$H_0 = -\frac{\hbar^2}{2m^*} \nabla^2 + V_c(z) + V_{el}(z)$$

(2.5)

and

$$H' = -eE.r$$

(2.6)
The Hamiltonian for the light-matter interaction is described usually using a Coulomb gauge, in which the vector potential $A(r, t)$ satisfy $\nabla \cdot A = 0$. The electric field is determined from $A$ by:

$$E = -\frac{\partial A}{\partial t}$$  \hspace{1cm} (2.7)

The time-harmonic electric field is then given by:

$$E = uE_0[-ie^{-i(kr-\omega t)} + ie^{i(kr-\omega t)}] = 2uE_0\sin(kr - \omega t)$$  \hspace{1cm} (2.8)

where $k$ is the wave vector, $\omega$ is the pulsation, and $u$ is the unit vector colinear to the direction of the optical electric field. The rate of transitions induced between two quantum-mechanical electronic states of $\hat{H}$, $|i\rangle$ and the final $|f\rangle$ is then [45]:

$$W_{i\rightarrow f}(k_i, k_f) = \frac{2\pi}{\hbar} |< f | V' | i >|^2 \delta(E_f(k_f) - E_i(k_i) \pm \hbar\omega)$$  \hspace{1cm} (2.9)

The transition rate depends on the optical matrix element $V'_{fi} = < f | V' | i >$ which is related to the perturbed Hamiltonian by $H' = V'e^{\pm\omega t}$. The optical matrix element gives a measure of the interaction between the final and initial states with the perturbation. From equations (2.8) and (2.6), the optical matrix element becomes:

$$V'^{\pm}_{(f,k_{||,f})(i,k_{||,i})} = ieE_0u < f, k_{||,f} | e^{\pm kr}.r | i, k_{||,i} >$$  \hspace{1cm} (2.10)

with $k_{||,f/i}$ is the in-plane wavevector of an electron in the final or initial subband. Within the electric-dipole approximation $e^{kr} \sim 1$ and $< f, k_{f,||} | r | i, k_{i,||} >= zz_{fi}\delta^{kr}(k_{||,f}, k_{||,i})$, the equation (2.10) is rewritten:

$$V'^{\pm}_{(f,k_{||,f})(i,k_{||,i})} = \pm ieE_0u zz_{fi}\delta^{kr}(k_{||,f}, k_{||,i})$$  \hspace{1cm} (2.11)

where the Kronecker delta function $\delta^{kr}$ denotes the conservation of the in-plane momentum and $z_{fi} = < \psi_j | z | \psi_i >$ is the dipole matrix element between the z component and the wavefunctions. Only the z polarization of the electromagnetic field couples the electronic states. The dipole-matrix element is written by:

$$z_{fi} = < f | i >= \int_{-\infty}^{+\infty} \psi_f^*(z)z\psi_i(z)dz$$  \hspace{1cm} (2.12)

From equation (2.9), a mode of wavevector $k$, the transition rate from state $i$ to state $f$ is given by:

$$W^{\pm}_{(i,k_{||,i})\rightarrow(f,k_{||,f})} = |u.z|\frac{2\pi}{\hbar} e^2E_0^2|z_{fi}|^2\delta(E_f - E_i \pm \hbar\omega)\delta^{kr}(k_{||,f} - k_{||,i})$$  \hspace{1cm} (2.13)
Chapter 2. Basic Operating Principles of THz QCLs

The notation $\pm$ in (2.13) denotes that the transition between the two electronic states of the unperturbed Hamiltonian $H_0$ is either an absorption or a stimulated. The Dirac function represents then the linewidth of the transition, which can be replaced by a Lorentzian function to account for broadening of the linewidth. The net optical transition rate per unit active region volume $V$ between the initial state $i$ and final state $f$, while factoring in the quasi-Fermi distribution, can then be deduced for stimulated emission and absorption processes by summing the rate over all the in-plane wavevector:

$$W^{\text{em-abs}}_{i\rightarrow f,k} = \frac{2}{V} \sum_{k_f} \sum_{k_i} \{W^+_{(i,k_{\parallel},i)\rightarrow (f,k_{\parallel},f)} f(E_{f,k_{\parallel},f}) (1 - f(E_{f,k_{\parallel},f})) - W^-_{(f,k_{\parallel},f)\rightarrow (i,k_{\parallel},i)} f(E_{i,k_{\parallel},i}) (1 - f(E_{i,k_{\parallel},i})) \}$$

$$= |u.z|^2 \frac{2\pi}{\hbar^2} e^2 E_0^2 |z_{ji}|^2 \delta(E_f - E_i + \hbar \omega) \frac{2}{V} \sum_{k_f} \sum_{k_i} \delta(k_{\parallel,f} - k_{\parallel,i}) (f(E_{i,k_{\parallel},i}) - f(E_{f,k_{\parallel},f}))$$

$$= |u.z|^2 \frac{2\pi}{\hbar^2} e^2 E_0^2 |z_{ji}|^2 \delta(E_f - E_i + \hbar \omega) (N_i - N_f)$$

(2.14)

where $f(E_{n,k_{\parallel,n}})$ is the Fermi function which determines the occupation probability of an electronic state in the $n^{th}$ subband. $N_i$ and $N_f$ are the number of electrons in the initial and final subbands. The rate transition equation (2.19) has been derived from the equation (2.13).

The above equation $z_{ij}$ is related to the oscillator strength $f_{ij}$ by:

$$f_{ji} = \frac{2m^* \omega}{\hbar} |z_{ji}|^2$$

(2.15)

$f_{ij}$ in (2.15) is the relative strength of the optical emission from the state $i$ to $j$. $f_{ij}$ fulfills the sum rule $\sum_j f_{ij} = 1$.

2.3 Miniband electronic transport

The resonant tunnelling transport mechanism is critical to QCLs operation. The elastic scattering, i.e. electron-electron scattering and electron impurity scattering can be treated in the Hartree approximation but this latter scattering mechanism rate has been difficult to evaluate and to measure. A numerical approach using the non-equilibrium Green’s function theory has been proposed to calculate the intra-subband and inter-subband quantum transport [46]. In a superlattice structure, the electronic transport can be treated in the limit of miniband transport. Kleinert and Bryksin in [47] have studied the non equilibrium carrier kinetics by an equation-of-motion analysis of the density matrix from which they have identified the contribution of semiclassical intraband and quantum-mechanical
tunneling. At vanishing electric field, the superlattice is described by two minibands whose dispersion relation is expressed by:

\[
E_1(k) = \frac{\hbar^2 k_x^2}{2m} + \frac{\Delta_1}{2} (1 - \cos(k_z d)) \\
E_2(k) = \frac{\hbar^2 k_x^2}{2m} + E_{\text{gap}} + \frac{\Delta_2}{2} (1 - \cos(k_z d))
\]

(2.16)

where \(\Delta_i\) for \(i = 1, 2\) are the widths of the minibands and \(E_{\text{gap}}\) is the gap between the minibands. \(d\) denotes the superlattice periods and \(k_{xy}\) is the in-plane wave-vector. The current density along the superlattice axis is then determined from two-time miniband transport model [48]:

\[
J(F) = q n \frac{d\Delta}{2\hbar} \frac{I_1(\Delta/2k_B T)}{I_0(\Delta/2k_B T)} \frac{eF\hbar/\tau_e}{(eF\hbar)^2 + \hbar^2/\tau_e\tau_m}
\]

(2.17)

Where \(F\) is the external electrical field applied to the superlattice (the electric field is applied perpendicularly to the superlattice layers), \(\Delta\) the miniband width, \(d\) the period of the superlattice, \(I_1\) and \(I_0\) are the modified Bessel functions. Here \(\frac{qF\hbar}{\hbar}\) is the Bloch frequency. \(\tau_e\) and \(\tau_m\) are the intraminiband scattering time. The current density in equation (2.17) depends on the width of the miniband and the electron temperature. In the case of a vanishing miniband width (decoupled quantum wells with wide barriers), the scattering-induced current, but also the tunnelling current vanishes. Chirped superlattice and bound-to-continuum heterostructures were among the first design to show a lasing effect in the THz where the radiative transition occurs between minibands where the miniband gap is determined by the well width and the miniband width by the barriers thickness.

2.4 Rate equations

The QCL active regions can be well described by a rather simple rate-equation model with constant scattering times [49]. The gain region of a QCL can be described by a three level system where the photonic transition takes place between the states \(n=3\) and \(n=2\) with respective population densities \(n_3\) and \(n_2\). The injection efficiency into the level \(i\) is noted \(\eta_i\).
\[
\frac{dn_3}{dt} = \eta_3 \frac{J}{q} - \frac{n_3}{\tau_3} - S\Gamma_p v_g g_c (n_3 - n_2)
\]
\[
\frac{dn_2}{dt} = \eta_2 \frac{J}{q} + \frac{n_3}{\tau_3} - \frac{n_2}{\tau_2} + S\Gamma_p v_g g_c (n_3 - n_2)
\]
\[
\frac{dS}{dt} = \Gamma v_g g_c (n_3 - n_2) S - \frac{S}{\tau_p} + \beta \frac{n_3}{\tau_{sp}}
\]

where \(\tau_3\) and \(\tau_2\) are the total lifetimes of the electrons in the upper and the lower laser level \(g_c\) is the gain and is a function of the population inversion, \(S\) the photon density and \(\tau_p\) the photon lifetime. \(\beta\) is the the spontaneous emission coupling coefficient. \(\Gamma\) is the confinement factor in the active region and \(\Gamma_p\) the confinement factor per periods. \(v_g = \frac{c}{n_g}\) is the group velocity, \(g_c\) the gain cross section. The rate equations can be easily solved in the steady state case for \(\frac{dn_i}{dt} = 0\). Below the threshold, the photon density is negligible and the solutions are obtained by setting the derivatives and \(S = 0\) in (2.19):

\[
\Delta n = n_3 - n_2 = \frac{J\tau_3}{q} (1 - \frac{\tau_2}{\tau_3})
\]  

(2.19)

The gain cross section \(g_c\) is estimated from the intersubband gain and the confinement factor \(\Gamma\) according to [50]:

\[
g_c = \Gamma \frac{4\pi q^2}{n_r \varepsilon_0 \lambda} \frac{z_{32}^2}{2\gamma_{32} L_p}
\]

(2.20)

where \(n_r\) is the refractive index, \(z_{32}\) is the dipole matrix element and \(\gamma_{32}\) denotes the half width at half maximum of the transition, \(L_p\) the period length and \(\lambda\) the wavelength. At threshold, the photon density is negligible \((\frac{dn_i(t)}{dt} = 0)\) and the modal gain \(g_c \Delta n\) is equal to the total losses \(\alpha_{tot} = \alpha_w + \alpha_m\) where \(\alpha_w\) are the waveguide losses and \(\alpha_m\) the mirror losses. With equation (2.20), the threshold current density is then written:

\[
J_{th} = q \left( \frac{\alpha_w + \alpha_m}{g_c} \right) \frac{1}{\tau_3 (1 - \frac{\tau_{32}}{\tau_3})}
\]

\[
= \frac{1}{\tau_3 (1 - \frac{\tau_{2}}{\tau_{32}})} \frac{2\varepsilon_0 n L_p \lambda \gamma_{32}}{4\pi q \Gamma \frac{z_{32}^2}{2}} \left( \alpha_w + \alpha_m \right)
\]

(2.21)

From (2.21), \(J_{th}\) decreases when the dipole matrix element \(z_{32}\) increases and when \(\gamma_{32}\) decreases. From (2.19) below threshold \((\frac{dS}{dt} = S = 0)\), in order to achieve a population inversion \((n_3 - n_2)\), it is mandatory to have a large \(\eta_3 \tau_3\) while keeping \(\eta_2 \tau_2\) low. One method to achieve the latter condition is depopulation via longitudinal optical phonon scattering [51]. These strategies have been used to optimise the lifetime and depopulation.

18
The design of the heterostructure used in this work (L1395) is adapted from a bound-to-continuum design with resonant phonon assisted depopulation mechanisms. The idea is to use the fast relaxation mechanism of the optical longitudinal phonon which prevents a parallel thermal injection of carriers onto the lower radiative level. Contrary to the first THz QCL design based on chirped superlattice (CSL) design and the Bound-to-Continuum designings from Scalari et al. [54], the structure proposed by Williams [51] uses
longitudinal optical phonon scattering for fast depopulation of the lower lasing level. As it was a major step toward higher working temperatures and higher efficiency, it is interesting to have a quick insight into this design. The heterostructure is constituted of four wells (figure 2.1). The figure 2.1 shows also the squared moduli wave-functions obtained using a 1-D Schrödinger equation solver. In this design, the level 1’ is anticrossed with level 5 through the injector barrier: an electron from the injection level 1’ is tunneling to the upper level 5 of the photonic transition (for a bias of 12kV/cm); The level 5 to 4 is the radiative transition with a frequency $\nu=3.2$ THz ($E_{54}=13$ meV) and depopulation via LO-phonon scattering between level 3 and level 2 with an energy difference of $E_{32}=39$ meV (energy activation for the LO-phonon in GaAs is 36 meV) and quickly (relative to the upper state lifetime $\tau_{21}$) depopulates the lower radiative level. The important feature of the “anticrossed-diagonal” design relies in the increased spatial separation between the injector state and the lower radiative level. This decouples the two wavefunctions and prevents diagonal electron transitions between these two states. As mentioned before, this coupling has detrimental effects on the population inversion and slope efficiency. A drawback of the diagonal design is caused by the additional well in the active region which increases the number of interfaces “seen” by the wavefunctions involved in the optical transition. This broadens the luminescence width and therefore reduces the peak gain. Moreover, these additional interfaces can increase the in-plane momentum scattering, thus shortening the value of $\tau_{\perp}$.

Figure 2.1 – Conduction band of heterostructure active region with a LO-phonon de-population scheme with a bias of 64mV/module from Williams et al. [4]. The radiative transition takes place between the level 5 to 4 with a $E_{54}=14$ meV (3.4 THz).

The design L1395 used in this work is an adapted design of a CSL from Wiedold et al. [43] consisting on 135 periods of nine quantum wells. The conduction band profile and the
Table 2.1 – Calculated parameters of the structure L1395 where $\hbar \omega$ is the photon energy, $z$ the dipole matrix element, $f$ the oscillator strength and $f'$ the normalized oscillator strength, $\Delta$ the miniband width, $F$ the electrical field strength at the anti-crossing and $\tau$ the upper lasing state lifetime.

\[
\begin{array}{ccccccc}
\hbar \omega & z & f & f' & \Delta & F & \tau \\
11 \text{ meV} & 5.2 \text{ nm} & 2.5 & 0.35 & 9.88 \text{ meV} & 4.4 \text{ kV/cm} & 1.6 \text{ ps}
\end{array}
\]

square moduli of the wavefunctions are shown in figure 2.2 with an applied electrical field of 4.4 kV/cm, which corresponds on a processed device of 6.6 V (the threshold has been measured experimentally to be around voltage 5.5 V). The upper level of the radiative transition is either the level 3 or level 2 while the lower level is within the miniband. The extraction from the lower radiative transition is made from the miniband which is aligned resonantly with the states 2 or 3 of a large coupled quantum well. Also, the larger number of wells allows to isolate the phonon well from the injector which reduces the parasitic current channel. The alignment condition is also somehow relaxed thanks to the width of the miniband $\Delta$. In this way, the threshold voltage is also reduced. The oscillator strength is concentrated mainly on the transition between the 2 or 3 and levels in the lower miniband of the active region. The normalised oscillator strength $f' = m^*.f$ is ranging from 0.34 to 0.36 (cf table 2.1).

The L1395 heterostructure was grown with 136 modules and a 15$\mu$m thick active region. The MBE growth was performed on semi-insulating GaAs holder substrates. The structure consists of a 300nm-thick GaAs buffer layer, 700 nm-thick highly doped GaAs layer followed by 136 periods of the active region and an 80 nm-thick doped GaAs contact layer. One period of the heterostructure consists of:

\[
\frac{3}{8.6}/\frac{3}{3}/\frac{3}{9.5}/\frac{3}{11.8}/\frac{3}{12.9}/\frac{3}{16.2}/\frac{2}{10.1}/\frac{3}{4}/\frac{3}{14.5}/\frac{3}{17}/\frac{3}{8.6},
\]

where the thickness is given in nanometres, and the Al$_{0.15}$Ga$_{0.85}$As barriers are indicated in bold. This device was first fabricated in a single-plasmon waveguide and then for the following work in metal-metal waveguide using Au-Au thermo compression bonding. For the single-plasmon device, the lower contact is a 700 nm thick doped layer with a doping of 1.9.10$^{18}$cm$^{-3}$. Fabry-Pérot emission spectra for a 4 mm long and 150 $\mu$m wide waveguide is shown in figure 2.3. At low temperatures, the laser emits at threshold within the range of 0.1 THz around a central frequency of 2.7 THz.

2.6 Waveguides for the THz QCLs

The design of the laser cavity is of main importance for the realization of high performance THz lasers. The resonators used for the THz QCLs are in a way similar to what the radio-
Figure 2.2 – Conduction band diagram of heterostructure active region L1395 with squared moduli wavefunctions for a bias of 57 mV/module. The heterostructure consists of a chirped super-lattice assisted with LO-phonon depopulation scheme. The radiative transition takes place between the level 2 or level 1, or both at the same time to the underneath miniband, while the phonon transition occurs from the miniband to the level 3 or level 2.

frequency domain calls the micro-strip line: the laser cavity, in its simplest type, is a strip line with cleaved mirror facets. There are two main families of resonator depending on the confinement of the propagating optical modes inside the cavity. In a Metal-Metal waveguide (M-M), the light is confined between two metallic claddings while for a Single-Plasmon waveguide (SP) the optical mode is confined between a superior metallic cladding and a Si-doped layer for the bottom cladding. The confinement factor of the optical modes, which quantifies the overlap of the mode with the active region, changes between the type of resonators and depends on the frequency of the modes, the order of the mode and on the width of the waveguide (figure 2.6b). The limited thickness of the heterostructure enables only the propagations of the longitudinal transverse magnetic $TM$ modes as there is a frequency cut-off for the TE modes. The waveguide’s $TM$ mode satisfies the polarization rule of inter-subband transitions, which require the electric field to be orthogonal to the quantum wells (growth directions) while the longitudinal magnetic field oscillates in the transversal direction, in the plane of the quantum wells. For quasi-TEM modes (cf figure
Figure 2.3 – Lasing spectra of a 4 mm long and 150 μm wide laser ridge of L1395 just after threshold and maximum output power at 10 K in pulsed operation.

2.4), the components along z and x of the electrical field are nil: \( \mathbf{E} = E_y \mathbf{y}, E_z = E_x = 0 \). The confinement factor \( \Gamma \) can be obtained by the integration over the area of the active region (AR) and over the whole space:

\[
\Gamma = \frac{\int_{\text{AR}} |E_z|^2}{\int_{\text{tot}} |E_z|^2}
\]  

(2.23)

The confinement factor is then used to determine the lasing threshold of the laser [8]:

\[
g_{th} \propto \frac{\alpha_m + \alpha_i}{\Gamma}
\]  

(2.24)

with \( \alpha_m \) and \( \alpha_i \) respectively the mirror losses and the intrinsic losses of the waveguide. For single plasmon and metal-metal waveguides, \( \alpha_i \) are mostly due to the free-carrier absorption in the doped layer and the active region. A high confinement factor and low metallic loss allow then to lower the intrinsic loss \( \alpha_i \) of the waveguide, a higher modal gain \( \gamma_{\text{mod}} \) and a lower lasing threshold \( g_{th} \). At higher temperature, the rate of population inversion is lower due to other competitive mechanism such as longitudinal optical phonon
emission from thermally activated carriers. Nonetheless, with lower intrinsic losses and lower threshold, the MM waveguides have demonstrated higher operating temperatures than SP waveguides, up to 200K [55]. The lack of bottom metallic cladding on SP waveguide can’t prevent the leakage of the optical mode into the substrate which leads to a confinement factor inferior to 0.5 [42]. The determination of the mirror losses $\alpha_m$ also differ between the SP and MM waveguides. While the calculation of the mirror losses for the SP waveguide is rather straightforward, it is slightly more tricky with the MM waveguides. The general expression of $\alpha_m$ is written below:

$$\alpha_m = -\frac{\ln(R_1 R_2)}{2L}$$  \hspace{1cm} (2.25)

The determination of the facet reflectivity $R$ on SP waveguide is done using the effective index method which use the Fresnel reflection coefficient and is determined by the effective refractive index of the waveguide material as described in [45]. For an interface GaAs/Air, the facet reflectivity, without anti-reflection coating $R = \frac{(n_{GaAs} - 1)^2}{(n_{GaAs} + 1)^2} \sim 0.32$. Unlike the SP waveguide, the mirror reflectivities on MM waveguides depends on the frequency of the optical mode and on the waveguide geometry. Mirror reflectivity for MM waveguides is greater than those for SP waveguides; the mirror losses contribute less to the overall loss on the MM waveguides. The method to determine the facet reflectivity of MM waveguide as in [45] is to use the S-parameters: the $s_{11}$ parameter which is the power of the reflected modes at the facets divided by the power of the incoming modes. For wide waveguides, where higher order transverse modes are allowed to propagate, the reflected power at the waveguide facets is then carried by all the modes excited at the facets: the facet reflectivity parameter is relevant only for the modes of interest and not the reflection.
2.6. Waveguides for the THz QCLs

<table>
<thead>
<tr>
<th>Layer</th>
<th>Thickness</th>
<th>n</th>
<th>( \tau_{\text{Drude}} )</th>
<th>( \varepsilon_{\text{core}} )</th>
<th>( m^* )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Top/Bottom Au</td>
<td>300 nm</td>
<td>5.9x10^{22} cm(^{-3})</td>
<td>120 fs at 77 K</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>Active Region</td>
<td>14 ( \mu )m</td>
<td>2.0x10^{15} cm(^{-3})</td>
<td>500 fs at 77 K</td>
<td>12.96</td>
<td>0.067</td>
</tr>
</tbody>
</table>

Table 2.2 – List of the M-M waveguide material parameters used for the simulation. The Drude relaxation times are found in [7] and [8].

into higher order modes; The mirror reflectivity in metal-metal waveguide is ranging from \( R_{MM} \sim 0.7 \) to 0.9 [8]. Another consequence of the strong confinement in THz QCLs with M-M waveguides is the very wide and highly divergent beam emission due to the high confinement of the optical mode inside the cavity and the impedance mismatch of the mode with the free space propagation. The emission pattern of a MM waveguide is not ideal for application which requires quasi-optical coupling with detectors and highly collimated beam. Three-dimensional finite element simulations have been carried out with commercial solver for the calculation of the effective index, the confinement factor and the waveguide/modal losses. The waveguide material was modeled as bulk lossless GaAs while the free-carrier contribution to the complex relative permittivity of the metal claddings has been taken into account by choosing a complex \( \varepsilon_r \) using the Drude–Lorentz approximation. According to the Drude model in [56], the complex relative permittivity of a medium can be calculated with:

\[
\varepsilon_r = \varepsilon_{\text{core}} + \frac{ine^2\tau}{\varepsilon_0\omega m^*(1 - i\omega\tau)}
= \varepsilon_{\text{core}} \left(1 - \frac{\omega_p^2\tau^2}{1 + \omega^2\tau^2} + i\frac{\omega_p^2\tau}{\omega(1 + \omega^2\tau^2)}\right)
\]

(2.26)

where \( \varepsilon_{\text{core}} \) is the relative permittivity of the material, \( m^* \) the effective mass of the electron, \( n \) the electron density, \( \omega \) the frequency of the mode, \( \omega_p = \sqrt{\frac{ne^2}{\varepsilon_0\varepsilon_{\text{core}}m^*}} \) the plasma frequency in the material and \( \tau \) the effective scattering time for the electron. For the Au claddings, \( \omega_p/2\pi \sim 2000 \) THz and below this frequency, the real part of \( \varepsilon_r \) is negative and a surface plasmon mode can propagate at the GaAs-Metal interface (exponential decay of the mode in the metal). The material parameters used for the simulations are shown in table 2.2. The conductivity of the metal through the Drude model is given by:

\[
\sigma = \frac{ne^2\tau}{m^*(1 + \omega^2\tau^2)} \quad [45].
\]

Figures 2.5a and 2.5b show the plots of the real and imaginary parts of gold (Au) over the THz frequency range calculated using the Drude models.

Figure 2.6 shows FEM simulations of the effect of the M-M waveguides ridge width on the modal losses, the confinement factor and the mode effective index for the fundamental transverse magnetic mode \( TM_{00} \) and first higher order mode \( TM_{01} \). For waveguide’s
Figure 2.5 – Calculated a) relative permittivity and b) conductivity of Au versus the frequency.

Figure 2.6 – Calculated a) modal losses, b) confinement factor and c) effective index for \( TM_{00} \) and \( TM_{01} \) modes at 2.7 THz in function of the waveguide width for a metal-metal waveguide with a thickness of 10 \( \mu m \).

width \( \leq \lambda_{GaAs} \), an important fraction of the guided mode in the cavity leaks outside of the waveguide which therefore reduces the mode effective index to values smaller than the index in the material (\( n_{GaAs} = 3.59 \)) and increases the modal losses (figure 2.6a). Since carrier absorption in the active region has been neglected, the waveguide losses ranging from 4 to 4.5 \( cm^{-1} \) for ridges width varying from 10 to 45 \( \mu m \) are mainly caused by the
2.7 Intrinsic linewidth of QCLs

The linewidth of a laser diode is an important parameter for applications which require a long coherent length and high spectral resolution. It depends mostly on the spontaneous emission and the fluctuation of the refractive index. The Schawlow-Townes equation gives the FWHM linewidth \( \Delta \nu \) of a single mode laser diode [57][58]:

\[
\Delta \nu = \frac{h \nu v_g (\alpha_i + \alpha_m) \alpha_m n_{SP}}{8 \pi P_{out}} (1 + \alpha_H^2)
\]  

(2.27)

with \( h \) the Planck constant, \( \nu \) is the laser frequency, \( v_g \) the group velocity of the optical mode inside the active region, \( \alpha_i \) the intrinsic loss of the material, \( \alpha_m \) the mirror losses, \( n_{SP} \) the population inversion factor, \( P_{out} \) the output power and \( \alpha_H \) is the “Henry linewidth enhancement factor.” Schalow-Townes determined the very "limitation" for the linewidth of a laser to be infinitely narrow with the above formulation (2.27) the spontaneous emissions in the active region are limiting the sharpness of the linewidth as the spontaneous emission is adding up a random phase contribution to the electromagnetic field inside the cavity.

The equation (2.27) shows that the linewidth is inversely proportional to the cavity length and the output power of the laser. Yamanishi et al. theory [59] predicts an intrinsic linewidth for QCLs below the Schawlow-Townes limit. QCLs in the mid-infrared regime have been reported with linewidth as narrow as few hundred Hertz [58] and THz regime, sub-Hz linewidth [60] when the QCL is stabilised with a phase lock loop system. DFB and DBR laser diodes have narrower linewidth than Fabry-Pérot cavity as shown in [61]: the DFB’s linewidth differs the Schawlow-Townes formulation in the expression of the cavity loss (threshold gain), which is lower for DFB than for Fabry-Pérot thanks to the
distributed feedback mechanism.
Chapter 3

Third Order Distributed Feedback Lasers

3.1 Distributed Feedback Lasers

The development of the distributed feedback lasers has been possible thanks to the works of Bragg in the beginning of the 19th century on periodical systems selecting the wavelength of light through the phenomena which are described in [62]. As described on the cartoon figure 3.1a, an incident plane wave with an angle $\theta_i$ on a periodic grating structure, $\Lambda$ being the grating periodicity, is diffracted at the $M$ - th order at an angle $\theta_d$. When the Bragg conditions are met, equation (3.1), the reflected waves from adjacent grating corrugations are constructively adding up in phase with an angle $\theta_d$. The order of grating is also defining the order of the diffraction between the diffracted waves from adjacent grating elements.

$$(\sin(\theta_i) - \sin(\theta_d))\Lambda = M\lambda$$ (3.1)

As the Bragg grating can select specific frequencies, Kogelnik and Shank [63] embedded

Figure 3.1 – a) Reflection on Bragg gratings. b) Waveguide with embedded Bragg gratings. For an embedded grating onto a laser, $\theta_i = \pi/2$ and for the diffracted wave $\theta_d = \pi/2$. For $p = 2$, $\theta_d = 0$ and the diffracted waves couple out from the guide.
this diffracting grating in a semiconductor laser. To obtain a constructive interference between two following grating corrugations as presented in figure 3.1b, the total optical path must be, in radians, an integer multiple of $2\pi$. This condition can be expressed, with the wavelength inside the waveguide $\lambda_i$ and outside the medium $\lambda_d$.

\[
\frac{\Lambda \sin(\theta_i)}{\lambda_i} - \frac{\Lambda \sin(\theta_d)}{\lambda_d} = M \tag{3.2}
\]

The equation (3.2) shows also that the propagating mode in the DFB laser waveguide cannot have an angle of incidence smaller than the critical angle for total internal reflection \[64\] $\sin(\theta_i) \geq \frac{n_d}{n_i}$. Combining (3.1) with the condition on the critical angle limit:

\[
\sin(\theta_d) \geq \frac{\lambda_m M}{\Lambda} - 1 = \frac{2M}{p} - 1 \tag{3.3}
\]

With $\lambda_m$ the effective wavelength in the waveguide and $p = \frac{2\Lambda}{\lambda_m}$ the order of the grating. Usually it is required that for a DFB waveguide laser with gratings all along the waveguide to have $\theta_d = \theta_i = \pi/2$. In that case, the order $p$ is equal to the order $M$ of diffraction. With the grating periodicity $\Lambda$, the optimum free-space wavelength $\lambda_B$ which is most strongly reflected is called the Bragg wavelength and satisfies the Bragg condition:

\[
M \frac{\lambda_B}{2n_{eff}} = \Lambda \tag{3.4}
\]

But gratings can be found also with a diffracted emission angle $\theta_d = 0$ or $\sin(\theta_d) \ll 1$ so that the diffracted wave can couple outside of the waveguide, for example for the second order of gratings $p = 2$: in this case, the light couples out of the laser into the perpendicular directions of the waveguide. Table 3.1 presents the influence of grating order $p$ and diffraction order $M$ on the feedback and out-coupling mechanisms experienced by a propagating mode. A third order grating couples the light outside of the waveguide in the first and second diffraction orders at symmetrical angles within the normal of the waveguide, but provides the feedback in the third diffraction order. Actually, it can be noticed that the feedback is obtained in the $N$-th diffracting order for a grating of order $N$ and the $N$-1 other orders are diffracted and represent the optical losses of the grating.

For a DFB grating at diffraction order $M = 3$ and period $\Lambda$, the vacuum Bragg wavelength is $\lambda_B = \frac{2n_{eff}\Lambda}{M}$. The surface emission of the DFB grating is illustrated in figure 3.1b. The radiated wave at angle $\theta_d$ has a propagation constant $k_d = \frac{2\pi}{\lambda_0}$, while
3.2. Overview of the coupled modes analysis

The coupled mode analysis has proved to be a powerful tool for the understanding of DFB lasers. Many text books have already presented this theory. Only a small insight of the theory will be given below, and this discussion has been widely inspired by [49] or [65].

The index coupled DFBs are here studied but gain coupled DFB structures are solved in a similar way.

Let’s consider a uniform periodical grating implemented along the z-axis of a waveguide. The dielectric constant of the waveguide is then changed along z and unchanged in the xy plane. We consider a TE wave propagating along the z-axis so \( k_x = k_y = 0 \) and \( E_x = E_z = 0 \) and \( E_y \neq 0 \). \( E_y \) verifies the equation:

\[
\left( \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z^2} + k_0^2 n^2(x, z) \right) E_y(x, z) = 0
\]

(3.6)

As an infinite structure is considered, the dielectric function of the waveguide \( \varepsilon \) can be

<table>
<thead>
<tr>
<th>order of diffraction</th>
<th>M=0</th>
<th>M=1</th>
<th>M=2</th>
<th>M=3</th>
</tr>
</thead>
<tbody>
<tr>
<td>p=1, 1(^{st}) order grating</td>
<td>Forward propagation</td>
<td>Feedback</td>
<td></td>
<td></td>
</tr>
<tr>
<td>p=2, 2(^{nd}) order grating</td>
<td>Forward propagation</td>
<td>Radiation at ( \theta_d = 90 )</td>
<td>Feedback</td>
<td></td>
</tr>
<tr>
<td>p=3, 3(^{rd}) order grating</td>
<td>Forward propagation</td>
<td>Radiation at ( \theta_d = \sin^{-1}(-1/3) )</td>
<td>Radiation at ( \theta_d = \sin^{-1}(1/3) )</td>
<td>Feedback</td>
</tr>
</tbody>
</table>

Table 3.1 – Feedback and radiation losses mechanism for different gratings order M and diffracting order p

In the propagating mode inside the cavity has a propagating constant \( k_i = \frac{2\pi n_{eff}}{\lambda_0} \) and one grating vector \( k_g = \frac{p2\pi}{\Lambda} \) with p still the grating order. With the continuity of the electromagnetic field at the boundary condition, the parallel propagation vector \( k_\parallel \) must be conserved:

\[
sin(\theta_d) = n_{eff} - \frac{\lambda_0}{\Lambda} p
\]

(3.5)

For the third order of diffraction, it can be noted that if the effective refractive index inside the cavity is kept at \( n_{eff} = 3 \), the emission angle of the diffracted mode will occur at \( \theta_d = 0 \). This property is of main importance for the THz MM QCLs with 3\(^{rd}\) order DFB to improve the out-coupling of light from the semiconductor into the free space.
expanded in Fourier series as following.

\[ n^2(x, z) = \sum_{m=-\infty}^{+\infty} A_m(x) e^{2i\pi m z / \Lambda} \]  \hspace{1cm} (3.7)

with

\[ A_m(x) = \frac{1}{\Lambda} \int_0^\Lambda n^2(x, z) e^{-2i\pi m z / \Lambda} \, dz \]  \hspace{1cm} (3.8)

The electrical field in the cavity considered can be decomposed in the sum of the propagating A and counter-propagating B waves in the waveguide and \( E_y \) can be written:

\[ E_y(x, z) = A(z) e^{-i\beta_A z} E_A(x) + B(z) e^{i\beta_B z} E_B(x) \]  \hspace{1cm} (3.9)

with A and B the respective amplitude of the propagating and counter-propagating mode and \( \beta_{A,B} \) the Bragg propagation constants along z for the corresponding modes and \( E_{A,B}(x) \) are the two propagating and counter-propagating mode’s amplitude, solutions of the two dimensional waveguide for the reference waveguide. Supposing that the amplitude A and B are slowly varying along z, \( \frac{\partial A^2}{\partial z} \) and \( \frac{\partial B^2}{\partial z} \) can be neglected. Then:

\[ \frac{\partial^2 E_y(x, z)}{\partial x^2} = \left[ -2iA'(z)\beta_A - \beta_A^2 A(z) \right] e^{-i\beta_A z} E_A(x) + \left[ -2iB'(z)\beta_B - \beta_B^2 B(z) \right] e^{i\beta_B z} E_B(x) \]  \hspace{1cm} (3.10)

Replacing (3.7) (3.8) (3.9) in (3.6) results in:

\[ \frac{\partial^2 E_A(x)}{\partial x^2} + \frac{\partial^2 E_B(x)}{\partial z^2} = -A(z) e^{-i\beta_A z} + B(z) e^{i\beta_B z} \]

\[ + \left[ -2iA'(z)\beta_A - \beta_A^2 A(z) \right] e^{-i\beta_A z} E_A(x) + \left[ -2iB'(z)\beta_B - \beta_B^2 B(z) \right] e^{i\beta_B z} E_B(x) \]

\[ + k_0^2 \left[ n^2(x) + \sum_{m=-\infty, m\neq 0}^{+\infty} A_m(x) e^{2\pi m z / \Lambda} \right] (A(z) E_A(x) e^{-i\beta_A z}) \]

\[ + B(z) E_B(x, y) e^{i\beta_B z} = 0 \]  \hspace{1cm} (3.11)

(3.11) is organized in a system of two equations as following:

\[
\begin{align*}
\frac{\partial^2 E_A(x)}{\partial x^2} A(z) + [-2iA'(z)\beta_A - \beta_A^2 A(z)] E_A(x) + k_0^2 [n^2(x)A(z)E_A(x) + A_{-m}(B(z)E_B(x))] &= 0 \\
\frac{\partial^2 E_B(x)}{\partial x^2} B(z) + [-2iB'(z)\beta_B - \beta_B^2 B(z)] E_B(x) + k_0^2 [n^2(x)B(z)E_B(x) + A_m A(z)E_A(x)] &= 0
\end{align*}
\]  \hspace{1cm} (3.12)
The condition for the forward and backward modes to interact in the waveguide is then:

\[ \beta_A + \beta_B = \frac{2\pi m}{\Lambda} \]  

(3.13)

This condition is known as the Bragg condition. \(E_{A,B}\) are also solutions of the equation:

\[ \left( \frac{\partial^2}{\partial x^2} + k_0^2 n^2(x) + \tilde{\beta}_{A,B}^2 \right) E_{A,B} = 0 \]  

(3.14)

where \(\tilde{\beta}_{A,B}\) are the propagation constants of the non-perturbed field in the cavity. For a weak coupling, and supposing that the periodical grating is not changing much the electrical field so that (3.14) is conserved, the coupled mode equations are written:

\[
\begin{cases}
(-2i\beta_A A'(z) + \left[ \beta_A^2 - \beta_B^2 \right] A(z)) E_A(x) + k_0^2 A_{-m} B(z) E_B(x) = 0 \\
(-2i\beta_B B'(z) + \left[ \beta_B^2 - \beta_A^2 \right] B(z)) E_B(x) + k_0^2 A_m A(z) E_A(x) = 0
\end{cases}
\]  

(3.15)

We define the detuning number \(\delta_{A,B} = \tilde{\beta}_{A,B} - \beta_{A,B}\). For small perturbation, we assume \(\beta_{A,B} \sim \tilde{\beta}_{A,B}\) and \(\beta_{A,B}^2 - \beta_{A,B}^2 \sim 2\beta_{A,B}^2 \delta_{A,B}\). From (3.15), we have:

\[
\begin{cases}
(-A'(z) - i\delta_A B A(z)) E_A(x) = \frac{i k_0^2}{2\beta_A} A_{-m} B(z) E_B(x) \\
(-B'(z) - i\delta_B B B(z)) E_B(x) = \frac{i k_0^2}{2\beta_B} A_m A(z) E_A(x)
\end{cases}
\]  

(3.16)

If \(E_A\) and \(E_B\) are two eigenmodes of the cavity, they are orthogonal and their scalar product is nil:

\[ \int_{-\infty}^{+\infty} E_A E_B = \delta_{A,B} \]  

(3.17)

Multiplying (3.16) by \(E_A^*\), the only solution different to zero is obtained for \(E_A = E_B\). We further simplify the notation with \(E_A = E_B = E\), \(\delta_A = \delta_B = \delta\) and \(\beta_A = \beta_B = \beta_{\text{Bragg}}\). The equation (3.13) defines the Bragg wave vector as:

\[ \beta_{\text{Bragg}} = \frac{m \pi}{\Lambda} \Rightarrow \lambda_B = \frac{2 n_{\text{eff}} \Lambda}{m} \]  

(3.18)
Finally, the coupled wave equation is obtained:

\[
\begin{align*}
\frac{\partial A(z)}{\partial z} - i\delta \beta A(z) &= i\kappa B(z) \\
-\frac{\partial B(z)}{\partial z} - i\delta \beta r B(z) &= i\kappa A(z)
\end{align*}
\]

with \( \kappa \) the coupling coefficient defined as:

\[
\kappa = \frac{k_0^2}{2\beta_B} \left| \int_{cell} \Delta n^2(x, z)|E_y(x)|^2 \right| \int_{-\infty}^{\infty} |E(x)|^2
\]

The grating induces periodical changes in the effective index of the waveguide and the constant \( \kappa \) represent the intensity of the coupling between the propagating and counter-propagating modes. The next section will deal with the calculation of this constant for different designs.

### 3.3 Calculation of the coupling coefficient for lateral DFB gratings

The coupling constant, as found in the section 3.2, is an important parameter of distributed feedback laser: it measures the intensity of the coupling between the propagating and the counter-propagating modes inside the cavity. Being able to evaluate \( \kappa \) from electromagnetic simulations or through analytical studies is an interesting asset for the design of DFBs with single mode emission features.

As discussed later in this report, the approach chosen for the 3\(^{rd}\) DFB QC lasers is to use laterally-corrugated ridge-waveguide gratings. There are several advantages in the use of lateral corrugation for our application. The first of them is that there is a limited interaction between the corrugation and the carriers unlike with the usual gratings on the top metallisation [66]. The second advantage, which will be further discussed later in the chapter, is that the deep lateral corrugation can be used to lower down the effective index of the laser cavity, as it is mandatory to obtain, for a 3\(^{rd}\) order DFB, a favourable phase coupling between the propagating mode and counter propagating mode. The previous section gave an expression for the coupling coefficient in a distributed feedback laser. Nevertheless, \( \kappa \) is slightly underestimated for the situation where lateral corrugations are used [66]. A calculation method of the coupling coefficient for the distributed feedback QC lasers with lateral corrugations will be presented in this section. It will further allow to select the right \( \kappa L \)-product where \( \kappa \) is the coupling constant and \( L \) the length of the DFB laser cavity. It has been experimentally shown that an optimal value for \( \kappa L \) for a
3.3. Calculation of the coupling coefficient for lateral DFB gratings

DFB laser is in the range of 1.0-2.0 [49]. When the $\kappa L$ value is too low, there is no mode selection and on the contrary when $\kappa L$ is too high, the strong longitudinal variation of the field inside the cavity cause spatial hole burning.

Since $\Delta \varepsilon$ is periodic along $z$ with the grating period $\Lambda$, it can be expanded in a Fourier series:

$$\Delta \varepsilon(x, y, z) = \sum_{m \neq 0} \varepsilon_m(x, y) e^{-\frac{2\pi im}{\Lambda} z}$$

From equation (3.21), the Fourier coefficient $\varepsilon_m(x, y, z)$ for the $m$th-order Bragg diffraction is given as [67]:

$$\Delta \varepsilon_m(x, y, z) = \frac{1}{\Lambda} \int_0^\Lambda \Delta \varepsilon(x, y, z) e^{-\frac{2\pi imz}{\Lambda}} dz$$

$$= \left( n_2(x, y)^2 - n_1(x, y)^2 \right) \frac{\sin(\pi m \gamma)}{\pi m}$$

where $n_1$ and $n_2$ are the transverse refractive index in the different dielectric slab of the grating, $m$ is the grating order, and $\gamma$ is the grating duty cycle, defined as the ratio between the length of the waveguide’s section with effective refractive index $n_1$ within the waveguide’s section of index $n_2$. For deep etched corrugation, the top metal layer is also corrugated and the effective refractive index changes need to be taken into account for the calculation of the coupling coefficient. The coupling coefficient $\kappa$ of the rectangular grating can be written as:

$$\kappa = \frac{k_0}{2n_{\text{eff}}} \left\{ \int \int_{\text{grating}} |n_2(x, y) E_y(x, y)|^2 \right\} \left[ \int \int_{-\infty}^\infty |E_y(x, y)|^2 \right] - \left[ \int \int_{-\infty}^\infty |E_y(x, y)|^2 \right] \frac{\sin(m \pi \gamma)}{m \pi}$$

where $n_1(x, y)$ and $n_2(x, y)$ are the transverse refractive index distributions for wide-ridge and narrow-ridge sections. This approach can be applied for the conventional DFB for which the grating corrugations are embedded. The refractive index of the different section of semiconductor materials are close to the effective index of the propagating mode in the cavity.

However, for the laterally corrugated waveguides, where the lateral mode profiles change between the wide and narrow sections of the waveguide, this standard formula underestimates $\kappa$ compared to experimental results (shorter photonic band-gap, multimodal behaviours of the laser). Following [49], we rewrite the coupled mode equation.
obtained in the precedent section by:

\[
\frac{d}{dz} \begin{pmatrix} A(z) \\ B(z) \end{pmatrix} = i \begin{pmatrix} \Delta \beta & \kappa \\ -\kappa^* & -\Delta \beta \end{pmatrix} \begin{pmatrix} A(z) \\ B(z) \end{pmatrix}
\] (3.24)

Where we have defined \( \Delta \beta = \beta_0 - \beta_{\text{Bragg}} \) and \( \beta_0 = \frac{\omega_{\text{eff}}}{c} \). The general solution of the coupled mode equation are then:

\[
\begin{pmatrix} A(z) \\ B(z) \end{pmatrix} = C_1 \begin{pmatrix} -\kappa \\ \Delta \beta - is \end{pmatrix} e^{iqz} + C_2 \begin{pmatrix} -\kappa \\ \Delta \beta + is \end{pmatrix} e^{-iqz}
\] (3.25)

where \( s \) is the eigenvalue define as:

\[
s = \sqrt{|\kappa|^2 - \Delta \beta^2} = -iq
\] (3.26)

From the equation (3.9), the dispersion relation of the propagating and counter-propagating mode inside the cavity is now:

\[
\omega_{\pm}(q) = \frac{c}{n} \left( \frac{\pi}{\Lambda} \pm \sqrt{q^2 + |k|^2} \right)
\] (3.27)

The situation where \( q = 0 \) correspond to the eigen-modes propagating at the \( \gamma \) point of the Brillouin zone of a one-dimensional photonic crystal. With \( \kappa = K + iK_g \), the complex frequencies of the band edge modes are:

\[
\omega_{\pm}(0) = \frac{c}{n} \left( \frac{\pi}{\Lambda} \pm k \right) = \frac{c}{n} \left( \frac{\pi}{\Lambda} \pm K + iK_g \right)
\] (3.28)

The coupling coefficient is then:

\[
| K + iK_g | = \frac{n_g}{2c} (\omega_+ - \omega_-)
\] (3.29)

with \( n_g \) the refractive index of the reference waveguide. The coupling coefficient is now obtained through the calculation of the frequencies of the eigenmode at the Bragg wavevector. The calculation of the coupling coefficient through the two methods proposed has been investigated: the method for conventional DFB with embedded grating and the second one assuming the DFB as a 1-D photonic crystal and using the dispersion relation. For the first method, evaluating \( \kappa \) using equation (3.23), we have used a commercially available finite element software which solves for the eigenmode to determine separately...
the effective refractive index values and the field distribution in the x-y plane of the waveguide in the two different sections of the grating. For the second method, we evaluated $\kappa$ using the formula in equation (3.29). For a DFB structure with lateral grating, it requires to solve for the Maxwell’s equations in the volume of a DFB unit cell using Bloch periodical boundary conditions in the ridge direction. The coupling coefficient values calculated with the dispersion formula (eq 3.29) and using the conventional approach (eq 3.23) are illustrated in figure 3.2. The coupling coefficient was calculated for a third-order distributed feedback with lateral corrugation grating on a laser cavity made of GaAs (refractive index $n = 3.55$ at 2.7 THz). The waveguide is 10 $\mu m$ of thickness, while its wide section is kept at 15 $\mu m$. The width of the narrow part of the waveguide, corresponding to the grating section, is then tuned to observe the influence of depth of the corrugations on the coupling coefficient of the DFB. The duty cycle, i.e. the ratio between the length of the narrow section of the waveguide with the length of the wide section, is kept at DC = 12%. In the simulation, an infinitely long DFB with symmetrical lateral corrugations on both sides of the cavity has been considered. As depicted on figure 3.2, the magnitude of the coupling coefficient increases with the depth of the lateral corrugation compared to the wide section of the waveguide. The conventional method gives a smaller coupling coefficient than the alternative method. The value of $\kappa$ is 4 times larger when calculated with the eigenfrequencies than calculated using the approximation $\Delta \varepsilon = n_{eff,2} - n_{eff,1}$. Equation (3.23) under-estimates the coupling factor in the case of deep etched lateral corrugation gratings.
Chapter 3. Third Order Distributed Feedback Lasers

3.4 Engineering the Distributed Feedback Structure

It will be discussed in this section the different parameters in the design of DFB cavity which will help to obtain the desired specification, i.e. a single mode operation at targeted frequency. The coupling factor dependence on the filling factor of the grating is analysed here. The filling factor affects the optical confinement factor in the grating region and the effective refractive index values. These supplementary influences result from the fact that the longitudinally transverse refractive index distribution changes with the filling factor according to equation 3.23 and the optical field distribution (x, y) changes correspondingly, influencing the optical confinement factor in the grating region. The dependence of the coupling coefficient \( \kappa \) with the grating duty cycle has been analysed using only the formula which take into account the eigen-frequency of the propagating modes (equation (3.29)) as it gives a better approximation for DFB with lateral corrugations. Figure 3.3c illustrates a calculated coupling coefficient in function of the grating filling factor. The calculation has been made on an infinitely long waveguide (periodical boundary condition) for waveguides with deep lateral corrugations. The width of the waveguide is 15\( \mu \text{m} \) at the wide section and 5\( \mu \text{m} \) at the narrow section. The parameters used for the materials in the simulation have been summed up in table 2.2 in chapter 2 (page 23).

Coupling coefficient values up to 100 \( \text{cm}^{-1} \) can be obtained with laterally corrugated ridge waveguides due to the big refractive index difference between the grating region and the non-corrugated region (figure 3.3). Smaller ridge cavity can then be designed in order to obtain a reasonable \( \kappa L \) value. This also can be done by increasing the width of the wide section of the ridge \( W_{\text{wide}} \) or by decreasing the width of the narrow section of the ridge \( W_{\text{narrow}} \). The highest coupling coefficient is obtained for a grating duty cycle of \( DC = 21.5\% \), \( DC = 57\% \) and \( DC = 86.5\% \). As the simulation is solving for the eigen-modes with a complex eigen-frequency \( \omega \), the out-coupling losses of the waveguides can be evaluated with \( \text{Im}(\omega) = \frac{\alpha c}{n} \) where \( \alpha \) are the optical losses. The losses of the symmetrical modes at the Γ-point of the photonic band structure are shown in figure 3.3.

The two band edge modes of the 1-D photonic structure will be present in the spectrum of the laser emission as they both exhibit low optical losses and can lase along the current-voltage range of the gain amplification structure. The mode lower than the band gap is called the dielectric mode for which its energy is concentrated in the high-\( \varepsilon \) regions (figure 3.5c). The mode above the band gap is called the air mode and has a larger fraction of its energy (not necessarily a majority) in the low-\( \varepsilon \) regions (figure 3.5d). A way to prevent one the band edge mode to lase is to ingeniously adapt the grating duty cycle and the depth of the lateral corrugations. As depicted in the figure 3.3b, it is possible to increase the lasing threshold of one mode by increasing the optical losses of
3.4. Engineering the Distributed Feedback Structure

Figure 3.3 – a) Schematic of the DFB structure with lateral gratings simulated. Simulations of (b) the damping of the two band edge modes of the photonic band structure and (c) the coupling coefficient $\kappa$ of the lowest transverse magnetic mode $TM_{00}$ calculated with the eingen-frequency formula. $\kappa$ is calculated for laterally corrugated waveguides with $W_{\text{narrow}} = 5\mu m$, $W_{\text{wide}} = 15\mu m$, $h = 14\mu m$, grating periodicity $\Lambda = 55 \mu m$ and the grating duty cycle $DC$ being varied.

Figure 3.4 – Eigen-frequency simulated of the symmetrical band edge modes of a laterally corrugated $3^{rd}$-order DFB laser varying with the grating duty cycle $DC$. The grating periodicity is $\Lambda = 55 \mu m$. 
the opposite mode in order to obtain the desired single mode operation. The losses of
the two band edge modes change with the grating duty cycle: the grating duty cycle fixes
the waveguide length of the narrow region and modifies the confinement of the air mode.
It is possible then to obtain more than 200 $cm^{-1}$ of optical losses for the lower band
mode while the upper band mode has optical losses lower than 10 $cm^{-1}$, which will lead
to an increased lasing threshold for the lower band mode compared to the upper band
mode one. Another aspect of the laser specification for heterodyne application implied
to obtain single mode operation at a desired frequency, of special interest in the THz
frequency range lies the rotational transition of the HD molecule at 2.7 THz (90 $cm^{-1}$).
The figure 3.4 shows the frequency of the band edge modes in function of the grating duty
cycle. The HD rotational transition is reachable with the configuration of a 3$^{rd}$-order DFB
cavity approximately for $DC = 4.5\%$ with the lower band edge mode and for $DC = 29\%$
with the upper-band edge mode. On the other hand, a grating DC of 4.5\% would lead
to slits width smaller than 3 $\mu m$ which would have been difficult to fabricate. We have
therefore modified the initial design of 3$^{rd}$-order DFB cavity so the lower band mode of
the photonic structure is reached with a higher DC. The figure 3.5b presents the band
structure of our second 3$^{rd}$-order DFB cavity design: the width of the waveguide is 20$\mu m$
at the wide section and 5 $\mu m$ at the narrow section. The frequency of interest is reached
for a grating DC of 12\%. Both of these designs have been investigated experimentally.
The figure 3.5 represents the simulated loss of an infinite cavity (periodical boundary
condition). At the Γ point of the band gap (edges band), the eigenmode of the upper
band edge branch (figure 3.5d) corresponds to modes where the energy density is more
localized in the air with consequently higher losses, while for the solutions at lower band
edge branch (figure 3.5c) the energy is more localized in the semiconductor region with
lower losses.

### 3.5 Emission

The beam profiles of a 3$^{rd}$ Order THz DFB QC laser have been oftenly described by the
antenna model and more precisely by the model of an end-fire antenna array. Within
this model, the grating slits are considered as a dipole array and the resulting far-field
is given by the calculation of the far-field emission of one dipole multiplied by an array
factor [68]. Amanti et al. in [69] have shown a good agreement between the calculated
beam of the 3$^{rd}$-order DFB using the antenna theory and the experimental measurements.
A 2$^{nd}$ method for the calculation of the beam pattern of such devices is to calculate the
electromagnetic field distribution using a finite element method (FEM) solver and then
to obtain the farfield beam pattern using a Fourier transformation of the nearfield. This
Figure 3.5 – Simulated band structure of infinite $3^{rd}$-order DFB cavities using a finite element method where in a) the width of the waveguide is $15\mu m$ at the wide section and $5\mu m$ at the narrow section and in b) respectively $20\mu m$ and $5\mu m$ for the wide/narrow section. In c) is presented the $E_z$ profile of both higher and lower modes $\omega_+$ and $\omega_-$ for the second design. The mode profiles of the band edge modes for the first design is similarly distributed in the waveguide.
method has the disadvantage to require the calculation of the electric and magnetic field distribution inside the laser cavity, implying heavy calculations. Simulations of beam-pattern in far-field regime of 3\textsuperscript{rd}-order DFB QCL using the two different methods are presented: the first one using the antennas theory and the second using the Fourier transform. These simulations are extremely useful as they allow fast designing of devices which require specific characteristics such as narrow beam emissions.

3.5.1 Antenna model

As it is well described in the Antenna Theory \[70\], the total field of an array is determined by the vector addition of the fields radiated by the individual elements. To obtain a narrow emission beam, it is mandatory that the fields from all elements of the array interfere constructively in the desired directions and interfere destructively in the remaining space. In figure 3.7a is represented the equivalent dipole model of the 3\textsuperscript{rd}-order DFB cavity: the radiation coming from each corrugation of the DFB is represented by a dipole of a length equal to the waveguide width (figure 3.6) according to \[71\]. The array of dipoles is represented over an infinite metallic ground, which is taken into account in the calculation of the far-field. All the elements are considered to have similar amplitudes of the electrical field and an emission of an aperture antenna as presented on figure 3.6. The dipoles are all linearly spaced so the phase shift between each neighboring dipoles is uniform along the array and is always 3\pi. Using the coordinate defined in the figure 3.7a, the emission of one emitter is, according to \[70\]

\[
E_0 = j \frac{w^2 k E_0 e^{-jkr}}{2\pi r} e^{\left\{ \frac{\sin(\phi)}{X} \right\}^2} \tag{3.30}
\]

with

\[
X = \frac{kw}{2} \sin(\theta)\cos(\phi) \tag{3.31}
\]

where \(w\) is the width of the dipole, \(k\) the wavevector, \(E_0\) a constant and \(r\) the distance from the origin to the observation point. To obtain the overall radiation of the antenna array, this term has to be multiplied by the array factor of the system:

\[
(AF)_N = \frac{\sin\left(\frac{N\varphi}{2}\right)}{N\sin\left(\frac{\varphi}{2}\right)} \tag{3.32}
\]

where \(\varphi\) the geometric phase shift between each corrugation:

\[
\varphi = k\Lambda\sin(\theta)\cos(\phi) \tag{3.33}
\]
where $\Lambda$ is the grating periodicity and $k$ the wavevector. An effective index for the dipole array is defined as $n_{eff} = \frac{3\lambda_0}{2\Lambda}$ with $\lambda_0$ the free space wavelength. The $n_{eff}$ is used as figure of merit to evaluate the phase matching of the 3$^{rd}$ order DFB.

We multiply (3.30) and (3.32) to obtain the radiated field of the array.

$$F = \frac{jw^2kE_0e^{-jkr}}{2\pi r} \sin(\phi) \left| \frac{\sin(X)}{X} \right|^2 \left| \frac{\sin \left( \frac{N\varphi}{2} \right)}{N \sin \left( \frac{\varphi}{2} \right)} \right|^2$$

(3.34)

To account for the metallic ground beneath and all around the waveguide, we add a supplementary array, symmetrical to the first one with respect to the $z$ plane at $z = 0$ with a height $h$ corresponding to the height of the waveguide. The metallic plane acts as a reflector for the waves above the plane. This effect can be modeled by using the method of images ([72]): a supplementary coefficient $1 + e^{\frac{2\pi R \sin(\theta)}{\lambda}}$ is added. Figure 3.7 presents the simulated far-field emission of dipole antenna array for $n_{eff}$ varying from 3.0 to 3.4. The beam profile is calculated in the $\theta/\Phi$ coordinate as represented in figure 3.7. When $n_{eff} = 3$, the arrays give relatively concentrated beams with very weak side-lobes as the electric field of each adjacent dipoles have a $3\pi$ phase shift and add up constructively. This situation is then refereed as the perfect phase matching [73]. However, when $n_{eff} \neq 3$ the arrays results in a multi-lobed beam, as a phase "error" will be accumulated all along the array. In these cases, the main lobe contains a fraction of the total energy within the

Figure 3.6 – Schematic of a rectangular antenna aperture on an infinite electric ground plane with the defined referential for the far-field calculation.
plotted beam area and the remaining energy is diffracted.

The simulations show that the control of the effective index of the waveguide is a crucial parameter in realizing a single-lobed narrow beam. The number of periods in the array also plays an important role into the beam emission: for longer devices, the emissions constructively add up and create narrower beam patterns when the perfectly phase-matching condition is obtained (Figure 3.8). On the contrary, it is evident that destructive superposition happens under the mismatched condition and the total output power will decrease as the number of periods is increased as seen on figure 3.9.

3.5.2 Finite Element Method Simulations

Finite Element Method (FEM) simulations have also been conducted in order to further analyse the emissions pattern of the waveguide structure investigated using a Fourier transform calculation. The characteristics of the waveguide investigated with the FEM simulations have been described in the section 3.4. Defining a surface S which extends over the surface of the 3\textsuperscript{rd}-order DFB waveguide cavity and the ground plane, the equivalent electric and magnetic currents \( J_S \) and \( M_S \) flowing tangentially on the surface S are then considered as source of secondary waves which produce distant wavefronts, according to Huygens principle. Knowing the electric and magnetic \( E \) and \( H \) fields tangential to the aperture, we can replace the original fields by equivalent electric and magnetic surface currents \( J \) and \( M \):

\[
J_S = n \wedge H = -H_z x + H_z y \\
M_S = n \wedge E = E_z x - E_z y
\]

where \( n \) is a normal to the surface S. The far-field is therefore arising from the equivalent current at the surface S. Following the treatment of Balanis [70], the magnetic potential vector induced by the electric current \( J_S \) and \( F \) the electric potential vector induced by the magnetic current \( M_S \) can be expressed by:

\[
A = \frac{\mu}{4\pi} \int_S J_S e^{-jkr} R dS \approx \frac{\mu e^{-jkr}}{4\pi r} N
\]

\[
F = \frac{\epsilon}{4\pi} \int_S M_S e^{-jkr} R dS \approx \frac{\epsilon e^{-jkr}}{4\pi r} L
\]

with \( N = \int_S J_S e^{ikr} dS \) and \( L = \int_S M_S e^{ikr} dS \) which are the Fourier transform of the current densities. \( R \) is the distance between the observations point M and a point on the surface S (as on figure 3.7a). The E field components in the far-field can then be written
3.5. Emission

(a) Aperture antenna array schematic

(b) $n_{eff} = 3.0$

(c) $n_{eff} = 3.1$

(d) $n_{eff} = 3.2$

(e) $n_{eff} = 3.4$

Figure 3.7 – (a) Schematic of the aperture antenna array and simulated beam patterns of different third-order DFB lasers using the dipole antenna array modeling. (b) in perfect phase-matching condition $n_{eff} = 3$, (c) $n_{eff} = 3.2$, (d) $n_{eff} = 3.4$ and (e) $n_{eff} = 3.6$. The grating period is $\Lambda = 55 \, \mu m$. 
Figure 3.8 – Simulated beam patterns of different third-order DFB lasers using the dipole antenna array modeling in perfect phase-matching condition \((n_{\text{eff}} = 3)\). (a) 15 periods, (b) 50 periods, (c) 150 periods

as:

\[
\begin{align*}
E_r &= 0 \\
E_\theta &\simeq \frac{-jke^{-jkr}}{4\pi r} \left( N_\theta + \sqrt{\frac{\mu_0}{\varepsilon_0}} N_\phi \right) \\
E_\phi &\simeq \frac{jke^{-jkr}}{4\pi r} \left( -\sqrt{\frac{\mu_0}{\varepsilon_0}} N_\phi \right)
\end{align*}
\tag{3.38}
\]

The calculated beam emissions for the 3rd-order DFB QCL are presented on figure 3.10. For similar parameters \((\Lambda, \text{width of the ridge})\), the antenna theory and the Fourier transform theory are in good agreement in terms of beam divergence. In a similar way to the last section on the antenna model, we use \(n_{\text{eff}} = \frac{3\lambda_0}{2\Lambda}\) as figure of merit were the geometric parameter \(\Lambda\) is tuned with regard of \(\lambda_0\). For 3rd-order DFB cavity where the perfect phase matching is reached \((n_{\text{eff}} = 3)\) the resulting beam divergence is fairly concise with a FWHM (Full Width Half Maximum) of 32° × 16° with a grazing angle emission at 16° from the direction of the waveguide (figure 3.10b). It can be further analysed that when departing from the optimised value for the grating periodicity \(\Lambda\) and duty cycle DC for a specified \(\lambda_0\), the beam pattern is highly diverging and presents a multi-lobe emission. In a similar way to the antenna theory, when the effective index of

Figure 3.9 – Simulated beam patterns of different third-order DFB lasers using the dipole antenna array modeling with \(n_{\text{eff}} \neq 3\). (a) 15 periods, (b) 50 periods, (c) 150 periods
Figure 3.10 – Simulated beam emission pattern using FEM simulations and Fourier transform model at 2.7 THz. The simulated waveguide consists of 15 periods with 15 $\mu m$ width for the wide section and 5 $\mu m$ for the narrow section. a) Schematic of the simulated 3$^{rd}$-order DFB with deep lateral corrugations and showing the coordinates system used for the farfield calculation. The grating duty cycle was fixed to be 29% and the grating periodicity was for (b) $\Lambda = 55 \mu m$, (c) $\Lambda = 55.5 \mu m$ and (d) $\Lambda = 56 \mu m$. 
the propagating mode is different from phase matching condition \( n_{\text{eff}} \neq 3 \), diffraction patterns arise in the beam patterns and the emission angle is tilted toward the normal of the waveguide (figures 3.10c and 3.10d). The FWHM is then > 40° × 30°. Upon these later simulations, the next chapter is dealing on the experimental results for the 3rd-order DFB QCLs and the implemented design explored in the last sections.
Chapter 4

Third Order distributed feedback: results

This chapter presents the results on the 3rd-order DFB obtained with the designs that have been investigated in chapter 3. The interests with the 3rd-order DFB waveguides is that they can not only achieve single mode operation at a specified target frequency but also have a low DC power consumption compared to what is usually known for the metal-metal QCL waveguide and a low divergence beam emission. The last property of this design is crucial for quasi-optical coupling with other opto-electronic devices and especially detectors as for instance in heterodyne receiver. All the devices which will be presented in this chapter have been built in the clean room facility of C2N and a new etching process has been required to obtain smooth side walls and highly defined corrugations on waveguides.

4.1 Results on similar devices in the literature

The use of photonic structures with THz QCL has drawn a lot of interest in the past few years to solve the problem of beam shaping. Laser applications typically require also stable and single-mode emission at a precisely defined frequency. DFB structures and photonic crystals have been reported with features such as single mode emission, narrow and low divergence beam emission, that make these QCLs usable in various applications such as spectroscopy and imaging. DFB gratings are implemented by slits in the top metallization of the laser ridge as in [74] and [75]: the top metallisation slits provide a strong feedback due to modal refractive-index modulation. 2nd-order DFB THz QCL have been demonstrated with surface emission and single mode operation and the grated photonic structures yield surface emission with a record high peak power superior to 100mW and with single-lobed emission patterns [33]. Lateral corrugations have been more recently used on 1st-order and 3rd-order DFB waveguides [76] [77]. Each time, the teams report single mode emission and narrow beam emission. The particular work of
Chapter 4. Third Order distributed feedback results

Amanti et al. in [34] has made a breach through the usual photonic structures with a DFB design of a MM waveguide QCL that features a grating resonant with the third-order Bragg condition: the improved extraction efficiency into a narrow beam with a divergence lower than $10^\circ \times 10^\circ$, the control of the emission wavelength and an enhanced output power compared to a similar MM Fabry-Pérot waveguide.

The fabrication of the THz QCLs with lateral DFB grating is similar to that of MM waveguide THz QCLs except for the contact pad for the electrical contacting as it is very difficult to wire-bond on a $15\mu m$ wide waveguide. In the first step, an insulating layer of $SiO_x$ prior to the top-metallisation is deposited to avoid electrical pumping of the active region beneath the contact pad. The top Ti/Au metallisation is then patterned on the active region with Plassys processes and lift-off. The top metallisation process is followed by the dry etching of the ridges with a ICP-DRIE (Inductive Coupled Plasma-Deep Reactive Ion Etching with a STS Surface Technology System) tool. An investigation of the parameters for an anisotropic and directive etching has been carried out. This work is more detailed in [78]. The detailed fabrication steps of the devices are reported in appendix A. Once the ridges are defined, the GaAs substrate beneath the heterostructure is polished to reduce its thickness. After processing, laser chips are soldered with indium to copper mount and contacted by wire bonding.

4.2 Experimental measurements

Prior to the DFB fabrication, standard single plasmon QCLs with $150 \mu m$ width and $4 mm$ long were fabricated on the $L1395$ structure by wet etching and Ti/Au top metallisation. Figure 4.2 shows the light-current-voltage (L-I-V) characteristics of the SP QCL in pulsed operation for different temperature. Lasing was observed up to 90 K in pulsed operation. The threshold current density is $210 A.cm^{-2}$ at 10K, increasing to $400 A.cm^{-2}$ at 90K, with a total electrical driving power in the range of 5W-25W. The Fabry-Pérot SP QCL although exhibits multi-modal spectrum as seen on figure 4.4c.

As presented in the previous chapter 3, two different designs have been implemented for the realization of 3rd-order DFB waveguides. In this section the results for the structures with the deeply etched lateral corrugations are presented. The SEM pictures of the devices are presented on figure 4.1a and figure 4.1b. A specially tailored lithographic mask has been realised where the grating duty cycle and the grating periodicity have been varied in order to ensure that the target frequency is reached in one running batch. The waveguides include 15 periods with a grating periodicity of $\Lambda = 55 \mu m$, so the overall length of the waveguides are $\sim 800 \mu m$. The wide part of the waveguide is kept at $15 \mu m$ for the first design and 20 $\mu m$ for the second one, while in the narrow section, for the
(a) Array of 3rd-order DFB waveguides   (b) Detail of a 3rd-order DFB waveguides

(c) Detail of the contact pad connected to the ridge through the thin bridge.

Figure 4.1 – SEM pictures of second generations devices: the contact pad is optically decoupled from the waveguide to avoid any perturbations for the light extraction
Figure 4.2 – L-I-V characteristic in pulsed operation of a 150 \( \mu m \times 4 \ mm \) Fabry-Pérot SP QCL.

corrugation on both designs, the width of the waveguide is 5\( \mu m \). The ridge section of the second design 20 \( \mu m \) has been chosen to be wider than the first design as the band edge mode at 2.7 THz is under the photonic gap and require grating duty cycle rather low (< 6%) and small lateral corrugation difficult to process. For the first generation of devices, the last periods at both ends of the waveguide were connected to bonding pads for electrical contacting. For the second generation of devices, the contact-solder pad is spatially separated from the waveguide. The electrical connection is provided by a thin bridge 4.1c, which minimizes the optical coupling between the cavity formed by the contact pad and the 3\(^{rd}\)-oder DFB waveguide.

The typical light-current-voltage curves for the first design are presented in figure 4.4a and have been performed both in pulsed and continuous wave operation. The lasing threshold current density is similar to the one measured for the Fabry-Pérot SP waveguide in figure 4.2 (\( \simeq 200 \, A/cm^2 \)). The operating currents are in the range from 20 to 30 mA while the DC power dissipation stays below 200mW. These values are much lower than for standard devices owing to the reduced surface area of the device. The low voltage threshold of the active-region design (chirped supper-lattice with phonon-resonant depopulation scheme) associated with the low driving current densities allowed good performances in continuous wave operation thanks to the low DC power dissipation of the lasers. At the
4.2. Experimental measurements

lowest temperature, a maximum output power of 850 µW measured in CW with a absolute power meter Thomas Keating (the QCL output power hasn’t been corrected for the collecting efficiency or the absorption of the dewar window). Lasing was observed up to 77K but the output power decreases down to 400 µW at liquid nitrogen temperature. Due to misalignments between the top metallisation and the dry etching of the lateral corrugations, it is difficult to ensure that the predicted design by simulations has similar experimental behaviour once processed. A specially customized lithographic mask has been designed where the grating periodicity and grating duty cycle are slightly changed between each laser so that the desired target frequency is obtained on at least one device. The tunability of the grating duty-cycle (not presented) of 0.2% between each laser for a fixed frequency showed only small changes in the lasers spectra and L-I-V. In figure 4.3a and 4.3b, typical lasing spectra measured in pulsed operation at 77K of both design are shown for different grating periods. The grating periodicity has been varied from $\Lambda = 54 \mu m$ to 56 $\mu m$ with $\Delta \Lambda = 0.2 \mu m$. For the second design, a finer tuning of 0.1 $\mu m$ has been chosen in order to have a better control on the frequency emission. The objective here wasn’t to study the tunability of the 3$^{rd}$-order DFB structure as in [79], rather was to calibrate the grating periodicity $\Lambda$ for a specific duty cycle (12% and 28% respectively for the first and second design) and therefore obtaining the perfect phase matching condition in one running batch. The spectra contain only one or two lasing modes, typical signature of DFB structures, which can be identified as the $\omega_+$ or $\omega_-$ of the band edge modes of the photonic band structure presented in figure 3.5. The corresponding band gap width is about respectively for the first design 10 $cm^{-1}$ and 9.1 $cm^{-1}$ for the second design, which is in a good agreement with the simulated photonic band structure of the fundamental $TM_{00}$ mode (cf figures 3.5a and 3.5b page 39). Single-mode operation is observed for both designs for a respective first and second design grating periodicity of 55 $\mu m$ and 54.6 $\mu m$. Generally, these lasers operate in single mode operation just after threshold, while a second mode appears for higher driving current, which can be explained by the higher lasing threshold of the opposite band edge modes as it has been explained in the chapter 3. For grating periods outside the range of 54.6$\mu m$-55.0$\mu m$, the spectra also contain an additional mode, which can be explained by a better overlap of the band edge modes with the gain range. For the first design of DFB, the spectra presented on 4.3a show lasers with modes inside the photonic gap, or laser modes with small frequency spacing, which can be explained by the optical coupling of the DFB waveguide with the contact pad: the longitudinal propagating fundamental $TM_{00}$ mode of the DFB could couple to the cavity beneath the contact pad and gives rise to other resonances.
Figure 4.3 – Spectra of 3\textsuperscript{rd}-order DFB from (a) first design and (b) second design. The grating periodicity is changed between each lasers while the grating DC is kept at 29\% for (a) and 12\% for (b). The lasers were driven in pulsed operation at 77K. The insets shows the mode wavelength in function of the grating periodicity.
4.3 Single mode emission at the target frequency

Of particular interest for the heterodyne detector are the lasers that exhibit single mode emission at the targeted frequency, or with high side modes suppression ratio. The corresponding emission spectrum of such device for the first design of 3rd-order DFB is presented in figure 4.4b in continuous operation at 4K and 4.4c in pulsed operation at 77K. The latter spectrum is compared with the emission spectrum in same operating condition of a SP waveguide. While the SP QCL exhibits a typical multi-modal Fabry-Pérot spectrum in the range of 2.5THz and 2.75THz at 77K in pulsed operation, the single mode operation at the target frequency has been obtained for a device with a grating periodicity of 55µm and a grating duty cycle \( DC = 28.8\% \) (figures 4.4b and 4.4c). As for the second design, the target frequency is reached with a grating periodicity of 54.8µm and a duty cycle \( DC = 12\% \) but the spectrum presents a second lasing mode as it is shown in figure 4.5d.

4.4 Far-field emission of 3rd-order DFB lasers

Figures 4.5c and 4.5e show the measured emission spectra and the corresponding beam profiles for both the 3rd-order DFB design. Both of the presented lasers show the main lasing mode at the target frequency with an effective refractive index respectively for the laser in figure 4.5b \( n_{\text{eff}} = 3.02 \) and \( n_{\text{eff}} = 3.05 \) for the laser in figure 4.5d. The condition \( n_{\text{eff}} = 3 \) is then called the perfect phase matching as the EM waves from each corrugation 'adding up' at both end of the waveguide, in a similar way to an end-fire antenna array as depicted in chapter 3. In the first design configuration, the laser exhibits a single lobe, rather narrow oblong emission diagram with a full width at half maximum (FWHM) of \( 30^\circ \times 20^\circ \). The second one presents two lasing modes but the main one is lasing at the targeted frequency. The second mode is nevertheless 10 dB less powerful than the main lasing mode and the farfield emission measured in figure 4.5e is mostly attributed to the \( \omega_+ \) mode. The measured far-field exhibits a circular Gaussian like emission with a FWHM divergence of \( 10^\circ \times 10^\circ \). To evaluate the farfield quality of the 3rd-order DFB, the directivity coefficient \( D \), from antenna theory, can be used as described in [80] and [81].

The directivity is defined as \( D = 10\log_{10}(4\pi I_{\text{max}}/I_{\text{tot}}) \) where \( I_{\text{max}} \) is the maximum intensity measured in farfield bring back to the elementary solid angle of the measurements (W/sr) and \( I_{\text{tot}} \) the total intensity (W). The first design measured directivity is 9 dBi while for the second design the directivity is 20 dBi.

Fairly good laser performances have been shown for third order devices with single lobe beam emission at the target frequency. The optimal far-field beam profile is obtained
Figure 4.4 – Typical electro-optical characterisation of the 3\textsuperscript{rd} order DFB: the ratio of the wide-narrow section of the waveguide is 20\textmu m-5\textmu m and the waveguide is composed of 15 periods with $\Lambda = 55\mu m$. (a) Light-current-voltage characterisation of the 3\textsuperscript{rd} order DFB. The threshold current is as low as 25mA and the overall DC power dissipation is below 200mW for the whole current-voltage range. (b) Spectrum of the same device as above which shows a single mode operation at 90cm\textsuperscript{-1} at 4K in pulsed operation (c) Emission spectra of 3\textsuperscript{rd}-order DFB QCLs of the first design at 77K in continuous wave operation compared with emission from a similar laser heterostructure L1395 in single-plasmon waveguide (4 mm x 150 \mu m).
4.4. Far-field emission of 3rd-order DFB lasers

(a) schematic of the angle notation for the measurements

(b) Measured spectrum for a 3rd-order DFB QCL with the first design with $\Lambda = 54.6\mu m$ and DC = 28.8% with a single mode emission on the lower-band edge mode $\omega_-=90.73$ cm$^{-1} \pm 0.25$ cm$^{-1}$

c) Measured far field corresponding to (a).

d) Measured spectrum for 3rd-order DFB QCL with the second design with $\Lambda= 54.6$ $\mu m$ and DC = 12% with two lasing modes corresponding to the upper (main mode) $\omega_+ = 89.99$ cm$^{-1}$ and lower band edge mode and

e) Measured far field corresponding to (d).
Figure 4.6 – Spectra and corresponding beam emission for devices from the first design with different value of $n_{eff}$. Figures a) and b) correspond to a $n_{eff} = 2.96$, figures c) and d) $n_{eff} = 2.94$ and figures e) et f) $n_{eff} = 2.92$
when the effective refractive index of the lasing mode is exactly or the nearest possible \( n_{\text{eff}} = 3 \). If one departs from this phase matching condition, for which there is perfect constructive interference in the far-field, the output beam will present diffraction patterns and high divergence angles. In figure 4.6 and 4.7 and is presented 3\(^{rd}\)-order DFB QCL for which the so called phase matching conditions is not reached. The influence of the effective refractive index \( n_{\text{eff}} \) on the beam emission, as shown in the later chapter with the antenna theory, has been experimentally verified.

4.5 3\(^{rd}\)-order DFB on L1390

The same 3\(^{rd}\)-order waveguide process has been applied on the heterostructure L1390 which is a Bound To Continuum heterostructure that presents interesting features for the heterodyne application with a low current density threshold at low temperature thanks to the phonon assisted depopulation mechanism of the lower level of the photonic transition. The first design of 3\(^{rd}\)-order has been implemented. The L1390 structure has been first tested with a Fabry-Pérot waveguide 80\( \mu \text{m} \) wide and 1\( \text{mm} \) long in pulsed operation at 4K. The lasing threshold is as low as 155\( \text{A.cm}^{-2} \) and shows a characteristic spectrum with a dominant mode around 86 \( \text{cm}^{-1} \) and lower energy modes with a spacing of 1.6 \( \text{cm}^{-1} \) that corresponds to the resonator length of 1 \( \text{mm} \). The 3\(^{rd}\)-order DFB waveguide on the same structure shows single mode operation at the desired target frequency (90 \( \text{cm}^{-1} \)). The electrical characteristic shows a low driving current at lasing threshold but the threshold current density, which is the intensity of the current driving the laser divided by the surface of the top-metallisation is much higher than for the MM waveguide with same heterostructure (figure 4.8b). The DFB structure seems to deteriorate the normal laser operation. It is probably caused by the mode frequency selection which is not at the maximum gain frequency of the heterostructure. The output power of the DFB on L1390, which has not been measured with a absolute power meter, nevertheless appears to be much less than for the standard DFB fabricated on L1395. The measured far-field (figure 4.9) presents similar characteristics to the one obtained on figure 4.5c with an oblong single lobe and an emission angle at 20 degree above the laser ridge. This experience shows that to obtain the desired target frequency with a 3\(^{rd}\)-order DFB structure and an optimal lasing operation, the selection of the laser heterostructure is of main importance and the maximum gain frequency band must include the target frequency.
Figure 4.7 – Spectra and corresponding beam emission for devices from the second design with different value of $n_{eff}$. Figures a) and b) correspond to a $n_{eff} = 3.05$ and figures c) and d) $n_{eff} = 3.09$.
Figure 4.8 – (a) Measured spectrum for a Metal-Metal Fabry-Pérot waveguide of the L1390 heterostructure and (b) corresponding Light-Current-Voltage characteristic in pulsed mode at 4K for the same laser. (c) Measured spectrum for the 3rd-order DFB waveguide with $\Lambda = 55 \mu m$ and $DC = 29\%$, the ridge section ratio between the narrow and wide part of the waveguide is 15/5$\mu m$ and (d) corresponding L-I-V.
Chapter 4. Third Order distributed feedback results

Figure 4.9 – Measured farfield emission pattern of the 3\textsuperscript{rd} order DFB on L1390. The laser was operated in pulsed mode at 4 K.

4.6 Broad Band Extractor

The previous sections on 3\textsuperscript{rd}-order DFB QCL have shown that a single mode operation at a targeted frequency with a narrow beam emission are both controlled by the grating duty cycle and the grating periodicity. In the perspective to have different LO frequency lines while keeping a concise beam emission for each frequency, we propose to disentangle the frequency selection mechanism from the light extraction mechanism. The QCL waveguide is here divided in two sections: the first section is composed of 50 periods of 1\textsuperscript{st}-order DFB with deep etched lateral corrugations for the frequency selection and the second section is composed of 6 periods of 3\textsuperscript{rd}-order DFB with a chirped grating periodicity for the light extraction (figures 4.11). The strong frequency selection and tunability of the 1\textsuperscript{st}-order DFB THz QCL with lateral corrugation have been already widely studied in [76] and [41]. Similar frequency specifications to the later 3\textsuperscript{rd}-order DFB (cf section 4.3) have been used for the design of the 1\textsuperscript{st}-order DFB waveguide section (\textit{i.e.} single mode at 90 cm\textsuperscript{-1}-2.7 THz). The ridge first section were 50 \( \mu \text{m} \) wide at the widest sections, with 10 \( \mu \text{m} \) lateral corrugations in both sides. The corrugation duty cycle was 50\% and the 1\textsuperscript{st}-order DFB section was roughly 880 \( \mu \text{m} \) long. The light extraction section has been inspired by [82] which was based on a 3\textsuperscript{rd}-order DFB with a chirped grating periodicity and is only used to shape the far-field of the extracted light. It consists of 5 periods of 3\textsuperscript{rd}-order Bragg gratings with an increased grating period by 2 \( \mu \text{m} \) every period enabling light extraction from 2.5 THz to 2.8 THz. A design consisting of deep etched lateral gratings for the light extraction has been investigated with squared teeth (figures 4.11a and 4.11c). The widest section of the 3\textsuperscript{rd}-order grating section was 80 \( \mu \text{m} \) while the narrowest section was

62
4.6. Broad Band Extractor

5 µm. The grating duty cycle was 45% which relaxes the Bragg condition. The extractor section is then coupled independently to different 1st-order DFB waveguides. We have performed Finite Element simulations on FEKO to estimate the beam divergence of the design for various frequencies between 2.5 THz and 2.8 THz. The calculation of the extractor farfield at different frequencies are presented in figures 4.10b, 4.10c, 4.10d and 4.10e. For the frequency range of 2.5 THz to 2.8 THz, the extractor farfield presents a central lobe with a FWHM 35° × 25° but presents pronounced side lobes at 2.5 THz and 2.8 THz. The proposed extractor design was first investigated coupled with a standard Fabry-Pérot waveguide. Figure 4.12a presents the spectra measured in pulsed operation at 10 K at different voltage bias. Far after threshold, the spectrum shows discrete comb like features with a frequency range from 75 cm⁻¹ to 91.6 cm⁻¹, which is broader than the cleaved facets Fabry-Pérot waveguide’s spectrum. The corresponding far-field however represented in figure 4.12b shows a wide beam pattern with a peak power at emission angle θ = 25° and −10° < φ < 12°. The different frequencies composing the spectrum at higher bias seem to radiate at different angles which broaden the beam emission of the QCL, similar to a non-phase matched 3rd-order DFB.

The extractor design has been on a second time investigated when coupled to a 1st-order DFB for which the grating periodicity, i.e. the mode frequency selection, has been tuned from 2.5 THz to 3 THz. Quasi-CW operation (DC ~ 98%) spectra taken at high injection current at a heat-sink temperature of 77 K from two devices with the 1st-order DFB section with grating periodicity Λ = 15.2 µm and 16 µm are shown in figures 4.13a and 4.13c. These devices lased predominantly at 2.49 THz and 2.72 THz which scales well with the grating period. As the bias was increased far beyond threshold, most devices lased on additional modes due to spatial hole burning but with an amplitude reduced by 10 dB compared to the main mode. A longer device would have had a more robust single-mode operation. The far-field patterns were recorded for these devices. The lasers were driven close to their maximal operation point in quasi-CW operation. The QMC pyroelectronic detector (cf appendix A) was scanned around the laser facets with the angle reference sketch in figure 4.10a. The measured far-field are presented in figures 4.13b and 4.13d. Both lasers have a central lobe with a FWHM of approximately 30° × 25°. A secondary lobe is observed on the figure 4.13d which could be attributed to the secondary lasing mode. The measured farfield are in good agreement with the simulated one on FEKO software using Fourier transform for farfield calculation. With a single broad extractor design, light could be extracted in fairly narrow beam from a 1st-order DFB waveguide. The beam diagram is much narrower than for standard Fabry-Pérot waveguide. This could be an asset for the fast designing of LO sources at different frequencies while keeping a narrow beam diagram as only the grating periodicity of the 1st-order DFB section needs
Figure 4.10 – Calculated far-field pattern of the extractor for different frequencies.
Figure 4.11 – SEM pictures of the fabricated devices showing on a) an array of extractor devices, b) and c) showing details on the waveguide section with the 1st-order DFB with deep-etched lateral corrugations and the second section for the light extraction mechanism with 3rd-order DFB lateral corrugations.
Figure 4.12 – a) Measured spectra and b) farfield beam emission of Fabry-Pérot waveguide coupled to broad extractor with squared teeth. The measurements were recorded at 10 K in pulsed operation. The spectra were measured just after lasing threshold (black curve) and at maximum output power (red curve) and the far field was measured at the maximum output power.

to be tuned.
Figure 4.13 – a) and c) Measured spectra for broad extractors on the L1395 heterostructure with different 1st-order DFB grating periodicity Λ. b) and d) corresponding beam diagram in pulsed operation at 77 K for the same lasers.
Chapter 4. Third Order distributed feedback results
Chapter 5

Dielectric hollow core waveguides

Hollow core waveguides (HW) are composed of a highly reflective wall with an inner central hole. We will study especially in this section the HW with a circular cross section with inner walls made of Pyrex and the core filled with vacuum. These waveguides can transmit a broad range of wavelengths with low losses. They are also of particular interest for the transmission of THz light where it is difficult to find materials that have low optical losses required for the propagation in a solid core fiber as used in the telecommunication frequency range. We have also explored the use of this type of waveguide as a way to perform the beam shaping of our 3rd order distributed feedback quantum cascade lasers. A transverse cross-section of the hollow core waveguide is shown in figure 5.2. It has a large air-core with a radius $a$ and a ring cladding with thickness $t$. The medium outside the cladding is also air. The material of the cladding, Pyrex glass for our application, is assumed to be dielectric with a non-dispersive refractive index $n = 1.75$. The first theoretical work on hollow core waveguide (HW) has been made by Marcatili and Schmeltzer [83]. They solved the Maxwell-Gauss equation for electromagnetic propagation in an air filled core cylindrical waveguide. Under the assumption that the bore diameter is much higher than the wavelength, the propagation constant $\beta$ is equal to the wavevector $k$. The optical loss of the transverse electrical modes, transverse magnetic mode and hybrid modes depend on the complex refractive index $\nu_n$ of the waveguide wall. The attenuation can be written as:

$$\alpha_{nm} = \left(\frac{u_{nm}}{2\pi}\right)^2 \frac{\lambda^2}{a^3} \begin{cases} 
\frac{1}{\sqrt{\nu^2 - 1}}, & \text{for } TE_{0m} \text{ modes (n=0)} \\
\frac{\nu^2}{\sqrt{\nu^2 - 1}}, & \text{for } TM_{0m} \text{ modes (n=0)} \\
\frac{1}{2 \sqrt{\nu^2 + 1}}, & \text{for } EH_{nm} \text{ modes (n=0)}
\end{cases} \quad (5.1)$$

Where $\lambda$ is the wavelength, $2a$ the bore diameter of the hollow core waveguide, and
Chapter 5. Dielectric hollow core waveguides

\[ u_{nm} \] the \( m^{th} \) roots of the \( J_{n-1}(x) \) Bessel function:

\[ J_{nm}(u_{nm}) = 0 \quad (5.2) \]

The important consequences of (5.1) is that the attenuation constants are proportional to \( \lambda^2/\alpha^3 \) [84]: the losses can be made small by choosing the radius of the hollow core sufficiently large compared to the wavelength \( \lambda \). The other implication is that the lowest loss is obtained for the lowest order modes (for which the value of \( u_{nm} \) are small): for a fixed value of \( n \), the attenuation increased with \( m \). The refractive index is also affecting the 3 types of modes. For the \( TE_{0m} \) modes, the higher the refractive index is, the lower is the attenuation. This is especially true for hollow core metallic waveguides. For \( TM_{0m} \), if the refractive index of the tube material is close to unity, the attenuation is high while the minimum of attenuation occurs for \( \nu = \sqrt{2} \). For the hybrid modes \( HE_{nm} \), the attenuation \( \alpha_{nm} \) has mixed dependance on \( \nu \) between the \( TE_{0m} \) and \( TM_{0m} \). The value that minimizes the attenuation for hybrid modes is \( \nu = 1.73 \). Modal intensity distributions of the first lowest modes are shown in figure 5.1. The fields \( E_r \) and \( E_\phi \) are both complex. The fundamental mode of the pipe waveguide is labeled as the \( HE_{11} \) mode for which its electric field distribution of the fundamental hybrid mode is very similar to that of the fundamental free space \( TEM_{00} \) mode. Modal intensity distributions of the \( TM_{01} \), \( TE_{01} \) and \( HE_{21} \) modes are identical. Nevertheless, their imaginary parts are much different which lead to different attenuation constants. The attenuation constant of mode \( TE_{01} \) is smaller than that for \( TM_{01} \) while the attenuation constant of mode \( HE_{21} \) is in-between \( TE_{01} \) and \( TM_{01} \) due to its mixed nature of TE and TM modes.

5.1 Ray optics model for Hollow Waveguide

Ray optical analysis as proposed by Miyagi in [85] [86] of the modal loss in HW is far less complex than the electromagnetic approach cited above but yet provides with the modal loss characteristics of any HW. This simple technique allows to solution the propagation losses of core modes in the pipe waveguides and for more complicated designs of hollow core such as the hollow bragg fiber. This approach is used here as it can clarify the understanding of the propagation in the hollow core fibers and later the coupling efficiency of the QCL with the HW. The approach of Miyagi et. al. is to characterise the incidence angles of all modes in the HW corresponding to their propagation constants. This method necessitates to calculate the reflectivity of the waveguide’s cladding for each mode’s polarisation and grazing angle. The attenuation loss of each mode can then be determined given the number of reflections of the wavelets propagating inside the HW.
Figure 5.1 – Modal intensity distributions of the first transverse electric and hybrid modes of a hollow core dielectric waveguide with a bore diameter of 4 mm and calculated for a wavelength of $\lambda = 110\mu m$
Chapter 5. Dielectric hollow core waveguides

Figure 5.2 - Cross-section of the hollow core waveguide, where the core is filled with air and the claddings is glass

and then how much intensity is lost at each interaction with the waveguide’s side. This technique has been proven to be accurate compared to the equations that were derived with electromagnetic wave theory [87]. The ray optics theory can only solve for modal losses while electromagnetic wave theory can also provide the dispersion of the modes. [88] used a FDFD mode solver (finite-difference frequency-domain) for the evaluation of the attenuation constant of the first propagating modes and non-propagating modes and their simulations are consistent with the calculated field profile with the wave optics theory and the ray optical model.

Considering the schematic on figure 5.2, the ray are coupled to the HW of bore radius a with an angle $\theta_z$ within the normal of the reflective wall. The distance between two consecutive bounces along the waveguide side is then:

$$L = 2.a.tan(\theta_N)$$

(5.3)

where a is the inner radius of the HW. L depends on the bore size and the propagation angle: L increases when the propagation angle decreases and decreases as the bore size is increased. As the waveguide’s material is not perfect, the ray is loosing intensity at each reflection on the walls. As Miyagi’s ray optical treatment of HWs in [86] correctly yields their modal loss, provided that the Goos-Hannchen lateral shift is neglected, the power attenuation coefficient $2\alpha$ for each ray coupled to the HW is simply given by the equation:

$$2\alpha = \frac{1 - R}{2.a.tan(\theta_N)}$$

(5.4)

R the reflectivity of the waveguide wall and depends on the polarisation on the mode TE,
TM or HE considered but also depends on the propagation angle. When the orientation of the electric field of the light ray is parallel to the propagation plane, the ray is said to be p-polarised and the wall reflectance is then called $R_p$. When the orientation of the electric field is perpendicular to the direction of the propagation, the ray is said to be s-polarised and the reflectivity is then called $R_s$. When the light is not polarised, then an average reflectivity is used $R_{avg} = \frac{R_p + R_s}{2}$. For the TE modes, the reflectivity is $R_s$ while for the TM modes the reflectivity $R_p$ is used. $R = R_{avg}$ for the hybrid modes $HE_{mn}$.

$\theta_{nm}$ and $R$ need to be defined for each mode to evaluate the modal losses. For a HW of given bore size, the corresponding propagation angle for a given propagating mode of a specific wavelength is approximated by [85]:

$$\theta_{nm} = \cos^{-1}\left(\frac{\lambda_0 u_{nm}}{2\pi n_0 a}\right)$$

(5.5)

where $n_0$ is the refractive index of the waveguide’s wall, $\lambda_0$ the free space wavelength, the mode parameter $u_{nm}$ which is the $m^{th}$ roots of the $J_{n-1}(x)$ Bessel function. The lower $\theta_{nm}$ is (and then lower the mode order), the lower $\alpha_m$ will be. In figure 5.3, the $HE_{11}$, $HE_{12}$ and $HE_{13}$ angle is plotted for $\lambda_0 = 110 \mu m$.

The task then is to calculate the reflection coefficient of the HW’s wall. The calculation of the reflectivity of an incident wave on a material can be treated as well as with the wave-theory or with the ray optic theory: the incident wave excites the charges on the material which acts as a secondary radiating source and is called the reflected wave. The superimposition of the incident wave and the reflected wave is called the transmitted wave. In the case of two half infinite spaces separated by an interface, the calculation of the reflectivity at the interface can be achieved by resolving the differential equations verified by the fields at the boundary conditions. The explanation of these calculus can be found in the Electromagnetic Waves of D.H. Staelin [89]. The reflection coefficients of a plane wave incident at the air-dielectric interface are different for TE (electric field in the x-z plane) and TM (electric field in the y-z plane) polarizations, and are given by:

$$r_{TE} = \frac{\sqrt{1 - \sin^2(\theta_N)} - \sqrt{1 - \sin^2\left(\frac{\theta_N}{n_2^2}\right)}}{\sqrt{1 - \sin^2(\theta_N)} + \sqrt{1 - \sin^2\left(\frac{\theta_N}{n_2^2}\right)}}$$

$$r_{TM} = \frac{\sqrt{1 - \sin^2(\theta_N)} + \sqrt{1 - \sin^2\left(\frac{\theta_N}{n_2^2}\right)}}{\sqrt{1 - \sin^2(\theta_N)} - \sqrt{1 - \sin^2\left(\frac{\theta_N}{n_2^2}\right)}}$$

(5.6)
Figure 5.3 – $HE_{11}$, $HE_{12}$ and $HE_{13}$ mode angle $\theta_N$ in function of the bore diameter for $\lambda_0 = 110\mu m$ for a Pyrex HW.

The reflectivity (reflected power per incident power) of the mirror formed by the core-cladding interface for TE and TM polarization is:

$$R_{TE} = 1 - \frac{(1 - r_{TE}^2)^2}{(1 - r_{TE}^2)^2 + 4r_{TE}^2 \sin^2(\frac{\omega \sin(\theta_N)}{c})}$$  \hspace{1cm} (5.7)

where $\omega$ is the pulsation and $c$ the light celerity. The reflectivity for TM polarization is the same as in equation (5.7) but with $r_{TM}$ instead of $r_{TE}$.

With $\theta_N$ and $R$, it is now possible to determine the modal attenuation as a function of propagation angle. Combining the equation (5.4) and (5.5), the modal attenuation constant $\alpha_m$ can be expressed by:

$$2\alpha_m = \frac{1 - R}{2a\tan(\cos^{-1}(\frac{\lambda_0 \cdot u_{nm}}{2\pi n_0 a}))}$$  \hspace{1cm} (5.8)

In figure 5.4, the attenuation coefficients for the first four hybrid modes in a hollow-glass waveguide made of Pyrex ($n = 1.98 \pm 0.02$ in the THz frequency regime [90]) are calculated with $\lambda_0 = 110\mu m$ with the equation (5.1) which shows the $\frac{1}{a^3}$ variation of the losses and the increase of the losses with the higher order modes. In small bore diameter, it is then possible to transmit only the lowest order mode through a long distance. The larger bore waveguides can propagate on the contrary the higher order modes for
5.2 Launch condition of the lowest order hybrid mode

Figure 5.4 – Calculated attenuation coefficients for the first four hybrid modes in a Pyrex HW.

A longer distance while the total loss will be much less compared to the small bore diameter waveguides. It could be noticed also that the propagation losses are the lowest for grazing angle which correspond to the lower order modes. The losses are therefore increasing as the propagation angle increases from the optimal angles of propagation in HW $85^\circ < \theta_N < 90^\circ$.

5.2 Launch condition of the lowest order hybrid mode

The modal losses have been determined in the previous section through the ray optics formulation. To determine the coupling efficiency to the $HE_{nm}$ modes of a HW with an incoming Gaussian beam, it is easier this time to use the electromagnetic formulation from Marcatili and Schmeltzer in [83]. As shown in [91], it is important to know well the launching conditions to couple as much as possible a THz beam with the lowest hybrid mode $HE_{11}$ of the HW as it is the mode with the lowest theoretical loss. The electrical field repartition profiles of the $HE_{nm}$ hybrid modes can be expressed by [83].

$$
E_\varphi = J_{n-1}(\frac{umr}{a})\cos(n\varphi) e^{i(kz-\omega t)}$
$$
E_r = J_{n-1}(\frac{unm}{a})\sin(n\varphi)
$$

with k the propagation constant. Considering the $HE_{1m}$ modes, their field profile can
Chapter 5. Dielectric hollow core waveguides

Figure 5.5 – Calculated coupling coefficients for a HW with a bore diameter of 4 mm and \( \lambda_0 = 110 \mu m \)

be approximated by:

\[ E(r) = J_0\left(\frac{u_mr}{a}\right) \]  

(5.10)

and a Gaussian beam with a waist \( \omega \):

\[ E(r) = e^{-\frac{r^2}{\omega^2}} \]  

(5.11)

The power coupling efficiency from the incident beam to the \( HE_{1m} \) waveguide hybrid modes is evaluated by the expression proposed in [91]:

\[ \eta_m = \left| \frac{\int_0^a e^{-\frac{r^2}{\omega^2}} J_0(\frac{u_mr}{a}) dr^2}{\int_0^\infty e^{-\frac{r^2}{\omega^2}} r dr \int_0^a J_0^2(\frac{u_mr}{a}) rdr} \right|^2 \]  

(5.12)

Using (7.13), the coupling efficiency from a \( TEM_{00} \) mode with a waist \( \omega \) to the lowest order hybrid mode \( HE_{11} \) of a HW with bore diameter (figure 5.5) is maximum when the ratio beam waist/bore diameter \( \frac{\omega}{a} \approx 0.64 \). Within these conditions, the coupling efficiency is nearly 98.07%. The figure 5.5 shows the coupling efficiency of the first orders hybrid modes.
5.3 Total loss in hollow core waveguide

To estimate the total loss induced by the transmission through a HW, both the modal attenuation and coupling efficiency for each mode need to be taken into account. For a HW and a transmission in a length $L$, the power loss can be expressed by:

$$P_L = \sum \eta_m e^{-2\alpha_m L}$$  \hspace{1cm} (5.13)

where $2\alpha_m$ is the power attenuation coefficient and $\eta_m$ the power coupling efficiency of the incident beam with the hybrid modes of the HW. As many higher order modes can be excited at the same time for poor launching conditions, the power transmission through the HW is the sum of all the possible modes coupling efficiency multiplied by their respective modal losses. Figure 5.6 shows the power transmission through a 10 cm long HW with a bore diameter of $a = 4$ mm, 3 mm, 2 mm and 750 $\mu$m of a Gaussian $TEM_{00}$ mode with $\lambda = 110 \lambda m$. For the calculation, only the contributions of $HE_{11}$, $HE_{12}$ and $HE_{13}$ modes have been taken into account. The maximum transmission occurs for a smaller value $\frac{\omega}{a}$ than for optimum coupling to the $HE_{11}$ mode obtained in previous section as the transmission is made also through the higher order modes. As the bore diameter is increased, the maximal transmission through the HW is obtained for a higher $\frac{\omega}{a}$ ratio.

Figure 5.6 – Transmission of a HW in Pyrex with different core diameter for $\lambda_0 = 110 \mu m$ and a waveguide length of $L = 10$ cm in function of the waist size
5.4 Coupling of a 3\textsuperscript{rd} order DFB QCL to a HW

As it has been shown in the precedent section, one can couple with nearly 100% of efficiency a THz $TEM_{00}$ Gaussian beam into the hollow core glass waveguide lowest order hybrid mode $HE_{11}$. The main condition is to adapt the incoming beam waist with the bore diameter of the HW. The other advantage with the use of HW is the small output beam divergence. The divergence angle depends on the mode propagating in the HW [92].

The approach which has been studied first is a monolithic copper mount where a 3\textsuperscript{rd} order DFB QCL is "butt" coupled to a HW as shown on figure 5.7a. The HW has a bore diameter $a = 4 \text{ mm}$, and the optical losses for the lowest order hybrid mode $HE_{11}$ are estimated to be 0.1dB/m according to the wave-optics based formula proposed by Degnan [84]. The first transverse electric mode $TE_{01}$ has lower losses than the $HE_{11}$ for the HW with bore diameter of 4 mm. Nevertheless, as the 3\textsuperscript{rd}-order DFB QCL is linearly polarised as the $HE_{11}$ mode, this mode is preferentially excited. The $TM_{01}$ mode has much higher attenuation and is not linearly polarised so the coupling efficiency of QCL with this mode is very low. It is expected that $HE_{11}$ is the main mode for the propagation inside the HW but as the QCL is not optimally coupled to the HW, coupling to higher order lossy modes is possible and is inducting losses. We measured the output of a 3\textsuperscript{rd}-order DFB laser both with, and without, the dielectric HW. The light-current (LI) characteristics are shown in 5.7b, for three cases: no HW, with a 20-mm-long HW, and with a 30-mm-long HW. Although the laser threshold current is not affected by the presence of the HW, the output power drops to 15% of its initial value (8.2 dB loss), from $\sim 650\mu\text{W}$ to $\sim 100\mu\text{W}$, with the HW present. However, as this is almost independent of the HW length, this confirms that the coupling losses are the dominant factor.

The advantage of the dielectric HW lies in its beam shaping ability. Figures 5.8c and 5.8d show the far-fields emission patterns of the same two 3\textsuperscript{rd}-order DFB lasers, whose natural emission patterns are shown in figures 5.8a and 5.8b, once a 4-mm-bore-diameter dielectric HW is positioned at the laser facet. The beam pattern of the phase-matched device 5.8a is qualitatively improved over the situation with no HW: it has a more symmetric Gaussian shape, and its FWHM is reduced to $5^\circ \times 5^\circ$. However, the significant benefit is found in the case of the oblong and diverging QCL beam: the dielectric HW dramatically reshapes the non-ideal far-field profile into a Gaussian beam with a $5^\circ \times 5^\circ$ divergence (figure 5.8d). The directivity of the beam pattern in both cases is greatly enhanced to 55 dBi. Furthermore, the beam direction is now more predictable (along the HW direction) and so the optical alignment with other components such as mixers or detectors in-situ is simplified.

78
Figure 5.7 – Schematic diagram showing the copper block mount enabling the QCL to be coupled to the HW. b) CW light-current (L-I) characteristics of the 2.7 THz 3rd order DFB QCL with dielectric HW length of 20 mm and 30 mm, and in the absence of the HW.
Figure 5.8 – a) Typical farfield beam pattern obtained from a perfectly phase matched 2.7 THz 3\textsuperscript{rd}-order DFB QCL measured from 11 cm away from the centre of the laser (cf chapter 4) and b) the same measurements but for an oblong and diverging 3\textsuperscript{rd} order DFB QCL. c) and d) the corresponding far-field beam patterns for the quasi-Gaussian and oblong 3\textsuperscript{rd} order DFB QCLs measured with the HW in place, respectively.
5.4. Coupling of a 3rd order DFB QCL to a HW

Figure 5.9 – Physical optics simulations of dielectric hollow waveguides: a) The coupling loss as a function of the laser diode divergence angle for different HW radii. b) Beam pattern of a 4-mm-bore-diameter HW obtained 110 mm away from the waveguide aperture.

We performed ray tracing simulations to gauge the coupling losses using the commercially available software Zemax in non-sequential mode. The QCL system was mimicked using an LED, in which a divergent output power was coupled into a HW with a similar configuration to the actual experimental arrangement. The 3rd-order DFB THz QCL is therefore simulated as an LED with a $20^\circ \times 20^\circ$ divergent output. For a HW bore diameter of 4 mm, coupling losses of 8 dB were obtained numerically (Figure 5.9a), which is in good agreement with the experimental value of 8.2 dB. The simulation reveals that only the radiation coupled into the $HE_{11}$ mode is transmitted through the HW, while radiation coupled into higher order modes with higher propagation losses and should accounts for the 8 dB loss. A larger radius waveguide would reduce the coupling losses, but at the cost of increasing the spot size of the output beam pattern. The ray optics approach also permits simulation of the output beam pattern from HW. A divergence of $4^\circ \times 4^\circ$ is predicted (5.9b), in good agreement with the experimental findings.

The combination of a 3rd-order DFB QCL structure with low threshold current density, low DC power dissipation, and single mode emission at the desired frequency with a HW to reshape the emitted beam into a Gaussian beam with a $5^\circ \times 5^\circ$ divergence and an improved directivity of the beam pattern is ideal for compact integration in situ with a hot electron bolometer mixer. Although input losses into the HW reduce the transmitted power, the achieved 100 µW peak power is sufficient to pump a sensitive THz HEB mixer.

Despite being a compact and practical technique for coupling a QCL to a HW, the poor coupling efficiency can be further enhanced by reducing the aperture angle of the QCL beam. The optical coupling between the QCL and the HW was carried out by using parabolic mirrors according to the experimental scheme shown in figure 5.10. We
have used parabolic mirrors for re-collimating the QCL beam instead of lenses. With a combination of two parabolic mirrors, the THz beam of the 3rd-order DFB QCL is focused at one end of the HW. This focalisation plays two roles: it reduces the waist of the incoming beam which minimises the ratio \( \frac{w}{a} \). It also reduces the incidence angle so that the QCL beam is only coupled to the lowest order hybrid modes \((HE_{1n})\). On figure 5.11d is presented the light-current-voltage in the configuration depicted in figure 5.10. The implementation of the 30 mm long hollow dielectric (Pyrex) pipe does not affect the L-I-V curves and exhibit values and trends identical to the characteristic curves of the QCL without waveguides (figure 5.7b). The output power was measured with a Thomas Keating absolute power meter. The output power measured is nearly 100% of the total power measured without the HW in the same condition at 4K in continuous wave operation (\( \sim 650 \) \( \mu W \)). The spatial quality of the input laser is then a critical factor as it influences the input power percentages which is coupled into the HW according to (7.13). The optical losses of the hollow fiber waveguide haven’t been investigated as it would have required a much longer HWs and different HW length to see a sufficient attenuation on the TK absolute power meter. Here only the coupling losses have been optimised. The beam pattern of this coupling configuration is presented in figures 5.11a, 5.11b, 5.11c and has been measured at different distance from the HW aperture with a \( R - \theta - \varphi \) far field setup. It seems that the quality of the input QCL beam from the situation where the QCL is "butt coupled" or focused to the rear aperture of the hollow waveguide does not change its beam-shaping quality. The beam width is still \( \sim 5^\circ \times 5^\circ \) for both coupling method.
Figure 5.11 – a), b) and c) Mode profiles of the 3rd-order DFB QCL coupled to the HW with two parabolic mirrors at distances of 120 mm, 160 mm and 200 mm from the exit of the tube measured with a $R - \theta - \varphi$ far field setup. d) Light-current-voltage of the 3rd-order DFB coupled with a HW of 4 mm bore diameter in the configuration depicted in 5.10. The measurements were recorded at 10 K in continuous wave operation and the output power of the QCL was measured with a Thomas Keating absolute power meter (courtesy of M. Douared).
5.5 Further Developments: Diagonal Dielectric Horn Antenna

Below we shortly present an investigation on a design of dielectric diagonal horn THz antenna for the shaping of the QCL beam pattern. The original idea and first design have been proposed by Haotian Zhu [93] and have been further adapted to work with M-M QCL waveguides at 2.7 THz. It consists of a HR-silicon based substrate integrated image guide (SIIG) as demonstrated in [94] coupled to a diagonal horn antenna in silicon as it is depicted on figure 5.12b. It is in some ways similar to the work of Lee et al. in [95]. They fixed a hemispherical high resistivity silicon lens to a facet of the M-M waveguide QC-laser and have shown a measured collection efficiency 16 times more than the collection efficiency of a bare facet THz M-M waveguide. They also measured FWHM of $\geq 4^\circ \times 4^\circ$.

**Design and simulation** In our case, the QCL M-M is coupled through its facet to the SIIG line of the horn antenna which has the same cross-section as the QCL waveguide (cf 5.12a). The horn antenna structure needs to lay on a metallic surface. The design of the horn antenna consists of 7 pyramidal grades of $\geq 25 \, \mu m$ each. The total height of the antenna is $164 \, \mu m$. The SIIG is $15 \, \mu m$ thick and $80 \, \mu m$ wide which yields to losses of $\sim 0.02 \, dB/mm$ at 2.7 THz. The loss induced by the SIIG is then relatively low considering its small dimension (length $\sim 240 \, \mu m$). The structure around the antenna is used to ease the handling of the antenna after its fabrication for the final assembly with the QCL. As the silicon effective refractive index is close to the index of GaAs, the reflection at the interface between the QCL end facet and the SIIG is small: the transmission from the QCL to the SIIG is attenuated by $-0.32 \, dB$. The highly confined $TM_{00}$ mode of the QCL couples to the $E_{11}^0$ mode of the SIIG and propagates to the silicon horn antenna where the mode is gradually adapted to the propagation in free space. FEM simulations have been conducted for the horn antenna. The beam diagram of the horn antenna presented on figure 5.12c has a FWHM of $\sim 15^\circ \times 20^\circ$ with a near Gaussian shape. The simulations have been optimised for a single $TM_0$ mode at 2.7 THz. Ideally, the dielectric horn antenna would be coupled to a M-M waveguide with a facet emission and 1st-order DFB grating which provide the mode selection at 2.7 THz.

**Fabrication process** The 2 main steps for the process of this structure is the lithography patterning of the antennas and the transfer of the resist patterns on silicon using dry etching techniques. The complex 3D pattern and the relatively important dimensions of the diagonal horn antenna make its processing impossible with standard photo-lithography or electron-beam lithography. Instead, we used two-photon lithography with the commer-
5.5. Further Developments: Diagonal Dielectric Horn Antenna

Figure 5.12 - a) Schematic for the coupling scheme of the QCL with the Dielectric Diagonal Horn Antenna and b) Initial design of the Dielectric Diagonal THz Horn Antenna in Si. c) Calculated Farfield pattern of the horn antenna coupled to a M-M waveguide
Chapter 5. Dielectric hollow core waveguides

Figure 5.13 – SEM picture of the SU-8 model of the horn antenna processed with a 2-photons lithography from Nanoscribe (courtesy of Dominique Decanini, C2N).

cialized “nanoscribe” machine. The 2-photons lithography is based on direct laser writing process, i.e. a non-linear two-photon absorption process. The resist polymerized when two photons of near-infrared light are absorbed simultaneously. The high light intensity is provided by a femtosecond laser. Dominique Decanini processed the horn antenna on the Nanoscribe machine. First, a perfect 1:1 photo resist (PR) mask in resist SU-8, as shown in figure 5.13 is built. The deep reactive ion etching (DRIE) is then performed to transfer the mask to HR-silicon wafer. The work on the ICP was made with the help of Etienne Herth. At the time where this manuscript was written, the work was still in progress to obtain an ICP etching recipe on silicon which could achieve highly vertical side-walls on the whole height of the horn antenna (164 µm). The combination of complex patterning with 2-photons lithography and transferring these patterns on silicon with a dry etching process is a great asset for designs applied in the THz where a precision in the µm range is required which cannot be achieved with standard micro-machining techniques.
Chapter 6

Superconduction Hot Electron Bolometer Mixer

6.1 Introduction

In this thesis, hot electron bolometers (HEB) based on niobium nitride (NbN) on thick silicon substrate have been used as mixers for heterodyne experiments. The realisation of the THz HEB mixer, as used in the experimentations whose results are presented in the next chapter, is relatively complex requiring special means and skills in nano-fabrication processes, mechanical engineering and quasi-optical system. Considering the time allowed for the completion of the thesis, no works have been done specifically on the HEB itself. HEB mixers developped at LERMA during previous R&D projects [96] [40] have been used in this thesis. Nevertheless, a good understanding of the physics and the theory of this kind of devices is mandatory to elaborate a compact heterodyne receiver for future space-borne missions. The goal of this chapter is to present a short introduction but rather exhaustive on the HEB basic working principles, and to describe the HEB used in this work.

6.2 Bolometer effects

A bolometer is a type of detector which is sensitive to the heat it absorbs. Bolometers are composed of a thermal absorber with a thermal capacity $C$ which is connected to a heat sink through a temperature dependant resistor (figure 6.1). The absorber is heated up by the incoming radiation power and is cooled down to an equilibrium temperature, balanced with the heat transfer to the cold temperature reservoir across the resistor. Knowing the heat capacity of the absorber $C$ and the thermal conductance $G_{th} = \frac{1}{R_{th}}$, the thermal relaxation time is defined as:

$$\tau_{th} = \frac{C}{G_{th}} \quad (6.1)$$
From the above equation, fast relaxation time is obtained for large thermal conductance and small heat capacity. At thermal equilibrium, the thermal balance of a bolometer can be simply written by:

$$C \frac{dT_{\text{bolometer}}}{dt} + G_{\text{th}}(T_{\text{bolometer}} - T_{\text{bath}}) = P_{\text{radiant}}(t)$$

(6.2)

When considering a sinusoidal change of radiant power at frequency $\nu$ and amplitude $\Delta P$, the above differential equation is solved and the temperature oscillation of the bolometer is:

$$\Delta T_{\text{bolometer}} = \frac{\Delta P_{\text{radiant}}}{G_{\text{th}}} \frac{1}{\sqrt{1 + (2\pi \nu \tau_{\text{th}})^2}}$$

(6.3)

Eq. 6.3 shows that at high frequencies, the temperature response to the radiant power is reduced: at the roll-off frequency (or cut-off frequency) $\nu_{roll-off} = \frac{1}{2\pi \tau_{th}}$, the response is reduced to $\frac{1}{\sqrt{2}}$. $\tau_{th}$ is the characteristic time required for the bolometer to react to a change in the input signal.

### 6.3 Superconducting Hot Electron Bolometer

In radio-astronomy observations, the signal power from the ISM in the THz frequency range is very weak and difficult to detect. To detect this kind of signals, highly sensitive detectors are required such as the hot electron bolometers (HEB). Superconducting HEB can be seen through the simple model proposed in the previous section. A HEB is composed of an ultra-thin micro-bridge in superconducting material (Nb, NbN, MgB$_2$) connected to two metal pads which play the roll of the thermal heat sink. The superconducting material is usually only few nano-meter thick. The micro-bridge is the incoming
6.3. Superconducting Hot Electron Bolometer

![Diagram of a superconducting hot electron bolometer](image)

Figure 6.2 – Thermalisation schematic in hot electron bolometer adapted from [5]

Radiant power absorber. The heat transfer from the micro-bridge can also be achieved via the substrate on which the superconducting material is relying on. The incoming radiant power is heating up the electron gas which is cooled down by diffusions or phonon emission of the "hot electron" with the cold temperature sink (normal metal pads, substrate). The thermal relaxation time constant of the HEB defines the bandwidth for which the difference between the LO and RF signals can be detected: only spectra with frequencies in this range can be observed. This bandwidth can be increased by reducing $\tau_{th}$, implying to reduce the thermal resistance or to reduce the thermal capacity. The electron temperature is completely decoupled from the lattice temperature and an increase of the electron gas temperature can be recorded with electrical measurements. A small heat capacitance and a large conductance can result in a large HEB response: $S_{HEB} \propto \frac{dR}{dT}$.

The hot electron effects are due to the large differences in interaction times between electrons-electrons ($\tau_{e\rightarrow e}$) and the electrons-phonons ($\tau_{e\rightarrow ph}$) and also the difference between their heat capacitances [5]: a large electronic heat $C_e$ capacity and a slow electron-phonon interaction time $\tau_{e\rightarrow ph}$, implying that when an incoming radiant power heats the electrons above the temperature of the lattice, the interaction time between electrons and phonons is too slow to 'evacuate' the heat off the bridge. The heating of the lattice’s electron higher to the lattice temperature is called "hot electron" effect. Hot-electrons have been first used to describe non-equilibrium electrons in semiconductors and since have been observed in metals and superconductors [97]. The thermodynamic scheme of electrons gas and phonons in a HEB is represented in figure (6.2).
6.3.1 Heterodyne mixing with Hot Electron Bolometer

The phase information of an incoming signal can be measured using a heterodyne detection scheme. The heterodyne principle gets involved a local oscillator signal (LO) and a RF signal of interest. These two signals are overlaid with a diplexer or beam-splitter and further coupled on the heterodyne mixer. The RF signal is mixed with the LO and is down-converted to frequencies (intermediate frequency) accessible with standard electronics systems (amplifiers, spectrum analyser...). The RF spectral range which is accessible with the heterodyne measurements mostly depends on the LO frequency and the intermediate frequency (IF) bandwidth. HEBs used as heterodyne mixer have an IF bandwidth of several GHz, allowing measurements on a large spectral range. For the mixing, the HEB is DC biased and kept in a bath temperature below the superconducting bridge’s critical temperature. The LO is also used to optically pump the HEB to its resistive state. The combination of LO power and DC bias leads to the formation of a hot spot. When the RF and LO signals are close in frequency, the hot-spot length variates at the intermediate frequency. The power dissipated by the bolometer through the thermal resistance is written:

\[
P(t) = \frac{V(t)^2}{R} = \frac{1}{R} (V_{LO} \cos(\omega_{LO} t) + V_{RF} \cos(\omega_{RF} t))^2
\]

\[
= \frac{1}{R} (V_{LO}^2 \cos^2(\omega_{LO} t) + V_{RF}^2 \cos^2(\omega_{RF} t) + 2V_{LO} V_{RF} \cos(\omega_{RF} t) \cos(\omega_{LO} t))
\]

\[
= P_{LO} + P_{RF} + \frac{1}{R} (V_{RF} V_{LO} \cos((\omega_{RF} - \omega_{LO}) t) + V_{RF} V_{LO} \cos((\omega_{LO} + \omega_{RF}) t)
\]

\[
(6.4)
\]

where \(V_{LO}\) and \(V_{RF}\) are the voltage of the LO and RF signals, \(R\) the resistance of the HEB and \(\omega_{LO/RF}\) the pulsation. \(P_{LO}\) and \(P_{RF}\) are respectively the power dissipated at frequencies \(\omega_{LO}\) and \(\omega_{RF}\) and are determined by:

\[
P_{LO} = \int \frac{V_{LO}^2 \cos^2(\omega_{LO} t)}{R} = \frac{V_{LO}^2}{2R}
\]

\[
P_{RF} = \int \frac{V_{RF}^2 \cos^2(\omega_{RF} t)}{R} = \frac{V_{RF}^2}{2R}
\]

\[
(6.5)
\]

The RF signal is much weaker than the LO signal so \(P_{RF}\) can be neglected from the total dissipated power and the signal components at frequencies \(2\omega_{LO}, 2\omega_{RF}\) and \(\omega_{LO} + \omega_{RF}\) are much higher than the roll-off frequency of the bolometer and only the signal \(\omega_{RF} - \omega_{LO}\) is dominating. The average power dissipation in the HEB is then written:

\[
P(t) \approx P_{LO} + \frac{V_{LO} V_{RF} \cos(|\omega_{RF} - \omega_{LO}| t)}{R}
\]

\[
(6.6)
\]
6.3. Superconducting Hot Electron Bolometer

The frequency band above $\omega_{LO}$ is called the upper side-band and the band underneath $\omega_{LO}$ the lower side-band. The intermediate frequency (IF) oscillation $\nu_{IF} = |\nu_{RF} - \nu_{LO}|$ leads to temperature oscillations only if the IF is the range of the $\nu_{roll-off}$. The responsivity of a superconducting hot electron bolometer as expressed in [98], which has been found using a lumped element model, gives the bolometer voltage compared to the absorbed energy. It is expressed in V/W and is written:

$$R_V = \frac{\Delta V}{\Delta P} = \frac{\Delta R I_{bias}}{\Delta T G_{th}} \frac{1}{\sqrt{1 + (2\pi f \tau_{th})^2}}$$

(6.7)

Where $I_{bias}$ is the bias current of the bolometer. The voltage responsivity due to the temperature variation for a heterodyne mixing is therefore written:

$$V_{IF} = R_{V,IF} \Delta P = R_{V,IF} \sqrt{P_{LO} P_{RF}} = \frac{\Delta R I_{bias}}{\Delta T G_{th}} \frac{\sqrt{P_{LO} P_{RF}}}{\sqrt{1 + (2\pi f \tau_{th})^2}}$$

(6.8)

The mixer conversion gain can be estimated by evaluating the ratio between the incoming RF power and the IF power response through a resistive load $R_L$. $P_{IF}$ is expressed by:

$$P_{IF} = \frac{V_{IF}^2}{2R_L} = \left(\frac{\Delta R I_{bias}}{\Delta T G_{th}}\right)^2 \frac{P_{LO} P_{RF}}{2R_L (1 + (2\pi f \tau_{th})^2)}$$

(6.9)

The mixing gain conversion for a single side band mixer is:

$$\eta_{mix}^{SSB} = \frac{P_{IF}}{P_{RF}} = \left(\frac{\Delta R I_{bias}}{\Delta T G_{th}}\right)^2 \frac{P_{LO}}{2R_L (1 + (2\pi f \tau_{th})^2)}$$

(6.10)

From (6.10), $\eta_{mix}^{SSB}$ is proportional to the LO radiant power but is clamped to the saturation level of the bolometer: if the LO radiant power heat up all the superconducting bridge to its normal state, there will be no mixing possible.

6.3.2 Current-Voltage Characteristics of the Superconducting HEBs

The temperature at which the resistivity of the material disappears is called the critical temperature and is noted $T_C$. The most used superconducting materials for bolometers or HEB are the nobium (Nb), niobium nitride (NbN) or $MgB_2$ with a respective $T_C$ of 6 K, 9.2 K and 39 K (ultra thin layers). When below $T_C$, the superconductor allows a current to flow through it without any resistance until attaining a critical current where
Figure 6.3 – Current-versus-voltage curve of a diffusion cooled HEB with NbN superconducting material grown on HR-Si and measured in a liquid helium dewar at 4.2 K. For the blue curve, the LO is not radiating on the HEB while for the pink, red and black curves, the LO optical power is increased until the negative differential resistance has vanished.

the superconductivity is abruptly destroyed: the superconducting material presents then a normal resistance. It can be seen with the Silsbee criterion [99]. The current flowing through the material produces a self-field $H$ at its surface which is proportional to the current intensity. When reaching the critical field $H_C$ of the material, it destroys the superconductivity. A typical I-V characteristic of HEB is presented in figure (6.3)

From a zero bias voltage, the current is increased with the voltage with nearly a vertical slope. The remaining resistance (few ohms) observed is due to the electrical wiring and the finite resistance of the gold contact pads of the HEB. The current reaches then the critical current $I_c$. When reaching the resistive state, pairs of vortex-anti-vortex exist below the Kosterlitz-Thouless temperature $T_{KT}<T_c$. Above $T_{KT}$, free vortices appear but are not fixed due to the Lorentz force. The bias current flowing through the bridge breaks the vortex-antivortex pairs. The vortices seemingly induce a voltage drop across the bridge which reduce by the same way the current below the critical current. Then the vortices vanish and cycles of oscillations between superconducting and transitional state of the material start. It is a very unstable region on the HEB I-V curve. Increasing again the voltage finally increases the electrons temperature which results in the creation of a hot-spot where electrons exceed the critical temperature (this latter point will be further disused in section 6.3.4). The hot spot balances the negative differential resistance, dissi-
6.3. Superconducting Hot Electron Bolometer

Pates DC power and a stable state is reached where usually the HEB is the most sensitive. A further increase in the bias voltage destroys completely the superconductivity and the material becomes normal. A Meissner state region is also found on the I-V curve for very small bias voltage [100] which is decreased when the RF radiant power is increased. The oscillation region becomes smaller as the RF power on the HEB is increased until the oscillations are completely suppressed by the RF heating. Another model of the I-V characteristic of NbN based mixer has been further proposed in [101] which used the Joule-self heating mechanism to describe the HEB bi-stability. The bi-stability of the mixer means that two different states can co-exist on the bridge: the normal resistance hot-spot and the superconducting state. Oscillations are observed from the superconducting state and resistive state when the critical current is reached due to the Joule sustained self heating: the Joule heat released by the hot-spot becomes sufficient to raise the temperature of the bridge. The critical current decreases as the temperature is increasing in this region and a larger part of the bridge would have a normal resistivity and dissipate power. With this simple model, the self-heating mechanism would be sufficient to heat-up the whole bridge above the $T_C$. Actually, the bistable region has not enough power available to sustain a stable hot spot which will eventually shrink, reducing the self-heating; the current subsequently increases which leads to the hot-spot oscillating back and forth until the device goes superconducting, etc..

6.3.3 Cooling mechanism in HEBs

There are two dominant cooling mechanisms in superconducting HEB. The first is the phonon cooled mechanism: the non-equilibrium electrons loose energy to the lattice by interactions with phonon. The second is the diffusion cooled: the energy is removed by diffusion of hot electrons to the normal metal pad heat sinks. This later mechanism is only efficient for small dimension bridges. Depending on the HEB configuration and the superconducting material of the bridge, either the phonon cooling or the diffusion cooling will dominate the cooling mechanism in the HEB.

1 Phonon Cooling

Hot electrons in the superconducting bridge interact with the lattice phonons which then escape into the substrate. The electron-phonon interaction time $\tau_{ep}$ has been observed in different materials including composite materials to be proportional to $\propto T_e^{-1.6}$ [102]. For NbN for instance, $\tau_{ep} = 13$ ps at 4.2 K for a 5 nm thick film [103]. The thermal constant as defined in the section 6.2 depends also on the escape time of the phonon into
the substrate $\tau_{\text{esc}}$ and the electron and lattice phonon heat capacities $C_e, C_{ph}$ [104]

$$\tau_{th} = \tau_{ep} + \frac{\tau_{\text{esc}}C_e}{C_{ph}} \quad (6.11)$$

To avoid any back-filling of energy in the electron gas, $\tau_{\text{esc}}$, which depends on the film thickness, needs to be much smaller than the phonon-electron time. Some approaches to enhance the IF bandwidth have been proposed using buffering layers of MgO beneath the superconducting layer to improve the acoustic matching [96].

### 2 Electron Diffusion

Electron diffusion as cooling mechanism has been proposed in [98] for transition edge micro-bolometer for nano-scale bridges. The thermal time constant depends then on the electron diffusion constant $D$ and the length of the micro-bridge $L$. It can be written by [105]

$$\tau_{th} = \frac{L^2}{\pi^2 D} \quad (6.12)$$

If the bridge length $L$ is inferior to the characteristic length $\sqrt{D\tau_{ep}}$, the out-diffusion of the hot electrons into the normal metal pads dominates over the phonon diffusion cooling mechanism [106] [98].

### 6.3.4 Hot Spot Model

Several models have been studied in the literature to determine the characteristic of the HEBs. The first of them, the lumped model, was used to describe niobium and $YBa_2Cu_3O_7$ superconducting mixers [107] and could predict the conversion efficiency of the heterodyne mixers, their noise temperature and fluctuation sensitivity. In [108] Karasik et al. used another simplified model, the broken-line transition model, to estimate the mixer gain and its noise performances. The approach which will be discussed in this section is the hot spot model. The following discussion has been widely inspired by the work of J. Baubert et al. [6], W. Floet [109] and H. Merkel [110] which dealt with superconducting micro-bridges and their applications for HEBs employed for heterodyne mixer. They explain that the hot-spot formation by a thermal heating in the center of the superconducting bridge by absorption of LO signal and DC bias power. The hot-spot model assumes that the electron temperature in the microbridge is not uniform but instead follows a gradient of temperature along the bridge length (see figure 6.5 b).

The Andreev reflection has an important effect on the hot electrons (fig 6.5) as it confines them in the center of the superconducting bridge and limits their diffusions. The
Figure 6.4 – a) Schematic of the superconducting HEB showing the different diffusion mechanisms: electron-electron scattering, electron-phonon scattering and phonon-phonon scattering. The Andreev reflections marked by the curved arrow on the schematic limit the heat diffusion to the gold pads (normal metal heat sink) and form the hot-spot delimitations. The figure is adapted from [6]. b) Schematic of a 1-D electron temperature in the HEB as a function of the bridge length.
effects of Andreev reflection have been more deeply treated in [111]. In the hot spot model, the highest electron temperature is in the center of the superconducting bridge. When the temperature is higher than the critical temperature of the superconductor, the bridge has locally a normal resistance. The bolometer resistance is proportional to the length of the hot-spot. While the DC current going through the bridge can only dissipate energy at the part of the bridge with normal resistance, the incident radiant power from the local oscillator $P_{LO}(t)$ can bring enough energy to break the electron Cooper pairs. The hot-spot size will depend on the bias current of the HEB and the incident radiant power $P_{RF}$: its size is varying at the beat frequency IF. In the hot-spot model, where the lowest temperature is near the heat sink gold pads where $T_e$ is equal to the heat sink temperature and the hottest temperature is in the center of the bridge. One-dimensional heat balance equations for electrons and phonons can be solved to obtain the electron temperature profile in the bridge.

$$\frac{dT}{dx} = \frac{\lambda_e}{m \tau_{ep} T_e} \left( T_e - T_n \right) + P(t) + j^2 \rho_N = 0 \tag{6.13}$$

$$\frac{dT}{dx} = \frac{\lambda_p}{m \tau_{esc} T_p} \left( T_p - T_n \right) + \frac{C_e}{n \tau_{ep} T_e} \left( T_e - T_p \right) = 0 \tag{6.14}$$

where $\lambda_e/p$ is the thermal conductance of the electrons and phonons, $\tau_{ep}$ is the electron-phonon relaxation time and $\tau_{esc}$ is the escape time of the phonon in the superconductor bridge into the substrate, $C_e/p$ are the electron/phonon heat capacity and $n$ is chosen to be equal to 3.6 for the NbN and $m = 4$ in agreement with [112]. Subscript $e$ and $p$ denote electron and phonon parameters respectively. The DC bias dissipation energy is represented by the term $j^2 \rho_N$ where $j$ is the current density and the $\rho_N$ the normal resistance of the hot-spot. The first terms in (6.13) and (6.14) represent the thermal conduction in the direction of the microbridge while the second term in (6.14) stands for power per unit length flowing from electrons to phonons. In (6.13), the power per unit length is flowing from electrons to phonons (cooling mechanism) while in (6.14), the phonons are heated up by electron phonon interaction while the last term in (6.14) describes phonon escape to the substrate. The incoming radiant power $P(t)$ resulting of the superimposition of the local oscillator and the RF power, causing a change in the superconducting film resistance is expressed as following

$$P(t) = P_{LO} + 2 \sqrt{P_{LO} P_{RF} \cos(\omega_{IF} t)} \tag{6.15}$$

The LO power is assumed to be absorbed uniformly over all the length $L$ of the bridge while the DC current is only absorbed by the hot spot. Solving these equations for electron
temperature $T_e$ gives the I-V curve, the mixer gain, and the noise temperature curves for HEB on substrate, assuming that the resistance of the bridge depends on the electron temperature. Miao et al. [113] has proposed an improved hot spot model where the absorption of the LO is not uniform along the microbridge and is mainly absorbed in the part of the superconducting bridge where the energy gap $2\Delta(T)$ is less than the incoming photon energy. The term $P(t)$ in the heat balance equation (6.13) would be multiplied by a radiation absorption efficiency term $\eta(T_e, \omega)$ which depends on the electron temperature and the photon energy.

The HEB mixer conversion gain is then given by [108]:

$$ G = \frac{2(I_0 C_{RF})^2 P_{LO} R_L}{(R_L + R_0)^2 (1 - C_{DC} R_0^2 R_L - R_0) R_L + R_0 \left[ 1 + \frac{(2\pi f \tau_{mix})^2}{1 + (2\pi f \tau_{mix})^2} \right]} $$

(6.16)

where $C_{RF}$ and $C_{DC}$ represents the heating capacity of the RF and DC incoming power, $R_0$ and $I_0$ are the HEB DC resistance and current at the bias operation point and $R_L$ the load resistance. The bolometer with resistance $R_0$ is capacitively coupled to the IF amplifier load with resistance $R_L$ (figure 6.5). We differentiate the heating from the RF power and the DC current by separating the capacity into $C_{DC}$ and $C_{RF}$. The differential resistance can be determined with the I-V curve of the pumped HEB $R_D = \frac{dV}{dI}_{DC}$ for a set bias point. It is possible to find experimentally the gain bandwidth of a HEB using 2 LO radiant power [106]. One LO is fixed in frequency and its output power level is adjusted to keep a constant optimal pumping level of the HEB while the second LO is
frequency tuned to obtain an IF signal. The relative mixer gain versus the IF frequency
is presented in figure 6.6 which has been obtained using (6.16) for a NbN film with a
thickness of 4 nm on Si at bath temperature of 6 K.

6.3.5 Noise in HEBs

The noise mechanisms in superconducting HEB have been widely studied in [114] and
[115]. The major noise sources in HEB mixer are known to be the Johnson noise and
thermal fluctuation noise. There are uncorrelated noise sources due to electrical bias
power. The quantum noise $T_{QN}$ has also an unavoidable contribution to the mixer noise,
even though the mixer is at 0K; it is called the zero-point fluctuation and its value is $\frac{h\nu}{2k_B}$
for a double side band mixer. This is the quantum limit of the noise temperature for a
heterodyne system when performing measurements in DSB. The quantum noise depends
on the frequency and becomes important in the THz frequency range. An extended study
of the quantum noise is reported in [16]: the quantum noise takes over the classical noise
at 5 THz. The aim here is to find the equivalent mixer input noise temperature defined
by $T_{in}^{\text{noise}} = \frac{T_{out}^{\text{noise}}}{G_{\text{mixer}}}$ with $T_{out}^{\text{noise}}$ the output noise temperature of the mixer and $G_{\text{mixer}}$ the
mixer gain.

1 Johnson Noise

Johnson noise describes the emitted noise by a resistor $R$ which is caused by thermal
agitations of charge carriers. The Johnson noise equivalent voltage $V_{in}$ in a bandwidth $B$
can be expressed by:

$$< V_{JN}^2 > = 4k_B T_e R B $$ (6.17)

The equivalent noise temperature at the output of the HEB due to the Johnson noise is
[108]:

$$T_{JN}^{out} = \frac{4R_L R_0 T_e}{(R_L + R_0)^2 (1 - \frac{I_0^2 C_{DC}}{R_L + R_0})^2} $$ (6.18)

The equivalent noise temperature from the Johnson noise at the input of the HEB mixer
can be obtained by dividing (6.18) by the conversion gain (6.16).

$$T_{JN}^{in} = \frac{R_0 T_e}{I_0^2 C_{RF} P_{LO}} (1 + \omega_{IF}^2 \tau_{mix}^2) $$ (6.19)
From (6.19), the Johnson noise depends on the intermediate frequency: \( T_{JN}^{\text{in}} \) increases for higher \( \nu_{IF} \) and lower G.

## 2 Thermal Fluctuation Noise

Thermal fluctuation is the largest noise contribution below \( \nu_{roll-off} \) in a bolometer. It is caused by random energy exchanges between electrons, phonons in the bolometer and the heat sink. This fluctuation leads to a variation in the electrons temperature and further to a change of the bolometer resistance, therefore implying a variation in the output noise power. The thermal fluctuation equivalent input noise can be written as:

\[
T_{FL}^{\text{in}} = \frac{2T_e^2 \tau_\theta}{C_e V C_{RF}^2 P_{LO}} \left( \frac{\partial R}{\partial T} \right)^2 \frac{1 + \omega_{IF}^2 \tau_{mix}^2}{1 + \omega_{IF}^2 \tau_{th}^2} \tag{6.20}
\]

\( C_e \) is the electron heat capacity and \( V \) the volume of the bolometer. The mixer time constant \( \tau_{mix} \) is usually close to the value of the electron thermal time constant \( \tau_{th} \). The thermal fluctuation noise is proportional to the square of the electron temperature (contrary to the Johnson noise which varies with \( T_e \)) but is independent of the intermediate frequency.

## 3 Noise Bandwidth

The gain bandwidth of a HEB has been treated in the section 6.3.4 and refers to the points where the conversion efficiency is reduced from its maximum by 3 dB. The noise bandwidth, in contrary, is defined as the points where the noise temperature is increased by a factor 2. The noise bandwidth is usually larger than the gain bandwidth and mostly due to the Johnson noise which is frequency independent [106]. In [116] the authors have determined experimentally that the noise bandwidth is 1.7 times larger than the gain bandwidth for a diffusion cooled HEB in Nb. For a heterodyne receiver, the noise contribution of the IF chain \( T_{IF} \) needs to be taken into account. The equivalent input noise temperature of the receiver including the IF amplifier noise contribution is plotted on figure 6.6 and has been deduced with the fact that \( T_{out}^{\text{noise}} = T_{noise}^{\text{in}} G_{mixer} \) and with equations (6.16), (6.19) and (6.20). The parameters used for the plots in 6.6 have been chosen with measured typical mixer properties: \( T_C = 9.5 \) K, \( \Delta T_C = 1.5 \) K, \( R_L = 50 \Omega \) (IF amplifier), \( R_0 \) is around 80 \( \Omega \). The Johnson output noise is small but constant. On the contrary \( T_{FL} \) is large at low IF frequency and decreases in for high IF. An approximation of
Figure 6.6 – Relative mixer gain versus IF frequency calculated for a 4 nm thick NbN films and bath temperature of 6 K with the calculated noise temperature contribution from the Johnson noise and the thermal fluctuation noise, and the receiver noise temperature versus IF frequency.

the noise bandwidth of HEB mixer related to the gain bandwidth can be written by [117]:

\[
f_{\text{noise}} = f_{\text{gain}} \sqrt{\frac{T_{\text{FL}}^{\text{out}} + T_{\text{IN}}^{\text{out}} + T_{IF}}{T_{\text{JN}}^{\text{out}} + T_{IF}}} \tag{6.21}
\]

6.4 Requirements for the HEB

To obtain highly sensitive HEB for heterodyne measurements in the THz frequency range, few characteristics of the HEB need to be properly optimised. The critical temperature of the superconducting bridge is fixed to be above the liquid helium temperature (4.2 K) so that standard cryogenic liquid dewar can be used. The IF bandwidth is required to be few GHz at least. Also, the desired frequency range of interest sets the dimension and the type of planar antenna used for the quasi-optical coupling of the incoming signal to the antenna coupled bolometer (either log-spiral or double slots for instance). The length and the width of the bridge has to be chosen so that its impedance is matching to antenna impedance. As it has been explained in a previous section, the choices for
the superconductor material and the film thickness used have a large influence on the HEB characteristics: relaxation time, IF bandwidth, critical temperature. The cooling mechanism in a HEB, as treated in a previous section, is performed by either phonon transfer or electron diffusion from the hot-spot to the heat sink. The electron-phonon interaction time $\tau_{ep}$, phonon-electron time constant $\tau_{pe}$ and the phonon escape time $\tau_{esc}$, the time that the non-equilibrium phonon escape to the substrate, are primordial parameters for phonon cooled HEB and depend on the type of superconducting material used. To have a fast response mixer, $\tau_{esc}$ must be very short and the film thickness must be very thin in order to evacuate the heat to the substrate for phonon cooled HEB. To avoid a back-exchange of energy to the electrons, it is mandatory to have $\tau_{esc} < \tau_{pe}$.

Considering the non-equilibrium electron diffusion to the normal metal and $\tau_{diff}$ being proportional to the length of the device, the design of the HEB bridge needs to be in the sub-micron dimension to exploit also the electron diffusion to the heat sink as cooling mechanism. A device length in the order of 200-300 nm was first reported with a bandwidth of 2 GHz [98]. Both electron diffusion and phonon cooling mechanism can cohabit, so the IF bandwidth can be further increased. These later characteristics need to be traded for an optimum configuration of the HEB used as heterodyne mixer. NbN as superconducting material has been widely used for HEB because of its short $\tau_{ep}$ (13 ps) and its high $T_c$ (9.2 K) which correspond to an IF bandwidth of 10 GHz. The phonon escape time $\tau_{esc}$ is 40 ps and the electron diffusion coefficient of the NbN is in the order of $D = 0.5 \text{ cm}^2 \text{s}^{-1}$. The dominant cooling mechanism on our NbN HEB with a micro-bridge length of 200 nm is phonon cooling.

6.5 LO requirements

For the compact integration of the HEBs in a heterodyne receiver for space-borne mission, the LO output power required needs to be taken into account. Several types of LO are available today in the THz with frequency dependant output power (cf chapter 1). For frequencies below 1.8 THz, Schottky diode based chains provide dozens of $\mu W$ of output power [118] which is plenty to fully pump a single pixel HEB. The maximum output power from the Amplifier Multiplier Chain (AMC) VDI 390 for a frequency range of 1310 GHz and 1380 GHz is about 60 $\mu W$. The output power of an AMC is nevertheless greatly decreased for frequency above 2 THz. FIR gas lasers as used in [29] and backward wave oscillator (BWO) can be used for frequencies above 1 THz but the limited available power and space in a satellite make them difficult for integration in an embedded heterodyne receiver. The QCLs stand out to be a good trade-off between the available output power and compactness. Nonetheless, the LO output power require to pump the HEB needs
to be defined to elaborate the specification of the LO. The LO radiation as mentioned in 6.3.1 is used to bring the HEB into the region on the I-V curve where the detector reaches its highest sensitivity, that is to say where the mixer conversion gain is the highest and the equivalent noise temperature is the lowest. A less pumped I-V curve would give a higher receiver equivalent noise temperature but as a trade-off would require less LO power. The factor which sizes the LO power requirements is the dimension of the HEB bridge. Another factor is the critical current of the superconducting material $I_c$. With a higher $I_c$, more energy is required to break the Cooper pairs into quasi-particles to have an optimum temperature distribution in the HEB [119]. The LO power requirement can be summarised by [120]:

$$ P_{LO} = \frac{\gamma V(T_n^c - T_0^c - nT_n^c \Delta T_c)}{n\tau_{th}T_n^c} \quad (6.22) $$

where $\gamma$ is the Sommerfeld constant, $T_0$ is the bath temperature, $\tau_{th}$ the thermal time constant $\sim 35\,\text{ps}$ and $n = 3.6$ for NbN as determined experimentally [112]. The bath temperature plays an important role in the LO power requirements. A lower $T_0$ would require a higher LO output power (20% increase of LO power is required when $T_0$ is changed from 4 K to 3 K) but at the same time the equivalent receiver noise temperature would be decreased. Of course, another important factor for the LO power requirement is the efficiency of optical coupling between the LO and the mixer. The beam pattern of the LO needs to overlap without "spill-over" with the receiver’s beam diagram. A way to determine the absorption of LO optical power by the HEB using the isothermal technique is presented in [96]. Typically for the NbN HEB mixer used in LERMA with a bridge size of $2 \times 0.2\,\mu m^2 \times 3.5\,\text{nm}$ and a critical current of 260 $\mu A$, the LO power required is about 270 nW.

### 6.6 Coupling to the HEB

Several ways are available to couple the HEB to the LO and RF signals for the heterodyne measurements. A classic way which has been used in several instruments [121] [122] is to use a waveguide probe for the coupling with the HEB. The RF signal is overlaid with the LO signal and is led through the waveguide with a horn antenna. The waveguide itself is coupled to the HEB mixer. Usually, rectangular hollow core metallic waveguides are employed and are designed to have very low optical losses and low divergence even for THz frequency range. It has been also demonstrated [123] an integration of a free-standing thin membrane HEB in waveguide. The waveguide horn antenna shapes the incoming RF beam and transform the free space wave mode into a waveguide mode. Despite the
difficulties to construct waveguides adapted to the sub-mm wavelength, the waveguide horn allows a quasi perfect coupling efficiency (98%) with an incoming Gaussian beam (the telescope signal for instance) and acts as a filter for frequencies below the cut-off frequency of the waveguide. Another technique is the quasi-optical coupling and it is this later technique which has been employed for this work. Contrary to the visible optics, sub-mm signal propagation can’t be approximated with the usual geometrical optic theory. A quasi-optical approach as described in [124] needs to be employed. The critical point for this technique is the coupling of the incoming signals to the HEB. It can be done by using focusing mirrors or lenses. One concept widely used in quasi-optical HEB mixer is to use the HEB’s substrate (silicon in our case) lens with an integrated planar antenna. The silicon lens is an attractive solution as this material offers extremely low losses in the THz frequency range and the effective refractive index of the lens can be matched with the one of the HEB substrate, avoiding interface reflection. This section will be focused on this latter quasi-optical technique as it is the one used in the experimental heterodyne set-up in LERMA.

6.6.1 Planar Antennas on dielectric substrate

Planar antenna is a key element for the coupling of RF/LO signals to the superconducting bridge. Main characteristics of planar antennas deposited on a dielectric substrate have been described in [125]. The modes of the antennas can be as well as considered backward or forward using the reciprocity theorem [124] [126]. A planar antenna laying on a dielectric substrate radiates most of its power into the dielectric side in contrary to a planar antenna in free space which radiates symmetrically. Considering a simple slot or dipole antenna, the power radiated from a slot antenna into the dielectric is proportional to \( \varepsilon^{3/2} \) or for a dipole the radiated power is proportional to \( \sqrt{\varepsilon} \) [127]. They have experienced that for \( \varepsilon_r > 1 \), the radiation field in the interface plane drops discontinuously to zero when \( \varepsilon_r \) increases and strong maxima appear at the critical angle in the dielectric. Actually, for \( \varepsilon_r > 8 \), the antenna beam diagram in the dielectric approximates to a cone with a semi angle of 45°. The antennas on dielectric being more sensitive from the substrate side, it is easier to couple the incoming signal to the HEB from this side. In their Schottky diode based receiver at 110 GHz, Clifton et al. [128] proposed to mount the mixer at the aperture of a hollow-metallic waveguide; the incoming signal couples through the dielectric substrates to the coupling antennas. They measured a DSB noise temperature of 339K at 110 GHz. A similar coupling scheme has been used with SIS junction mixer reported in [24]. The solution which has been widely explored by the HEB mixer community is to mount the planar antenna on a dielectric lens. Engheta et al. in
[129] proposed a way to calculate the radiation pattern of an infinitesimal dipole laying at the interface of two dielectric half spaces. They found that in the upper half space for which the dielectric refractive index is lower than for the lower half space, the diagram is constituted with a single lobe which has a maximum at the normal of the interface while the side with higher index of refraction, the pattern is formed by 3 lobes. Most importantly, they found that the power distribution in the upper half space ($n_{\text{upper}} < n_{\text{lower}}$) is decreasing as the refractive effective index of the lower half space is increased.

An estimation of the proportion of the antenna gain repartition between the air side and the dielectric side has been obtained using the finite element solver FEKO. In figure 6.7, it is clear that for the case of the planar antenna surrounded by free space (6.7a), the antenna’s gain is symmetrical along the antenna’s plan and shows two single lobes at $\theta = 0^\circ$ and $\theta = 180^\circ$: this situation is representative of the HEB on membrane when there is no back-short mirror. As the refractive effective index of the lower half space is increased from 1 (6.7b and 6.7c), the antenna’s gain in the air side is reduced while the gain in the dielectric side is increased. When the planar antenna is lying on silicon, the gain distribution in the air is ten times smaller than on the dielectric side (10 dB of difference). This simulation doesn’t take into account the effect of the lens on the beam diagram of the planar antenna. Simulation of an integrated log-spiral antenna with a hemi-spherical silicon lens have been already conducted but at 600 GHz [111]. The dimension of the lens (1 cm in diameter) compared to the wavelength at 2.7 THz makes it complicated to do similar simulations. In the case of a convex lens in silicon, the directivity of the antenna is greatly increased.

### 6.6.2 Quasi-optical Coupling

As it has been mentioned earlier, an efficient solution to focus the incoming signal onto the HEB is the use of lenses in the same material as the HEB substrate (usually high resistivity silicon). Attaching a lens on the antenna substrate will increase the antenna gain on the dielectric side. The HEB substrate is glued to the flat side of the hemispherical lens. The dielectric lenses can be hemispherical, hyper-hemispherical, or ellipsoidal. The extended hemispherical lens system with a double-slot antenna, for instance, shows an increase of 10 dBi in directivity of the planar antenna [130][111]. At the crossing of two different optical or electromagnetic approach, the effect of the lens on the incoming signal can be seen through the ray optic formulation: the RF incoming signal is concentrated/focused on the antenna: it is mandatory that the antenna is carefully placed at the focus point of the lens to get an optimal coupling, a precision of few dozen $\mu$m is required. Ray-tracing techniques are used to compute the characteristic of the integrated lens antenna.
Figure 6.7 – Simulated beam patterns at 2.7 THz of the log-spiral antenna in the plane $\phi = 0^\circ$, $\phi = 45^\circ$, $\phi = 90^\circ$ when a) the antenna is surrounded by free space, b) for the case where the antenna is at the interface of the free space and an infinite dielectric half space with the effective refractive index $n = 2$ and c) for the case where the antenna is lying on silicon with $n = 3.42$. 

105
Figure 6.8 – Physical Optic simulations on Zemax of the extended hemispherical lens with the HEB and its planar antenna placed at the focal point of the lens the planar antenna. on dielectric as shown on figure 6.8.

The extension length L of the hemispherical lens is then optimized in the simulation depending on the required beam.

6.7 HEB developed at LERMA

The HEB used in this work is similar to the one used in [29]. The ultra-thin superconducting NbN film has been provided by the Russian company SCONTEL. The niobium nitride was selected as superconducting material. The 3.5 nm thick NbN film was deposited with reactive DC-magnetron sputtering at substrate temperatures of 700 °C on high purity silicon substrate. This first step in the fabrication process of the HEB is of main importance as it will further condition the performances of the devices. The further patterning of the HEB’s electrodes, antenna and bonding pads were performed at C2N Marcoussis by R. Lefevre and reported in [131]. The fabrication process of our HEB requires only electron beam lithography which gives a high flexibility to test different designs of antenna (double-slots or log-spiral antenna, operating frequencies). The HEB device has a critical current of 260 μA at 4.2K. The critical temperature $T_c$ is around 9.2 K. The DC resistance measured at room temperature is 74 Ω. For this work, a quasioptical coupling has been used for the HEB. The superconducting bridge is in the middle of a double helix log spiral planar antenna (see figure 6.9). The HEB bridge is 2 μm wide and 200 nm long and 3.5 nm thick. The advantages in the use of the log spiral antenna is that they couple with circular and linear polarization radiations. The log-spiral antenna is almost frequency independent for a broad frequency range. The frequency range of the
Figure 6.9 – Scanning electron microscope pictures of a niobium nitride (NbN) HEB on silicon substrate. Picture a) shows the double helix log spiral antenna and picture b) shows the micro-bridge with a passivation layer on the top. (photos from Roland Lefevre)

The spiral antenna actually depends only on the outer and inner diameter of the spiral. The HEB chip is further coupled to an hyper-hemispherical HR-Si lens by gluing the HEB substrate on the back of the lens. The lens with HEB is then mounted into a copper block with the bias circuits and the SMA connector for the IF signal extraction. The IF bandwidth is about 3 GHz. A similar device as the one used in this work has shown an uncorrected DSB noise temperature of 790 K at a radiation frequency of 2.5 THz which is so far the state of the art $T_{\text{noise}}$ in this frequency range [29].
Chapter 7

Heterodyne measurements

This chapter presents the heterodyne measurements with QCLs as local oscillators (LO) operating at 2.7 THz and the NbN HEB presented in the previous chapter for the mixing. A large number of measurements have been performed to characterise the receiver noise temperature using the Y-factor method. We also performed beat measurements between QCL and an amplifier-multiplier chain in order to characterise the signal emitted by the QCL. This chapter will conclude with the demonstration of an original solution to integrate the QCL as LO in a heterodyne detection system. A novel optical coupling scheme is demonstrated where the LO signal is coupled through the air side of the planar HEB antenna, while the signal to be detected is still coupled through the lens. In this way, there is no use of diplexers to overlay the RF and LO signal as it is usually employed in conventional heterodyne detection system. A very low double side band receiver noise temperature below 1000 K at 2.7 THz has been obtained thus confirming the potential for ultra-high sensitive heterodyne detection of this scheme.

7.1 Sensitivity of a heterodyne receiver system

The minimum difference of temperature $\Delta T$ which can be detected in an integration time $\tau$ is expressed with the radiometer equation (7.1). $\Delta T$ is function of the receiver equivalent noise temperature, the integration time $\tau$ and the IF bandwidth $\Delta v_{IF}$ [132]:

$$\Delta T = \frac{T_{rec}}{\sqrt{\Delta v_{IF} \tau}}$$  \hspace{1cm} (7.1)

From the equation above, $\tau$ is proportional to $T_{rec}^2$ for a fixed $\Delta T$, hence one needs to reduce the receiver equivalent noise temperature to reduce the observation time.

The overall equivalent noise temperature of the heterodyne detector system can be
Chapter 7. Heterodyne measurements

evaluated by summing the noise contribution of all its elements. In this model, every component of the detection chain is modelled by a noise-less amplifier with a noise at its input. For an incoming signal power $P_{in}$ and $P_i, G_i$ the input noise and gain for the $i$ component so an input signal is amplified to $P_{out} = G(P_{in} + P_{noise})$, the output power of the overall receiver is written:

$$P_{out} = P_{in} \times G_1 \times G_2 \ldots \times G_n + P_1 \times G_2 \ldots \times G_n + P_2 \times G_3 \ldots \times G_n + \ldots + P_{n-1} \times G_n \quad (7.2)$$

With two known incoming signal’s powers and the corresponding output power from the heterodyne receiver, as the noise contribution from each element is independent from the input signal, it is possible to deduce the receiver gain $G_{sys}$ but also the noise’s power contribution of the receiver $P_{noise}$. The calibration of the receiver sensitivity is performed by measuring the IF output power while 2 black bodies are applied alternatively as RF signal sources at the input of the receiver: a cold load at liquid nitrogen temperature 77 K and a hot load at room temperature 300 K. The IF output power is then for the cold or hot load:

$$P^{hot/cold}_{IF} = G_{sys}(P^{hot/cold}_{in} + P_{noise}) \quad (7.3)$$

From (7.3), the unknown parameters in the measurements of the IF output power are $G_{sys}$ and $P_{noise}$. The consecutive measurements of $P^{hot}_{IF}$ and $P^{cold}_{IF}$ are then sufficient to evaluate $P_{noise}$. This calibration method is known as the Y-factor method. The Y-factor is defined as the ratio between the IF output power when the hot and the cold blackbody are presented at the input of the receiver:

$$Y = \frac{P^{hot}_{out}}{P^{cold}_{out}} = \frac{G_{sys}(P^{hot}_{in} + P_{noise})}{G_{sys}(P^{cold}_{in} + P_{noise})}$$

$$P_{noise} = \frac{P^{hot}_{in} - Y.P^{cold}_{in}}{Y - 1} \quad (7.4)$$

The radiant power emitted per mode over a bandwidth $B$ from the hot and cold load can be calculated with Planck’s law formula:

$$P_{Planck}(\nu, T_{phys}) = \frac{h\nu}{e^{h\nu/k_BT_{phys}} - 1} \quad (7.5)$$

where $k_B$ is the Boltzmann constant, $h$ the Planck constant, $T_{phys}$ the physical temperature of the black body and $\nu$ the frequency. When $h\nu << k_BT$ the radiant power per mode of
the black body can be approximated by the Rayleigh-Jeans approximation. The radiant power emitted per mode over a bandwidth $B$ is then written $P_{RJ} = k_B T B$. For the frequency range of 2 to 5 THz, $\nu$ is in the order of magnitude of $k_B T$ for temperature range of 10 K - 300 K. It has been shown in [133], when using 77K and 300K black bodies for the measurements of the $Y$-factor above 1 THz, the Planck law (eq. (7.5)), for the hot and cold load equivalent power calculation, is inappropriate, and results in receiver noise temperatures higher by half a photon energy $\frac{\hbar \nu}{2}$. The dissipation-fluctuation theorem given by Callen and Welton ([134]) takes into account this noise fluctuation and is written:

$$P_{CW}(\nu, T_{\text{phys}}) = k_B T B \frac{\frac{\hbar \nu}{k_B T}}{e^{\frac{\hbar \nu}{k_B T_{\text{phys}}}} - 1} + \frac{\hbar \nu B}{2} \quad (7.6)$$

Using (7.6), an effective Callen and Welton temperature can be determined by $T_{CW} = P_{CW} k_B B$, where $k_B$ is the Boltzmann’s constant. $T_{CW}$ is further written:

$$T_{CW} = T \frac{\hbar \nu}{k_B T} \frac{1}{e^{\frac{\hbar \nu}{k_B T_{\text{phys}}}} - 1} + \frac{\hbar \nu}{2k_B} \quad (7.7)$$

The noise temperature can be deduced from the noise power transformation to Callen & Welton temperature (7.7),

$$T_{\text{noise}} = T^{\text{hot}}_{CW} - Y T^{\text{cold}}_{CW} \quad (7.8)$$

The different noise temperatures for a black-body radiator at 2.7 THz are presented in figure (7.1). The $Y$-factor method is also used for the calibration before and during astronomical observations.

The overall heterodyne receiver noise temperature is the sum of all the noise contribution coming from mixer, from the optics and from the amplifier of the IF channel. $T^{\text{DSB}}_{\text{receiver}}$ is written by:

$$T^{\text{DSB}}_{\text{receiver}} = T_{\text{optics}} + \frac{T_{\text{mixer}}}{G_{\text{optics}}} + \frac{T_{\text{IF}}}{G_{\text{optics}} G_{\text{mixer}}} \quad (7.9)$$

In figure 7.2 the schematic of the signal path with the gains and noise contributions of the receiver components is presented.

where $G_{\text{optics}}$ is the gain of the optical components that connect the superconducting bolometer to the input of the receiver respectively at room temperature and at 4 K including the beam splitter, the dewar windows, the IR filter, etc. To get the intrinsic noise temperature of the mixer, the measured receiver noise temperature using the Y-factor
Figure 7.1 – Noise temperature as a function of the physical temperature for a black body radiator at 2.7 THz, according to the Planck, Rayleigh-Jeans and Callen-Welton formulas.

Figure 7.2 – Block diagram of the heterodyne receiver showing the various noise contributions and gain of the optics guiding the RF signal, the mixer which down-converts the RF signal to the intermediate frequency and the IF chain.
7.1. Sensitivity of a heterodyne receiver system

method needs to be corrected from the noise temperature contributions of all optical components and the IF chain. $G_i$ of each optical element can be determined experimentally using a Fourier Transform Spectrometer (FTS) at room temperature. In this way also the total $G_{opt}$ is determined, multiplying all individually measured optics losses. Once the gain $G_i$ is known, the equivalent input noise temperature of passive components, for example optics and filters, can be calculated by:

$$ T_{optics} = \frac{1 - G}{G} T_{phys} $$

(7.10)

where $T_{phys}$ is the physical temperature or the Callen and Welton temperature $T_{CW}$ of the optical components. We have investigated through electromagnetic simulations on FEKO software the transmission and reflectivity of Mylar film with different thickness used as beam splitter which is presented later in this section. The transmission spectra of high-density polyethylene (HDPE) used for the dewar windows or Zitex used as infra-red filter can be found in [135] or [136].

**Beam splitter**  Beam splitters are used to split an incident beam into two output beams and are crucial elements for heterodyne measurements, most of interferometers or time domain spectrometers. There are different kinds of beam splitters according to the needs of the experiments. For heterodyne measurements in the THz frequency range, the beam splitter is usually constituted with a dielectric material with high transmission coefficient in the THz frequency regime, as silicon or mylar films for example. Beam splitter could be obtained with a wide band beam polariser constituted of thin metallic parallel wires in similar way as it is used for the Martin-Puplett interferometer diplexer [137]. Other beam splitters are formed by a thin layer of conductors deposited on a dielectric substrate. All the above techniques of beam splitters show on the other hand different characteristics depending on the polarisation of the incoming beam either in the plane perpendicular or in the plane parallel to the plane of incidence. The figures 7.3a, 7.3b, 7.3c and 7.3d show the simulated reflection and transmission coefficient for radiation parallel to the plane of incidence ($E_P$) or perpendicular to the plane of incidence ($E_S$) with an incidence angle of 45° and for a Mylar film (refractive effective index $n = 1.8$ at 2 THz) with a thickness of 50$\mu$m and 13$\mu$m. The transmission and reflexivity is calculated with the Fresnel coefficient and the imaginary part of the effective refractive index of the material. Reflectivity and transmission of plane waves on the beam splitter depend a lot on the incidence angle. The thickness of the film is also critical as the presence of two interfaced air-dielectric and dielectric-air produce constructive and destructive interferences: figure 7.3 shows that transmission and reflectivity are
dramatically changed depending on the thickness of the Mylar and the frequency either for the $E_S$ and $E_P$ polarised beams.

7.2 Direct detection with the HEB

The HEB mixer can also be used as a direct detector. This is an issue for the heterodyne receiver calibration as the input radiant power is changed from the hot and cold blackbody load for a constant LO radiant power (cf Y-factor method). Due to the high sensitivity of the HEB, as the mixer gain is a function of the bias current, a change in the I-V characteristic of the HEB between the hot and cold load will modify the IF output power: with the hot load, the bias current of the HEB is smaller (the HEB is better pumped) than with the cold load which affects the mixer gain, and then the IF’s output power with the hot load. A clear evidence of the direct detection is shown on the voltage-current measurements on a HEB pumped with a LO and with a cold (77 K) and hot (300 K) blackbody at the input of the mixer in figure 7.4.

The direct detection effect may falsely decrease the equivalent noise temperature using the Y-factor method. The change of the bias current leads to a change in the mixer gain and falsely reduces $T_{rec}$. The direct detection can be compensated by adjusting the LO power level when changing the loads to keep the DC bias current identical for each load input. The intensity of the direct detection depends on the temperature of the black body used for the calibration and on the bandwidth of the HEB mixer.

7.3 Linewidth characteristics of THz QCL

The work presented in this section is focused on the study of the intrinsic linewidth of a free running QCL and is a step forward to achieve the phase locking of the QCL with a frequency reference. The THz QCLs used for this experimentation have been presented in chapter 4, that is to say a 3rd-order DFB QCL which shows single mode operation characteristics in continuous wave operation at 4K. The THz QCL produces $\sim 1$ mW at 2.7 THz at 4K. The threshold current of the QCL is approximately 30 mA for a voltage of 5.5 V leading to a DC power dissipation lower than 200 mW which allowed a compact integration of the QCL with the superconducting detector in the same cryostat. The testing scheme used is based on the heterodyne principle to down-convert the THz radiation of the QCL under study to the intermediate frequency by the mixing with a precisely known frequency. The mixer (the detector) used here is the hot electron bolometer as presented in chapter 6 and the local oscillator an Amplifier Multiplier Chain (AMC) from Virginia Diode Inc. (VDI), a solid state source based on a microwave
7.3. Linewidth characteristics of THz QCL

Figure 7.3 – Calculated a) reflection and b) transmission coefficient for an incident plane wave at 45° with polarisation in the plane of incidence ($E_P$) and perpendicular to the plane of incidence ($E_S$) for mylar thickness of 50$\mu$m. c) and d) Reflection and transmission coefficient respectively for an incident polarized $E_P$ and $E_S$ plane wave at 45° for mylar thickness of 13$\mu$m.
Figure 7.4 – I-V curves of a pumped HEB at 2.7 THz with a QCL as LO and with hot and cold black body load: the HEB is more pumped with the hot black body compared to the situation with the cold black body.

frequency amplifier and a cascaded series of frequency multipliers at 2.7 THz (generically called Amplifier-Multiplier-Chain) with an output power $\sim 2\mu W$ (figure 7.5) [118]. The QCL was mounted on a copper block with a HCW as shown in chapter 5. The quasi-optical coupling of the QCL with the HEB is made in situ by adapting the HCW aperture at the air side of the HEB planar antenna, on the opposite to the silicon lens. Further information about using this experimental scheme is presented in the next section of this chapter. This coupling scheme allows to remove the beam splitter (beam diplexers) from conventional heterodyne scheme. The nearly Gaussian beam of the AMC’s waveguide allowed an ideal coupling with the HEB’s lens, and the AMC chain was itself sufficiently powerful to fully pump the HEB without particular effort on the alignments. The QCL THz signal is here the signal to be detected while the AMC’s signal is the local oscillator which pumps the HEB into its heterodyne state. The intermediate frequency IF $|\nu_{QCL} - \nu_{LO}|$ produced on the HEB by the mixing of the signals of the QCL and the AMC is amplified by two amplifiers, one cryogenic placed just at the IF output of the HEB and one outside the cryostat at room temperature for a total amplification of 80 dB. The intermediate frequency bandwidth of the NbN HEB being limited to 2-3 GHz, the QCL and AMC’s emission frequencies have to be separated by less than this limitation so that $|\nu_{QCL} - \nu_{AMC}|$ is within the IF bandwidth. As the QCL mode’s frequency has been measured with a precision of 0.25 $cm^{-1}$ on a Fast Fourier Transform Spectrometer (FTS),
7.3. Linewidth characteristics of THz QCL

Figure 7.5 – Schematic of the experimental set-up showing the THz 3rd-order DFB QCL and the solid state source with a frequency emission close to that of the QCL.

the tuneability of the solid state source which is driven by a frequency synthesiser allows to bring $\nu_{AMC}$ close enough to $\nu_{QCL}$ to observe the beat note.

The I-V curve of the HEB is presented on figure 7.6. The HEB was pumped to its most sensitive state with the help of both QCL and AMC radiant power. The bias point of the HEB was fixed, usually just after the negative resistance region on its I-V characteristic. The QCL was driven in continuous wave operation far above threshold. A scan of the beat note on a spectrum analyser is recorded using different resolution bandwidth (RBW) and video bandwidth (VBW). The beat note is shown in figures 7.7. Thermal and electrical fluctuation limits a long time analysis (sweeping time $\sim$10 ms) of the beat note because the narrow instantaneous free running QCL line randomly fluctuates on a wide band (several MHz). Stabilizing the QCL temperature and the bias current can limit these low frequency fluctuations. Repeating several times the beat note measurements and averaging the results, we have measured a beat note with a linewidth $\Delta \nu_{QCL} = 120 kHz \pm 10 kHz$. This value represents an average value of the linewidth of the QCL. The linewidth of the beat note can be mainly attributed to the QCL linewidth as the AMC spectral purity is given by the ultra stable, low phase noise frequency synthesiser. We have been able to measure precisely the QCL single mode frequency at $\nu_{QCL} = 2.66437417$ THz. The measured QCL linewidth can be compared with the laser intrinsic linewidth found with the Schawlow-Townes linewidth limit as expressed in chapter 2. Here the QCL frequency is $\nu_{QCL} = 2.66437417$ THz, the output power was $\sim$850 $\mu W$ measured with a Thomas Keating absolute power meter. The cavity lifetime $\tau_c$ is expressed by $\tau_c = \frac{2\pi n_{eff}}{\alpha c}$ with $\alpha$ the mirror losses taken to be 5.5 cm$^{-1}$ and the effective index $n_{eff}$=3. The upper and lower radiative transition population factor is chosen to be neglected. The intrinsic linewidth is calculated to be in the order of $\Delta \nu_{QCL} \sim 4 kHz$. At
Chapter 7. Heterodyne measurements

Figure 7.6 – I-V curves of the superconducting HEB mixer at 4 K for different radiant power of the local oscillator and the QCL.

Figure 7.7 – Heterodyne beat note of the QCL with the solid state AMC for a resolution bandwidth of 10 kHz and video bandwidth of 10 kHz. The sweep time for this scan is 3 ms.
the time this work was done, the laboratory is developing the system in order to operate a phase lock loop of the QCL beat note with the AMC.

7.4 Noise temperature measurements

The experiment described below allows the characterisation of the double side band noise temperature of the receiver $T_{DSB,rec}$ in LO frequency range $\nu_{LO}$ of 2.7 THz. The receiver is composed of the HEB for the mixing and the QCL as LO. The test set-up for measuring noise temperature $T_{rec}$ is presented in figure 7.9. The QCL was mounted in continuous flux liquid helium dewar at 4K. An elliptical metallic mirror with one focal length of 45 cm and a second focal length of 7 cm is used to focus the QCL beam onto the HEB mixer. The QCL used here was a surface emitting QCL with graded heterostructure resonator demonstrated by Xu et al. [33]. The electro-optic characteristic of the 2nd-order DFB QCL is shown in figure 7.8. The laser was operating in CW mode and presented a single mode operation at 2.7 THz with an output power in the mW range. The other specificity of this 2nd-order DFB is its single-lobed far-field beam patterns, which allows a nicer quasi-optical coupling with the HEB. The NbN HEB ($T_C \approx 9.2$ K) was mounted on a separate dewar at 4 K and was similar to the HEB used for the heterodyne measurements reported in [29] where the state of the art receiver noise temperature was obtained at 2.5 THz. The mixer consists of a double helix log spiral gold patch antenna with a superconducting NbN HEB processed on a silicon substrate. The NbN micro bridge was 2 $\mu$m long and 200 nm wide. The silicon substrate with the HEB and its antenna was glued at the back of a hyper-hemispherical silicon lens with an anti-reflection coating designed for 2.5 THz. The low-noise cryogenic amplifier is directly attached to the IF output of the HEB, on the same cold plate in the dewar. The first and second stage amplifier at room temperature have an amplification gain of 43 dB for the 0.5-4 GHz frequency band. The out-put power of the IF is measured with a power-meter. In addition to the optical elements, loss was induced by the optical path under atmosphere conditions from the RF sources to the mixer cryostat window. In our test setup, the distance between the cold blackbody and the mixer was about 25 cm and the distance between the LO and the mixer was about 55 cm. The noise contributions of the different optical elements and the optical path are summed up in the table 7.1.

The dewar window in HDPE was made with a thickness of 2 mm. The RF and LO signals were overlaid using a beam splitter in mylar with a thickness of 25 $\mu$m. The QCL maximum output power was about 2 mW but due to the atmosphere absorption and Mylar reflection of TM polarisation, only a fraction of this available power is actually coupled to the superconducting bridge. We can adjust the LO power coupled to the HEB
### Optical losses

<table>
<thead>
<tr>
<th>Components</th>
<th>Loss (dB)</th>
<th>Temperature (K)</th>
<th>Noise temperature contribution (K)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Beam splitter (Mylar 13 μm)</td>
<td>0.9</td>
<td>300</td>
<td>69</td>
</tr>
<tr>
<td>2 mm HDPE windows</td>
<td>1.25</td>
<td>300</td>
<td>100</td>
</tr>
<tr>
<td>Zitex G 108</td>
<td>0.80</td>
<td>4</td>
<td>1.0</td>
</tr>
<tr>
<td>QMC filter</td>
<td>0.75</td>
<td>4.2</td>
<td>1.4</td>
</tr>
<tr>
<td>25 cm optical path</td>
<td>0.87</td>
<td>300</td>
<td>101</td>
</tr>
</tbody>
</table>

Table 7.1 – Optical components with their respective calculated losses.

Figure 7.8 – a) Spectrum and b) farfield measurement of the 2nd-order DFB QCL used as LO for the heterodyne measurements (courtesy of G. Xu for the measurements of the farfield).
by a fine tuning of the optical alignment between the QCL and the HEB or directly by tuning the bias current of the QCL (figure 7.10a). The isothermal method [96] allows to estimate that $\sim 270$ nW is actually absorbed by the HEB when it is pumped to its optimum sensitive state. The receiver sensitivity is measured using the Y-factor method which has been presented in 7.1. The Callen-Welton temperatures of the black body at 2.7 THz are $T_{\text{cold}} = 91$ K and $T_{\text{hot}} = 295$ K ($T_{\text{phys}} = 77K$ and $300K$ respectively). The intermediate frequency output power was swept as a function of the bias voltage of the HEB while the QCL was driven in continuous wave operation. The sweep step is 0.01 mV from -5 to 5 mV. At every bias point the DC current through the HEB and the IF power at the receiver output are measured. The radiant output power of the LO was then fixed but with small fluctuations due to the free running, not stabilized operation of the QCL. As mentioned in the section 7.1, the direct detection effect between the hot black and cold body induces an error on the Y-factor measurement. The bias current has to be the same for the hot and cold load to compensate this effect. Experimentally, the QCL output power needs to be slightly tuned down when measuring the hot black body, i.e. by adjusting the QCL bias voltage so that the I-V curves for the hot load and cold load were super-imposed. The I-V curves of the HEB at 4K with and without the LO are presented in figure 7.10a and the noise temperature measurements together with the IF output power responding to the hot and cold load are shown in 7.10b. The best noise temperature $T_K$ was found to be 2000 K for a bias voltage around 0.5 mV. A mixer noise temperature with a similar HEB mixer for the frequency range of 2.5 THz has been measured to be $T_{\text{mixer}} = 790$ K [29] where the LO source was an optically pumped far infrared laser adjusted at 2.5 THz and its output optical power was regulated by a rotating wire grid. The optical path was also purged. We can further apply corrections of optical losses summed up in 7.1. We can re-write equation (7.9) into:

$$T_{\text{mixer}} = G_{\text{optics}}(T_{\text{DSB\ receiver}} - \frac{T_{\text{IF}}}{G_{\text{mix}} G_{\text{optics}}} - T_{\text{optics}})$$

(7.11)

$T_{\text{DSB\ receiver}}$, $G_{\text{optics}}$, $G_{\text{mixer}}$ are known and $T_{\text{IF}}$ has been measured to be $\sim 6$ K. We can then determine the mixer noise temperature $T_{\text{mixer}}$. $T_{\text{optics}}$ have been estimated at $\sim 200$ K. Using (7.11), we find $T_{\text{mixer}} \sim 900$ K which is near the previously established result cited above. This measurement proved that a free running THz QCL could be experimentally used as a LO for a heterodyne measurement with HEB and provide with fairly good performances compared to the FIR pumped gas laser set-up with much more compact features. The last result (receiver noise of 2000 K at 2.7 THz) is not optimised and could be improved: the cold and hot load could be integrated into the mixer dewar to avoid optical losses, the beam splitter used could be thinner (3 $\mu$m) to favour the transmission

121
Chapter 7. Heterodyne measurements

of the RF signal.

7.5 Heterodyne measurements with a novel quasi-optical coupling system

The planar antenna presented in chapter 6 shows a non-symmetrical radiation pattern depending on the effective refractive index of the antenna’s substrate. For a substrate in HR-Si substrate ($n = 3.42$), the antenna’s gain in the air side is 10 dB lower than on the dielectric side. The use of silicon lenses enhances the directivity of the antenna through the dielectric. Nevertheless, we have investigated the possibility to couple the LO power to the planar antenna through the air side to avoid using the beam-splitter with the aim of achieving highly compact receiver configurations for future embedded observation missions. This study was motivated by the results obtained with the 3rd order DFB QCL coupled to a HW, giving a collimated nearly Gaussian beam with a directivity of 55 dBi and by the low power dissipation of the newly developed QCL. This section presents a novel quasi-optical concept where the LO and RF signals are coupled to the HEB from both sides of the planar antenna: LO from the air side and RF from the silicon lens. Due to the poor coupling efficiency to the antenna through the air side, it is important to estimate the optimal beam shape of the LO.

7.5.1 Quasi-Optical Coupling with the Planar Antenna

Another key parameter of the double-helix log-spiral planar antenna needs to be taken into account to optimize the quasi-optical coupling the LO and RF signals on the HEB: the effective aperture $A_{\text{eff}}$. The antenna’s aperture is the surface through which the power is radiated or received. This parameter is important when determining the waist dimension of the incoming Gaussian beam which needs to be coupled to the antenna. The effective aperture of an antenna is given by [70].

$$A_{\text{eff}} = \frac{\lambda^2}{4\pi}G$$  \hspace{1cm} (7.12)

where $G$ is the directional antenna gain considered. The aperture efficiency of an antenna determines the amplitude of the power received by the antenna which depends on both the distribution of the energy radiated by the source and the beam pattern of the receiving antenna. For an extended source of energy, and especially if the angular dispersion of the beam is larger than the main lobe of the antenna, the coupling efficiency of the incoming beam to the antenna needs to be optimised. The coupling efficiency can be calculated using a normalized inner product between the antenna beam pattern and the incoming
Figure 7.9 – Experimental set-up for heterodyne sensitivity measurements with HEB and QCL.
Figure 7.10 – Heterodyne measurements with 2\textsuperscript{nd}-order QCL and HEB a) I-V curves of the HEB for different level of $P_{LO}$ coupled to the HEB and b) output power of the intermediate frequency measured with a power-meter for two different black body temperatures and noise temperature of the overall heterodyne detector calculated with the Y factor technique.
radiation pattern resolved in the angular domain \([124][138]\) and is expressed by:

\[
\eta_{\text{coupling}} = \left| \frac{\Psi_{\text{antenna}}}{\Psi_{\text{beam}}} \right| \tag{7.13}
\]

where \(\Psi_{\text{antenna}}\) and \(\Psi_{\text{beam}}\) are the normalised patterns of the planar antenna and the incoming beam. This calculation is similar to the one presented in chapter 5 for the coupling efficiency between an incoming Gaussian beam and a HW. Following the calculus in \([139]\) and introducing the aperture radius \(a\) and the beam waist \(\omega\), the optimal ratio to get the maximum coupling efficiency is when \(\frac{\omega}{a} = 0.6435\) which gives \(\eta_{\text{coupling}} = 0.9792\). The later result is only true at the aperture as the beam may diverge and will not preserve the properties of a Gaussian beam. Nemoto et al. \([140]\) verified that for the ratio \(\frac{\omega_m}{\lambda} > 0.9003\) where \(\omega_m\) is the beam waist at the aperture and \(\lambda\) the wavelength, the Gaussian beam preserves its properties and the above relations can be used. The beam waist of the 3\textsuperscript{rd}-DFB QCL coupled to the HW has been estimated in the chapter 5 with \(\omega_x = 5.32\) mm and \(\omega_y = 5.28\) mm. The planar antenna’s gain on its air side was estimated through FEM calculation at 0.71 dB which gives an effective area of 0.7 mm\(^2\) using the equation (7.12) for the antenna’s aperture calculation in its focal plane. Since our beam area is 35 time wider, alignment is relatively easy, but the signal to noise could be significantly improved with a more focused beam pattern. For the ease of the alignment inside the cryostat, we first focus the QCL in the HW following the optical scheme found in chapter 5 and further adapt the end aperture of the HW to the HEB: the wide bore diameter (4 mm) of the HW allows to easily cover completely the HEB antenna’s radiation beam and doesn’t require a precise alignment.

The heterodyne experimental set-up for the double-side band noise temperature \(T_{DSB}\) measurement is shown in figure 7.11. The coupling of the RF radiant power to the HEB was made with standard technique using a hyper-hemispherical HR-Si lens with an anti-reflection coating designed for 2.5 THz. The QCL with HW and the HEB is then mounted in a pulsed tube cryo-cooler operating at 4 K. In this way, the beam splitter to overlay the RF and LO signals on the HEB wasn’t used, the signals were directly coupled to the mixer. Despite the low DC power dissipation of the 3\textsuperscript{rd}-order DFB QCL, the cooling capacity of the pulsed tube wasn’t sufficient for the cold plate of the dewar to remain at 4 K with the QCL switched on and biased at sufficient voltage to pump the HEB. Therefore, the QCL copper block was physically mounted on the 4 K cold plate with the HEB but was thermally decoupled from it with pieces of HDPE beneath the QCL block for the thermal insulation. We have used the cooling capacity of the 50 K stage of the cryo-cooler instead to cool down the QCL using a specially designed cold finger: the QCL was operating at nearly 60 K for the whole experimentation. Despite our effort, when the
Figure 7.11 – Schematic picture of the measurement setup for the ultra compact heterodyne receiver with QCL as LO and HEB as mixer. Notice the absence of beam diplexer for the overlaying of the RF and LO signals. The incoming signals are symmetrically coupled on the HEB bridge and the QCL can be compactly integrated into the cryostat with the mixer block.
Figure 7.12 – HEB current versus voltage characteristics with different level of LO pumping. The HEB’s temperature was 6.1 K without QCL and was 6.49 K when the QCL was switched on.

QCL was operating in CW, the HEB cold plate was heated up to 6.49 K. Nevertheless, the HEB was still below the $T_C$ and could still perform heterodyne mixing. The QCL used as LO has been presented with its optical and electrical characteristics in the chapter 4 and was driven in CW operation. The QCL was not frequency stabilized using PLL neither power stabilised using PID circuits as in [38]. The HEB mixer’s IF output signal is extracted through a bias-T and amplified with a cryogenic HEMT amplifier from Caltech with a amplification bandwidth from 0.5-4 GHz and followed by a room temperature amplifier. We use the voltage current curve of the HEB to monitor the LO pumping level (figure 7.12). We were able to pump the HEB into a sensitive state and to completely suppress the negative resistance where the characteristic of the HEB is perturbed (cf figure 7.12). The LO power is fixed while the HEB’s bias voltage is swept in 0.015 mV steps from 0 to 3 mV. At every voltage point the DC current through the HEB and the IF power are recorded. In figure 7.12, the I(V) curves for an un-pumped (no LO power), optimum, under-pumped and over-pumped LO power levels are shown.

The receiver noise temperature of the overall system has been measured using the Y-factor technique. It is exactly the same measurement as depicted in the previous section. The receiver noise temperature was calculated with the equation (7.8). For the under and over pumped levels of the detected power, the noise temperature and current are shown versus the bias voltage in figure 7.13. For the situation where the HEB is over pumped,
Figure 7.13 – HEB mixer for two different pumping levels: under pumped and over pumped. In a) IF output power in $\mu$W as a function of the bias Voltage, the red curves correspond to the measurements of the hot and the blue curves of the cold blackbody. b) the DSB receiver noise temperature as a function of bias Voltage.
the response of the HEB is reduced (lower IF output power figure 7.13b). The equivalent receiver noise temperature is between 1000 K-200K for the voltage bias range from 0.5 to 1.0 mV and rapidly increased beyond the bias voltage of 2 mV. Similar behaviour on the $T_{\text{rec}}$ is observed for the situation of the under pumped level. The minimum low noise temperature at 2.7 THz has been recorded for the optimal pumping level with receiver noise temperature $T_{\text{rec}} \sim 900K$ for HEB bias voltage around 0.6 mV (figure 7.14b). For a wide bias range from 0.5 to 2 mV a nearly constant $T_{\text{rec}}$ can be achieved which is typical of NbN mixers. This uncorrected noise temperature is relatively close to the state of the art in [29]. To account for the optical losses, only the 2 mm thick HDPE dewar window, the zitex infrared filter and the optical path of only few cm have to be taken into account which give $T_{\text{mix}}$ similar to the one in [29] (790 K). Nevertheless, this experimental set-up is much compact as the LO and mixer were integrated in the same dewar, without the use of beam splitter.

In conclusion, we have introduced an original optical scheme to realize THz heterodyne measurements. The LO is a 3rd-order DFB THz QCL, and the mixer is a HEB without using a beam splitter. A receiver noise temperature $T_{\text{rec}}$ as low as 881 K has been obtained. The result could be further enhanced with a better cooled HEB as all the above measurements where performed at 6.49 K, better sensitivity could be expected for a HEB temperature kept at 4K as the mixer noise should be reduced. These results open a new horizon for THz coherent compact and sensitive imagery in future space missions.
Figure 7.14 – a) Measured DC-IV curve for the NbN HEB with an optimal LO power for a room temperature and 77 K radiating black bodies. b) Measured DSB receiver noise temperature as a function of the voltage bias of the HEB versus bias voltage with an IF bandwidth of 100 MHz centred at 1.5 GHz.
Chapter 8

Conclusion and Future Work

8.1 Conclusion

In this thesis, we have developed a compact and ultra-sensitive heterodyne receiver. The first building block of our heterodyne receiver is the QCL used as local oscillator. We have first worked on the design and the fabrication of 3rd-order DFB THz QCL. We have obtained devices with single mode operation at the targeted frequency with hundreds of $\mu W$ of output power in continuous wave operation. The devices have a DC power dissipation lower than 200 mW which make them suitable for a compact integration in a cryogenic dewar. Furthermore, we have achieved 3rd-order DFBs with a perfectly phase matching and their beam patterns have a FWHM $\leq 10^\circ \times 10^\circ$.

In a second time, we have worked on the beam shaping of THz QCLs using dielectric hollow core waveguides. We have designed a specially customized copper block mount to adapt the QCL facet at the rear aperture of a hollow core glass waveguide with a bore diameter of 4 mm. The beam patterns of the phase-matched and non phase-matched 3rd-order DFB QCL are significantly improved with the HW. The resulting beams had a more symmetric and near Gaussian shape with a FWHM reduced to $\sim 5^\circ \times 5^\circ$. We have also studied different ways to couple the QCL beam into the HW. When the optimal ratio between QCL beam waist and HW bore diameter is reached, the QCL light is coupled to the lowest order hybrid mode of the HW and can propagate with low losses. Furthermore, the beam direction is now more predictable (along the HW direction) and so the optical alignment with other components such as mixers or detectors in-situ is simplified.

We have then implemented our 3rd-order DFB QCL into a compact heterodyne receiver. The low power dissipation of the device permits its integration in the same dewar as the HEB. Upon the sufficient QCL output power, the high directivity of the beam with the use of HW, we have proposed a novel coupling scheme for the heterodyne measurements where the QCL signal is coupled through the rear side of the HEB while the RF
signal is still coupled through the HEB silicon lens. This permits to avoid the use of a
beam splitter to overlay the LO and RF signals. Receiver noise temperature as low as
881 K has been recorded. This LO coupling method doesn’t reduce the performance of
our mixer. On the contrary, it suppresses the equivalent noise temperature brought by
the beam splitter and simplifies the heterodyne measurement scheme. The compactness
is also much increased.

8.2 Future Work

A lot of work still needs to be done. HEB receivers are extremely sensitive to instabilities
in the received LO power, it is mandatory to stabilise the QCL phase with a phase lock
loop (PLL) or amplitude stabilisation systems in order to enhance the receiver sensitivity
and spectral resolution. One idea which could be explored for the phase locking of the
QCL would be to divide the IF channel in two parts. The first IF channel would be
used to perform the heterodyne measurements (spectroscopy measurements, black body
calibration...) while the second IF channel would be used to recover the beat note between
the LO and an ultra-stable AMC source. The resulting beat note could be used as input
in a phase lock loop system. Also we could further use a voice coil attenuator with a
proportional–integral–derivative (PID) feed-back loop to stabilize the power output of
the QCL.

Another challenge for HEB heterodyne receiver is to have a multi-pixel receiver. Up to
now, only a modest number of pixels (7 or 14 pixels for SOFIA) have been developed. HEB
receivers with large number of pixels are complicated above 2 THz because of the lack
of sufficiently powerful LO to pump the array and efficient techniques for quasi-optical
coupling between LO and HEB array. The development of a nano-fabrication process
using silicon dry etching techniques would further help to obtain special structures for
THz phase-gratings (beam dividers) where the standard micro-machining techniques can’t
give the sufficient precision.

For very large arrays (128 pixels or more), the development of THz dielectric trans-
mission line with T-junction power splitters could be an attractive solution for the LO
power coupling between QCL and HEB mixers.
Appendix A

Metal-Metal Waveguides Fabrication Process

This process was used in the fabrication processes of the M-M QCL.

1. Wafer-bonding step

- Cleaning the sample (aceton+isopropanol+N2 dry).
- Plasma O$_2$ during 15 minutes.
- Desoxidation: HCl:H$_2$O= 1:4 during 40''.
- Evaporation of Ti/Au = 5 or 8 /500nm.
- Waferbonding.
- Dicing.

2. Substrate removal

- Substrate polishing.
- Removal substrate : acid citric etching.
- Stop layer etching : RIESTS O$_2$ plasma for 10' + HCl:H$_2$O (1:1) 2' + HF 2'.

3. Wet etching of the doped layer

- Shallow etch mask : Etching of the doped layer
  - Dehydration 3' at 110°C.
  - Spin coating: HMDS+ Shipley S1805.
  - Bake 3’ at 110°C.
  - Lithography for edge resist removal : soft contact 20”+5” break+20”
  - Development MF319 60”
  - Shallow etch mask : hard contact (HC) 4”
Appendix A. Metal-Metal Waveguides Fabrication Process

- Development MF319 15" (check under the microscope)
  - Plasma O$_2$ 3'

4. SiN deposition for insulation of the contact pad

- Plasma O$_2$ 3'.
- Desoxidation HCl:H$_2$O 1:4.
- PECVD SiN deposition 300 nm.
- Mesa definition
  - Dehydration 3' at 110°C.
  - Spin coating: HMDS + Chipley S1818
  - Bake 3' at 110°C.
  - Lithography for edge resist removal: soft contact 20" +5" break +20".
  - Development MF319 60"
  - Mesa definition: Hard contact mask 12".
  - Development MF319 15" (check under the microscope)
- Plasma O$_2$ 4'.
- Desoxidation HCl:H$_2$O 1:4.
- RIE plasma etching.

5. Top Contact Deposition

- Top contact pattern definition
  - Dehydration 3' at 110°C.
  - Spin coating: HMDS + MicroChemicals AZ5214.
  - Bake 3' at 110°C.
  - Lithography for edge resist removal: soft contact 20" +5" break +20".
  - Development AZ400K:H$_2$O = 1:4 60".
  - Top contact mask: Low Vacuum Contact (LVC) 2".
  - Post bake 2’ at 120°C.
  - Flood exposure 40".
  - Development AZ400K:H$_2$O = 1:4 20-25".
- Plasma O$_2$ 3'
- Desoxidation HCl:H$_2$O 1:4.
• Top contact deposition Ti/Au = 5/250nm.
• Lift off with aceton – clean with isopropanol + N2 dry.
• Plasma O$_2$ 4’.

6. **ICP etch lithography**

• Ridges definition
  - Dehydration 3’ at 110°C.
  - Spin coating: HMDS+ MicroChemicals AZ9260.
    * 30’ at 6000 rpm with an acceleration of 3000 rpm/s.
  - Bake 4’30” at 110°C.
  - Lithography for edge resist removal: soft contact 45”+5” break+45”.
  - Development AZ400K:H$_2$0 = 1:4 300”.
  - Deep etch mask LVC 16”.
  - Development AZ400K:H$_2$0 = 1:4 180-240”.
  - Plasma O$_2$ 3’.
  - Desoxidation HCl:H$_2$O 1:4.

• ICP etching following the recipes described in [78]
  - Sample cleaning (aceton+isopropanol+N2 dry)
  - RIE plasma O$_2$ 4’.

7. **Back side polishing by polishing of the sample after bonded to a glass slide**

8. **Back side deposition Ti/Au = 5/250nm**

9. **Sample cleaning**

10. **RIE plasma O$_2$ 4’**
Bibliography


[37] Heiko Richter, Martin Wienold, Lutz Schrottke, Klaus Biermann, Holger T Grahn, and Heinz-Wilhelm Hübiers. 4.7-thz local oscillator for the great heterodyne


[45] Shun Lien Chuang and Shun L Chuang. Physics of optoelectronic devices. 1995. 2.2, 2.6, 2.6


144


[67] NK Dutta and GP Agrawal. Semiconductor lasers, 1993. 3.3


[70] Constantine Balanis. Antenna theory analysis and design, 1982. 3.5.1, 3.5.2, 7.5.1


Bibliography


[75] Lukas Mahler, Alessandro Tredicucci, Fabio Beltram, Christoph Walther, Jérôme Faist, Harvey E Beere, and David A Ritchie. High-power surface emission from terahertz distributed feedback lasers with a dual-slit unit cell. *Applied physics letters*, 96(19):191109, 2010. 4.1


148

Bibliography

[94] Andreas Patrovsky and Ke Wu. Substrate integrated image guide (siig)-a planar
dielectric waveguide technology for millimeter-wave applications. *IEEE transactions
on Microwave Theory and Techniques*, 54(6):2872–2879, 2006. 5.5

and high-temperature thz quantum-cascade lasers based on lens-coupled metal-

[96] Grégory Gay. *Mélangeurs à bolomètres à électrons chauds sur membranes fonction-
6.1, 1, 6.5, 7.4

[97] N Perrin and C Vanneste. Response of superconducting films to a periodic optical

[98] Daniel E Prober. Superconducting terahertz mixer using a transition-edge mi-

6.3.2

[100] Harald F Merkel, Erik L Kollberg, and K Sigfrid Yngvesson. A large signal model
for phonon-cooled hot-electron bolometric mixers for thz frequency applications. In
Laboratory, 1998. 6.3.2

6.3.2

[102] Yu P Gousev, GN Gol’Tsman, AD Semenov, EM Gershenzon, RS Nebosis,
MA Heusinger, and KF Renk. Broadband ultrafast superconducting nbn detector

[103] Boris S Karasik, William R McGrath, and Rolf A Wyss. Optimal choice of material
for heb superconducting mixers. *IEEE transactions on applied superconductivity*,
9(2):4213–4216, 1999. 1

[104] S Cherednichenko, P Yagoubov, K Il’in, G Gol’tsman, and E Gershenson. Large
bandwidth of nbn phonon-cooled hot-electron bolometer mixers on sapphire sub-
1


Bibliography


[124] Paul F Goldsmith et al. Quasioptical systems. Chapman & Hall, 1998. 6.6, 6.6.1, 7.5.1

[125] K Button. Chapter 10 of infrared and millimeter waves, 1983. 6.6.1


[135] JS Tydex. Co.,“thz materials.”. 7.1


[137] Fabien Defrance. *Instrumentation of a 2.6 THz heterodyne receiver*. Theses, Université Pierre et Marie Curie - Paris VI, December 2015. 7.1


151


Financements

Ces travaux de recherches ont été financés par le Centre National d’Etudes Spatiales, par l’Observatoire de Paris et par l’Université Paris-Sud. Les travaux ont été conjointement réalisés entre le Laboratoire d’Etudes du Rayonnement et de la Matière en Astrophysique et Atmosphères (LERMA) et au Centre de Nanosciences et de Nanotechnologies (C2N).
Publications

• *A Compact and Sensitive Heterodyne Receiver at 2.7 THz exploiting a novel quasi-optical HEB-QCL Coupling Scheme.*
  
  **F. Joint,** P-B. Vigneron, T. Vacelet, G. Gay, S. Pirotta, L. H. Li, A.G. Davies, E. H. Linfield, Y. Delorme, R. Colombelli

  Submitted

• *Advanced and reliable GaAs/AlGaAs ICP-DRIE etching for optoelectronic, microelectronic and microsystem applications.*
  
  PB Vigneron, **F Joint,** N Isac, R Colombelli, E Herth. Microelectronic Engineering, 2018

• *330 GHz and 600 GHz schottky heterodyne systems for QPSK terahertz telecommunication.*
  

• *Development of thz quantum cascade lasers and hot electron bolometers for ultra-sensitive and ultra-compact heterodyne detection in astronomy applications.*
  

• *Development of quantum cascade lasers at 2.7 thz for heterodyne detection.*
  
Résumé
Nous avons développé un récepteur hétérodyne terahertz (THz) compact et ultra-sensible à base de laser à cascade quantique (QCL) comme oscillateur local et de bolomètre à électron chaud (HEB) comme mélangeur. Le récepteur est basé sur un nouveau concept pour le couplage quasi-optique entre l'oscillateur local et le mélangeur ce qui a permis de ne pas utiliser de lame semi-réfléchissante pour la superposition du signal provenant du QCL et du signal à détecter. Le mélangeur utilisé est un HEB en nitrure de niobium avec une antenne planaire formée d'une double hélice log-spiral. Le HEB est monté sur la partie plane d'une lentille convexe en silicium. L'oscillateur local est un QCL que nous avons développé avec un système de contre-réaction répartie du troisième ordre avec une faible dissipation thermique, un faisceau peu divergent et un fonctionnement mono-mode à la fréquence cible de 2.7 THz. Le couplage entre l'oscillateur local et le mélangeur HEB a également été amélioré en couplant le QCL avec une fibre creuse en diélectrique ce qui a permis d'améliorer la directivité du faisceau laser à 55 dBi. Grâce aux précédents résultats, nous avons obtenu un récepteur THz hétérodyne compact qui présente une sensibilité proche de l'état de l'art à 2.7 THz.

Mots Clés
Térahertz, Hétérodyne, Laser à Cascade Quantique, Bolomètre à Électron Chaud

Abstract
We demonstrate an ultra-compact Terahertz (THz) heterodyne detection system based on a quantum cascade laser (QCL) as local oscillator and a hot electron bolometer (HEB) for the mixing. It relies on a new optical coupling scheme where the local oscillator signal is coupled through the air side of the planar HEB antenna, while the signal to be detected is coupled to the HEB through the lens. This technique allows us to suppress the beam splitter usually employed for heterodyne measurements. The mixer is a Niobium Nitride HEB with a log-spiral planar antenna on silicon and mounted on the back of a plano-convex silicon lens. We have developed a low power consumption and low beam divergence 3\textsuperscript{rd}-order distributed feedback laser with single mode emission at the target frequency of 2.7 THz to be used as local oscillator for the heterodyne receiver. The coupling between the QC laser and the HEB has been further optimized, using a dielectric hollow waveguide that reliably increases the laser beam directivity up to 55 dBi. Upon the high beam quality, sufficient output power in a single mode at the targeted frequency and low power dissipation of our local oscillator, we have built an ultra compact THz heterodyne receiver with sensitivity close to the state of the art at 2.7 THz.

Keywords
Terahertz, Heterodyne, Quantum Cascade Laser, Hot Electron Bolometer