Novel mid-infrared quantum cascade devices for applications in free-space optics, data security and microwave photonics
Pierre Didier

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Novel mid-infrared quantum cascade devices for applications in free-space optics, data security and microwave photonics

Thèse de doctorat de l’Institut Polytechnique de Paris préparée à Télécom Paris

École doctorale n°626 École doctorale de l’Institut Polytechnique de Paris (ED IP Paris)
Spécialité de doctorat : Electronique et Optoélectronique

Thèse présentée et soutenue à Palaiseau, le 11 Octobre 2023, par

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I would like to express my heartfelt gratitude to all those who have contributed to the completion of this Ph.D. thesis. This journey has been filled with challenges and milestones, and I couldn’t have reached this point without the unwavering support and encouragement of many individuals and institutions.

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IV Novel MIR QC Devices for applications in FSO, data security and microwave photonics
Les communications espace libre offrent une alternative intéressante pour la transmission de données lorsque la fibre optique n’est ni pratique ni réalisable. Cette technologie s’est imposée comme un candidat solide avec un large potentiel d’applications allant de l’internet à large bande dans la vie quotidienne aux liaisons par satellite. La disponibilité de transmetteurs et de détecteurs de haute qualité fonctionnant dans la fenêtre du proche infrarouge fait de la longueur d’onde optique de 1.5 µm choix naturel pour les systèmes optiques en espace libre. Néanmoins, deux autres domaines de longueur d’onde peuvent être envisagés. Tout d’abord, la fenêtre infrarouge à ondes moyennes (MWIR) (3−5 µm), et ensuite la fenêtre de l’infrarouge à ondes longues (LWIR) (8−14 µm). Les longueurs d’onde MWIR et LWIR sont réputées pour leurs performances de transmission supérieures à travers les phénomènes atmosphériques défavorables, tels que le brouillard, les nuages et la poussière. En outre, le domaine LWIR assure la furtivité du signal de communication grâce au rayonnement thermique du corps noir. L’environnement extérieur présente en effet une forte émissivité à ces longueurs d’onde, ce qui réduit considérablement la probabilité que des adversaires interceptent un signal laser LWIR. L’objectif de cette thèse est d’examiner les capacités de transmission de données à l’aide de l’optoélectronique quantique opérant dans les plages spectrales MWIR et LWIR. Pour ce faire, nous étudions les lasers interbande et inter-sous-bande à cascades, permettant une émission stimuliée dans les fenêtres 3−5 µm et 8−14 µm, ainsi que sur leur dynamique de modulation respective. Trois aspects clés sont explorés en utilisant des lasers à cascade interbande et inter-sous-bande émettant dans les plages MWIR et LWIR : la modulation rapide (directe ou externe) de ces sources, la sécurisation des données par synchronisation de chaos et l’intégration des technologies optique et radiofréquence (RF). Dans ce travail, j’ai accordé une attention particulière à l’optimisation des paramètres du système tels que les formats de modulation, les techniques de codage et de filtrage. Plus spécifiquement, des algorithmes d’égalisation sont systématiquement mis en œuvre pour améliorer les débits de transmission. Dans une première partie, j’ai étudié les lasers interbandes à cascade pour les communications FSO à faible consommation dans la plage de 3 à 5 µm. Ces lasers combinent les avantages des technologies interband et de diode laser classique. Une étude du bruit d’intensité a révélé leur fréquence de relaxation de l’ordre du GHz et leurs capacités à être modulés...
directement à des fréquences de plusieurs GHz. En utilisant un laser interbande à cascade modulé directement et un détecteur interbande en cascade, j’ai démontré des débits allant jusqu’à 14 Gbit/s. En ce qui concerne les transmissions dans la bande $8 - 14 \mu m$, l’utilisation d’un laser à cascade quantique optimisé en termes de courant de seuil et de longueur de cavité a permis d’atteindre un débit proche de 5 Gbit/s en modulation directe. Afin d’accroître la capacité de transmission, ce travail présente un nouveau système de transmission à $9 \mu m$ fondé sur une optoélectronique unipolaire intégralement réalisé en technologie inter-sous-bande. Ce dernier composé d’un laser, d’un détecteur et d’un modulateur Stark a permis d’atteindre des débits records allant jusqu’à 30 Gbit/s sur une distance de plus de 30 mètres. A la fin de ce chapitre, je montre que l’utilisation d’une seconde génération de modulateur à base de métamatériaux permet d’accroître le débit jusqu’à 68 Gbit/s. Dans une seconde partie, j’ai exploré la synchronisation de chaos pour les communications privées. Les propriétés non linéaires des lasers à cascade quantique, notamment la génération d’hyper-chaos, sont exploitées. Je démontre la possibilité d’une transmission privée dans la fenêtre infrarouge thermique en cryptant les données dans un signal chaotique à 8 Mbit/s. Les résultats montrent qu’un utilisateur malveillant atteint un taux d’erreur binaire (BER) d’environ 40 %, tandis qu’un utilisateur légitime obtient un BER de seulement 4 %. Je confirme également la viabilité d’un tel système en augmentant la distance de propagation du signal à 30 mètres grâce à l’utilisation d’une cellule de Herriott. La dernière partie de ma thèse aborde l’intégration des systèmes FSO avec la technologie RF. Un signal RF est généré en combinant de manière hétérodyne un signal modulé et un signal référence provenant de deux lasers à cascade quantique à $9 \mu m$ avec une séparation en longueur d’onde de 26 GHz. Ce signal RF est ensuite transmis via une antenne adaptée à la bande de fréquence cible. Les récents résultats obtenus sur les détecteurs à cascade quantique permettent d’envisager cette technologie pour la génération RF à très haute fréquence (> 100 GHz). En conclusion, ce travail met en lumière, les avantages des technologies moyen infrarouge à cascade quantique pour le développement de liens FSO à très haut-débit, sécurisés et intégrés avec des systèmes de communications RF. De manière notable, les récents progrès dans les plate-formes intégrées MWIR et LWIR favorisent l’avancement des systèmes de communication dans les fenêtres de longueur d’onde du moyen infrarouge. Cela englobe la transmission QAM et l’expansion potentielle vers des technologies de transmission multiplexées.
Summary

Free-space communications offer an attractive alternative for data transmission when optical fiber is neither practical nor feasible. This technology has emerged as a strong contender with a wide range of potential applications, from broadband internet in everyday life to satellite links. The availability of high-quality transmitters and detectors operating in the near-infrared window makes the optical wavelength of 1.5µm a natural choice for free-space optical systems. Nevertheless, two other wavelength ranges can be considered. Firstly, the medium-wave infrared (MWIR) window between 3 and 5 µm, and secondly the long-wave infrared (LWIR) window between 8 and 14 µm. MWIR and LWIR wavelengths are renowned for their superior transmission performance through adverse atmospheric phenomena, such as fog, clouds and dust. In addition, the LWIR domain ensures the stealthiness of the communication signal thanks to the thermal radiation of the black body. The external environment has a high emissivity at these wavelengths, which considerably reduces the probability of adversaries intercepting an LWIR laser signal. The aim of this thesis is to investigate the data transmission capabilities through quantum optoelectronics operating in the MWIR and LWIR windows. To achieve this, interband and intersubband cascade lasers, enabling stimulated emission in the 3−5µm and 8−14µm windows, are studied along with their respective modulation dynamics. Three key aspects are explored using interband and intersubband cascade lasers emitting in the MWIR and LWIR windows: the rapid (direct or external) modulation of these sources, data security through chaos synchronization, and the combination of optical and radiofrequency (RF) technologies. In this work, I paid particular attention to optimizing system parameters such as modulation formats, coding and filtering techniques. In particular, equalization algorithms are systematically implemented to improve transmission rates. In the first part, I studied cascaded interband lasers for low-power FSO communications in the 3 to 5µm range. A study of the intensity noise revealed their relaxation frequency in the GHz range and their ability to be modulated directly at frequencies of several GHz. Using a directly modulated cascaded interband laser and a detector with cascaded interband transitions, I demonstrated data rates of up to 14 Gbit/s. For transmissions in the 8−14µm band, the use of a quantum cascade laser optimized in terms of threshold current and cavity length enabled data rates close to 5 Gbit/s to be achieved with direct modulation. To
increase transmission capacity, this work presents also a new 9µm unipolar system based entirely on intersubband technology. This system, comprising a laser, a detector, and a Stark amplitude modulator, has achieved record data rates of up to 30 Gbit/s over a distance of more than 30 meters. At the end of the chapter, I show that the use of a second-generation metamaterial-based modulator can increase throughput to 68 Gbit/s. In a second part, I explored the use of chaos synchronization for private communications. The nonlinear properties of quantum cascade lasers, in particular the generation of hyper-chaos, were extensively exploited. I have demonstrated the possibility of private transmission in the thermal infrared window by encrypting data in a chaotic signal at 8 Mbit/s. The results show that a malicious user achieves a bit error rate (BER) of around 40 %, while a legitimate user achieves a BER of just 4 %. I also confirmed the feasibility of such a cryptographic system by increasing the signal propagation distance to 30 meters using a Herriot cell. The final part of my thesis deals with the integration of FSO systems with RF technology. An RF signal is generated by heterodyning a modulated signal and a reference signal from two 9µm quantum cascade lasers with a wavelength separation of 26 GHz. This RF signal is then transmitted via an antenna adapted to the target frequency band. Recent results with quantum cascade detectors mean that this technology can be used for RF generation at very high frequencies (> 100 GHz). It could result in the development of a highly efficient microwave photonic system to capitalize on the rapid dynamics of intersubband transitions. In conclusion, this work highlights the advantages of quantum cascade mid-infrared technologies for the development of high-speed, secure FSO links integrated with RF communications systems. Notably, the recent advancements in MWIR and LWIR integrated platforms will enable the progression of communication systems within mid-infrared wavelength windows. This includes QAM transmission and the potential expansion towards multiplexed transmission technologies.
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Acronyms

AO  Adaptative Optics.
AOM  Acousto-Optic Modulators.
BER  Bit Error Rate.
DFB  Distributed Feedback.
DFE  Decision-Feedback Equalizer.
DFG  Difference-Frequency Generation.
EOM  Electro-Optic Modulators.
ESA  European Space Agency.
FEC  Forward Error Correction.
FFE  Feed-Forward Equalizer.
FP  Fabry-Perot.
FSO  Free-Space-Optics.
FSOoRF  Free Space Optics on Radio Frequency.
ICIP  Interband Cascade Infrared Photodetector.
ICL  Interband Cascade Laser.
ISI  inter-symbol interferences.
LE  Lyapunov exponent.
LED  Light-Emitting Diode.
LEF  Linewidth Enhancement Factor.
LEO  Low Earth Orbit.
LMS  Least Mean Squares.
LO  Local Oscillator.
LWIR  Long Wavelength Infrared Radiation.
MCT   Mercury-Cadmium-Telluride.
NRZ   Non-Return-Zero.
OAM   Orbital-Angular-Momentum.
OFC   Optical frequency combs.
OOK   On-Off-Keying.
PAM-4 Pulse Amplitude Modulation 4.
PAM-8 Pulse Amplitude Modulation 8.
PIC   Photonic Integrated Platform.
PPLN  Periodically Poled Lithium Niobate.
PRBS  Pseudo-Random Binary Sequence.
QCD   Quantum Cascade Detector.
QCL   Quantum Cascade Laser.
QCSE  Quantum Confined Stark Effect.
QKD   Quantum Key Distribution.
QPSK  Quadrature-Phase-Shift-Keying.
QW    Quantum-Well.
QWIP  Quantum Well Infrared Photodetector.
RF    Radio-frequency.
RIN   Relative Intensity Noise.
RLS   Recursive Least Squares.
RRC   Root-Raised-Cosine.
RZ    Return-to-Zero.
SEM   Scanning Electron Microscope.
UTC   Uni-Traveling Carrier.
WDM   Wavelength-Division-Multiplexing.
WPE   Wall Plug Efficiency.
"And as we let our own light shine, we unconsciously give other people permission to do the same." Nelson Mandela
The demand for high-speed internet access is on the rise, primarily driven by the advancements in mobile internet and the Internet of Things (IoT). Additionally, various initiatives led by governments, non-profit organizations, and private sector entities have been launched to foster digital inclusion and narrow the gap between those with internet access and those without. According to the World Bank Organization, the number of internet users increased from only 414 million in 2000 to over 4.7 billion in 2020. Nevertheless, 41% of the population still don’t have access to a broadband internet connection for their daily life, as shown in the Figure 1.1. Near half of the population is unable to enjoy the educational, healthcare, and financial advantages that come with having internet access. The underlying causes for this phenomenon include significant social and economic disparities, such as differences in language and education. In some cases, insufficient infrastructure and high investment costs also contribute to the issue. Additionally, Radio-frequency (RF) communication experiences service saturation due to the need to allocate frequencies to each operator, which is particularly sensitive to interference. Furthermore, RF wavelengths encounter inherent limitations, such as limited data rates and sensitivity to atmospheric conditions, making ground-to-air communication problematic. Meanwhile, development of fiber network remains very complicated and expensive.

1.1 Optical communication: a long history of research

The origins of optical communication can be traced back to ancient times, when people communicated using signals from fires on mountaintops or using a lighthouse. Although these methods were rudimentary with a low data rates (1 bit/min), they still conveyed information through the use of light. In the late 19th century, a major breakthrough in optical communication was achieved with Alexander Graham Bell’s invention of the "Photophone". This device utilized the discovery of the refractive index of various materials, as well as the development of mirrors and lenses, to transmit voice information through
Figure 1.1: The map displays the percentage of the population that has used the Internet in the past three months, revealing the disparities in internet development across the globe. On the map, the dark blue color denotes regions where nearly all the population has internet access, whereas the light blue color signifies areas with a meager percentage of people who can use the internet. Courtesy of Our World in Data.

free space. The Photophone worked by modulating the position of the focus of the sunlight beam with the vibrations of a mirror caused by the voice.

It is interesting to note that during the early stages of optical communication, the preferred medium for light propagation was through free space. However, in the 1960s, Charles Kao and George Hockham proposed a revolutionary idea to use a different medium to confine and guide light. They engineered a glass fiber that could guide light with low loss, thus paving the way for the use of fiber optics in communication. Unfortunately, at that time, the manufacturing of glass was not yet optimized, which presented a challenge for their proposal. The challenge of producing high-quality glass for fiber optics was finally overcome in the 1970s when Corning was able to manufacture top-grade glass. Meanwhile, the development of laser technology offered a new possibility for efficient transmission of data information through light. In the 1960s, direct modulation was explored as a way to transmit information by turning the laser on and off. However, at that time, this approach quickly faced limitations in bandwidth and power due to the important losses in the channel link. In the 1970s, external modulation emerged as a more promising technique.
that could overcome these limitations and achieve higher output power. Despite the high cost and difficult implementation of these devices, improvements in both light sources and transmission media led to the first successful demonstration of a viable communication system. In 1976, engineers installed an experimental fiber optic system under the streets of Atlanta, Georgia, using a directly modulated gallium-arsenide semiconductor laser. Since then, there has been continuous improvement in fiber quality to minimize light losses and enhance transmission efficiency. During the 1980s, a novel technique named electro-absorption modulation was introduced by researchers to modulate the laser light. It was integrated into the laser package. This made communication systems more practical for use in commercial fiber optic systems. Simultaneously, advancements in parallel communication using radio-frequency signals were made. It is dating back to the 19th century when Heinrich Hertz demonstrated the existence of radio-wave. In 1901, Guglielmo Marconi achieved a groundbreaking milestone by successfully transmitting a radio signal across the Atlantic Ocean, marking the advent of the first practical wireless communication system. It was the demonstration of the possibility to transmit radio-wave over long distances. Radio communication became a significant part of military operations during World War I, enabling troops to communicate with one another and their commanders, and facilitating coordination of troop movements and tactics. The use of radio communications expanded in the interwar period with the development of broadcast radio and the establishment of commercial radio networks. In World War II, radio communications continued to play
a vital role in military operations, not only in facilitating troop communications and co-
ordination but also in radar, navigation, and other critical purposes. In the present day,
Radio-frequency (RF) communications has become an integral aspect of modern life, serv-
ing a myriad of purposes ranging from broadcasting of radio and television programs to
facilitating wireless internet and mobile phone networks.

1.1.1 Basic on optics

Light is a form of electromagnetic radiation characterized by the oscillation of an elec-
tric and magnetic field perpendicular to each other. It is identified by its wavelength or
frequency, denoted as $\lambda$. Photons make up light, and they possess both wave-like and
particle-like properties. The wavelength of light is essential because it determines the en-
ergy and color of the beam. Light of varying energies behaves distinctively in different
environments, often experiencing varying levels of losses or demonstrating unique behav-
iors, as emphasized in Figure 1.3. Each wavelength is generated using different sources and
detected using specific methods which will be described later in the manuscript.

![Figure 1.3: Electromagnetic spectrum of light. Courtesy of NASA](image-url)
1.2 From Fiber optic and Radio-Frequency network to Free-Space-Optics

Fiber optics are widely used for telecommunications and are poised to become the predominant means of communication. For instance, the partnership between Orange, a French company, and Google has enabled the seamless implementation of a remarkable 250 Tbit/s connection that spans over 6,400 km, connecting America and Europe (the Dunant link). However, this technology requires expensive infrastructure and high maintenance costs, making it difficult to provide access to regions with low population density. Even in areas where the infrastructure exists, the cost of maintenance and repairs can be prohibitively expensive, making it challenging to achieve a return on investment. For instance, only 50% of households in France are currently connected to a fiber broadband line, and reaching the 100% target will require an investment of at least 15 billion euros, pushing the completion date to 2030 according to Arcep.\textsuperscript{10} Radio-frequency communications in the 1 GHz-100 GHz spectrum has become the standard for communication with 4G and 5G networks. However, RF network still faces challenges such as limited data rates (100-1000 Mbit/s) resulting from the need for licensing to attribute different wavelengths to operators due to interferences. In addition, the data rates of RF systems decrease drastically with the spatial density of users on the network as well as degraded atmospheric condition. In the Table 1.1, a comparison of three different communication networks was conducted. It is evident that each application has its own unique advantages and disadvantages. Consequently, it becomes important to determine the specific contexts where each application is relevant.

Industrial programs, including Starlink, utilize the advantages of RF frequency in the 26 GHz band to provide internet connection worldwide. Nevertheless, the utilization of this system with numerous users is not advisable due to RF interference. Furthermore, it experiences substantial reductions in both system availability and efficiency when subjected to unfavorable conditions. To address these obstacles and expedite space programs such as SpaceX (Tesla), OneWeb, Lightspeed (Telesat), and Sentinel (ESA), there is a pressing requirement for advanced communication systems. Free-Space-Optics (FSO) systems could offer a promising solution.
Table 1.1: Comparison between different systems of communications. PP: Point to Point; A communication link refers to the connection or pathway between two communication endpoints or nodes. M: Mesh. PM: Point to Mesh. Courtesy of Buckley

<table>
<thead>
<tr>
<th></th>
<th>FSO</th>
<th>Fiber</th>
<th>RF (5G)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Deployment time</td>
<td>Day to weeks</td>
<td>4-12 months</td>
<td>Weeks to years</td>
</tr>
<tr>
<td>Provisioning time</td>
<td>Immediate</td>
<td>Complex</td>
<td>Complex</td>
</tr>
<tr>
<td>Initial investment</td>
<td>Low</td>
<td>High</td>
<td>High</td>
</tr>
<tr>
<td>Reliability</td>
<td>Medium</td>
<td>High</td>
<td>Medium</td>
</tr>
<tr>
<td>Topology/flexibility</td>
<td>PP, PM, Mesh</td>
<td>PP, PM, Mesh</td>
<td>PP, PM, Mesh</td>
</tr>
<tr>
<td>Data security</td>
<td>High</td>
<td>High</td>
<td>Low</td>
</tr>
<tr>
<td>Distance limitation</td>
<td>200-2000 m (for now)</td>
<td>200 km</td>
<td>&lt;km</td>
</tr>
<tr>
<td>Bandwidth/speed</td>
<td>TBit/s</td>
<td>TBit/s</td>
<td>&lt;1 GBit/s</td>
</tr>
<tr>
<td>Maintenance cost</td>
<td>Low</td>
<td>Very high</td>
<td>Low</td>
</tr>
</tbody>
</table>

1.3 FSO System

FSO communications refers to the transmission of information using optical signal through the air or vacuum without the use of any physical medium or guided wave structures. It would be able to provide an efficient solution to a worldwide development of high speed free space communications at very low cost and low infrastructure. FSO systems present some other advantages like the immunity to RF interference (no licensing), high data rates of several hundreds of Gbit/s and can lead to high distances link. As depicted in the Figure 1.4, the implementation environments are diverse and can be terrestrial, aerial, in space, and in deep-space between satellites. FSO communications via satellites offer several benefits. Firstly, they have the capability to cover expansive areas of the Earth’s surface, even encompassing remote and inaccessible regions. This makes it feasible to establish communication links with any location on the planet. FSO systems can also effectively address the density problem and building interferences in urban environments, thanks to their directional and narrow beam characteristics. Additionally, this technology presents an intriguing opportunity to minimize infrastructure demands.

In urban environments characterized by a high population density and obstacles posed by buildings, the proposed system holds immense value. As depicted in the Figure 1.4, a FSO system comprises a modulated optical source, such as an Light-Emitting Diode (LED),...
Figure 1.4: Alternative setup for a FSO system. The graph shows the differentiation implementation of an FSO system from short haul transmission between building to communication between ground and satellite. The fundamental components of the FSO system are also illustrated. Courtesy of Trichilli.

or a semiconductor laser, emitting a beam that can be shaped using adaptive optics to counteract channel effects. Various modulation techniques, including electrical, optical, or external modulation, can be employed to shape the signal. The signal traverses through the atmosphere, where it is received by a detector and subsequently demodulated to retrieve the transmitted message. FSO systems offer numerous advantages, such as:

- **Link capacity**: By utilizing visible and infrared wavelengths instead of the RF spectrum, it becomes possible to achieve higher data rates, reaching hundreds of Gbit/s.
- **Narrow beams size**: Narrow beam size allows having a very furtive signal (hard to detect)
- **Interference free**: Minimize interference with other FSO systems, it is possible to avoid the need for wavelength allocation, in contrast to RF communications.
- **Cheap and fast deployment**: Systems can be implemented in very complicated environment.
1.3.1 Transparency windows

The selection of appropriate wavelengths for FSO systems is mainly based on the transparency windows of the atmosphere. The existence of low absorption windows in the atmosphere is determined by the molecular composition of the air. The absorption of photons by atmospheric molecules leads to electron transitions to higher levels of vibrational or rotational electronic energy, with the specific transitions being molecule-dependent. As shown in the Figure 1.5, there is several low absorption band in the infrared spectrum. In particular, wavelength around 1.5\(\mu\)m, 2.2\(\mu\)m, 3-5\(\mu\)m, and 8-14\(\mu\)m, are considered suitable for FSO communication.

![Figure 1.5](image)

**Figure 1.5:** The atmosphere’s attenuation affects various ranges of the electromagnetic spectrum. There are three transparency windows that are particularly useful for data transmission: radio frequencies to millimeter waves, which are employed for short-range communication solutions, near-infrared at 1.5\(\mu\)m, and the mid-infrared transparency window ranging from 8 to 14\(\mu\)m. Courtesy of cFlow Project

1.3.2 Modulation/demodulation

The initial assessment of a device’s viability for use in a communication system involves investigating its response to direct modulation. Modulation is the process of changing one or more characteristics of a periodic waveform, known as the carrier signal, using a separate signal called the modulation signal. The modulation signal typically contains information...
that is to be transmitted, as illustrated in the Figure 1.6. In optical transmission, we have access to three parameters that can carry information: intensity, phase, and polarization. However, in our case, due to source and detection limitations, amplitude modulation is primarily used for transmitting signals over FSO links. Amplitude is related to the square root of intensity. The concept involves either directly modulating the output power of a laser by adjusting the bias current, known as direct modulation, or externally modulating a continuous wave signal, referred to as external modulation, to encode information. FSO sources need to be capable of high-frequency modulation.

**Figure 1.6:** A carrier signal can be modulated either in amplitude or frequency, allowing the generation of a signal where information can be encoded using bits. In each curve, the y-axis represents power of the signal, and the x-axis represents time.

### 1.3.3 Pseudo-Random-Binary-Signal

A Pseudo-Random Binary Sequence (PRBS) signal is a signal that combines randomness with determinism. It consists of a sequence of binary values that appear random but can be reproduced given the same initial conditions. PRBS signals find extensive application in telecommunications, data communications, and digital testing. They are a test patterns for evaluating the performance and quality of digital systems, including communication links, data transmission channels, and electronic devices. PRBS signals exhibit desirable characteristics such as a flat power spectrum, well-distributed transitions, and an equal distribution of ones and zeros. These properties make them highly effective in assessing system behavior, identifying potential errors, and verifying system functionality. Leveraging their pseudo-random nature, PRBS signals enable thorough stress testing and evaluation of digital systems in real-world scenarios, thereby ensuring robustness and reliability. The sequence length of the generated signal can vary from $2^N - 1$, where N typically ranges from 7 to 31. Having a longer sequence length ensures that all combinations of following bits have been tested in the process, enhancing the thoroughness of the testing.
1.3.3.1 Pulse Amplitude Modulation (PAM)

The variation in the power of the signal carries the encoded information. The modulated signal thus obtained contains frequency components at the carrier frequency and its sidebands, which can be transmitted over the FSO link. At the other end of the link, the receiver can demodulate the signal to extract the original message signal. One advantage of amplitude modulation is its relative simplicity of implementation, requiring only a nonlinear element to vary the carrier signal’s amplitude. AM is easier to detect as it primarily focuses on the amplitude of the received signal, making the phase unnecessary in the detection process.

**ON-OFF-KEYING (OOK)** One of the most basic kind of modulation is OOK (On-Off-Keying). As depicted in the Figure 1.7.a), bits are generated by toggling the laser on and off. Within this category, there are different keying techniques, such as Return-to-Zero (RZ) and Non-Return-Zero (NRZ). In the case of RZ, the signal returns to a zero state between individual bits. This type of modulation is categorized as such because, in practice, we do not turn off the laser, but rather change the output between two different values.

![Figure 1.7](image)

**Figure 1.7:** Various techniques are employed to transmit information encoded as binary bits, where each bit can have a value of either 0 or 1. The baud rate, which corresponds to the symbol rate, is set here at 10 Gbaud. The top of each figure illustrates the corresponding bits transmitted within a 2 ns time period. a) A schematic representation of OOK (On-Off Keying) modulation is shown. In this case, a data rate of 10 Gbit/s can be achieved. b) A schematic representation of PAM-4 modulation is depicted. In this case, a data rate of 20 Gbit/s is achieved by transmitting two different bits for each symbol.

**Higher Modulation Techniques** In practical, we can increase the number of level to increase the number of bits send per symbol. In the Figure 1.7.b) the number of level is 4, and then we can send 2 bit/symbol. In general, there exists a relationship between the number of $N_b$ bits per symbol and the modulation level $M$, given by, $N_b = \log_2(M)$. It is...
noteworthy that in higher modulation formats, the use of Gray coding becomes essential. Indeed, this coding allows only a single bit to be modified for successive coding levels, thus minimizing errors in the event of a misdetermination between two nearby levels.

1.3.3.2 Phase modulation (PM)

In phase modulation schemes, binary information is transmitted by modulating the phase of the signal, as shown in Figure 1.8. In this specific example, the phase of the signal is modulated while keeping the amplitude unchanged. This type of system is commonly referred to as Quadrature Phase Shift Keying (QPSK). However, in many systems, both the phase $\phi$ and the amplitude of the signal $r$ are independently modulated. This type of system is known as N-QAM, where N can vary (e.g., 2, 4, 16, etc.). In particular, Quadrature Amplitude Modulation (QAM) is a widely used scheme in which information is encoded simultaneously on two carriers that are phase-shifted by 90°. Coherent detection and modulation techniques, including phase modulation, offer improved performance in terms of data transmission and resilience against channel impairments.

![Carrier signal (ex. $\lambda = 14 \ \mu m$)](image)

**Figure 1.8:** A diagram is presented to illustrate the modulation scheme of 4-phase shift keying (4-PSK), where a carrier signal with a wavelength of 14 $\mu m$ is modulated at a baud rate of 10 Gbaud. The yellow curve represents the signal over a 0.4 ns time period. By employing this modulation scheme, a data rate of 20 Gbit/s can be achieved. The modulation of the carrier signal is depicted using a constellation diagram, which shows the amplitude of the modulated signal $r$ corresponding to the phase $\phi$.

There is a challenge with coherent detection. It is a technique used to detect and extract phase and amplitude from modulated optical signals using a local oscillator. In this method, the local oscillator generates a reference signal that is coherent with the incoming modulated signal. By combining the reference signal with the modulated signal, interference occurs, resulting in an output signal that contains both amplitude and phase information. They are extensively utilized in modern communication systems to achieve...
high-speed and reliable data transmission. However, the limited availability of reliable coherent emitters in the mid-infrared range currently restricts the widespread use of this technique.

1.3.4 Multiplexing

In an FSO link, one effective approach to significantly enhance the information rate is through the use of multiplexing techniques. Multiplexing enables the transmission of multiple streams of information simultaneously over the same communication channel by utilizing different carriers or properties of the signals. There are several methods of multiplexing that can be employed, including Wavelength Division Multiplexing (WDM), Polarization Division Multiplexing (PDM), Orthogonal Frequency Division Multiplexing (OFDM), and Mode Division Multiplexing (MDM). WDM involves transmitting multiple signals at different wavelengths, allowing them to coexist in the same optical FSO link. Each signal is modulated onto a different wavelength. PDM takes advantage of the polarization properties of light to transmit multiple signals simultaneously. Different polarization states are used to encode separate information streams, which can be separated and recovered at the receiving end. OFDM divides the available spectrum into multiple orthogonal subcarriers, each carrying a portion of the data. MDM exploits the spatial modes of light propagation to transmit multiple independent signals in parallel. These multiplexing techniques play a crucial role in maximizing the information rate and spectral efficiency in FSO links, enabling high-speed and reliable communication over optical channels. However, it is important to note that implementing these methods can be challenging and require careful design and optimization to achieve their full potential.

1.3.5 Evaluation of transmission quality

In any communication system, it is crucial to assess the feasibility of using the system in a real communication setup. Evaluating key information plays a vital role in determining its suitability.

1.3.5.1 Eye diagram

The eye diagram is a graphical representation of a signal and is widely employed in the analysis and evaluation of high-speed digital communication systems, such as Ethernet, fiber optic systems, and serial data transmission. It plays a crucial role in assessing various characteristics of the signal, including timing, amplitude, noise, jitter, and overall signal
integrity. By evaluating the quality and integrity of the signal, the eye diagram enables us to gauge the effectiveness of communication. The name "eye diagram" originates from the distinct shape formed by the representation of the signal.

Figure 1.9: Eye diagram for two different configurations: a) The diagram was plotted by overexposing 50 traces, and b) overexposing 500 traces on the same figure at a constant sample time. To emphasize the construction of the diagram, each sample of traces is plotted using different colors.

To generate an eye diagram, consecutive bits are overlaid, with each bit period divided into equal time intervals. The vertical axis represents the voltage of the normalized signal. In Figure 1.9.a), we can observe the superimposition of 50 traces that were sampled at the same time intervals. Additionally, we have included a diagram (Figure 1.9.b)) that depicts the construction of the system using 500 overlapping traces. By overlaying multiple bits, the eye diagram reveals several essential characteristics:

- **Eye-opening**: The central region of the diagram symbolizes the 'eye' of the signal. Ideally, this area should be wide and unobstructed. A narrower or distorted eye-opening suggests potential problems such as noise, interference, or timing errors.

- **Transitions**: The diagram displays both the rising and falling edges of the digital signal, facilitating the assessment of the signal’s timing performance.

- **Jitter**: It denotes the variability in timing between successive bits. It becomes apparent as fluctuations in the position of signal transitions within the eye diagram. Excessive jitter has the potential to introduce errors during data transmission.

- **Noise**: The eye diagram is capable of detecting the presence of noise or interference within the signal. Noise manifests as random fluctuations or irregularities in the signal’s amplitude, resulting in a reduction in the eye’s definition and closure.
Through the analysis of an eye diagram, signal integrity issues can be detected, leading to adjustments that enhance system performance. Optimization of parameters such as equalization, timing recovery, and signal conditioning enables the attainment of a well-defined and clear eye diagram, thereby ensuring dependable data transmission.

### 1.3.5.2 Bit Error rates

The metric that represents the probability of an erroneous bit being transmitted in digital communication is known as the Bit Error Rate (BER). It is calculated by dividing the number of bit errors by the total number of bits transmitted during a specific time interval. The BER is expressed as a ratio without units and plays a crucial role in evaluating the performance and reliability of communication systems. Therefore, it is essential to study and evaluate the BER value to establish the system’s efficiency and accuracy. The BER value in the case of a 2 level modulation can be defined as:

\[
BER = \frac{1}{2} \left( erfc\left( \frac{1}{\sqrt{2}} \frac{l_1 - l_0}{\sigma_1 + \sigma_2} \right) \right) \tag{1.1}
\]

where \(I_{0,1}\) is the photocurrent for the bits level 0,1, \(\sigma_{0,1}\) the noise for the level 0,1 and \(erfc\) is the complementary error function (erfc). This constant defines the bit error rate during a transmission.

### 1.3.6 Demodulation

Demodulation plays a significant role in communication systems by extracting the original bits from the detected signals. The specific demodulation method employed depends on the modulation scheme used. Techniques such as frequency discrimination, envelope detection, and phase detection are commonly utilized. By employing these methods, demodulation enables the retrieval of the corresponding bits, ensuring accurate data recovery at the receiver end.

#### 1.3.6.1 Laser noise

The investigation of noise in optical communication systems is extremely important since excessive noise can significantly affect the system’s performance. There are numerous factors that can result in a noisy signal, including the source, propagation medium, and detection. This section aims to analyze the various sources of noise and their direct impact on the system.\(^{13}\) Firstly, we will examine the noise generated by the laser and the detector. In our case, we will place ourselves in the situation of direct amplitude modulation.
The first noise contribution to consider is the noise generated by the laser itself. Relative Intensity noise (RIN) defines the ratio between the power spectral density of the intensity noise and the squared average optical power. If this noise is important, it can degrade the quality of the transmission. The RIN is defined by the fluctuation of the optical power $\delta P(t)$ to the mean power $<P>$ expressed in linear units per Hertz:

$$RIN = \frac{<\delta P(t)^2>}{<P>^2}$$

(1.2)

By using this formula, we can write the impact of intensity noise in the case of direct modulation on the detector:

$$\sigma_{laser}^2 = R^2P_R^210^{RIN/10}\Delta f \ [A^2]$$

(1.3)

where $R$ is the responsivity of the detector, $P_R$ the power received by the detector, $\Delta f$ the bandwidth of our system.

1.3.6.2 Detector noise

Thermal noise, also referred to as Johnson-Nyquist noise, is generated by the thermal fluctuations of electrons in the receiver circuit caused by the temperature-induced variations in the resistance of electronic components. This noise is considered frequency-independent or white noise.

$$\sigma_{th}^2 = \frac{4k_bT}{R_0}\Delta f \ [A^2]$$

(1.4)

where $k_b$ is the Boltzmann constant, $T$ the temperature (K), and $R_0$ is the differential resistivity of the circuit.

Shot noise is a type of noise that results from the quantum nature of light, and it describes the random fluctuations of the number of detected photons over time. It is also caused by the movement of electrons in a circuit or through a photodiode. In the case of large signals, shot noise follows a Gaussian distribution with a zero mean and a variance that is determined by the signal strength.
\[ \sigma_{sn}^2 = 2q(RP_R + I_d)\Delta f \ [A^2] \] (1.5)

where \( q \) is the charge of the electrons, \( R \) is the responsivity of the detector, \( P_R \) is the received power, \( I_d \) correspond to the dark current of the detector. The dark current can be attributed to various factors such as thermally excited carriers, surface recombination, and tunneling between the conduction and valence bands. It is typically minimized when the device is unilluminated, but it tends to increase with temperature and the applied reverse bias commonly used in photodiodes to sweep carriers out of the depletion region more rapidly.\(^{25} \) Therefore, cooling the photodetector can be beneficial in reducing dark current.

**BACKGROUND NOISE** Background noise in FSO communication can be defined as any light that passes through the optical filters beyond the receiver and is not the desired optical signal. Multiple sources contribute to this background noise in FSO, including blackbody radiation and light reflected from the Earth’s surface and atmosphere.\(^{26} \) The level of background power reaching the receiver primarily depends on the optical filter’s bandwidth and the receiver field of view. While one might assume that using a narrow field of view system with a narrow optical filter would minimize the background noise, it’s important to consider that narrow optical filters tend to have higher losses. First, the Planck’s law describes the distribution of spectral radiance emitted by a black body in thermal equilibrium at a temperature \( T \).

\[ L_{\Omega \lambda} = \frac{2hc^2}{\lambda^5} \frac{1}{e^{hc/(\lambda k_b T)} - 1} \ [W.m^{-2}] \] (1.6)

In the equation, \( h \) represents the Planck constant, \( c \) is the speed of light, \( \lambda \) denotes the wavelength, \( k_b \) stands for the Boltzmann constant, and \( T \) represents the temperature of the body. Now, we must assess the total background radiance received by the detector. As evaluated in\(^ {27,28} \), we can express the background radiation power as:

\[ P_{bg} = \underbrace{A_r \cdot FOV \cdot \Delta \lambda \cdot T_F \cdot Rad}_{\text{Solar radiation}} + \underbrace{\alpha FOV \cdot A_r \cdot T_A \cdot T_F \cdot \Delta \lambda \cdot L_{\Omega \lambda}(T, \lambda)}_{\text{Blackbody radiation}} \] (1.7)

In the given equation, \( A_r \) represents the receiver’s primary area, \( FOV \) denotes the receiver’s field of view, \( \Delta \lambda \) represents the optical filter’s bandwidth. \( Rad \) symbolizes the reflected solar radiance, \( T \) represents the Earth object’s blackbody temperature, \( T_A \) denotes the
atmospheric attenuation, $T_F$ represents the optical transmission of the filter. The factor $\alpha$ represents the radiant absorbance, which relates the perfect blackbody spectral radiance to the graybody spectral radiance. The power evaluation can be utilized to retrieve the current variance.

$$\sigma^2_{bg} = 2qRP_{bg}\Delta f$$

### 1.3.6.3 Sum of the different noise

We obtain the total noise power ratio $P_n$ using the sum of the different power spectral density $\sigma$:

$$\sigma^2_{total} = \sigma^2_{sn} + \sigma^2_{th} + \sigma^2_{font} + \sigma^2_{laser} + \sigma^2_{bg}$$

### 1.3.6.4 Electrical amplifier noise

When analyzing amplifier noise, it is essential to recognize that it not only multiplies the input noise but also introduces additional noise from the amplifier itself. Here $P_{a}$, $G$ is the noise and the gain of the amplifier respectively:

$$P_{total,amp} = P_nG + P_a$$

### 1.3.7 Beam through atmospheric propagation

To describe the evolution of the electric field during the propagation through the atmosphere, measured at the distance $r$, the signal shape can be approximated as a Gaussian beam. The expression is the following:

$$E(x,y,z) = E_0\frac{w_0}{w(z)}\exp\left(-\frac{r^2}{w(z)^2}\right)\exp\left(-ikz - ik\frac{r^2}{2R(z)} + i\arctan\left(\frac{z}{z_0}\right)\right)$$

The radial distance, represented as $r$, and the distance from the emitter, denoted as $z$, are illustrated in Figure 1.10 for visual clarity. The values $w_0$ and $w(z)$ respectively denote the waist size and the radius of the beam at distance $z$. $R$ corresponds to the radius of curvature of the beam’s wavefronts at $z$. For an FSO link ($z >> Z_R$), we can approximate $w(z) = \tan(\theta_{div}z)$ where $\theta_{div}$ is the angle of divergence of the beam. It can be approximated as $\theta_{div} = \frac{\lambda}{\pi w_0}$ far from the origin where $\lambda$ is the wavelength of the signal.
Figure 1.10: Schematic visualization of a Gaussian beam at the output of a laser. It shows the important parameters to describe the beam shape.

**Beam Divergence** Maintaining a narrow beam during propagation and ensuring sufficient power during detection are important for a FSO system, and beam divergence plays a critical role in achieving this. In the following, $L$ will correspond to the total length of the FSO link. By approximating the signal beam as a Gaussian beam, we can express the channel response factor $h_d$ as the ratio of emitted power to transmitted power\(^{30}\):

$$h_d = \frac{D_{Rx}^2}{(D_{Tx} + 2\theta_{div}L)^2}$$

(1.12)

where $D_{Tx}$, $D_{Rx}$ is respectively the diameter aperture of the emitter and the receiver. It allowed us to take into account the losses caused by the divergence of the beam.

### 1.3.7.1 Attenuation

During its propagation, the beam suffers from degradation due to the interaction with the medium, which is in our case the atmosphere. It may come from absorption by gas molecules like carbon dioxide, scattering, or wavefront deformation induced by turbulence. Neglecting wandering and the effect of turbulence in a first approach, losses caused by the propagation of our Gaussian laser can be written as\(^{19,31}\):

$$P_R = P_T h_d e^{-\gamma(\lambda)L}$$

(1.13)

where $P_R$ is the power at the reception, $P_T$ is the emitting power, $L$ the propagation dis-
tance, $\gamma(\lambda)$ represent the relative attenuation caused by atmospherics losses and is written as $\gamma(\lambda) = \alpha_m(\lambda) + \alpha_a(\lambda) + \beta_m(\lambda) + \beta_a(\lambda)$ where $\alpha_m$, $\alpha_a$, represent the absorption coefficient of the molecule and of the aerosol and $\beta_m$, $\beta_a$ their respective scattering coefficients. The next part will focus on evaluation of the $\beta$ parameters.

**SCATTERING** As our signal propagates, it will interact with atmospheric molecules and scatter due to their absorption and re-emission in random and anisotropic directions. The extent of scattering depends on the wavelength and size of the molecules, and can cause changes in the signal’s absorption. To quantify the size parameters of the signal, we use (1.14):

$$x = \frac{2\pi r}{\lambda}$$

where $r$ is the radius of the scattering molecule and $\lambda$ the wavelength.

**Figure 1.11:** Three different types of scattering depending on the incoming signal wavelength: from left to right Rayleigh scattering, Mie scattering and geometric scattering.

As depicted in Figure 1.11, there are three different types of scattering which depend on the value of $x$:

If $x \ll 1$, we observe Rayleigh scattering. This type of scattering is caused by the electrostatic dipole oscillation resulting from the deformation of the electrostatic cloud. This is considered insignificant for wavelengths beyond 800 nm. According to (1.15), we can express $\beta_m(\lambda)$ as:

$$\beta_m(\lambda) = 0.827.N_pA_p^3\lambda^{-4}$$

In the given context, $N_p$ represents the number of particles per unit volume, and $A_p$ is the
cross-sectional area of scattering.

If \( x \simeq 1 \), we observe Mie scattering, which describes elastic scattering in the direction of propagation without any change in wavelength. This type of scattering is mainly caused by fog and haze. The corresponding scattering coefficient can be expressed as \( \beta_a(\lambda) \), according to the empirical Kim model\(^3\):\(^1\)\(^2\):

\[
\beta_a(\lambda) = \frac{3.91}{V} \left( \frac{\lambda}{\lambda_0} \right)^{-q}
\]  
   \( \text{(1.16)} \)

where \( \lambda_0 \) is the visibility range reference wavelength, \( V \) the visibility range, \( q \) is the size distribution of the scattering particles. \( q = 0 \) for \( V < 0.5\,\text{km} \), \( q = V - 0.5 \) for \( 0.5 < V < 1\,\text{km} \), \( q = 0.58V^{1/3} \) for \( 1 < V < 6\,\text{km} \), \( q = 1.3 \) for \( 6 < V < 50\,\text{km} \) and \( q = 1.6 \) for \( V > 50\,\text{km} \).

If \( x \gg 1 \), the signal endures attenuation due to the presence of large particles like snow, rain or fog. This is geometric scattering. We describe the different value caused by geometric scattering \( \gamma(\text{dB}\,\text{km}^{-1}) \). We have\(^3\):

- **Rain**: \( \gamma_{\text{rain}} = K_R R^{a_R} \), where \( R \) is the precipitation intensity, \( K_R \) and \( a_R \) are empirical values that depend on raindrop size and rain temperature which can be found in the literature.
- **Snow**: \( \gamma_{\text{snow}} = a_S S^{b_S} \), where \( S \) is the snowfall rate (mm/hr), and \( a_S \) and \( b_S \) are snow parameters whose values change if the snow is wet or dry. We have \( a_S = 0.000102\lambda + 3.79 \) and \( b_S = 0.72 \) for wet snow and \( a_S = 0.0000542\lambda + 5.50 \), \( b_S = 1.38 \) for dry snow.
- **Dust**: \( \gamma_{\text{dust}} = K_d B^{b_d} \), where \( K_d \) and \( b_d \) are also empirical value whose values can be chosen as 52 and \(-1.05\), respectively.\(^3\)

One of the main challenges in FSO systems is the ability to withstand very foggy conditions, which result from the accumulation of water droplets in the atmosphere. This phenomenon causes a significant amount of scattering and attenuation, particularly affecting the 1.5\,\mu m systems. The publication\(^1\) provides information on the attenuation values around 850 nm and 1550 nm, as well as the visibility values. In order to provide a more comprehensive overview, we have added information on the values for the 3-5 \,\mu m and 8–14 \,\mu m windows to the Table 1.2.

These results specifically pertain to the influence of Mie scattering, which is contingent upon the specific conditions. Additionally, it is worth noting that Rayleigh scattering
Table 1.2: Beam attenuation caused by the mie scattering. Courtesy of Trichilli

<table>
<thead>
<tr>
<th>Visibility (km)</th>
<th>1550 nm</th>
<th>4000 nm</th>
<th>9000 nm</th>
</tr>
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<tbody>
<tr>
<td>20 (Clear)</td>
<td>0.2</td>
<td>0.06</td>
<td>0.02</td>
</tr>
<tr>
<td>5 (Haze)</td>
<td>1.2</td>
<td>0.47</td>
<td>0.2</td>
</tr>
<tr>
<td>2 (Mist)</td>
<td>4</td>
<td>2</td>
<td>1.1</td>
</tr>
<tr>
<td>1 (Fog)</td>
<td>9.3</td>
<td>5.4</td>
<td>3.3</td>
</tr>
</tbody>
</table>

becomes negligible for signals beyond the wavelength of 800 nm. Finally, if we want to include the potential losses due to the divergence, we can write the response factor thanks to the Equation 1.12. The received power can now be evaluated using the Equation 1.13, where γ is determined based on the aforementioned considerations.

**Example of losses in the band 3 of the atmospheric window**

We can apply the formulas from the Equation 1.16 and Equation 1.13 with a $L = 1$ km FSO link in an average visibility condition $V = 5$ km and basic optics with a $10$ cm aperture (transmitter and receiver). For a $9 \mu m$ beam, we calculate an absorption and scattering attenuation of $0.2 \text{dB km}^{-1}$ while the divergence of the laser beam induces an $8.6 \text{dB km}^{-1}$ attenuation. Note that these values could be easily decreased by using more advanced optics. It leads to total losses around $8.8 \text{dB}$ in this configuration. Assuming $100 \text{mW}$ of output power, we retrieve around $13 \text{mW}$ which is sufficient to detect a signal and receive the message. In a situation where the visibility goes down to $V = 1$ km, which corresponds to mist visibility, attenuation due to scattering jumps to $3.3 \text{dB km}^{-1}$ which remains small compared to equivalent scattering at $\lambda = 1.5 \mu m$ ($9.3 \text{dB km}^{-1}$) and at $\lambda = 4 \mu m$ ($5.4 \text{dB km}^{-1}$). In addition, if there is particular weather condition like rain, snow or large dust we have to take into account the geometric scattering for which the absorption coefficients can be approximated with the formula expressed in the previous part. The drawback at longer wavelength comes from optical element’s size, which needs to be increased accordingly to cope with the effect of divergence.

**1.3.7.2 Turbulence**

Given that our transmission channel is a medium that is constantly in motion, it is important to consider the impact of air movements on our signal’s characteristics. Atmospheric turbulence arises from the kinetic energy generated by the movement of air in eddies. Large eddies ($L_0$ accounts for the inverse of the spatial frequency) transfer their energy
to smaller ones, and this process continues until it reaches the viscosity regime ($l_0$) as depicted in Figure 1.12. In the inertial range, turbulence introduces fluctuations in the refractive index and mass density of the medium. As a consequence, the signal wavefront experiences distortion, and there are spatial variations in the intensity of the signal known as scintillation. These fluctuations in intensity can pose a considerable challenge for demodulating the signal. Kolmogorov was the first to study the influence of turbulence on signal propagation.\textsuperscript{36}

\textbf{Figure 1.12:} An energy cascade refers to the process of transferring energy from large scales of motion to smaller scales (known as a direct energy cascade) or from smaller scales to larger scales (known as an inverse energy cascade). Courtesy of Sauvage\textsuperscript{37} and Wheelon\textsuperscript{38}

\textbf{KOLMOGOROV TURBULENCE} To statistically characterize atmospheric turbulence, it is necessary to define key parameters for representation. By referencing the research conducted by Kolmogorov\textsuperscript{36} and modifications by Tatarski\textsuperscript{39} and Von Karman,\textsuperscript{40} a model for atmospheric turbulence can be established. The fluctuation of the refractive index can be described using a characteristic length scale. The parameter $C_n^2$ represents the refractive index structure parameter and indicates the strength of turbulences. It is defined as the structure function of the refractive index fluctuations and represents the vertical profile of...
turbulence strength. Utilizing this parameter, the Kolmogorov power-law can be employed to describe the refractive index fluctuations:

\[
\Phi_n^K(K) = 0.033 C_n^2 \exp\left(-\frac{K^2}{K_m^2}\right) \ast (K^2 + K_0^2)^{-11/3}
\]  

(1.17)

The Kolmogorov spectrum incorporates the spatial coordinate frequency, denoted as \(K\) \((1/l_0 \gg K \gg 1/L_0)\). Notably, the Tatarskii spectrum, a modified version of the Kolmogorov spectrum proposed by Von Karman, accounts for both the inner and outer scales of turbulence. The calculations involve the values of \(K_m = 5.92/l_0\) and \(k_0 = \frac{2\pi}{L_0}\). To determine the \(C_n^2\) coefficient, we utilize the Hufnagel-Valley model:

\[
C_n^2(h) = 0.00594 (v_{\text{wind}}/27)^2 (10^{-5}h)^{10} \exp(-h/1000) + 2.7 \times 10^{-16} \exp(-h/1500) + A \exp(-h/100)
\]  

(1.18)

In the given equation, \(h\) represents the altitude in meter, \(v_{\text{wind}}\) is the RMS wind speed, and \(A\) denotes the nominal value of \(C_n^2(0)\) at the ground. This equation provides us with valuable information on \(C_n^2\), which can be used to describe the spatial distortion of the signal due to turbulence. In this section, we have discussed the impact of turbulence and defined the spectrum of index fluctuations.

Turbulence induces a range of effects on the signal during its propagation:

- Scintillation
- Small scales eddies causes **beam broadening**
- Larger scales eddies causes **displacement of the center of the beam**

**Gamma-Gamma Model for Scintillation Estimation** When a laser beam propagates through optical turbules that are smaller than the beam’s diameter (small eddies), the quality of the beam wavefront deteriorates, leading to laser beam scintillation, as depicted in Figure 1.14. In the context of a weak turbulence regime, the intensity of scintillation is determined using the Gamma-gamma model, as described in the research conducted by Andrews et al. and Churnside. It is valuable to define the average aperture for a plane wave approximation in Kolmogorov turbulence. This average aperture represents the ratio of the scintillation index observed by a receiver \(A_s\) with a wave function \(k\) and diameter...
$D_{Rx}$ to that observed by a receiver with an infinitesimally small aperture when a plane wave propagates over a distance $L^{41,42}$:

$$A_s = \left[1 + 1.062\left(\frac{kD_{Rx}^2}{4L}\right)^7\right]^{-7/6} \quad (1.19)$$

Our objective now is to investigate the effects of these variations on the propagation of our signal. Several methods have been tested to model this influence accurately. The Split Step Fourier (SSF) method\textsuperscript{43} is widely recognized as one of the most precise techniques for simulating the propagation of optical waves through turbulent media. It is a numerical method that effectively captures the complex dynamics of turbulent environments. The SSF method involves dividing the propagation distance into smaller steps and performing calculations. At each step, the method applies a Fourier transform to the optical wave, converting it into its spectral representation. In the spectral domain, the wave is propagated by applying a phase factor that incorporates the effects of the turbulent medium, including phase distortions, scintillation (intensity fluctuations), and beam spreading. By considering these factors, the SSF method accurately models the behavior of the wave as it propagates through the turbulent medium. After propagating in the spectral domain, an inverse Fourier transform is applied to obtain the spatial representation of the wave at the next propagation step. This iterative process is repeated for each step, allowing for a detailed and precise simulation of the wave’s propagation through the turbulent medium. The phase screen is generated by employing the previous results, as given by (1.17).

Figure 1.13: This figure illustrates the method approximating the turbulence by a series of transverse random phase screens and using the numerically efficient fast Fourier transform (FFT) algorithm to propagate the signal.

The SSF method serves as a valuable tool for researchers and engineers in studying the impact of turbulence on optical systems, assessing system performance, and optimizing system parameters. It finds applications in diverse fields such as free-space optical communication, novel MIR QC Devices for applications in FSO, data security and microwave photonics.
communication, remote sensing, and imaging, providing insights and strategies for understanding and mitigating the effects of turbulence.

**BEAM WANDER ESTIMATION** When the optical turbules are larger than the beam diameter, also known as large eddies, they deflect the entire beam and induce random wandering around the aimpoint, which is referred to as "beam wander". In laser communication satellite uplinks, both beam wander and scintillation are significant factors because the laser beam diameter is usually smaller than the size of many encountered optical turbules. However, beam wander has less impact on satellite downlinks since the laser beam has already spread to some extent before entering the Earth's atmosphere.

Figure 1.14: The beam undergoes wandering as it propagates through a turbulent environment, causing the spot to vary in position with time. This can make the detection of the beam challenging. a) Beam wandering caused by large eddies. b) Scintillation caused by small eddies.

For a Gaussian beam propagating through a turbulent medium, the deviation of the center of the beam position can be defined with respect to the center of the beam in the absence of turbulence. This deviation leads to variations in the radius $w$ as defined in Equation 1.11:

$$w_{\text{long}}(z) = w(z) + \langle \beta_w^2 \rangle$$  \hspace{1cm} (1.20)

with $\langle \beta_w^2 \rangle = 1.52C_n^2z^3w_0^{-\frac{1}{4}}$ represent mean square lateral displacement of the beam, $z$ the distance of propagation.
1.3.7.3 Pointing error

In FSO communications, pointing error refers to the discrepancy or misalignment between the optical beam emitted by the transmitter and the aperture of the receiver. This misalignment causes a reduction in the received signal power. Pointing errors can arise from multiple factors, such as atmospheric turbulence, mechanical vibrations, optical misalignment, and inaccuracies in tracking mechanisms. Minimizing pointing errors is of utmost importance in FSO systems to ensure the reliable and efficient transmission of data. We can approximate pointing error for a Gaussian beam by using the reference 45:

\[ h_p = A_0 \exp \left( \frac{-2r^2}{w_{eq}^2} \right) \]  

(1.21)

where \( A_0 = [\text{erf}(v)]^2 \) correspond to the fraction of power collected without pointing error, \( \text{erf} \) is the error function. The equivalent beam width at the receiver side \( w_{eq}^2 = w(L)^2 \frac{\sqrt{2} \text{erf}(v)}{2v \exp(-v^2)} \) and \( v \) correspond to \( v = \sqrt{\frac{\pi (D_{Rx}/2)}{2w(L)}} \).

1.3.8 Final link budget

By defining all these concepts, we can then give a final approximation of the link budget of a propagation channel in the atmosphere:

\[ P_R = P_T \cdot \frac{D_{Rx}^2}{(D_{Tx} + 2\theta_{div})^2} \exp(-\gamma(\lambda)z)Ah_pL_t \]  

(1.22)

where \( A_h \) is the loss caused by scintillation phenomenon, \( h_p \) is the loss from pointing error, \( L_t \) is the loss from atmospheric turbulence.

1.3.9 Need for adaptive optics

As mentioned earlier, various disruptions can occur during the propagation of the signal. To establish an FSO system with a link distance of over 100 m, Adaptive Optics (AO) must be developed for both the emitter and receiver sides. As depicted in Figure 1.15, the beam at the receiver side is completely scattered, resulting in ineffective detection. It is decisive to explore real-time wavefront distortion correction algorithms based on approximating and correcting phase distortion induced by turbulence in real-time. AO is a closed-loop system designed to correct real-time wavefront distortions. The measurement
of distortions is performed using a wavefront sensor. The corrective element typically used is a deformable mirror. The signal produced is subsequently corrected to counteract any deformation caused by the transmission channel.

![Simulated beam intensity at the receiver](image)

**Figure 1.15**: The simulated beam intensity at the receiver is depicted in both linear (top row) and decibel (bottom row) units. The scenario includes a range of 20 km with a transmitter and receiver aperture diameter of 10 cm. The left column represents a wavelength of 1.55 µm, while the right column represents 10.6 µm. The solid lines illustrate the beam size at the receiver considering diffraction only. A $C_n^2 = 5 \times 10^{-14} \text{m}^{-2/3}$ has been used. Courtesy of Liu.

### 1.4 Applications
1.4.1 Ground to ground network

The primary objective of FSO technology is to create a ground-to-ground network that can provide a versatile and high-capacity mesh. Laser diodes operating at a wavelength of 1.5 µm have proven to offer the highest data rates and ranges for FSO links due to the small absorption coefficient and the availability of reliable sources and detectors with low power consumption and interesting modulation bandwidth. To increase data rates significantly, multiplexer links can be used. Several commercial FSO systems have been developed, such as MRV Connection’s where they demonstrate a 10 Gbit/s link over 350 m. The state-of-the-art FSO system currently available is the Nexus-10G™, a bi-directional and fully ruggedized solution for terrestrial point-to-point communication. It boasts impressive data rates of up to 10 Gbit/s and can establish FSO links spanning up to 5 km. Towards research aera, the German Aerospace Center has achieved a 1.72 Tbit/s link over 10.45 km. The current record for data rates was set in a laboratory bench by the German Aerospace Center at a rate of 1.036 Pbit/s using 54.139 Gbit/s OFDM-8QAM signals multiplexed. However, these systems face reliability issues in atmospheric conditions like cloud and fog, and data rates can drop to very low data rate in some atmospheric scenarios. Research is increasingly focused on mid-infrared sources due to their advantages in resisting degraded atmospheric conditions. These sources are being studied in this manuscript to overcome limitations in FSO systems. The underlying concept of such ground to ground transmission involves the development of multi-hop FSO systems to ensure both high-speed and long-distance links.

1.4.2 Hybrid FSO/RF links

A hybrid system combining FSO communication and RF transmission offers a robust solution for high capacity and uninterrupted communication. This system leverages the benefits of both FSO and RF technologies to ensure seamless communication even in challenging conditions. In cases where FSO communication is interrupted due to obstacles or adverse weather conditions, the system seamlessly switches to RF transmission, ensuring uninterrupted connectivity as shown in Figure 1.16.a). The FSO system can serve as the primary communication method, while the RF system acts as a backup system in case of disruption or adverse weather conditions that can limit the data rate. The hybrid system finds application in various domains, including military communications, cellular network backhaul, and high-speed internet access in urban areas. Its ability to provide high-bandwidth communication and adaptability to different environmental conditions makes it a preferred choice in scenarios where reliability and continuous connectivity are
Figure 1.16: Presentation of two type of FSO system configurations. a) The dual-hop link that incorporates both RF and FSO technologies and b) the heterogeneous link that utilizes both FSO and RF technologies together are being described.

In the second approach, as depicted in Figure 1.16.b), the data is initially transmitted using an FSO link. However, to overcome the limitations of FSO links, such as the requirement for line-of-sight communication, RF transmission is employed for the final leg of the communication to deliver the data to the end user.

1.4.3 Satellite FSO transmission

1.4.3.1 Ground to space

Satellite FSO transmission involves the use of FSO technology to establish a communication link between the ground and space allowing high speed and low latency. However, satellite FSO transmission also presents unique challenges, such as atmospheric interference and the need for precise pointing and tracking of the communication beams. As a result, careful system design and optimization are critical for ensuring the reliability and effectiveness of satellite FSO transmission. Currently, companies such as SpaceX Starlink and Telesat utilize RF signals at 26 GHz to transmit information from low Earth orbit (LEO) satellites to the ground for their worldwide internet service. However, RF frequencies are vulnerable to atmospheric conditions, resulting in restricted data rates (kbit/s) when the weather is unfavorable. The National Institute of Information and Communic-
tions Technology successfully demonstrated the first optical downlink and uplink in 1994, achieving an information exchange rate of 1 Mbit/s. In 2014, NASA’s Lunar Laser Communication Demonstration (LLCD) achieved bidirectional communication between a satellite orbiting the moon (which is 400,000 km away from Earth) and several ground stations located in Spain and the USA. The LLCD reported a maximum uplink throughput of 20 Mbit/s and a downlink throughput of 622 Mbit/s.\(^5\) Although FSO systems are considered as a promising solution for high-speed communication, they still suffer from pointing errors and the impact of atmospheric conditions even if FSO is less affected by atmospheric disturbances. However, several projects are focusing on developing those types of systems. In partnership with MIT Lincoln Laboratory, NASA is presently engaged in the development of a 200 Gbit/s optical communication system for deployment on a Low Earth Orbit (LEO) CubeSat, enabling the transmission of over 50 terabytes of information daily to a ground station.\(^5\)

### 1.4.3.2 Inter-satellite communication

Ensuring transmission reliability in inter-satellite optical communication is a significant challenge due to pointing errors caused by satellite vibrations. Nevertheless, optical links have been found to be highly suitable for satellite-to-satellite communication, as compared to RF microwave links.\(^6\) In line with this, companies like SpaceX Starlink and Telesat are planning to incorporate FSO satellite crosslinks within their satellite constellations, with the objective of providing global broadband internet access.\(^7\) Notably, an inter-satellite demonstration achieved a successful error-free transmission of 5.6 Gbit/s for several seconds between two LEO satellites (NFIRE and TerraSar-X).\(^8\) FSO technology proves to be highly advantageous for applications requiring rapid and efficient data transfer between satellites, including satellite-to-satellite communication, satellite data relay, and inter-satellite links.

### 1.4.4 Quantum-Key-Distribution FSO (QKD-FSO) system

With the advancement of quantum computing, unprecedented computational speeds are achievable, raising concerns regarding the security of cryptographic methods.\(^9\) Indeed, conventional cryptography methods may become less secure, necessitating the exploration of new approaches for secure key exchange. There is a growing interest in developing free-space data transfer, particularly through quantum key distribution (QKD) protocols, as current free-space communications are susceptible to eavesdropping due to their unidirectional nature. Unlike classical cryptography, continuous-variable (CV) QKD protocol is
a practical application of quantum information theory, enabling secure key distribution over a public channel, even against unlimited attackers.\textsuperscript{60} This involves encoding classical information into quantum signals exchanged between two parties connected by a quantum channel, supported by an authenticated classical channel. Recently, this protocol, already deployed in fiber optic systems, has been theoretically analyzed in the context of free-space communications.\textsuperscript{61,62} In that way, the implementation of FSO systems can address these concerns by facilitating the transmission of QKD. High amount of research are going to this direction. In 2018, Pang et al. conducted a laboratory test-bench experiment, achieving a quantum transmission rate of 10 Mbit/s over a distance of 1 meter.\textsuperscript{63} Currently, FSO-QKD is still hindered by the performances of low-speed (1 kbps), low-yield, expensive transmission apparatus, despite the great security provided by the intrinsic properties of photon entanglement.\textsuperscript{64,65} A recent study demonstrated that effective entanglement can be maintained for successful transmission of up to 100 bit/s over a distance of 10 kilometers between ground-to-ground stations. It is interesting to note that the losses encountered in an urban environment were found to be comparable to those observed during communication with a low Earth orbit satellite. However, recent findings indicate that entanglement is highly sensitive to deteriorated atmospheric conditions, particularly in the presence of rain.\textsuperscript{66} Ongoing research is being conducted to verify the effectiveness and robustness of QKD in FSO links.


Novel MIR QC Devices for applications in FSO, data security and microwave photonics


Mid-infrared sources for free-space optics applications

The development of new mid-infrared sources, such as super-continuum generation\(^1\-^3\), optical parametric oscillator (OPOs)\(^4\), and cascade laser technology\(^5\), offers promising prospects in fields like spectroscopy\(^6\), medicine\(^7\), and free-space laser communication\(^8\), as they replace bulky CO\(_2\) and lead-salt lasers\(^9\). One of the most promising technologies is the cascade laser technology which is presented hereinafter.

### 2.1 Intersubband transition for mid-infrared devices

![Diagram showing interband transition and intersubband transition in a single quantum well.](image)

**Figure 2.1:** Principle of interband transition and intersubband transition in a single quantum well. Courtesy of Lee\(^10\)

\[
\mathcal{H} = \left( \frac{\vec{p} + q \vec{A}}{2m^*} \right)^2 + V(z) = \frac{\vec{p}^2}{2m^*} + \frac{q}{2m^*} (\vec{A} \cdot \vec{p} + \vec{p} \cdot \vec{A}) + \frac{q^2 \vec{A}^2}{2m^*} + V(z) \tag{2.1}
\]

Since we are considering infrared wavelength, the spatial dependence of A can be neglected,
as $\frac{1}{\lambda} \ll \frac{1}{a}$ where $a$ is the unit cell constant, which is the dipole approximation. Here we choose the coulomb gauge, so we have $\vec{A} \cdot \vec{p} = \vec{p} \cdot \vec{A}$ and we can neglect the $\vec{A}^2$ term.

$$\mathcal{H} = \frac{\vec{p}^2}{2m^*} + \frac{q}{m^*} \vec{A} \cdot \vec{p} + V(z)$$

We consider an initial state $|i\rangle$ and a final state $|f\rangle$ defined by the Bloch function. Here, $k$ and $r$ are the two-dimensional wave and spatial vector, respectively. The $u_{c0}(\vec{r})$ corresponds to the periodic part of the Bloch function and $\chi_{c,i,k}(\vec{r}) = e^{i \vec{k} \cdot \vec{r}}$.

$$\psi_{c,i,k}(\vec{r}) = \chi_{c,i,k}(\vec{r}) u_{c0}(\vec{r}) - \text{Initial state}$$

$$\psi_{c,f,k'}(\vec{r}) = \chi_{c,f,k'}(\vec{r}) u_{c0}(\vec{r}) - \text{Final state}$$

We can evaluate the transition rate between those two states $|i\rangle, |f\rangle$ using the Fermi golden rules, where $\mathcal{H}$ is the perturbation and $g$ is the density of states in a quantum well in the final state $f$.

$$W_{i\rightarrow j} = \frac{2\pi}{\hbar} \left| \langle i | \mathcal{H} | f \rangle \right|^2 g(\hbar\omega)$$

$$= \frac{2\pi}{\hbar} \left| \langle i | \frac{qE_0}{2m^*\omega} \vec{e} \cdot \vec{p} | f \rangle \right|^2 g(\hbar\omega)$$

$$= \frac{2\pi}{\hbar} \frac{q^2 E_0^2}{4m^*\omega^2} \langle i | \vec{e} \cdot \vec{p} | f \rangle \delta(E_f - E_i - \hbar\omega)$$

Specifically, we consider the slowly varying envelope functions of the initial and final states, denoted as $\chi_{c,i,k}(\vec{r})$ and $\chi_{c,f,k'}(\vec{r})$ respectively:

$$\langle i | \vec{e} \cdot \vec{p} | f \rangle = \left\langle \chi_{c,i,k}(\vec{r}) u_{c0}(\vec{r}) \right| \vec{e} \cdot \vec{p} \left| \chi_{c,f,k'}(\vec{r}) u_{c0}(\vec{r}) \right\rangle$$

By evaluating the integral relative, we can show that the contribution of the terms of polarisation $\vec{e}_x$ and $\vec{e}_y$ is equal to zero due to the absence of crystal periodicity in the $z$ direction within a heterostructure, the light-matter interaction becomes directionally
preferential. Therefore, the final expression for the transition probability per unit time in an intersubband transition in a quantum well can be written as:

$$W_{i \rightarrow j} = \frac{2 \pi}{\hbar} \frac{q^2 E_0^2}{4m^* \omega^2} \left\langle \chi^c_{i, \mathbf{k}} | e_z p_z | \chi^c_{f, \mathbf{k}'} \right\rangle \delta(E_f - E_i - \hbar \omega)$$

(2.6)

This demonstrates the polarization selection rules for intersubband transitions, which means that the polarization of the transition for intersubband devices is always parallel to the z-axis. This can be disadvantageous for devices such as detectors, where special attention must be paid to the polarization of the incoming light.

2.1.1 Quantum Cascade Laser (QCL)

QCL complete their transition within the conduction band by using a cascading engineering structure consisting of several quantum wells. This is illustrated in Figure 2.2.a), which shows how the use of these thin layers of semiconductor material allows electrons to cascade down an energy staircase and emit photons, enabling a wide range of frequencies to be reached, from mid-infrared to the THz domain. Pioneering research has introduced the concept of stimulated emission generated in intersubband transitions based on heterostructures. Several studies followed this works and were able to demonstrate various transport properties of superlattices, including tunnelling in superlattices, resonant tunnelling in semiconductor double barriers. These studies led to experimental evidence of discretization in quantum wells. After a decade of attempts, the first functional QCL was developed by Frederico Capasso in 1994 at Bell Labs, enabled by advancements in molecular beam epitaxy (MBE) and band-gap engineering, which eliminated detrimental defect in the heterostructure. They demonstrated a direct evidence of laser action at a wavelength of 4.2 μm with peak powers exceeding 8 milliwatts in pulsed operation, as indicated by a significant narrowing of the emission spectrum. The significance of this finding stems from the scarcity of resources available on the mid-infrared spectrum.

Following those pioneer researches, they have focused on improving the performance of quantum cascade lasers (QCLs) by exploring the possibility of room temperature operation, increasing output power, and broadening the accessible output wavelength. However, QCLs have been limited by significant phonon scattering loss, resulting in high heat limitation and important losses in the structure. To address this issue, researchers have focused on reducing heat generation and increasing heat dissipation in the following quantum
Figure 2.2: Pioneer works on Quantum Cascade Laser showing the band alignment to achieve intersubband transition. a) Conduction band energy diagram of portion of the 25 period (active region plus injector) section of the quantum cascade laser. The injector ('Digitally graded alloy') facilitates the transportation of electrons to the upper level of the radiative transition via non-radiative processes and resonant-tunneling. The electron then relaxes to the lower state by emitting a photon before tunneling into the right miniband, referred to as the extractor. This process is repeated multiple times, with the extractor serving as the injector for the next radiative transition. b) Schematic representation of the dispersion of the n = 1, 2, and 3 states parallel to the layers. Courtesy of Faist.5

wells by using materials with higher thermal conductivity.16 Additionally, improving the doping profile has been found to significantly enhance carrier transport and reduce the effect of carrier leakage. After dealing with the heat problem, research was able to focus on designing smaller devices. Only one year after its discovery, the first demonstration of a room temperature operation of a pulsed-mode quantum cascade laser (QCL) was published.17 Subsequently, further advancements led to the successful implementation of continuous wave operation.18,19 These groundbreaking achievements showcased the rapid progress in the field of QCLs and solidified their potential as viable light sources for various applications. Recently, QCL has gathered attention because of their ability to generate easily frequency comb in the mid-infrared.20 Optical frequency combs (OFC) are coherent dynamic states emitted by a light source, which consist of a series of evenly spaced spectral lines in the frequency domain. These states are distinguished by low amplitude and phase noise. The ultrafast gain dynamics and broad gain spectrum of QCLs makes efficient mode-locking and broad comb generation. This discovery opens a very wide field of application from dual-comb spectroscopy21 to lidar technology.22
2.1.1.1 Rate equation for Quantum Cascade Laser (QCL)

The fundamental model used to describe single-mode QCL is based on rate equations, which consider three energy levels: lower, middle, and upper, as described in the work by Hamadou et al.\textsuperscript{23} In these equations, \( N_1, N_2 \) and \( N_3 \) represent the population densities of the corresponding energy levels in the QCL cavity, while \( S \) represents the number of photons in the cavity.

\[
\begin{align*}
\frac{dN_3}{dt} &= W l J - \frac{N_3}{\tau_3} - \Gamma p \frac{\sigma_{32}}{n_{\text{eff}} V} (N_3 - N_2) S \\
\frac{dN_2}{dt} &= \left( \frac{1}{\tau_{32}} + \frac{1}{\tau_{sp}} \right) N_3 - \frac{N_2}{\tau_{21}} + \Gamma p \frac{\sigma_{32}}{n_{\text{eff}} V} (N_3 - N_2) S \\
\frac{dN_1}{dt} &= \frac{N_3}{\tau_{31}} + \frac{N_2}{\tau_{21}} - \frac{N_1}{\tau_{out}} \\
\frac{dS}{dt} &= m \Gamma p \frac{\sigma_{32}}{n_{\text{eff}} V} (N_3 - N_2) S + m \beta \frac{N_3}{\tau_{sp}} - \frac{S}{\tau_p}.
\end{align*}
\] (2.7)

In the Equation 2.7, \( m \) is the number of stage of the QCL cavity, \( n_{\text{eff}} \) the effective refractive index, \( W \) and \( l \) are the lateral widths and length of the cavity respectively, \( c \) the velocity of light, \( \Gamma p \) the optical confinement factor, \( \beta \) spontaneous emission factor which corresponds to the rate of electrons taking part in the stimulated emission, \( \tau_{32}, \tau_{21}, \tau_{31} \) the times constant of the different transition respectively, from the upper level to the middle level, from the upper level to the lower level and from the middle level to the lower level. \( \sigma_{32} \) the stimulated emission cross-section. Only the spontaneous emission between the upper and the lower level is taken into account \( \tau_{sp} \). After performing calculation in the steady state configuration and above threshold, the system of equation in Equation 2.7 can be resolved:

\[
\begin{align*}
J_{th} &= \frac{1}{1 - \frac{1}{\tau_{32}} - \frac{1}{\tau_{sp}}} \frac{L p \alpha_w + \alpha_m}{\Gamma p} \\
S &= \left( \frac{J}{J_{th}} - 1 \right) + \frac{1}{(1 + \frac{1}{\tau_{31}})} \frac{\beta}{\tau_{sp}} J_{th} \frac{J}{J_{th}} S_{ph, sat}.
\end{align*}
\] (2.8)

Here, \( S_{ph, sat} = \frac{1}{(\frac{1}{\tau_{32}} + \frac{1}{\tau_{31}} + \frac{1}{\tau_{sp}})(1 + \frac{1}{\tau_{31}}) \beta \frac{\tau_{31}}{\tau_{sp}} V} \) and \( \alpha_w \), \( \alpha_m \) are the losses caused by the waveguide and the mirror.

The simple modeling presented here presents some drawback:

- Assumption of a single mode laser.
- Insufficient description of the lower state which doesn’t participate in any physical
process leading to an accumulation of charge in the lower level and an unrealistic Coulomb potential.

- The model’s incapacity to replicate the conditions necessary for resonant tunneling of carriers through the injector, as well as the associated behavior of the (I-V) characteristic.
- By omitting the temperature dependency of phonon-assisted rates, significant aspects of laser functionality are being disregarded.

Unfortunately, the optimization of this model was outside the scope of my study.

2.1.1.2 High power QCL

Developing high power QCLs is crucial for enhancing the performance of various applications including sensing, imaging, military, and communications. The quest for high power QCLs began with the invention, but it remains a challenging task. Achieving high power operation in a QCL requires optimizing the Wall Plug Efficiency (WPE), which represents the ratio of optical power to electrical power. The WPE is a crucial factor that needs to be prioritized in order to maximize the laser’s performance and power output while minimizing the current injection. One of the main obstacles is the need for a minimum voltage (usually above 10 V) to align the cascade structures before any gain behavior can be observed, which causes significant thermalization. Increasing the number of stages \( m \), improving the confinement factor \( \Gamma \), reducing the waveguide loss \( \alpha_w \), and decreasing the current threshold \( J_{th} \) are some ways to increase the output power without changing the QCL’s band structure. However, the number of wells cannot be increased too much, as it would cause an increase in voltage and more thermal effects that degrade the performances.

In 2011, improvements in heat dissipation and nanofabrication led to achieving 27% wall plug efficiency at RT and pulsed mode for a maximum power of 12 W at 4.9 \( \mu \)m as shown in the Figure 2.3. More recently, impressive result demonstrated 14 W pulse operation at RT while handling very high lattice mismatch on InP.

The development of high-power laser fabrication has continued to progress, resulting in impressive output power for continuous wave (CW) operation, reaching several watts. Recently, a 5 W output at room temperature in CW operation has been successfully demonstrated. They achieve to optimize the device geometry for counter heating process in the structure, which drag back the WPE due to enhanced non-radiative relaxation and reduced transport efficiency. This result constitutes a significant milestone for long-haul free-space communication and sensitive spectroscopy applications.
2.1. Intersubband transition for mid-infrared devices

2.1.1.3 Distributed Feedback (DFB) QCL

Due to the significant heat generation caused by the structure, developing the Distributed Feedback (DFB) architecture was not a straightforward task. The first demonstration of DFB structure happened 4 years after the initial demonstration of QCL. The idea of DFB is to be able to select one lasing wavelength within the cavity. In a wide range of applications, including multi-wavelength communication systems and spectroscopy, the utilization of monomode laser is essential for several reasons. Indeed, DFB lasers offer valuable advantages in precise spectroscopic systems due to their high stability and exceptional spectral purity. Their single-frequency operation is particularly beneficial for implementing multi-wavelength channel links without encountering interference issues. Furthermore, the requirement of monomode operation is crucial for various components like external modulators as it guarantees their optimal performance.

Nanofabrication has enabled the development of high-quality gratings as shown in Figure 2.4, resulting in reliable distributed feedback (DFB) devices. For example, Razeghi et al. have demonstrated impressive high-power single-mode operation of 2.4 W at 4.8 µm. Furthermore, DFB lasers have also been developed in the 7.2 µm range with an impressive output power of 550 mW. Long-wavelength (λ > 17 µm) DFB quantum cascade lasers have also been demonstrated to operate in continuous wave at room temperature, providing high output power. However, for high speed modulation purposes, processing short ridges in DFB QCLs is challenging as the optical gain significantly reduces and cannot
Figure 2.4: a) Structure of a Distributed-Feedback-Laser Quantum Cascade Laser. The grating allows reflection in the cavity and select a particular wavelength for monomode operation. b) Scanning electron microscopy picture of a DFB QC laser array\textsuperscript{30} and output laser spectra.\textsuperscript{31} Courtesy of Yao\textsuperscript{32}

compensate for ridge losses below 1 mm, hindering room-temperature operation. However, the implementation of high-efficiency external modulator systems greatly benefits from the use of high-power continuous wave DFB lasers.

### 2.1.2 External modulators

Research on mid-infrared devices builds upon the advancements made in near-infrared devices during the last few decades. However, it is crucial to adapt these techniques to operate effectively at higher wavelengths. In the near-infrared region, there are already several options for optical modulation using external modulators, employing various approaches:

- **Pockels Effect** By applying an electric field to a material, it becomes possible to manipulate its refractive index, enabling phase modulation. This modulation technique can be employed in a tuned waveguide or waveplate. By combining the modulated beam with the original beam in a Mach-Zehnder interferometer setup, it becomes possible to generate interferences and convey information through amplitude modulation\textsuperscript{36}.

- **Acousto-Optic Modulators (AOM)** AOMs utilize sound waves to modulate laser output by manipulating the refractive index of crystal or glass materials. The AOM system consists of an acousto-transducer piezoelectric device that generates the modulated sound wave. This longitudinal wave scatters a portion of the light, resulting
in a decrease in the intensity of the transmitted and measured signal, thereby achieving amplitude modulation. It is worth noting that the bandwidth of these devices is limited due to the use of piezoelectric components.

- **Electro-Optic Modulators** These modulators utilize either the Quantum Confined Stark Effect (QCSE)\(^37\) or the Franz-Keldysh Effect\(^38\) to modify the light absorption at a specific wavelength by applying an external electric field. This modification enables amplitude modulation of the light signal by modulating the absorption by changing the position of transition energy in a well for QCSE or in a bulk for Franz-Keldysh Effect.

Numerous techniques have been explored to develop mid-infrared modulators that can operate with multi-GHz bandwidths, building on previous work with cryogenic lead-salt lasers and Ti-indiffused guided-wave interferometric modulators.\(^39\) However, conventional unipolar modulators based on the transition between weak and strong light-matter coupling regimes have shown limited bandwidths of 750 MHz, limited by drift transport velocity, with only 10% modulation depth.\(^40\) In this thesis, we focus on an approach based on Stark effect in quantum wells for building an external modulator. The Stark effect, discovered in 1913 by Johannes Stark by observing the splitting of spectral lines of atoms under an external electrical field, has been observed in quantum wells since 1987.\(^41\) As demonstrated in the thesis of Thomas Bonazzi,\(^42\) the equation \(\Delta E_{\text{stark}} = 2.74 \frac{L^4 m^*}{\pi^2 \hbar^2} F^2\) can be used to calculate the variation in energy of the absorption peak, where \(L\) is the size of the wells, \(e\) is the charge of the electron, \(m^*\) is the effective mass of the electron, and \(F\) is the applied electric field. This equation shows that an electrical field of 50 kV/cm only leads to a variation in energy of 1 meV of the absorption peak. To increase the Stark effect shift, several ideas have been investigated. The more straightforward is to increase the size of the QW.

The manufacturing of very large quantum wells has been found to result in low Stark shifts and high non-linearity in the devices.\(^43\) Therefore, in 1991, the first demonstration of modifying the absorption of a quantum well using electron tunneling between asymmetric coupled quantum wells was shown.\(^44,45\) By applying a strong electric field, it becomes possible to modulate the absorption peaks of the quantum wells, transitioning from a strong to a weak coupling regime. However, this method requires the use of two asymmetric quantum wells to maximize the Stark shift, as the effect of the electric field on an electron’s energy transition alone is weak. It is worth mentioning that this method has a limitation in terms of the diagonal transition strength from the low-energy state of the large quantum wells to the high-energy state of the thin one. This requires the application

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Figure 2.5: Structure of the energy band of the Stark effect modulator. By applying a modulated electrical field, the bands in the asymmetric quantum wells are bent, allowing for the tuning of the absorption peak.

of a very high bias voltage to compensate for the lower transition strength compared to the vertical transition. This can pose challenges in manufacturing structures for high-speed applications.

Figure 2.6: a) The potential and bandstructure of our Stark modulator is simulated at electric field strengths of -60 kV/cm, 0 kV/cm, and 60 kV/cm. b) The evolution of energies $E_{12}$ and $E_{13}$, as well as the oscillator strength $f_{12}$ associated with the 1-2 transition, as a function of the bias applied to the structure. There is a linearity of the Stark shift over a range of 200 kV/cm and 75 meV. Courtesy of Thomas Bonazzi\textsuperscript{42}.

The modulator utilized in this study is based on cascaded asymmetric coupled quantum wells, as depicted in Figure 2.5. The modulation of the electric field results in the modu-
lation of the band and alters the energy transition $E_{12}$, as depicted in Figure 2.6.a). By varying the applied electric field, the band alignment between the two quantum wells is adjusted. This, in turn, leads to a modification of the oscillator strength $f_{12}$ between the two quantum wells and the corresponding absorption frequency peak, as shown in Figure 2.6.b). Consequently, the incident laser’s absorption is modulated accordingly.

2.1.2.1 Phase modulation with Stark effect external modulator

Significant advancements have been made in the field of phase modulation using Stark effect modulators, as evidenced by recent studies. At ENS, they successfully demonstrated phase modulation capabilities of approximately $15^\circ$. (Hamza Dely and Thomas Bonazzi). However, a recent breakthrough has demonstrated an highly modulated phase of around $60^\circ$. It has been shown that local phase tuning can be achieved by applying voltages ranging from -3 to +3 V, enabling precise electrical control over the polarization state and wavefront of the reflected wave using metasurfaces as shown in the Figure 2.7.c).

![Figure 2.7: Electrical control of reflection spectra and phase. a) The unit cell structure is schematically depicted, consisting of a top plasmonic resonator with a width (W), a multiple quantum well (MQW) layer. b) The measured reflection spectra for different bias voltages, spanning from -3 V to +3 V with a step size of 1 V, are presented. c) The phase shift at 6.49 µm, measured using a Michelson-type interferometer, is displayed for applied bias voltages ranging from -3 V to +3 V with steps of 0.5 V. Courtesy of Chung.

This technology enables the realization of high-bandwidth devices operating in a strong coupling regime. It represents a crucial advancement towards enhancing modulation schemes like Quadrature Amplitude Modulation (QAM) or phase modulation.
2.1.3 Quantum Well Infrared Photodetector (QWIP)

The initial concept for QWIP was proposed in the late 1980s, following the discovery of intersubband transition and its potential for generating stimulated emission, as discussed earlier. In a QWIP, the active region consists of multiple layers of semiconductor materials with different energy bandgaps, typically based on GaAs/GaAlAs. The key component is the quantum well layer, a narrow and thin layer sandwiched between wider bandgap barrier layers. This quantum well layer acts as a trap for infrared photons, efficiently absorbing them and generating electron-hole pairs, as illustrated in Figure 2.8. The application of an external voltage tilts the energy levels, enabling the excited carriers to overcome the potential barrier and escape from the well, resulting in the generation of photocurrent. The short photon lifetime of intersubband transitions enables the device to achieve a high bandwidth of approximately 100 GHz. By carefully selecting the well width, barrier height, barrier width, well doping density, and the number of wells, this detector can achieve rapid detection for wavelengths up to 20 µm. To improve absorption, the proposed design incorporated multiple quantum wells, as the absorption of a single well is limited. In order to achieve low dark current and high responsivity, a QWIP detector requires biasing with multiple volts. However, a notable characteristic of QWIP detectors is the intersubband selection rule, which dictates that only the z (TM)-component of the incident light’s polarization can induce electron scattering between subbands, leading to the detection process.

First effort of developing QWIP was restricted to very low temperature operation since the dark current was very important. Although the bandwidth was already important around 26 GHz. A very important breakthrough was attributed to Grant et al. where they demonstrated a very high heterodyne detection bandwidths achieved at room temperature. They utilized a 100 quantum wells QWIP operating at ~ 10 µm. The device was fabricated in a 16 µm side square mesa and illuminated from a 45° polished substrate, which is not practical for real-world applications. Despite the device’s 3 dB cutoff being around 25 GHz, it is noteworthy that it exhibited a flat response up to 110 GHz, as demonstrated by a relatively low attenuation of around 10 dB. This observation highlights the impressive high-frequency capabilities of QWIPs, as it indicates that the device’s performance remained relatively consistent even at such elevated frequencies. Next advancement with this technology will come with the implementation of an antenna as shown in the Figure 2.9. In fact, by coupling the detector element to an antenna, it becomes possible to reduce the active volume of the detector without compromising the radiation collection area and then do not compromise the quantum efficiency of the detector. In 2001, QWIPs
Figure 2.8: A diagrammatic illustration depicting the absorption and emission of electrons when subjected to light is presented. In the structure, the dark current and photocurrent combine. The photocurrent is produced by the photoemission from the wells, subtracting the recaptured photoelectrons in the wells.

Based on arrayed patch antenna resonators (PARs) were proposed.\textsuperscript{53} PARs is particularly interesting for QWIP as they enable normal incidence illumination, which offers clear advantages compared to facet illumination.\textsuperscript{51,52} Additionally, they also have the advantage of confining the electromagnetic field within a subwavelength volume. By incorporating an antenna into the QWIP structure, it becomes possible to reduce the size of the radiation collection area. As a result, the number of quantum wells can be reduced, leading to a smaller device capacitance and then high speed performance. Despite the reduction in size, the QWIP still maintains a high detectivity, making this approach highly desirable for high-performance applications.

The figure shown in Figure 2.9 illustrates the highest bandwidth achieved to date. The researchers were able to demonstrate a flat response of heterodyne detection up to 70 GHz at room temperature, associated with a responsivity of 150 mA/W. This impressive result was achieved using a 5 x 5 array of patch antennas. As we will see later in this manuscript, PAR is not limited to just QWIPs but can also be implemented for other mid-infrared devices like external modulators. Recently, a breakthrough was achieved in patch-array QWIP photodetectors, showcasing a remarkable frequency response of up to 220 GHz through mid-infrared photomixing. At room temperature, these detectors demonstrated
Figure 2.9: High speed capability of the QWIP. a) SEM image shows a $5 \times 5$ array of PARs integrated with a coplanar waveguide. The image provides a detailed view of the structure, illustrating the arrangement of the PARs and their integration with the QWIP. b) Heterodyne detection at 10µm with a nearly flat frequency response up to 70 GHz at room temperature, solely limited by the measurement system bandwidth. Courtesy of Hakl.$^{54}$

an impressive 3 dB bandwidth of approximately 100 GHz and a remarkable responsivity of 80 mA/W.$^{55}$

2.1.4 Quantum Cascade Detector (QCD)

Despite their similarity, a notable distinction from the QWIP is found in the inclusion of an electron injector in the QCD as depicted in Figure 2.10. QCDs are composed of coupled quantum wells, such as InAs/AlAsSb or GaAs/AlGaAs. When a photon is absorbed in the quantum well, an electron moves from the ground energy level to a higher energy level. This excited electron undergoes a cascade of quantum wells through longitudinal-optical phonon scattering and resonant tunneling until it reaches the ground level in the next optical quantum well, as depicted in Figure 2.10. The cascade extractor’s quantum wells have progressively increasing thicknesses, creating an asymmetric potential that enables unilateral carrier transport. As the number of cascade stages increases, a substantial photocurrent is generated without the need for bias, leading to a dark current-free operation. In consequence, unlike a QWIP, QCDs can operate passively without the need for bias current, and it is not affected by the typically high dark current observed in QWIPs. Another significant feature is the intersubband selection rule, where only the $z$ (TM)-component of the incident light’s polarization can induce electron scattering between subbands, as previously demonstrated.
The concept of Quantum Cascade Detectors (QCDs) was first proposed in 2002 by the research group led by Jerome Faist. Their initial design utilized an InGaAs/AlInAs lattice matched on InP, employing a Quantum Cascade Laser (QCL) structure as the detector. One intriguing feature of QCDs is the inclusion of an electron injector, which enables the possibility of zero bias operation. Consequently, these detectors are not limited by dark current at high temperatures and can operate at room temperature. However, their initial responsivity was relatively low, on the order of $\mu A/W$. To enhance the detector’s performance, researchers recognized the need for a specific device structure. In 2005, Thales successfully implemented the first enhanced QCD on GaAs/AlGaAs. This development achieved an impressive responsivity of $35 mA/W$ at a wavelength of approximately $9.2 \mu m$ and a temperature of 50 K. Subsequently, researchers focused on the development of high-speed QCDs, demonstrating a bandwidth of 23 GHz with a responsivity of $33 mA/W$ at $5.6 \mu m$. Meanwhile, other research focuses have been investigated, such as lower wavelengths, higher wavelengths, and broadband wavelength detection. The development of QCD detectors has reached a mature stage, demonstrating impressive responsivity of up to $111 mA/W$ at room temperature. Also, these detectors achieve high bandwidth, with a notable 25 GHz, while benefiting from a responsivity of 85 mA/W at room temperature. Additionally, alternative geometries, such as double QCD detection for spectroscopy applications, have been investigated. Moreover, the need for monolithic integration has led to the exploration of QCD implementation on different platforms, such as Si or GaN.

### 2.2 Interband transitions in cascaded devices

Interband laser diodes have been dominating the industry of telecommunication for decades because they are very reliable and efficient. It operates by transferring electrons between a semiconductor material’s conduction and valence bands, as shown in Figure 2.1, resulting in the absorption and emission of electrons when subjected to light.
in laser emission. The frequency of the laser ($\nu$) is determined by the energy gap ($E_g$) between these two bands, which satisfies the inequality $E_g \leq h\nu \leq F_c - F_v$ for a bulk, where $F_c$ and $F_v$ are the quasi-Fermi levels associated with the conduction and valence bands, respectively. However, the use of this principle for mid-infrared diodes is not viable since there are no materials with an energy gap corresponding to wavelengths above 2 $\mu$m in the III-V. In the case of Indium Arsenide (InAs) or indium antimony (InP), the energy gap encompasses wavelengths beyond 2 $\mu$m in the mid-infrared range. While these semiconductors may not be suitable for laser emission because of high losses, they remain significant for detection purposes.\(^{67}\)

2.2.1 Interband Cascade Laser (ICL)

Following the successful demonstration of Quantum Cascade Lasers (QCL), R.Q. Yang proposed the concept of Interband Cascade Laser (ICL).\(^{68}\) ICLs combine the advantages of QCLs and interband diode lasers. ICLs exploit radiative transitions between the conduction band and the valence band. The structure of ICLs is similar to that of QCLs, employing a heterostructure band cascaded design, as shown in the Figure 2.11. However, due to the nature of interband transitions, both electrons and holes need to traverse the heterostructure, making ICLs intrinsically slower compared to QCLs. Despite this limitation, ICLs can operate in the mid-infrared wavelength range of 3–5 $\mu$m at room temperature, and with a low power consumption which is advantageous for various applications like space industry.\(^{69}\) The first experimental demonstration of ICLs was conducted by Lin et al. in collaboration with R.Q. Yang.\(^{70}\) The device, consisting of 20 active stages, exhibited a high threshold current density at temperatures ranging from 80-120 K, when used for generating short pulses. However, the device structure was not yet optimized and lacked strong electron injection in the QW region. Over the years, significant improvements were made by introducing multiple QW configurations in the ICL design. Initially, ICLs only operated in pulse mode at low temperatures below 200 K.\(^{71-74}\) The breakthrough came with the demonstration of pulse mode operation at room temperature by R.Q. Yang’s group.\(^{75}\)

The development of CW operation in ICL was facilitated by the design of W-shaped quantum well, which significantly increased electron-hole recombination and lowered the current threshold. The quantum well (QW) structure consists of a hole confinement region surrounded by two electron confinement regions. The first demonstration of CW operation in ICL was achieved in 2008.\(^{77}\) This achievement was made possible through various modifications to the device. Firstly, the number of stages was reduced to five to minimize the
threshold current. Secondly, separate-confinement layers (SCLs) made of n-doped GaSb were introduced on both sides of the active core. Thirdly, the doping concentration and thickness of the confinement layer were optimized to reduce heat generation. These developments have resulted in the creation of highly efficient devices that can achieve an output power of 300 mW for FP cavity or 50 mW for DFB cavity in CW operation. These devices operate at room temperature and emit in the $3-5\,\mu\text{m}$ wavelength range. A recent discovery is the possibility of implementing ICLs on silicon substrates. Furthermore, there have been recent efforts to increase the lasing wavelength of ICL to the $8-14\,\mu\text{m}$ range, but it faces two challenges. Firstly, the InAs/AlSb superlattice (SL) utilized for the optical cladding layers needs to be significantly thicker to accommodate the longer decay length of the optical wave, which complicates the growth process using molecular beam epitaxy (MBE). Secondly, the SL exhibits low thermal conductivity, so increasing the overall thickness of the SL would also raise the thermal resistance of the device, thereby impacting its performance. Despite that, recent result showed continuously lasing ICL at $11\,\mu\text{m}$ at cryogenic temperature with an output power of $14\,\text{mW}$ using adapted structure. Finally, the low power consumption of ICL makes these lasers interesting for various applications that require integrated devices, such as spectroscopy, integrated free-space optics, and medical imaging. Moreover, simulations using a semi-classical rate equation model for ICLs have shown great potential for generating squeezed states and utilizing ICLs in mid-infrared

**Figure 2.11:** Interband cascade structure based on InAs/AlSb/GaSb type-II QWs. The band diagram is not drawn to scale for the purpose of emphasizing the interband coupling window. Courtesy of Yang.  

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quantum applications.\textsuperscript{82}

2.2.1.1 Rate equation of the Interband Cascade Laser (ICL)

In this section, we will examine the rate equations for the ICL. The model used here is a simplified version that considers only two levels. The equation used is derived from the research conducted by Deng et al.\textsuperscript{83} To simplify the modeling process, certain assumptions have been made. It is assumed that all stages of the cascaded devices are identical\textsuperscript{84} and that the electron and hole densities remain constant in the active region, which is true for an efficient rebalancing.\textsuperscript{85} It is assumed that the carrier lifetime is predominantly governed by Auger and spontaneous recombination. Interestingly, this differs from other laser diodes where Auger recombination is usually neglected since the timescale for Auger recombination is much longer than that of spontaneous recombination. However, in this case, Auger recombination plays a significant role due to its comparable timescale to spontaneous recombination.\textsuperscript{86} Furthermore, the effect of transport time in the injector region is neglected, considering that the carrier transport time (sub ps) is much shorter than the carrier lifetime (sub ns).\textsuperscript{78,87} The following equation allows us to determine the variation of carriers per stage $N$:

$$\frac{dN}{dt} = \eta \frac{I}{q} - \frac{N}{\tau_{sp}} - \frac{N}{\tau_{aug}}. \quad (2.9)$$

In this equation, $I$ represents the pump current, $S$ the photon number in the cavity, $q$ is the elementary charge, $\eta$ represents the current injection efficiency. The equation incorporates several factors such as $g$, which represents the material gain per stage, $v_g$, the group velocity, and $\Gamma_p$, the optical confinement factor. The time constants for stimulated and Auger recombination processes are denoted as $\tau_{sp}$ and $\tau_{aug}$ (supposed constant here), respectively. This equation gives rise to a photon number rate equation, where $m$ represents the number of stages, $\tau_p$ is the photon lifetime, and $\beta$ is the spontaneous emission factor accounting for the rate of electrons involved in the stimulated emission process:

$$\frac{dS}{dt} = m\Gamma_p v_g g S - \frac{S}{\tau_p} + m\beta \frac{N}{\tau_{sp}}. \quad (2.10)$$
The gain $g$ can be represented by the equation
\[ g = \frac{g_0}{1 + \varepsilon N_p} \ln\left(\frac{N + N_s}{N_0 + N_s}\right) \] where $g_0 = a_0(N_{tr}/A)$. $N_s$, $N_{tr}$ number of carrier in steady state and at transparency respectively, $\varepsilon$ is the gain compression factor and $a_0$ is the differential gain. $A$ represents the area of the laser cavity and $m$ the mass of the electron. Using the Equation 2.9 and Equation 2.10, we can determine the number of carriers above and below threshold. However, it is important to note that the assumption that Auger recombination remains constant below threshold may not hold true. Classically, we have $\tau_{aug} = \frac{1}{\gamma_3 (N/A)^2}$, with $\gamma_3$ is the 2-D Auger coefficient.

\[ N = \begin{cases} \eta \frac{I}{q} \frac{1}{\tau_s + 1/\tau_{aug}} & \text{Below threshold} \\ \frac{1}{m \Gamma_p V_g a_0 \tau_p} + N_{tr} & \text{Above threshold} \end{cases} \]  

The current threshold can be expressed as:

\[ I_{th} = \frac{q}{\nu} \left( \frac{1}{m \Gamma_p V_g a_0 \tau_p} + N_{tr} \right) \left( \frac{1}{\tau_s \nu} + \frac{1}{\tau_{aug}} \right). \]  

The provided rate equation represents a simplified and basic representation of the dynamics of an ICL. However, it fails to consider several important factors that can significantly impact the behavior of an actual ICL. Neglected aspects include carrier transport, carrier leakage, and optical losses, which can have a substantial influence on device performance. Moreover, the assumption of constant carrier densities for Auger recombination oversimplifies the complex nature of the process. Auger recombination rates can vary with carrier densities, and a more sophisticated model should account for this dependence. To gain a deeper understanding of the ICL dynamics and obtain more accurate results, advanced methods based on stochastic approaches are being explored. These approaches take into consideration the inherent randomness and fluctuations associated with carrier transport and recombination processes. By incorporating these stochastic effects, more insightful information about the device performance can be obtained.

2.2.1.2 Small signal analysis for the ICL

The evaluation of the system's high-speed capability involves conducting a small signal analysis of the rate equation, as demonstrated by Deng et al. This small signal analysis requires expressing the different parameters as follows:
\[
\begin{align*}
  I(t) &= I_0 + \delta I e^{j\omega t} \\
  N(t) &= N + \delta N(\omega) e^{j\omega t} \\
  S(t) &= S + \delta S(\omega) e^{j\omega t}
\end{align*}
\]  

(2.13)

where \( \delta I, \delta N \) and \( \delta S \) represent the modulation of the amplitudes of the current, the number of carrier, and of the number of photons in the cavity. \( \omega \) is the modulation angular frequency. This analysis leads to the classical equation that expresses the transfer function for interband devices. The damping factor and the relaxation oscillation frequency is denoted as \( \gamma \) and \( \omega_R \):

\[
H(\omega) = \frac{\omega^2_R}{\omega^2_R - \omega^2 - j\omega \gamma}.
\]

(2.14)

where

\[
\begin{align*}
  \omega^2_R &= G_0(N - N_{tr})(mG_0S + m\frac{\beta}{\tau_p} + (G_0S + \frac{1}{\tau_p} + \frac{1}{\tau_{aug}})(\frac{1}{\tau_p} - mG_0(N - N_{tr})) \\
  \gamma &= G_0S + \frac{1}{\tau_p} + \frac{1}{\tau_{aug}} + \frac{1}{\tau_p} - mG_0(N - N_{tr})
\end{align*}
\]

(2.15)

For a detailed calculation, please refer to appendix A. This initial analysis demonstrates the potential of ICLs to be modulated at frequencies in the GHz range, as shown in the Figure 2.12.83

### 2.2.2 Interband Cascade Infrared Photodetector (ICIP)

Similar to the QCD, the ICL structure can be employed as a detector, known as an Interband Cascade Infrared Photodetector (ICIP) or Interband Cascade Detector (ICD). The convenience of this technology makes it particularly interesting. Indeed, unlike intersubband detectors, which require careful consideration of the incoming light’s polarization, ICIPs can operate with any polarization and can be illuminated from various directions. The initial efforts to utilize cascaded interband recombination for mid-infrared light detection involved using the same W-QW structure as ICLs.88 At this time, the detector was already working at room temperature and showed a very broad wavelength window up to 4\( \mu \)m. Unfortunately, the responsivity was still low in the order of tens of mA/W. The
challenge with this structure was that it necessitated the use of multiple photons in an ICIP to generate a single electron in the external circuit which limited a lot the responsivity. In subsequent years, efforts were made to enhance the efficiency of the ICIP through the development of new theories and approaches, such as the adoption of superlattice heterostructures. Experimental\textsuperscript{89} and theoretical\textsuperscript{90} studies have demonstrated that the limiting factors in previous designs, namely Johnson noise and shot noise, can be significantly reduced by utilizing a multistage absorber. This configuration involves the incorporation of alternating layers of different semiconductor materials with mismatched energy bandgaps within the ICIP structure. By confining the electron and hole wave functions to different materials, improved carrier confinement and reduced dark current can be achieved, leading to enhanced performance of the detector. Through studies, researchers have successfully optimized the thickness of individual absorbers and the number of cascade stages, taking into account the effects of light attenuation. By incorporating multiple cascade stages to minimize noise, these ICIPs have exhibited excellent device performance at high temperatures, surpassing commercial MCT detectors in terms of detectivities limited by Johnson noise.\textsuperscript{91} Regrettably, the large phonon scattering times, which are on the order of nanoseconds, imposes a limitation on the maximum bandwidth of these devices.\textsuperscript{92} This limitation arises from the effects of Shockley-Read-Hall and Auger recombination processes. However, it is worth noting that this characteristic enables the ICIP device to achieve higher peak responsivity and lower dark current compared to similar QCDs.\textsuperscript{93}

These advancements have opened up new possibilities for ICIP technology, enabling enhanced detection capabilities across a wider range of infrared wavelengths associated to

\textbf{Figure 2.12:} a) Simulation of modulation response of the ICL is shown for various pump currents, with a fixed number of quantum wells ($m = 5$). b) The 3-dB bandwidth is plotted as a function of the normalized modulated current. Courtesy of Deng\textsuperscript{83}
Figure 2.13: Pioneer demonstration by H. Lotfi et al., of an high speed Interband Cascade Infrared Photodetector. a) shows the great potential in terms of broad frequency response and responsivity. b) 3-dB bandwidth for an uncooled ICIP was 1.3 GHz, signifying the great potential of interband cascade structures for high-speed applications. Courtesy of Lotfi.  

a very high responsivity and a wide bandwidth in the GHz range, as depicted in the Figure 2.13a)-b). More recently, bandwidth up to 7 GHz associated with 60 mA/W and working up to 5 µm have been reported. Compared to the widely used MCT detectors, ICIPs also offer several advantages such as lower Auger recombination, reduced tunneling currents, and high uniformity in their growth. Using ICIPs in a FSO communication setup increases flexibility in design, allowing for the use of a wider range of wavelengths in transmitters. ICIPs offer a combination of fast detection, high responsivity, and a high saturation level when biased.

2.3 Mid-Infrared photodetector

2.3.1 Mercury-Cadmium-Telluride (MCT) detector

Mercury-Cadmium-Telluride (MCT) is a versatile compound semiconductor material widely used in various applications such as infrared sensing, communication, and imaging. It has maintained its position as the leading mid-infrared detector for many years, primarily due to its tunable bandgap, high optical absorption coefficient, excellent quantum efficiency, and long carrier lifetime. The concept of MCT was first proposed in 1954 by Lawson et al., who demonstrated the ability to control the bandgap of HgCdTe material in the infrared spectrum. By adjusting the composition of the HgCdTe alloy, the selectivity range of the detector can be tailored, ranging from near 1.5 µm to very-long-wavelength infrared...
at 30µm. The research also showed that this semiconductor material could provide both photoconductive and photovoltaic current in the infrared part of the spectrum. During the 1960s, the development of MCT detectors faced obstacles due to the confidential nature of research focused on the 8 – 12µm range, which was of high interest for military imaging. The first significant demonstration of MCT performance was made by Bartlett et al., who achieved background-limited performance of photoconductors operated at 77K in the long-wavelength infrared (LWIR) spectral region. However, MCT still faced challenges related to dark current noise. Subsequent improvements came with advancements in fabrication technology, particularly in the capacity to produce highly n-doped HgCdTe material. These developments led to enhanced device performance, including high quantum efficiency (>65 %) and fast response time (several ns, GHz range), allowing for operation across a broad frequency range. Nevertheless, HgCdTe infrared detector technology encounters various notable limitations, including limitations on array size, substantial dark currents at room temperature due to band-to-band tunneling, and material fragility. Despite these limitations, state-of-the-art Vigo peltier-cooled MCT detectors can achieve bandwidths of up to 1.2 GHz, along with responsivities of hundreds of mA per watt.

2.3.2 Uni-Traveling Carrier Photodetector Based on InAs/I-nAsSb Type-II Superlattice

Type-II superlattices (T2SLs) represent a significant advancement in mid-infrared detection, building upon the research on superlattice structures. These T2SLs are created by stacking InAs and GaSb monolayers on GaSb substrates. It utilizes interband transitions, making them suitable for mid-wavelength infrared detection up to 30µm. To achieve high-speed operations in the mid-infrared range, researchers have employed uni-traveling carrier (UTC) photodetectors based on InAs/GaSb T2SL. The UTC structure harnesses the rapid electron transport and minimal dielectric relaxation time of the hole carriers, resulting in exceptional device bandwidth performance. Consequently, this technology has found extensive application in the near-infrared band. In terms of performance, T2SLs and MCT detectors demonstrate similar capabilities, with the distinguishing factor being the ability of T2SLs to operate at higher temperatures. However, T2SL structures are prone to high non-thermal noise caused by trap-assisted tunneling. In recent research, Chen et al. have reported a high-performance T2SL InAs/GaAsSb detector operating under a -5 V negative bias. This detector exhibits a bandwidth of tens of GHz and achieves high responsivity of approximately 300 mA/W.
2.4 Mid-infrared detector for FSO transmission

Table 2.1: Comparison of most promising detectors for free-space communication. RT: Room Temperature. TM: Transverse Magnetic

<table>
<thead>
<tr>
<th></th>
<th>QWIP</th>
<th>QCD</th>
<th>ICIP</th>
<th>MCT</th>
<th>InAs/InAsSb T2SL</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power consumption</td>
<td>High (&lt;W)</td>
<td>None</td>
<td>High (&lt;W)</td>
<td>High (&lt;W)</td>
<td>High (&lt;W)</td>
</tr>
<tr>
<td>Operating temperature</td>
<td>RT (300 K)</td>
<td>None</td>
<td>High (&gt;W)</td>
<td>High (&gt;W)</td>
<td>High (&gt;W)</td>
</tr>
<tr>
<td>Cutoff wavelength (µm)</td>
<td>9</td>
<td>10.3</td>
<td>5.5</td>
<td>12</td>
<td>5</td>
</tr>
<tr>
<td>Responsivity (mA/W)</td>
<td>~ 100</td>
<td>~ 85</td>
<td>~ 240</td>
<td>~ 500</td>
<td>~ 600</td>
</tr>
<tr>
<td>Detectivity (Jones)</td>
<td>0</td>
<td>~ 10¹¹</td>
<td>3.10⁸</td>
<td>~ 10¹¹</td>
<td>10⁹</td>
</tr>
<tr>
<td>Bandwidth (GHz)</td>
<td>~ &gt; 90</td>
<td>~ 25</td>
<td>~ 7</td>
<td>1</td>
<td>11</td>
</tr>
<tr>
<td>Polarization section rules</td>
<td>Yes (TM)</td>
<td>Yes (TM)</td>
<td>No</td>
<td>No</td>
<td>No</td>
</tr>
<tr>
<td>References</td>
<td>55</td>
<td>63</td>
<td>97</td>
<td>105</td>
<td>110</td>
</tr>
</tbody>
</table>

Free-space communication relies on specific detectors to convert optical signals into electrical signals, and different types of detectors have distinct advantages and limitations in this context. In Table 2.1, we provide a summary of the state-of-the-art performances of the most promising mid-infrared detectors for free-space communication. Photodetectors such as MCT, InAs T2SL, and ICIP stand out for several reasons. Firstly, they exhibit high detectivity, which is crucial for building an effective FSO system. Due to high losses and dispersion in free-space transmission, the received power is typically low, making high detectivity essential for signal recovery. Secondly, these devices offer a broad frequency range, facilitating the implementation of wavelength division multiplexing (WDM) in FSO systems, enabling increased data rates. Furthermore, unlike intersubband devices, they do not have strict polarization selection rules, allowing for the development of polarization division multiplexing. However, these devices have limitations such as relatively high dark current (mA) and limited bandwidth (<12GHz). Additionally, ICIP and InAs T2SL detectors are currently unable to detect band 3 signals. On the other hand, devices based on intersubband recombination have unique features, including potential high bandwidth (100GHz) and wide wavelength accessibility. However, their polarization selection rules, relatively low responsivity, and limited frequency range make them suboptimal for free-space optical detection. Notably, QCDs have the advantage of passive operation and extremely low noise characteristics, making them highly suitable for FSO applications where noise is critical. It is important to consider the specific advantages and limitations of each detector. The choice of detector should be carefully evaluated based on the system requirements and
objectives.
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The advent of broadband communication in the early 2000s has revolutionized education, entertainment, and interpersonal communication by continuously enhancing data transfer speeds. However, certain regions still encounter challenges in accessing high-speed internet due to infrastructure limitations and the high costs associated with expanding wired networks. Free-Space-Optics (FSO) systems provide a rapid, dependable, and cost-effective solution for ground-to-ground communication and transmissions between the ground and low-orbit satellites. These advantages have been highlighted in the introduction. In this study, we will explore the methods we have implemented and the setup we have constructed for mid-infrared transmission.

### 3.1 Digital signal processing

In this section, we will provide a detailed description of the various tools we implemented to enhance the transmission quality. We will delve deeper into the algorithms that we developed and utilized for FSO transmission. A comprehensive overview of the processing setup, including every step, is presented in Figure 3.1.

Figure 3.1: Digital processing at the transmitter and receiver sides consists of several steps that aim to enhance and evaluate the quality of the system’s transmission.

Those algorithms have been developed on MATLAB and have been used for every following FSO setup that we built.
3.1.1 Signal generation and pulse shaping

The capacity of our system to transmit information is assessed using Pseudo-Random Binary Sequence (PRBS), which enables us to evaluate the performance of our system with various symbol sequences that are encountered in real-world communications. This approach ensures, by considering a wide range of symbol sequences, that we avoid overestimating the achievable data rates. In most of our studies, we generated $2^7$-bit and $2^{15}$-bit long PRBS signals for three different formats: Non-Return-Zero (NRZ), 4-level Pulse Amplitude Modulation (PAM-4), and 8-level Pulse Amplitude Modulation (PAM-8).\footnote{An example of PAM-4 PRBS signal at 10 Gbit/s can be seen in Figure 3.2.a).} Due to the limited bandwidth of our sources compared to the bandwidth of the generated signal, we have implemented filtering both before and after modulation. A commonly used pulse-shaping filter for this purpose is the Root-Raised-Cosine (RRC) filter, which pre-shapes our signal to reduce the bandwidth of the modulated signal.\footnote{The formula for the RRC filter is as follows:}

\[
H_{rrc}(f) = \begin{cases} 
\sqrt{T}, & (0 \leq |f| \leq \frac{1-p}{2T}) \\
\frac{T}{\pi} \left[1 + \cos\left(\frac{\pi f}{p} \left(|f| - \frac{1-p}{2T}\right)\right)\right], & \left(\frac{1-p}{2T} \leq |f| \leq \frac{1+p}{2T}\right) \\
0, & \left(|f| > \frac{1+p}{2T}\right)
\end{cases} \quad (3.1)
\]
where $\rho$ is the roll-off factor of the filter and $f$ is the frequency, $T$ is the reciprocal of the symbol-rate. The formula corresponds to the whole filtering process of the signal, with $\sqrt{H_{\text{rrc}}}$ implemented at the transmitter and at the receiver side. Different roll-off factors were used to reduce the spectral occupation of the transmitted signal and hence increasing the data rate of the transmission. The occupied bandwidth of our optical signal is defined by $B = \frac{1+\rho}{2} R_s$ in which $\rho$ is the roll-off factor of the filter and $R_s$ is the symbol rate. The roll-off factor satisfies the condition $0 \leq \rho \leq 1$ and an unfiltered signal corresponds to $\rho = 1$, thus, to the maximum bandwidth occupancy. In the Figure 3.2.b), we compare the spectral footprint of the unfiltered PRBS signal, an RRC filtering ($\sqrt{H_{\text{rrc}}}$) with a $\rho = 0.3$ and a $\rho = 0.9$. For the two latter, it means to decrease the bandwidth occupation of modulated optical signal by 35 % ($\rho = 0.3$) and 5% ($\rho = 0.9$), respectively, with the relationship: 
$$\frac{\Delta B}{\Delta B_{\text{unfiltered}}} = 1 - \frac{1+\rho^2}{2}.$$ 
As depicted in Figure 3.2, the eye diagram of the unfiltered PRBS signal differs from that of the RRC filtered signal due to the wider bandwidth of the unfiltered signal. The filtered signal exhibits overshoot between bits due to the significant bandwidth compression. It is important to note that at the receiver, the second square root of the RRC filter transfer function ($\sqrt{H_{\text{rrc}}}$) is used to compensate for this compression.

### 3.1.2 Feed-Forward Equalizer (FFE)

Feed-Forward Equalization (FFE) is a signal processing technique employed to address inter-symbol interferences (ISI) in communication systems. It involves applying a linear filter to the received signal, which compensates for the channel’s frequency response and mitigates the impact of ISI. The coefficients of the FFE filter are determined through adaptive algorithms such as the LMS (Least Mean Squares) or the Recursive Least Squares (RLS) algorithms. In collaboration with Elie Awwad from Telecom Paris, we developed an FFE algorithm. In our specific case, the filter coefficients are learned using a gradient descent algorithm with a convergence parameter ($\mu$) that controls the learning speed and accuracy. These adaptive algorithms continuously adjust the filter coefficients based on the received signal and the desired target response, often represented by known sent sequences. Once the filter is evaluated, the received sequence is filtered using the learned coefficients, which are configured to have a specific number of taps ($n_{\text{taps}}$). Each tap represents a specific delay and weight applied to the incoming signal. By equalizing the distorted signal, FFE enhances the signal-to-noise ratio, resulting in improved data transmission reliability and higher achievable data rates.

This filter can be seen in the top right part of, Figure 3.3 as well as the convergence of the filter (bottom). The efficiency of the transmission quality is optimized by adjusting
3.1.3 Decision-Feedback Equalization (DFE)

We also implemented a Decision Feedback Equalizer (DFE) that serves a similar purpose as the FFE but involves a more complex process. In the DFE, the received signal is initially equalized using an FFE to mitigate the effects of significant ISI. Subsequently, the filter is updated through a feedback loop. This feedback loop utilizes the previously detected symbol decisions to estimate and cancel the interference caused by post-cursor ISI. The feedback loop in the DFE leverages the past symbol decisions to generate an estimation of the interference affecting the current symbol. This estimation is then subtracted from the received signal to reduce the impact of ISI. By incorporating feedback, the DFE can...
Figure 3.4: Comparison of the two eye diagrams a) without and b) with FFE equalization $n_{\text{tap}} = 391$ and $\mu = 1 \times 10^{-5}$ at 30 Gbit/s with a $2^{15}$-bit long PRBS in back-to-back configuration. FFE corrects systematic channel defects and intersymbol interference (ISI), hence opening the eye.

Figure 3.5: Principle of the process of the Decision-Feedback Equalizer.

effectively cancel the interference caused by previous symbols, thereby improving the overall signal quality. The diagram of the DFE is depicted in the Figure 3.5. The choice of equalizer depends on several parameters, which will be listed in the following.

- **Complexity** The FFE is generally easier to implement compared to the DFE. Indeed the feedback loop introduces additional complexity and computational overhead, making the DFE more challenging to implement.

- **Latency** FFE exhibits lower latency compared to DFE. This is because FFE processes the signal in a single pass, while DFE introduces processing delay due to the feedback loop that relies on past decisions.

- **Performance** DFE offers superior performance in handling specific types of ISI, especially those arising from post-cursor intersymbol interference. By leveraging the
feedback loop, DFE can effectively eliminate interference from previous symbols, making it highly effective in challenging scenarios. Especially, DFE is particularly interesting when the characteristics of the channel are undergoing rapid changes.

- **Robustness** DFE can be more susceptible to errors in previous decisions, as inaccuracies in decision making can propagate through the feedback loop, resulting in error propagation. In contrast, FFE does not depend on past decisions and is generally more robust in handling such errors.

In summary, FFE is generally favored when simplicity and low latency are crucial, particularly for mitigating pre-cursor intersymbol interference. On the other hand, DFE is preferred in scenarios where post-cursor ISI is prominent and higher performance is desired, even at the expense of increased complexity and latency. The specific requirements and trade-offs of the communication system should be carefully evaluated to determine the most suitable equalization technique.

### 3.1.4 Forward error correction

Forward Error Correction (FEC) is an error correction method that can identify and fix a limited number of errors in transmitted data without requiring retransmission. This involves sending a redundant error-correcting code along with the data frame from the sender, with the receiver conducting necessary checks using the additional redundant bits. The transmission system’s design must account for the overhead introduced by the redundancy of these extra bits. There are two types of FEC algorithms: Soft-Decision (SD) and Hard-Decision (HD) algorithms. Usually, SD introduces more latency as well as a higher BER threshold compare to the HD one. In most of our study, we used the Hard-Decision 7% and 27% which respectively require BER under 0.4% and 4%. For a maximum pre-FEC BER of 4 %, a FEC code with a 27 % overhead and HD decoding is assumed to be used to achieve error-free communication, while a pre-FEC BER of 0.4 % requires an FEC code with an overhead of only 7 %. In the following, we define error-free communication when we achieve a BER below the maximum allowable limit set by the FEC algorithm on the recorded sequence. When the BER is below this threshold, the FEC algorithm can correct all errors, resulting in zero remaining errors after correction. It is important to note that errors are assessed over a finite time scale, which means that additional errors may occur when extending the recorded time traces.
Figure 3.6: Laboratory demonstrations presented in this summary highlight the successful utilization of quantum/interband cascade devices for enabling free-space transmissions. Among the experiments conducted, the ones represented by the red circle stand out as they involved the use of a source operating at room temperature.

3.2 Results on mid and long infrared free-space transmission

In this part, we will draw a picture of the current mid-infrared transmission results. The results obtained in the mid-infrared from 2001 to 2022 are summarized in Figure 3.6. Interestingly, results on Band 3 transmission emerged before the one in band 2 because of the military industry’s interest and focus on this particular wavelength range.

3.2.1 Band 2 transmission: 3 – 5 µm

In the early stages of utilizing cascaded technologies for FSO communication, a significant milestone was achieved in 2010. Researchers successfully implemented an FSO link using an ICL operating in the mid-infrared region. As depicted in Figure 3.7, they achieved a transmission rate of 70 Mbit/s for an ICL operating at 77 K with a BER below $10^{-8}$ using a 70 MHz 3dB-bandwidth MCT.\(^2\) They also used a QWIP detector, but MCT was preferred for the transmission experiment because of the high responsivity. Although this work was a significant first step towards high-speed transmission, it was strongly limited by the need for cryogenic temperatures, making it impractical for most applications. Although ICLs have lower output power and a limited achievable wavelength (up to 6 µm at room temperature), they gained attention for their potential in energy-efficient communication systems.

The optimization of link capacity came with the use of QCLs. In recent advancements, state-of-the-art transmission rates of 4 Gbit/s were achieved in the 3 – 5 µm wavelength range.
range using QCLs,\textsuperscript{4} which were later improved to 6 Gbit/s.\textsuperscript{5} These milestones were accomplished using a specific QCL operating at 4.65\,\textmu m with a 350 MHz 3 dB-bandwidth. However, achieving higher data rates remains a challenge due to the limited bandwidth of mercury cadmium telluride (MCT) detectors. Otherwise, researchers have directed their focus towards devices capable of generating and detecting high-bandwidth signals in the C-band, with the aim of performing wavelength up-and down-conversion to reach the mid-infrared band. The C-band devices have undergone extensive optimization over the years, enabling them to be modulated at high speeds\textsuperscript{6} and to support advanced modulation formats\textsuperscript{7} such as phase modulation or orbital angular modulation. Leveraging the well-established development of lasers, detectors, and modulators in the C-band, this approach also facilitates the use of wavelength division multiplexing (WDM) techniques to increase the capacity of mid-infrared links. To achieve wavelength conversion, various techniques are employed, including four-wave mixing (FWM),\textsuperscript{8} cross-phase modulation (XPM),\textsuperscript{9} and optical parametric amplification (OPA).\textsuperscript{10} One early successful demonstration of effective wavelength conversion from 1.5\,\textmu m to 3.8\,\textmu m was conducted by Cho et al.\textsuperscript{11} in 2009 using a Ti:PPLN-based nonlinear crystal. They studied the transmission of a 160 Mbit/s IQ RZ signal and subsequent research has shown the possibility of transmitting Gbit/s signals using these methods. A recent breakthrough showcased the utilization of the high-capability 1.5\,\textmu m range to transmit signals in the 3−5\,\textmu m windows by using the combination of WDM and OAM. In this particular study, three wavelengths around 3.4\,\textmu m (3.396\,\textmu m, 3.397\,\textmu m, and 3.398\,\textmu m) were employed on a single polarization, with each wavelength...
Mid-infrared (mid-IR) wavelength-division-multiplexing (WDM) signals can be generated by converting C-band signals into the mid-IR range using the difference-frequency generation (DFG) process. After this conversion, the signals are detected at the C-band again through another DFG process. In addition, mid-IR orbital-angular-momentum (OAM) beams can be generated by passing mid-IR Gaussian beams through spiral phase plates (SPPs). Courtesy of Zou.

Figure 3.8: Mid-infrared (mid-IR) wavelength-division-multiplexing (WDM) signals can be generated by converting C-band signals into the mid-IR range using the difference-frequency generation (DFG) process. After this conversion, the signals are detected at the C-band again through another DFG process. In addition, mid-IR orbital-angular-momentum (OAM) beams can be generated by passing mid-IR Gaussian beams through spiral phase plates (SPPs). Courtesy of Zou.

Carrying two orbital angular momentum (OAM) beams as depicted in the Figure 3.8. By transmitting 50 Gbit/s quadrature phase shift keying (QPSK) data on each frequency with different angular momenta, a total data rate of 300 Gbit/s was achieved. However, nonlinear wavelength conversion from near-infrared to mid-infrared still faces certain limitations highlighted in previous studies, including low output power (in the milliwatt range), high power consumption, low efficiency, operation in the sub-4 µm region, and the need for expensive and bulky periodically poled lithium niobate devices. Despite these challenges, ongoing research aims to overcome these limitations and unlock the full potential of nonlinear wavelength conversion in achieving extremely high data rates in the mid-infrared range.

### 3.2.2 Band 3 transmission: 8 – 14 µm

Despite experiencing lower detector efficiency and requiring larger optics for free space communication, band 3 has garnered considerable attention due to its resilience in degraded atmospheric conditions and its stealthy nature. Initial results in this band were achieved using cryogenically cooled QCLs. Martini et al. demonstrated a successful free space transmission of 2.5 Gbit/s at 8.1 µm at a temperature of 85 K, using a cryogenically...
cooled MCT. The result is depicted in the Figure 3.9. They have also successfully established a transmission link using QCL operating at a wavelength of approximately 9.3 µm over a distance of nearly 10 meters, achieving a data rate of 115 kbit/s. Notably, even when the link was extended to a distance of 300 meters under foggy conditions, the transmitted power output remained around 80% of the back-to-back value. This observation highlights the resilience of the 9.3 µm wavelength to adverse effects caused by humidity. However, the progress in this field has been hindered by the limitations associated with cryogenic operation of QCLs. In 2019, a significant milestone was achieved with the demonstration of 1 Gbit/s transmission at room temperature using electrical direct modulation at 10.6 µm. A recent advancement showcased the transmission of 11 Gbit/s at 9.6 µm using a directly-modulated QCL coupled with computer-assisted processing to enhance the performance of the QCD. To overcome the bandwidth limitations, external modulators have emerged as a key technology, which explains why the adoption of QCL-based communications for long-haul free-space transmissions has been limited. Prior to my research, pioneering work demonstrated the feasibility of utilizing a Stark effect external modulator for free space transmission, achieving a transmission rate of 10 Gbit/s over a distance of 1 meter at 9 µm. As depicted in Figure 3.6, there has been a significant upward trend in data rates achieved at room temperature since 2015. This suggests that in the coming years, we can anticipate the attainment of data rates in the hundreds of Gbit/s at room temperature.

**Figure 3.9:** Pioneer demonstration of a transmission with an high speed cooled Quantum Cascade Laser. a) 3 dB-bandwidth characterization estimated at 7 GHz at 85 K. b) Detected signal for a data rate of 2.5 Gbit/s in a back to back configuration. Courtesy of Martini.
3.3 Interband cascaded technologies (3-5 µm)

In the introduction, we have presented the basic principles of Interband Cascade Laser (ICL). This technology appears to be very suitable for various applications where there is the need of easy implementation, low current consumption and high reliability. Interestingly, integration of this mid-infrared technology can be further developed with ICLs on silicon for on-chip sensing technologies. Application of ICLs has been extended to free-space communication between 3 – 5µm. However, further investigations need to be conducted to thoroughly assess the feasibility of using these sources for high-speed optical transmission. Early efforts in experimenting with high speed based ICLs have demonstrated modulation bandwidths of over 3 GHz in a cryogenic ICL. Unfortunately, devices operating at cryogenic temperature are not very suitable for broad use of this technology on the real field. In this section, we investigated the suitability of this technology for high-speed mid-infrared transmission. Despite progress in experimental applications of ICLs, there is still limited understanding of their intrinsic properties and more research is needed. A recent study examined the differential gain and gain compression factor of ICLs. However, the conclusions drawn from this study were inconsistent with the aforementioned results regarding the high-speed capabilities of ICLs. In this study they exhibited a low 3dB-bandwidth around 300 MHz. Additionally, the study did not observe the expected relaxation oscillations, which are typically present in a class B laser.

3.3.1 Initial findings on transmission system using unoptimized RF ICL

In this first attempt of FSO transmission, we employed a non-optimized RF ICL. This device is developed at Université de Montpellier. The ICL under investigation in this study has a threshold current of 71 mA at 293 K and a maximum output power of 5.5 mW at 6.5 V and 250 mA. It has been mounted in a QCL mount LDM 4872 to be able to provide both current and temperature stabilization. The laser is operated in CW mode using a low-noise current source (Wavelength Electronics QCL2000). The DC and AC signals are combined using a bias-tee before connecting to the laser’s probes.

3.3.1.1 Bandwidth characterization

The first step was to evaluation the bandwidth of our system. We conducted optical and electrical rectifications. This method allows for the measurement of the electrical bandwidth without the need for a high-bandwidth optical detector, and it is commonly used
in the mid-infrared range where fast room-temperature detectors are not yet available. This method has been further described in the appendix B. We also evaluated the optical bandwidth by modulating the bias current with a sine wave and measuring the response with a 700 MHz 3dB-bandwidth mid-infrared detector. The schematic representation of the setup can be seen in Figure 3.10. We studied the optical and electrical response of the non-optimized ICL, as depicted in the Figure 3.11.a). We observed that the electrical contact limits the maximum possible bandwidth of the ICL. In fact, there are several drops in the electrical response and a cut-off bandwidth around 300 MHz, which causes drops in the optical response. This is mainly due to the high impedance mismatch caused by the non-RF optimized contact.

3.3.1.2 Transmission experiment

Despite the quite low bandwidth, we still performed a transmission for two different modulation formats: non-return-to-zero (NRZ) and return-to-zero (RZ) with a 40 % duty cycle. The NRZ signal is generated using a pattern generator (Hewlett Packard 70841 A) with data rates ranging from 50 Mbits/s to 3000 Mbits/s. On the other hand, the RZ signal is produced by an arbitrary waveform generator (AWG Tektronix 7122B) with a sampling rate of 12 GS/s and an analog bandwidth of 3.5 GHz. Each symbol in the RZ format is coded with 10 samples, resulting in a maximum symbol rate of 1200 Mbits/s. For both the NRZ and RZ formats, a 127-bit long PRBS sequence is used. The optical beam is directed towards an 400 MHz 3dB bandwidth MCT detector located one meter away from
Figure 3.11: Optical rectification (in red) and the electrical rectification (in blue): a) non RF-mounted ICL for a temperature 293 K. b) Rf mounted ICL for a supply current of 125 mA and a temperature of 293 K.

The AC output signal from the MCT detector is then captured by a low-noise oscilloscope (Tektronix MSO64) with a maximum sampling rate of 25 GS/s and an analog bandwidth of 2.5 GHz. In the case of the NRZ format, the signal is also analyzed using an error detector (Hewlett Packard 70004 A).

Figure 3.12: Eye diagram for two different modulation formats: a) 110 Mbit/s transmission using a 2 level format in NRZ configuration. b) 300 Mbit/s transmission using a 2 level format in RZ configuration. In both cases the eye is open, suggesting a good discrimination between bit 0 and bit 1.

The eye diagrams for NRZ and RZ transmissions can be seen in Figure 3.12. We achieved successful NRZ transmission at 110 Mbit/s with a low BER compliant with 7% HD FEC. Interestingly, the RZ modulation format yielded even better results, thanks to the laser’s...
response that exhibited significant attenuation drops around 40 MHz and 200 MHz, acting as a bandpass filter. We demonstrated state-of-the-art performance by achieving data rates of 300 Mbit/s using the RZ format while maintaining a BER compliant with 7% HD-FEC.

3.3.2 RF mounted Interband Cascade Laser

Due to the electrical limitations of the ICL, we encountered constraints in achieving high data rate for our FSO transmission. Consequently, a collaborative effort was undertaken with Laurent Cerutti in Université de Montpellier to investigate the implementation of an RF electrical connector, aiming to enhance the high-speed capability of the ICL. The 4.18 µm Fabry-Perot ICL being studied is designed and produced at Université de Montpellier. Band diagram of the laser can be found in Figure 3.13.a). The structure is grown using molecular beam epitaxy on a GaSb substrate. The active region consists of seven-stage cascade periods of type-II "W" Quantum-Well (QW). The active region is surrounded by GaSb optical confinement layer and a cladding of AlSb/InAs superlattices. The ICL was then attached to a copper block using indium and placed in a package with a direct connection between the SMA central contact and the laser ridge. A SMA-PCB end launch connector (Radiall R125539000) with a bandwidth of 18 GHz was used for this purpose, which is suitable for radio frequency modulation. The electrical and optical rectification responses of the RF optimized ICL are displayed in Figure 3.11.b). The two 3dB-bandwidths are consistent and are approximately 700 MHz at 288 K. Additionally, the correlation between the electrical and optical bandwidths suggests that the mount of the ICL does not limit the modulation frequency. However, we did not observe any clear indication of relaxation oscillations with these two methods. We opted to assess the Relative Intensity Noise of the RF-mounted ICL specifically to evaluate the relaxation oscillation frequency and intrinsic parameters. Remarkably, when comparing the results with Figure 3.11, it becomes evident that the optimization of the electrical contact has significantly improved the high-speed capability of the ICL.

3.3.2.1 Relative Intensity Noise experiment (RIN)

The evolution of the RIN as a function of the frequency has been studied. Spontaneous noise is the primary source of noise for optical sources, causing variations in the optical output power. Overcoming these noise limitations is crucial in order to enhance the quality of laser transmission as already discussed in the introduction. Moreover, the relaxation oscillations, which are a limiting factor for high-speed applications, are observed in ICLs due to their class B regime. In this regime, the upper state lifetime is typically much longer.
Figure 3.13: a) Schematic representation and simplified band diagram of our RF optimized Interband Cascade Laser at 4.18 µm. b) ICL Light-intensity-voltage (LIV) curve under continuous wave pumping and a temperature of 288 K. c) Optical spectrum of the ICL at 100 mA and at a temperature of 288 K. The multimode spectrum shows two separate lobes around 4.18 µm than the cavity damping time.\(^{27}\) RIN is defined as the ratio of the power spectral density of the intensity noise to the squared average optical power:

\[
RIN = \frac{\langle \delta P(t)^2 \rangle}{\langle P \rangle^2} \quad (3.2)
\]

As shown in Figure 3.14, the ICL was kept at a constant room temperature and a continuous bias was provided by a low-noise current source (LDX - 3232 CW QCL Driver). The ICL has a threshold current of 75 mA at 297 K and a maximum beam power of 15 mW per facet at 185 mA and 4.3 V as depicted in the Figure 3.13.b), resulting in a wall-plug efficiency of 4%. As experiments were conducted using a multimode laser (Figure 3.13.c)), the mode spacing was evaluated, yielding an average material index of \(n_g = 3.5\). The optical signal generated by the ICL is focused by a lens onto a Mercury-Cadmium-Telluride detector (MCT-VIGO UHSM-10.6) with a 3dB-bandwidth of 1.3 GHz and a 50 Ω loadout resistance. The 50µm × 50µm active area of this detector allows it to detect mid-infrared signals as low as a few hundreds of µW. The resulting electrical signal can be accessed.
Figure 3.14: Schematic representation of the setup. In the experiment, an ICL operating at room temperature is used in conjunction with an MCT detector positioned approximately 20 cm away from the emitting facet. The DC component of the signal is extracted to assess shot-noise, while the AC part enables the evaluation of the RIN spectrum. Analysis of the RIN signal allows for the determination of the relaxation frequency, which is typically in the GHz range.

through two SMA ports: a DC port for frequencies up to 300 Hz and an AC port for frequencies from 10 kHz. The AC signal is amplified by a low-noise amplifier with 25-dB gain and 25 GHz bandwidth (SHF-826H) before being analyzed by a 43 GHz bandwidth electrical spectrum analyzer (ESA-FSU Rohde & Schwarz). For each measurement, the RF response needs to be assessed to determine the noise contribution in both setups: one with illumination on the detector from the laser and the electronic spectrum analyzer (ESA), and the other without such illumination. After, we perform difference to extract only the contribution of the laser on the RF response and get rid of the thermal noise or other noises coming from the different components of the setup. This corresponds to the first term in Equation 3.3. On the other hand, the DC signal directly gives a first evaluation of the shot noise of the detector, which corresponds to the second term in Equation 3.3. In this equation, \( S_p(f) = S(f) - S_{dark}(f) \) represents the difference between the noise caused by the entire setup \( S(f) \), and the noise caused by the setup without illumination, \( S_{dark}(f) \), which is considered as white noise. \( B \) is the resolution bandwidth of the ESA, \( G \) corresponds to the gain of the amplifier, \( R \) is the resistance of the detector, \( e \) represents the electric charge carried by an electron, and \( I_{ph} \) is the photovoltaic current of the detector. We used a homemade MATLAB program to process the experimental data.

\[
RIN(f) = \frac{S_p(f)}{BGRI_{ph}^2} - \frac{2eB}{<I_{ph}>}. \tag{3.3}
\]
Figure 3.15: Relative intensity noise of the ICL at 295 K for various bias currents, from $1.8 \cdot I_{th}$ to $2.3 \cdot I_{th}$. Dashed green lines correspond to the fitting of the resonance associated with the relaxation oscillation. It leads to the determination of the relaxation frequency around the GHz.

Previous studies have reported RIN levels around -130 to -150 dB/Hz\textsuperscript{20,21} but due to a narrow detector bandwidth of 400 MHz, did not exhibit any clear relaxation oscillation frequency. The RIN analysis for a wide frequency range is shown in Figure 3.15. We observe the evolution of RIN for five different currents above $1.8 \cdot I_{th}$ at a constant temperature of 295 K. At low frequencies around 10-100 MHz, the RIN level is significantly high, in the range of -95 dB/Hz, and drops as frequency increases. As the ICL under study is multimode, various factors such as technical, thermal, or partition noise can impact the noise level at low frequencies.\textsuperscript{28} Additionally, for this frequency range, the RIN increases as bias current increases, likely due to the appearance of new modes in the optical spectrum leading to an increase in partition noise. Around 100-500 MHz, the impact of these factors diminishes, and the trend is reversed, with RIN decreasing as bias current increases, consistent with prior results.\textsuperscript{20,21} By further increasing the frequency, the relaxation oscillation of the ICL is clearly observed in the GHz range. As expected, Figure 3.15 shows that the oscillation frequency increases with bias current. In this case, the relaxation oscillation was not observed for pump currents below $1.52 \cdot I_{th}$ due to other substantial noise contributions and large damping factors. Let’s stress that the noise is not influenced by bias current above the relaxation oscillation peak. The estimated shot noise of our detector is around -135 dB/Hz. We will see hereafter the curve fitting of the RIN reveals some important intrinsic lasers parameters.

3.3. Interband cascaded technologies (3-5 µm)
Table 3.1: List of ICL’s parameters used in the numerical simulations and in the fitting of the RIN.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cavity Length L</td>
<td>3 mm $^i$</td>
<td>Photon lifetime $\tau_p$</td>
<td>7.6 ps $^i$</td>
</tr>
<tr>
<td>Cavity width W</td>
<td>10 $\mu$m $^i$</td>
<td>Spontaneous emission time $\tau_{sp}$</td>
<td>15 ns $^i$</td>
</tr>
<tr>
<td>Facet reflectivity R</td>
<td>0.32 $^i$</td>
<td>Auger lifetime $\tau_{aug}$</td>
<td>1.08 ns $^i$</td>
</tr>
<tr>
<td>Refractive index $n_r$</td>
<td>3.5 $^i$</td>
<td>Differential gain $a$</td>
<td>$6 \times 10^{-17}$ cm$^2$ $^{ii}$</td>
</tr>
<tr>
<td>Optical confinement factor $\Gamma_p$</td>
<td>0.23 $^i$</td>
<td>Transparent carrier number $N_{tr}$</td>
<td>$10^8$ $^i$</td>
</tr>
<tr>
<td>Internal loss $\alpha_i$</td>
<td>5 cm$^{-1}$ $^{iii}$</td>
<td>Spontaneous emission factor $\beta$</td>
<td>$10^{-3}$ $^{iv}$</td>
</tr>
<tr>
<td>Stage number $m$</td>
<td>7 $^i$</td>
<td>Current injection efficiency $\eta$</td>
<td>0.34 $i$</td>
</tr>
</tbody>
</table>

$^i$ ICL data, $^ii$ Evaluated from fitting, $^{iii}$ Extrapolated from a similar structure emitting at 3.3 $\mu$m $^{29,30}$, $^{iv}$ ICL data from literature

3.3.2.2 Determination of intrinsic parameters of the ICL

To determine the intrinsic properties of the ICL, we need to fit the relaxation oscillation frequency. This can be achieved by performing a small signal analysis on the rate equations of the laser, as outlined in previous works by Coldren et al. and Deng et al. $^{21,27}$ The phase dynamics are described by the following equation:

$$\frac{d\phi}{dt} = \frac{\alpha_H}{2}(m\Gamma_p v_g g - \frac{1}{\tau_p}).$$

(3.4)

where $\Gamma_p$ is the optical confinement factor, $v_g$ is the group velocity, $\alpha_H$ is the Linewidth Enhancement Factor (LEF). The Langevin noise model, as described in Appendix 13 of Coldren et al., $^{7}$ can be employed for the RIN analysis. The noise sources outlined in this model are derived from the shot noise model developed by McCumber $^{31}$ and Lax. $^{32}$ This model provides a simplified representation of noise in lasers by attributing it to the discrete random movement of particles entering and leaving the carrier and photon reservoirs. It is very complicated to simulate each contribution of different noise process. Within the Langevin formalism, each process contributes to the shot noise of the overall noise observed in the reservoir. To determine the overall Langevin noise spectral density or correlation strength $\langle F_i F_i \rangle$, we aggregate the contributions of shot noise from all particle flow rates into and out of reservoir i. Similarly, when evaluating the cross-correlation strength $\langle F_i F_j \rangle$ between reservoirs i and j, we consider only the particle flows that affect both reservoirs simultaneously. The particle flows in the photon reservoir caused by the carrier noise $F_N$, the photon noise $F_s$ and the photon phase noise $F_\phi$. $^7$ These noise sources are assumed to
be white noise and their amplitudes are considered small enough to allow for the use of differential rate equations:

\[
\begin{align*}
F_N.F_N^* &= 2(\frac{\beta_{NS}}{\tau_{sp}} + \frac{N}{\tau_{sp}} + \frac{N}{\tau_{aug}}) \\
F_S.F_S^* &= 2m\frac{\beta_{NS}}{\tau_{sp}} \\
F_{\phi}.F_{\phi}^* &= m\frac{\beta_{NS}}{2\tau_{sp}} \\
F_N.F_S^* &= -2\frac{\beta_{NS}}{\tau_{sp}} + G_0(N - N_{tr})S \\
F_N.F_{\phi}^* &= F_S.F_{\phi}^* = 0
\end{align*}
\]  

(3.5)

The parameters of this set of equations have already been defined in the Section 2.2.1.1. The Langevin noise sources described above cause disturbances in the laser system. We can apply small-signal analysis to rate Equation 3.4 and the following equation:

\[
\frac{dN}{dt} = \eta I - \Gamma_p v_g g S - \frac{N}{\tau_{sp}} - \frac{N}{\tau_{aug}}.
\]  

(3.6)

and

\[
\frac{dS}{dt} = m\Gamma_p v_g g S - \frac{S}{\tau_p} + m\beta N.
\]  

(3.7)

By considering the Equation 3.4, Equation 3.6, and Equation 3.7, we can derive the linearized rate equations system:

\[
\begin{bmatrix}
\gamma_1 + j\omega & \gamma_2 & 0 \\
-\gamma_1 & \gamma_2 + j\omega & 0 \\
-\gamma_3 & 0 & j\omega
\end{bmatrix}
\begin{bmatrix}
\delta N(\omega) \\
\delta S(\omega) \\
\delta \phi(\omega)
\end{bmatrix}
= \begin{bmatrix}
F_N(\omega) \\
F_S(\omega) \\
F_{\phi}(\omega)
\end{bmatrix}.
\]  

(3.8)

The matrix coefficients can be written as:

3.3. Interband cascaded technologies (3-5 \textmu m)
\[
\begin{align*}
\gamma_{11} &= G_0 S + \frac{1}{\tau_p} + \frac{1}{\tau_{aug}} \\
\gamma_{12} &= G_0 (N - N_{tr}) \\
\gamma_{21} &= m (G_0 S + \frac{\beta}{\tau_p}) \\
\gamma_{22} &= \frac{1}{\tau_p} - m G_0 (N - N_{tr}) \\
\gamma_{22} &= \frac{1}{2} m \alpha_H G_0 \tag{3.9}
\end{align*}
\]

By applying the Cramer’s rule in the Equation 3.8, the RIN can be expressed as a function of angular frequency \( \omega \). The detailed calculation to obtain the following result can be retrieved in the appendix A:

\[
RIN(\omega) = \frac{| \delta S(\omega) |^2}{S^2} = \frac{a_1 + a_2 \omega^2}{S^2 \omega_R^4} |H(\omega)|^2. \tag{3.10}
\]

Here, \( \omega_R \) is the angular relaxation frequency, \( S \) is the total photon number, \( a_1, a_2 \) depend on intrinsic parameters of the ICL. The \( a_1 \) and \( a_2 \) can be written as:

\[
\begin{align*}
a_1 &= \frac{2m \beta NS}{\tau_p} (G_0 S + \frac{1}{\tau_p} + \frac{1}{\tau_{aug}})^2 (G_0 S + \frac{1}{\tau_p} + \frac{1}{\tau_{aug}}) (2S \omega_R^2 + 2m \tau_p^2 N \omega_R^4) \\
a_2 &= \frac{2m \beta NS}{\tau_p} \tag{3.11}
\end{align*}
\]

\( H \) is the modulation transfer function of the ICL taking into account the damping factor \( \gamma \), such as:

\[
H(\omega) = \frac{\omega_R^2}{\omega_R^2 - \omega^2 + j \omega \gamma}. \tag{3.12}
\]

The use of Equation 3.10 and Equation 3.12 allows for the fitting of the experimental RIN data using a least-squares method. The values chosen for the fitting of the ICL curve can be found in the Table 3.1. This fitting can also be used to determine intrinsic parameters such as the angular relaxation frequency, \( \omega_R \), and the damping factor \( \gamma \). According to theory, the squared oscillation relaxation frequency linearly increases with the current offset, as outlined in the book by Coldren et al.\textsuperscript{27}:

\[
f_R^2 = \frac{\Gamma_p V_{gs}}{4\pi^2 eV} \eta_i (I - I_h). \tag{3.13}
\]
where \( a \) is the differential gain, \( e \) the charge of an electron, \( \eta \) the injection efficiency, and \( V \) the volume of the active area whereas \( I, I_{th} \) are the bias current and the threshold current of the laser, respectively. By plotting the squared resonance frequency against the current offset \((I-I_{th})\) as shown in the Figure 3.16.a) and performing a linear fit at 293 K, we were able to calculate the differential gain in the ICL. A high differential gain, coupled with a small active region, is a crucial material parameter for improving the laser’s high-speed performance. Our results showed a differential gain of \( 6 \times 10^{-17} \) cm\(^2\) at 293 K, which is consistent with values reported in previous studies of QW laser (for instance, \( a=1.2 \times 10^{-16} \) cm\(^2\) in Ref. 33). The \( \gamma \)-factor is defined as:\(^{27}\)

\[
\gamma = K f_R^2 + \gamma_0.
\]  

(3.14)

where \( f_R \) represents the relaxation oscillation frequency and \( \gamma_0 \) the damping offset, which is inversely proportional to the carrier lifetime. Equation 3.14 gives the possibility to evaluate the value of \( K \) and \( \gamma_0 \) for our ICL. In Figure 3.16.b), the damping factor is plotted as a function of the squared resonance frequency at 293 K. Another parameter of interest for laser high-speed operation is the K-factor, which is a marker of the damping response. Large \( K \) prevents high-speed modulation, but in some cases, appropriate damping suppresses overshoot due to relaxation oscillations. From the linear fit, \( K=3.7 \) ns which gives a maximum modulation bandwidth of 2.4 GHz according to the formula \( 2\sqrt{2\pi} / K \).\(^{21}\) This result shows that ICLs are not limited to low-frequency operation.\(^{22}\) The damping factor offset \( \gamma_0=1.5 \) ns\(^{-1}\) leads to an effective carrier lifetime of 0.6 ns. Furthermore, we can also derive the gain compression factor \( \varepsilon \). This parameter may be used to describe the reduction of the optical gain above threshold due to other processes, such as spectral hole burning and carrier heating.\(^{7}\) In order to retrieve the gain compression factor, the following relationship can be used:

\[
\varepsilon = \left( \frac{K}{4\pi} - \tau_p \right) v_g a.
\]  

(3.15)

The compression gain also accounts for the capability of a laser to be modulated at large output power. Here, using the previous findings, a value of \( \varepsilon = 4.7 \times 10^{-17} \) cm\(^3\) is extracted at 293 K. This result is in good agreement with a prior work about interband lasers where a compression factor \( \varepsilon \) of \( 1 \times 10^{-17} \) cm\(^3\) was exhibited.\(^{34}\) On top of that, it is relevant to highlight that the compression factor reported in this work is similar to that of QW lasers.

3.3. Interband cascaded technologies (3-5 \( \mu \)m)
3.3.3 Energy efficient mid-infrared free-space communication based on cascaded interband transition.

After obtaining these initial transmission results, which already represented a state-of-the-art data rate, we build a system of communication with the RF-mounted Interband Cascade Laser already studied in the previous part. The laser has been associated with an Interband Cascade Infrared Photodetector (ICIP). This detector was designed and manufactured at TU Wien by Hedwig Knötig under the supervision of Benedikt Schwarz. The ICIP was grown by molecular beam epitaxy (MBE) on a compensation-doped GaSb substrate \((n \leq 5 \times 10^{16} \text{ cm}^{-3})\) to facilitate reduced free-carrier absorption when illuminating the detector from the substrate side. A 500 nm thick GaSb bottom contact was grown with a doping of \(n = 1 \times 10^{18} \text{ cm}^{-3}\), since the compensation doping lowers the conductivity of the substrate. The bottom contact is followed by three active stages, that are each made of the electron barrier, the active region based on a type-II superlattice and the hole barrier, in this order, as shown in Figure 3.17.a). The absorbers consist of 54 superlattice periods of 1.5 nm GaSb/0.8 nm AlSb/1.5 nm GaSb/2.7 nm InAs, following a design previously reported by Lei et al.\(^{35}\) The entire active region was left undoped. In order to balance the strain in the structure, 0.15 nm thin InSb layers were grown at the InAs interfaces. This modification leads to a redshift of the cutoff wavelength of about 500 nm compared to the original design. The electron and hole barriers consist of four GaSb/AlSb pairs and eleven InAs/AlSb pairs, respectively. The highly doped \((n = 2 \times 10^{18} \text{ cm}^{-3})\) top contact was realized as a lattice matched InAs\(_{0.91}\)Sb\(_{0.09}\) layer of 50 nm thickness. Transition layers were employed towards the bottom and top contacts to avoid potential barriers at
the interface and allow for a smooth transition of the carriers. Circular detector mesas with a diameter of 75 µm were fabricated using reactive ion etching. A Si₃N₄ passivation layer is deposited and selectively removed on top, where the Ti/Au top contact is realized afterward. Forming an ohmic contact to the n-GaSb bottom contact, Ge/Au/Ni/Au is deposited and thermally annealed. Subsequently, the mesas are mounted on a copper substrate with a slit to facilitate back-side illumination and connected to a coplanar waveguide and an end-launch connector. The responsivity is displayed in Figure 3.17b), exhibiting a rather flat spectral response between 2 - 4.5 µm. At the lasing wavelength of 4.18 µm, the responsivity is evaluated to be around 130 mA/W. As shown in Figure 3.17c), the emission wavelength of the ICL is located near the peak detection of the ICIP. In addition, the Intensity-Voltage (I-V) response of the ICIP as a function of the size of the detector is shown in Figure 3.17b). The 75 µm-ICIP exhibits a dark current around 1.6 mA at 4.5 V.

Figure 3.17: Characteristics of interband cascaded devices. a) Schematic representation of the band alignment and SEM picture of the heterostructure of the ICIP detector. b) Dark current density characterization for four different mesa sizes from 25 µm to 100 µm. With a 4.5 V bias and at 20 °C, the 75 µm-ICIP shows a dark current of 1.6 mA. When illuminated by the 5 mW ICL beam, the photocurrent is around 2.3 mA. c) Measured responsivity of the ICIP under zero bias operation at room temperature (yellow curve) together with the normalized emission spectrum of the ICL at a temperature of 20°C and a drive current of 100 mA (red curve). The detector is well suited for the emission wavelength of the source around 4.18 µm.
3.3.3.1 Setup description and characterization

The experimental setup for free-space transmission is shown in Figure 3.18. First, we determined the optimal ICL parameters to minimize non-linearity and achieve the best modulation depth. This is a balance between the bandwidth and thermal parasitic effects, which both increase with the bias current of the laser, and a sufficient laser output power. The optimal configuration was found to be a bias current of 140 mA at a temperature of 20°C. For high-speed electrical modulation of the ICL, we used a Socionext arbitrary waveform generator (AWG) with a sampling rate of 120 GS/s and an analog 3dB-bandwidth of 30 GHz. We generated sequences for three different modulation schemes: On-Off Keying (OOK) with a Non-Return to Zero (NRZ) format, 4-level Pulse Amplitude Modulation (PAM-4) and 8-level Pulse Amplitude Modulation (PAM-8).\textsuperscript{1}

![Figure 3.18: Experimental setup for the high-speed transmission in the mid-infrared domain with interband cascade devices. A 4.18\(\mu\)m wavelength ICL emits around 8 mW of optical power. We modulate our laser with a 10dB-amplified signal produced by an AWG. The ICIP detects the modulated optical wave and converts it to an electrical signal, which is subsequently fed to a high-speed oscilloscope and then analyzed. The RF-optimized ICL can be seen on the left with its contacts connected to a coplanar waveguide and a 2.92 mm RF-launcher. On the right, the ICIP can be visualized. Short wire bonds are used to connect the contacts to a coplanar waveguide that is attached to an end-launch connector as well.]

The use of 2\textsuperscript{15}-bit long PRBS allows for a greater variety of bit sequence combinations to be studied and evaluated in the transmission system. Additionally, we used an RRC filter to shape our signal and decrease the spectral occupation, thereby increasing the data rate of the transmission, as already discussed. The modulated electrical signal on the ICL from the AWG had an amplitude of 1.5 Vpp which was amplified upstream using a 10 dB gain D837C SHF differential amplifier. After traveling approximately 2 meters, a silicon
lens focused the optical signal onto the ICIP detector. The ICIP detector is advantageous over QCDs in that it does not require a specific polarization for detection, making it very convenient. The ICIP was connected to a Pasternack bias tee for detection, where the DC component was used to bias the detector with a Keithley source, and the AC component was amplified by a 25dB-gain 826H SHF amplifier before being recorded by a 20 GHz oscilloscope with a sampling rate of 50 GS/s. After being recorded by the oscilloscope, the signal was processed offline using a MATLAB code. This included filtering it with a root-raised cosine filter and then using a feed-forward equalizer. The recovered bits were then compared to the transmitted ones to calculate the BER.

The bandwidth of a setup composed of amplifiers, laser, and detector was measured using a 40 GHz Rohde & Schwarz ZVK Vector Network Analyzer (VNA) with two different types of detector as shown in the Figure 3.19. The VNA was used to determine the input-output relationship between ports in the system as a function of the source signal frequency. The source signal of the VNA was plugged into the AC input of the laser’s bias tee, and the receiver was plugged into the 25 dB amplified response signal of the ICIP detector. The response was plotted for four voltage conditions for the ICIP and showed a flat response until the cut-off frequency. The bandwidth of the detector can be increased by one order of magnitude by increasing the bias voltage, from 140 MHz at 0 V to 1.5 GHz at 5 V bias. However, increasing the bias also increases the dark current, which can limit detection capabilities at low optical power. To provide a comprehensive analysis, we examine the
response of the complete system when employing a QCD detector with a 3dB bandwidth of 20 GHz, in contrast to an ICIP detector. The system 3dB bandwidth is 2.5 GHz, which is the bandwidth of the ICL structure itself, as the QCD detector is no longer the limiting component. This 2.5 GHz bandwidth aligns well with our predictions made using our previous relaxation oscillation frequency study.

3.3.3.2 Interband free-space transmission system.

In order to achieve the highest data rate possible, we tested the effect of various parameters of the equalization process on the transmission quality. In Figure 3.20.b) and d), the importance of optimizing the equalization coefficients is evident. For OOK, the optimal value for the convergence parameter $\mu$ is around $10^{-3}$ and the number of taps $n_{\text{tap}}$ should be above 11 to meet the tabulated threshold BER. This means that increasing the $n_{\text{tap}}$ value above 11 does not matter much since the ISI has no influence anymore. Only well-chosen parameters result in significant improvements in transmission quality. By optimizing these coefficients, we achieved an OOK transmission at a bit rate of 12 Gbit/s with a BER under 0.4% and 14 Gbit/s with a BER below 4%. We used optimized parameters to plot the relative eye diagrams in Figure 3.20 a) and c). For the 4-level format (PAM-4), we achieved a transmission at 14 Gbit/s with a BER of 0.23% and 16 Gbit/s with a BER below 4%. It is worth noting that the optimal value of the convergence parameter $\mu$ must be lower for multi-level modulation formats.

PAM-8 was also investigated, but due to noise from the amplifier and detector in our system, a bit error rate below 0.4% was not achieved. The parameters were optimized and the optimum $\mu$ factor was lower compared to lower multi-level modulation formats, as shown in Figure 3.21 b). A high $n_{\text{tap}}$ value was necessary to keep a low BER. The presented result in Figure 3.21 a) is a 15 Gbit/s transmission associated with a BER of 1.9%. This can be corrected by using an HD-FEC algorithm with a 27% overhead, which can accommodate a BER up to 4%. It should be noted that using a higher overhead may also introduce more complex decoding and higher latency.

The results of the transmission for the different modulation formats were summarized in Figure 3.22. The best performance was achieved when a $\rho$ value of 0.3 for the OOK and PAM-8 modulation formats and a $\rho$ value of 0.4 for the PAM-4 format was used. This means that the bandwidth occupation of the modulated optical signal was decreased by 35% and 30%, respectively. It allows achieving a BER below the 7% HD-FEC threshold for the OOK and PAM-4 configurations. In this setup, moderate voltage is required, and the power consumption of the overall IC system is still as low as 1 W at maximum conditions.
Figure 3.20: Eye diagrams of the transmission after a free-space propagation of approximately 2 meters were created for two different modulation formats. a) OOK format at 12 Gbit/s showing an error rate of 0.22%. c) PAM-4 format at 14 Gbit/s with an error rate of 0.23%. b) and d) show the optimization parameters of FFE equalization in order to achieve the lowest BER. In both cases, there is a particular set of parameters where the equalization allows one to highly optimize the quality of the transmission.

3.4 Intersubband cascaded technologies (8-14 µm)

3.4.1 Direct modulation of QCL

As mentioned in Chapter 2, the QCL has a theoretical modulation limit of up to 150 GHz, making it highly promising for high-speed communication. In 2017, researchers achieved 26.5 GHz modulation with a 9µm QCL, but did not perform any communication experiments. As the technologies surrounding quantum cascade lasers continue to advance, we have witnessed notable improvements in various aspects. These include higher laser output power, reduced power consumption, expanded accessible wavelengths, and increased...
modulation frequencies. We will present in this part results towards FSO transmission using direct modulation of QCL. A RF-optimized 8.1 µm QCL was employed for direct modulation in FSO communication. The experimental setup used was similar to the one presented in the previous section. The laser characteristics being investigated are depicted in Figure 3.23. It has a 8 µm ridge width and a 0.5 mm short cavity length that enhances the modulation bandwidth by minimizing parasitic capacitance and reduces electrical consumption to 1.4 W at maximum optical power. Although it has low output power, the
low losses at this wavelength in a free-space environment do not affect transmission quality. The electrical contact is made through an MMCX connector for high-speed electrical injection. The QCL is maintained at a constant temperature with a Peltier module and temperature controller and is powered by a low-noise current source (Wavelength Electronics, QCL2000). Figure 3.23.a) shows a threshold current of 110 mA at 293 K and a maximum beam power of 10 mW at 140 mA and 10 V in the light-intensity-voltage curve. Figure 3.23.b) displays monomode behavior for currents < 140 mA.

Figure 3.23: a) Light-intensity-voltage (LIV) curves for various temperatures. b) Optical spectrum of the laser for three different bias currents while keeping the temperature at 288 K: 110 mA, 125 mA and 140 mA. c) S12 measurements for four different bias currents: 105, 115, 125, 140 mA. One can see how the electrical bandwidth of the QCL increases with bias current and is then kept unchanged for values above 125 mA. The maximum 3-dB bandwidth is 900 MHz.

The experimental setup is very similar to the setup used for the ICL transmission. To generate the OOK, PAM4 and PAM8 signal, a pulse pattern generator (AWG Tektronix 7122B ) with pseudo-random binary sequences (PRBS) is used at sample rates up to 12.5 GS/s, with a sequence length of $2^7$. The modulated electrical signal is combined with the DC bias, at the AC connector of a bias-tee. This combined signal is then sent to the QCL, where it is converted into a mid-infrared beam. The beam is focused onto a detector (Vigo UHSM-I-10.6) with a bandwidth of 700 MHz. The electrical output of the

3.4. Intersubband cascaded technologies (8-14 µm)
MCT detector is analyzed using a high-speed oscilloscope (Tektronix MSO72004C) with a maximum bandwidth of 20 GHz and a sampling rate of 50 GS/s. To reduce high-frequency noise, the oscilloscope’s actual bandwidth is reduced to 3 GHz. The traces are then postprocessed to implement FFE algorithm, eye diagram plotting and BER evaluation. The eye diagrams for OOK, PAM-4 and PAM-8 at data rates of 2.4 Gbit/s, 4.8 Gbit/s and 4.2 Gbit/s respectively are shown in the Figure 3.24. For every configuration, the BER is low enough for error-free transmission using a FEC algorithm.

**Figure 3.24:** Eye diagrams of the transmission in B2B for three different modulation schemes and a sequence length of $2^7$ detected by a 700 MHz MCT detector a) OOK: 2.4 Gbit/s for a \( \text{BER} = 0 \) (no error in the analysed trace) and a \( n_{\text{tap}} = 51 \) and \( \mu = 10^{-5} \). b) PAM-4: 4.8 Gbit/s for a BER=0.18% and a \( n_{\text{tap}} = 51 \) and \( \mu = 10^{-4} \). c) PAM-8: 4.2 Gbit/s for a BER=0.14% and a \( n_{\text{tap}} = 391 \) and \( \mu = 2.510^{-5} \).

We also performed the optimization of the equalization parameters to achieve the lowest BER achievable as seen in the Figure 3.25.

**Figure 3.25:** Optimization of the FFE equalizer for a) the PAM-4 and b) the PAM-8 transmission. The optimization process involves fine-tuning parameters of \( n_{\text{tap}} \) and \( \mu \).
3.4.1.1 Live video broadcasting in free-space using QCL

We performed a live video broadcasting in FSO using an 8.1 µm QCL. The result is shown below with both pictures representing a frame of the video. The setup begins with a computer connected to an HDMI/SDI converter (MD-HX Decimator board) that enables the creation of a modulated signal. The converter transforms the HDMI signal into an on-off keying scheme and offers a convenient BNC output. To combine the DC current from a low-noise current source and the RF modulated signal from the board, we employ a Pasternack bias tee with a bandwidth of 12 GHz. The resulting signal is then collimated using a lens and directed towards the MCT detector, positioned 1.5 meters away. The MCT detector utilized in this experiment is capable of detecting mid-infrared AC signals with amplitudes as low as a few dozens of µW with a 3dB bandwidth of approximately 700 MHz. The electrical signal extracted from the detector is sent to another SDI/HDMI converter, which is connected to a TV monitor via an HDMI cable. This setup enables real-time monitoring of the live transmission on a remote screen.

![Experimental setup](image)

**Figure 3.26:** Experimental picture of the video broadcasting in free space using 8.1 µm QCL. The length of the link can be easily increased up to tens of meters.

To control the beam attenuation, a polarizer is introduced in the light path, allowing for...
adjustment by rotating the mid-infrared polarizer in the transmission channel. We have observed that the signal can still be successfully recovered even after applying up to 6 dB of attenuation. However, this limit is determined by the threshold requirements of the SDI format, which necessitates a signal with a peak-to-peak amplitude of 800 mV. To overcome this limitation, the addition of an amplifier after the MCT detector could be considered, facilitating even greater attenuation in the optical path. An interesting observation is that the link demonstrated resilience over time, with no observed disconnections for several hours. This highlights the reliability and stability of the setup. Moreover, considering that the absorption of the atmosphere at this wavelength is approximately 3 dB per kilometer for a visibility of 1 km, this type of setup holds promises for long-distance FSO transmissions. 39 Three video formats transmission are demonstrated in Figure 3.27: SD480i59.94, HD720p59.94 and 3G1080p59.94 (3G-1080p-59.94 = Video quality-Number of pixels-Frames per second). The audio is included in the HDMI format, allowing the sound to be heard through the monitor’s speakers. The broadcasting results are presented below, accompanied by images. In the upper portion of Figure 3.27, the generated signal is plotted for the various formats and compared with the signal obtained from the detector at the end of the transmission channel. For both the SD480i59.94 and HD720p59.94 formats, a flawless transmission is achieved. The wide bandwidth of our detector enables the recovery of a signal that closely resembles the original seed modulation, as illustrated in Figure 3.27.a) and b).

For the HD1080p59.94 format, which requires a data rate of 2.970 Gbit/s, the limitations of the detector and laser bandwidth become evident. As shown in the right panel of Figure 3.27, the signal becomes distorted in this configuration. High frequencies are attenuated, and errors occur when consecutive '1'-bits or '0'-bits are triggered. Indeed, in Figure 3.27.c), there are several errors with the 3G1080p59.94 format, as we estimate the BER to be around 10 % but we made the following observation: the recovered signal on the TV monitor still maintains a very high quality, as demonstrated in Figure 3.27.f) and the accompanying video footage. One possible explanation for this is that the Decimator receiver board applies robust equalization and error-correction processing to the signal. 38 Furthermore, our experiments show that when we transmit the signal in OOK using the same QCL, we were able to achieve an errorless recorded signal at a rate of 2.4 Gbit/s, as shown in Figure 3.24.a).

3.4.2 External modulation
3.4.2.1 Stark effect External modulator

In this section, we discuss our findings on FSO communication using an external modulator operating in band 3 of the atmospheric transparency windows. The technology for the external modulator and detector was developed at ENS PSL Paris by Thomas Bonazzi and Hamza Dely.\textsuperscript{17,40,41} This technology offers many benefits in terms of high modulated output power and bandwidth for long-distance free-space transmission. As previously discussed in Chapter 2, external modulators are a key technology for enhancing the bandwidth and modulated power of QCL-based communications.

**CHARACTERISTIC OF THE EXTERNAL MODULATOR** Building upon the research conducted by H. Dely et al.\textsuperscript{17}, we employed an external amplitude modulator utilizing the Stark effect. The design enables the elimination of charge depletion gates, resulting in a reduction of intrinsic parasitic capacitance. The modulator consists of an asymmetric quantum well-made of GaInAs/AlInAs material systems, n-doped at $1.5 \cdot 10^{18} \text{cm}^{-3}$ in the wider well. The Stark shift is due to the localization of electrons in state 1 in the large quantum

**Figure 3.27**: QCL video transmission at 8.1$\mu$m for three different formats a) SD480i59.94, b) HD720p59.94 and c) 3G1080i59.94 corresponding to different data rates: 270 Mbit/s, 1.485 Gbit/s and 2.970 Gbit/s. In black, and in red, we have respectively the generated signal and the received signal by the MCT. The transmission is error free for the two first formats mentioned. On the bottom of the figure, one can find the related image obtained on the screen at the reception for the three different formats in d), e) and f).
well and electrons in state 2 in the thin well. Under an applied bias voltage, the energy shift of the transition $E_{12}$ is equal to the potential drop between the barycenters of the two electron distributions, $L_{\text{bar}}$, which is approximately equal to the distance between the centers of the quantum wells.

$$E_{12}(V) = E_{12}(0) + e\Delta V_{qw}.$$  \hspace{1cm} (3.16)

with $\Delta V_{qw} = V_{\text{bar}}/L_{\text{struct}}$ and $L_{\text{struct}}$ the total thickness of the structure, $e$ the electron charge and $V$ the potential of the quantum well. As shown in the Figure 3.28.a) and b), by modulating the bias current of the modulator, we will modulate the absorption energy of the structure and then the output power of the modulator. By using the external modulator, we built our free-space communication setup shown in Figure 3.30. We used a commercial 9 µm-wavelength continuous wave QCL as the optical source. The QCL emits up to 97 mW of output power at room temperature and is expanded and focused using telescopes. The beam is then directed to a 50x50 µm external modulator, which is connected to a bias-tee and an Agilent 33500B generator for DC bias and a Socionext arbitrary waveform generator for RF signal. The signal is amplified to 15 V peak-to-peak before being applied to the
modulator, which loses two thirds of the optical power due to reflections. The modulated beam is then either directly collected by a MIR high-speed detector or passed through a 31 m Herriott cell before detection. The Stark-effect-based external modulator is designed for high-speed operation with quantum engineering that prevents charge displacement in the heterostructure, avoiding speed throttling due to electronic transport. The modulator has a 9-GHz bandwidth under RF probing, mainly limited by its geometric capacitance. The modulator’s modulation depth is measured using voltage pulses of various peak voltages. The peak voltage measured on the QCD is proportional to the peak intensity and is presented in Figure 3.29.b) with a Beer-Lambert law fit. From our analysis, we estimate a modulation depth of approximately 25%. The signal is acquired using a 33 GHz bandwidth oscilloscope and processed offline using Python and MATLAB scripts. Two detectors are used in this work: a passive, room-temperature QCD and a nitrogen-cooled 25x25 µm mesa QWIP at 77 K. We used the same signal shaping for transmitting the data. We investigate the possibility to send multi-Gbit/s information. In this experiment, data is transmitted using On-Off Keying (OOK) and 4-level Pulse Amplitude Modulation.
Figure 3.30: Experimental setup for 30 meters free-space transmission using an external modulator.

(PAM-4) with pseudo-random binary sequences of either $2^7$- or $2^{15}$-bit patterns. The modulator is biased at a DC value of 1.1 V to reduce distortion. Measurements are taken for back-to-back transmission and for 31 meters transmission using a commercial Herriott cell. The transmitted signal is collected by a QCD or QWIP detector, and the quality of the transmission is evaluated by the BER. Eye diagrams are presented to evaluate the quality of transmission for different configurations.

**Optimization of the Parameters of the Equalizer** As already depicted in the equalization paragraph, this part is mandatory to increase drastically the maximum data rate achievable. In consequence, every result is presented after equalization. Figure 3.31.b) provides further details on the optimization of the equalizer parameters for achieving this low error rate. By using the QWIP detector, we achieve a data rate of 30 Gbit/s in the back-to-back configuration, as illustrated in Fig. 3.31 a). The best BER value we were able to achieve for the B2B experiment at 30 Gbit/s is 0.23%. The objective is to identify a combination of these parameters that satisfies one of the two hard-decision forward error correction (HD-FEC) conditions. Ideally, the 7% HD-FEC case is preferred, as it incurs a data rate penalty of only 7%. When multiple options are available for $n_{\text{tap}}$ and $\mu$, it is advantageous to choose the one with the lowest value of $n_{\text{tap}}$. This choice simplifies the implementation process as lower values facilitate faster estimations. It is evident that only specific, carefully chosen parameters enable significant improvement in transmission quality through
Figure 3.31: a) Eye diagram of the QWIP-scheme transmission after a free-space propagation of approximately 30 meters for an OOK format at 30 Gbit/s, exhibiting an error rate of 0.3 %. b) Optimization of the FFE equalization parameters in order to achieve a BER compatible with either the 27% HD-FEC case or with the 7% HD-FEC case.

We performed this type of optimization of the equalization coefficient for each modulation format, considering different link configurations (2 meters or 31 meters), and utilizing either the QCD or QWIP detector. The figures related to those study can be found in appendix C for the OOK transmission.

OOK TRANSMISSION RESULTS AFTER 31 METERS PROPAGATION. In the following, we will present the results obtained with both the QWIP and the QCD detector. The maximum data rate for QCD transmission through the Herriott cell was 14 Gbit/s with a BER < 0.4 % and a sequence length of $2^7$ as shown in Figure 3.32.a). With QWIP detector, the data rate can be increased to 40 Gbit/s for a PRBS of $2^{15}$ bits and a BER < 4 %. Additionally, we obtained 30 Gbit/s data rate with a BER < 0.3 %.

PAM4 TRANSMISSION RESULTS As already discussed, one way to increase bit rate while keeping the symbol rate the same is by using higher-order modulation formats such as PAM-4. By applying RRC with a roll-off factor of 0.5 at the transmitter and FFE equalization at the receiver, it is possible to reach a bit rate of 70 Gbit/s with a BER of 3.6% in a back-to-back configuration. However, this result may overestimate the capacity of the link due to the limited sequence length. A more realistic result is a bit rate of 40 Gbit/s with a BER of less than 1% in both back-to-back and a 31-meter link, as shown in Figure 3.33

3.4. Intersubband cascaded technologies (8-14 µm)
Figure 3.32: Eye diagrams of the transmission through a 31 meters Herriott cell for three different data rates and a sequence length of $2^7$ for the QCD and $2^{15}$ for the QWIP. a) QCD: 14 Gbit/s for a BER=0.21%. b) QWIP: 30 Gbit/s for a BER=0.33% c) QWIP: 40 Gbit/s for a BER=2.5%. d) Evolution of the BER (equalized and non-equalized) with respect to the data rates for the QWIP transmission. The dashed lines recall BER limits for 7% HD-FEC (green) and 27% HD-FEC (cyan).

3.4.2.2 External modulator based on metamaterial antenna

We conducted experiments using a new modulator and detector design based on a patch antenna, as shown in Figure 3.34. This modulator was designed and fabricated by Thomas Bonazzi, and a detailed description of its design can be found in his thesis manuscript. The previous external modulator was directly connected to the coplanar waveguide, while the new design offers several advantages. The concept behind this new design was to enhance the intersubband absorption of the external modulator by incorporating an antenna stripe array. This approach aims to achieve a strong light coupling regime, thereby greatly increasing the interaction between the mid-infrared (MIR) electric field and the...
Secondly, the electrical surface is reduced, limiting the current noise produced by the device. This design is suitable for use at room temperature, achieving a high photocurrent density. Furthermore, by reducing the electrical surface, capacitance can also be minimized, thereby increasing the potential for achieving high-speed performance. This can be further optimized by establishing precise electrical connections between the resonators and the load.

**Specific design** The schematic representation of the striped structure antenna design is shown in Figure 3.34.a). It provides a top-down view of the device under study as observed...
through a scanning electron microscope (SEM). The device comprises a one-dimensional array of stripe cavities, as depicted in Figure 3.34.b). Each stripe is filled with an optically active region. These stripes extend for several tens of micrometers in the direction perpendicular to the array, with a spatial period. The incident light directly couples with the cavity’s electromagnetic mode.

![Figure 3.34](image)

**Figure 3.34:** The stripe-based modulator devices are shown in SEM pictures. a) It displays the device with its coplanar waveguide. b) It provides a zoomed-in view of the modulator. The stripes are 1.5µm wide and spaced 7µm apart from each other.

It’s worth noting that a striped pattern demonstrated greater efficiency in coupling compared to the patch structure analyzed in Bonazzi’s thesis. Both of the detector and modulator in our experiments were designed using the stripe antenna architecture, with the modulator operating in reflection to minimize insertion loss compared to the previous design.

**CHARACTERISTIC** The improved bandwidth of the setup is illustrated in Figure 3.35. The modulation depth has been studied, and we evaluate a modulation depth of around 28%. The modulated output power was around 20 mW. Furthermore, the responsivity of the QCD detector has been significantly increased up to 25 mA/W, enabling passive operation without the need for cooling. To construct a transmission system, we combined the external modulator with a 6 GHz 3dB-bandwidth QCD. While the system’s 3-dB bandwidth is limited to 6 GHz due to the detector, it demonstrates a relatively flat response up to 16 GHz. This characteristic proves particularly advantageous for the transmission experiment.

**IMPLEMENTATION OF A DECISION-DIRECTED DFE ALGORITHM** Similarly to the previous study, our investigation focused on the feasibility of transmitting information using two different
modulation formats and a PRBS with a length of $2^{15}$ using a RRC filtering. The setup remains very similar for the transmission presented above. We use an optimized $\rho = 0.3$ for the RRC filtering parameters. In collaboration with Elie Awwad, we have made significant improvements to our equalization process. Throughout our study, we observed substantial fluctuations in the channel, resulting in increased errors in the equalized traces as they passed through. To address this issue, we explored two approaches: periodic updating of the FFE filter along the trace or the implementation of a Decision-Directed DFE technique. The latter was preferred due to its superior performance. The values of the filter coefficient and equalization parameters are highlighted in Figure 3.36.a). Initially, we constructed a filter using FFE, employing a training sequence consisting of $N_{\text{training}} = 50000$ samples, which accounted for approximately 1% of the entire recorded trace length. After the convergence of the FFE filter, we utilized this filter as the initialization for the Decision-Directed DFE. In this approach, we adapted the filter coefficient based on the feedback from previous decisions made on the received signal. This method enabled us to maintain a stable error value across all the traces, even in the presence of significant channel variations over time. In Figure 3.36.b), it illustrates the gradual convergence of the traces to establish a clear distinction between the various levels. The convergence of both the FFE and DFE filters is highlighted in Figure 3.36.c). Interestingly, the average error was estimated even when considering the use of only the FFE evaluated filter (in light gray), thereby highlighting the additional benefits and improvements achieved through the utilization of the DFE. We also verified that the errors were evenly distributed along the traces, ensuring the stability of our equalization process, even with constant adaptation of the DFE filter. The graph is presented in the Figure 3.36.d). Overall, we were able to successfully transmit approximately 2 million bits at a rate of 60 Gbit/s with a BER of 0.028%. These achieve-
ments demonstrate the effectiveness of our enhanced equalization process and its ability to maintain reliable transmission under challenging channel conditions.

Figure 3.36: Description of the DFE equalization use in this part for transmission of $2^{15}$ PAM4 PRBS and a data rate of 60 Gbit/s. a) The coefficient of the FFE and DFE filters is illustrated. The FFE filter, depicted at the top, has parameters of $\mu_{\text{FFE}} = 5 \cdot 10^{-5}$ and $n_{\text{tap}} = 401$. In the lower section, the DFE equalizer is displayed with a value of $\mu_{\text{DFE}} = 5 \cdot 10^{-4}$ and $n_{\text{tap}} = 16$. b) The initial samples of the output of the FFE+DFE equalizer are presented. The training sequence or the FFE corresponds to 1% of the whole traces. Following the initialization of the filter around 0.1 $\mu$s, distinct markings between the four different levels became visible in the output. c) The graph displays the error over time for the first sample of the traces. We began by applying the FFE algorithm, which required some time to converge. Once the FFE reached a stable state, we transitioned to using the DFE algorithm. The graph highlights the convergence of both the FFE and DFE algorithms. Additionally, the light grey trace represents the average error of the traces using only the filter after the FFE convergence, showing the interest of adding a Decision Directed DFE along the trace. d) The number of errors along the traces is observed for groups of 1000 bits. The errors are evenly distributed throughout the acquisition, which consists of sending 3 million bits.

RESULTS The enhanced responsivity and increased bandwidth of the devices have enabled access to higher data rates while simultaneously maintaining a high modulated output power of approximately 30 mW. The results are presented in the Figure 3.37. With a two-level modulation, we achieved a data rate of 55 Gbit/s while keeping the BER of 0.26
%, and with a four-level modulation, we achieved 68 Gbit/s with a BER of 0.36 %. These results significantly surpass our previous mid-infrared results.

Due to the presence of noise in our system, achieving very high data rates with PAM8 modulation posed challenges. However, we achieved successful transmission at a rate of 24 Gbit/s with a BER of 0.14 %, as demonstrated in Figure 3.37.c).

### 3.5 Conclusions

In this chapter, we first described here the exhibition of the relaxation frequency of ICLs in the GHz range, which is of first importance for enhancing modulation capabilities of such lasers. For this purpose, differential gain, K-factor, gain compression, and carrier lifetime are also analyzed. This work is a key step in considering ICLs for versatile, light, and cheap
multi-Gbits/s transmission systems in the mid-infrared domain. After this study, knowing the high speed capability of the ICL, we investigated the possibility of building a free space communication system based on interband cascaded technologies. We presented the development of a full interband cascade system that operates in one of the transparency windows of the atmosphere at room temperature, specifically at a wavelength of 4.18 µm. The signals to be transmitted were shaped using an RRC routine before being injected into the laser to optimize the achievable data rate. The received signal detected by the ICIP was equalized using an FFE algorithm and the FFE parameters were optimized to minimize the BER. The system operates at 12 Gbit/s and 14 Gbit/s with a 2-level modulation format (OOK) with a BER suitable for 7 % and 27 % HD-FEC correction, respectively. Multilevel modulation formats were also investigated, and we demonstrated data transfer at 14 Gbit/s and 16 Gbit/s with a 4-level modulation format (PAM-4) with a BER suitable for 7 % and 27 % HD-FEC correction, respectively. This shows that moving from a 2-level to a 4-level modulation format improved the achievable data rates by 2 Gbit/s while ensuring comparable performance. An 8-level modulation format (PAM-8) was also investigated, which led to transmission at 15 Gbit/s with a BER suitable for 27 % HD-FEC correction. These results are very encouraging for direct-modulation in the 3 – 5 µm window and benefit from 10-times lower power consumption of the IC devices compared to their quantum cascade counterparts. Furthermore, it opens up opportunities for compact and integrated designs on silicon platforms. This research is critical in creating an energy-efficient system that utilizes cascaded interband devices for communication purposes, particularly in energy-limited scenarios such as data exchange between satellites or rovers exploring distant planets. Alternatively, it is intriguing to shift our focus to the wavelength range of 8 – 14 µm, with the aim of harnessing the potential to enhance the transparency window and enhance resistance to scattering. We successfully demonstrated high-speed communication at multiple Gbit/s rates by leveraging an RF optimized QCL in combination with advanced digital signal processing techniques. The QCL used in our experiments is notable for its compact short-ridge design, which contributes to its low power consumption and GHz bandwidth. To further enhance our transmission capabilities, we used an external modulator, enabling us to achieve an impressive transmission rate of 30 Gbit/s over a distance of 30 meters with a BER below 0.4 % for an OOK modulation using a cryogenic cooled QWIP. Similar to our earlier study involving ICLs, we implemented advanced algorithms to optimize and enhance the quality of our transmission. It is worth highlighting that this particular work represents a groundbreaking achievement, as it marks the first successful implementation of high-speed transmission over an extended distance. Moreover, we also introduced new devices developed at ENS Paris, featuring stripe an-
tenna Stark external modulators. These innovative devices further pushed the boundaries of data rates and significantly improved the practicality of the setup. In particular, we achieved a data rate of 68 Gbit/s in a back-to-back configuration with a BER under 0.4 % and with PAM4 modulation. Remarkably, this achievement was achieved using a passive and uncooled detector, showcasing the practical advantages of these devices. It is worth noting that in our previous study on the external modulator setup, the best results were obtained using a cryogenically cooled detector. The new devices exhibit immense potential for the development of long-haul transmission systems due to their ability to generate high modulated output power (20 mW). In the Table 3.2, we summarized the different types of mid-infrared sources for free space communication. Advances in the RF design of mid-infrared technologies have enabled the consideration of direct modulation for both ICL and QCL transmission applications. Currently, the only sources capable of providing high-speed transmission in the $8 - 14\,\mu m$ wavelength range are QCLs that support direct modulation or can be used with an external modulator. Indeed, long-wavelength ICL are still working at cryogenic temperature. Considering an external modulator for high-speed long-haul transmission is particularly interesting because of the output power associated with a high modulation power. It can be used for atmospheric layers crossing or between space to ground link. Therefore, despite recent effort to decrease the power consumption of QCL, it still quite important. Otherwise, ICL is a highly energy-efficient source which can provide high-speed free-space transmission. For the sake of completeness, we provide a comprehensive overview of frequency conversion as a potential solution for high-speed mid-infrared transmission in Table 3.2. While this technology shows promise with its ability to access high frequencies and support various modulation formats, it still faces certain limitations. One of the key challenges is the insufficient mid-infrared output power, which restricts its practical implementation. Moreover, frequency conversion setups often exhibit high power consumption, which hampers their widespread adoption. However, researchers are actively working to address these limitations and unlock the full potential of this intriguing technology\textsuperscript{12,44,45}.

Despite the numerous advantages of FSO communication, it is crucial to address the inherent security concerns associated with the physical layer. It makes it vulnerable to eavesdropping and interception. However, in recent years, researchers have been exploring methods to address this security challenge and establish private communication on FSO links. By employing advanced encryption techniques and secure key exchange protocols, it is possible to enhance the confidentiality and integrity of data transmitted over FSO links. In the following sections, we will delve into these possibilities and explore the po-
Table 3.2: Comparison between different mid-infrared modulating sources. DM: Direct Modulation, * correspond to the work presented in this manuscript.

<table>
<thead>
<tr>
<th></th>
<th>DM ICL (*)</th>
<th>DM QCL (*)</th>
<th>External modulator (17, *)</th>
<th>Frequency conversion</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power consumption</td>
<td>Low (&lt;W)</td>
<td>Medium (~ W)</td>
<td>High (&gt;W)</td>
<td>Very High (&gt;W)</td>
</tr>
<tr>
<td>Band 3 accessibility</td>
<td>No</td>
<td>Yes</td>
<td>Yes</td>
<td>No</td>
</tr>
<tr>
<td>Modulated output power</td>
<td>Low</td>
<td>Medium</td>
<td>High</td>
<td>Very low</td>
</tr>
<tr>
<td></td>
<td>(~ 10mW)</td>
<td>(~ 50mW)</td>
<td>(~ 100mW)</td>
<td>(~ 1 mW)</td>
</tr>
<tr>
<td>Reliability</td>
<td>High</td>
<td>High</td>
<td>Low</td>
<td>Medium</td>
</tr>
<tr>
<td>Integration</td>
<td>Easy</td>
<td>Medium</td>
<td>Very complex</td>
<td>Very complex</td>
</tr>
<tr>
<td>Potential bandwidth (GHz)</td>
<td>Low</td>
<td>Very large</td>
<td>Very large</td>
<td>Very high</td>
</tr>
<tr>
<td></td>
<td>(~ 5 GHz)</td>
<td>(~ 100GHz)</td>
<td>(~ 100GHz)</td>
<td>(~ 100GHz)</td>
</tr>
<tr>
<td>Modulation format</td>
<td>AM</td>
<td>AM</td>
<td>AM/PM</td>
<td>AM/FM/OAM/WDM</td>
</tr>
<tr>
<td>Maximum data rates</td>
<td>14 Gbit/s (*)</td>
<td>11 Gbit/s</td>
<td>68 Gbit/s (*)</td>
<td>300 Gbit/s (*)</td>
</tr>
</tbody>
</table>

tential for establishing secure and private communication in FSO systems using chaos synchronization.


Chapter 3 • Mid-infrared devices for high speed applications

"Et si les cyclones étaient la somme de tous les derniers soupirs de la Terre ? Et si les tremblements de terre étaient la somme de tous nos premiers pas ?" Premier Pas, Nekfeu
Despite its many benefits, traditional free-space communication lacks data protection from the physical layer, leaving sensitive information vulnerable to cybercrime. Free-space optical Quantum Key Distribution (QKD) offers great security through the intrinsic properties of photon entanglement, but is hindered by low-speed (1 kbps), low-yield, and expensive transmission equipment.\textsuperscript{1,2} Recent studies have demonstrated effective entanglement maintenance for successful transmission of up to 100 bits per second over a 10-kilometer distance between ground-to-ground stations, with similar losses incurred as communication with a low-earth orbit satellite in an urban environment. However, the technology is not yet available at MIR wavelengths due to the lack of efficient single-photon detectors and sources. A promising approach for secure free-space optical communication is the use of chaos synchronization of coupled lasers for ciphering sensitive messages.\textsuperscript{3-5}

### 4.1 Introduction

Linear dynamics is a common feature of most dynamical systems, where a small change in initial conditions produces a proportional change in the outcome. However, in the late 20th century, researchers discovered the existence of other systems that behave “strangely”.\textsuperscript{6} A system where very slight changes in the initial conditions can yield entirely distinct outcomes. Henri Poincaré was one of the early advocates of chaos theory. His work on the three-body problem in the 1880s led to the discovery of non-periodic orbits that neither continually increase nor converge to a fixed point. This discovery marked a significant shift in our understanding of complex systems and paved the way for the development of chaos theory. The study of chaotic systems has broadened our understanding of complex phenomena, from the behavior of fluids to the dynamics of financial markets.
4.2 "Does the flap of a butterfly’s wings in Brazil set off a tornado in Texas?"

The pioneer to observe such a system was Lorenz in 1963 during a well-known conference on chaos theory, where he famously asked “Does the flap of a butterfly’s wings in Brazil set off a tornado in Texas?”. Lorenz’s meteorological simulation system was composed of just three simple equations (Equation 4.1), yet even the tiniest variation in the initial conditions could lead to vastly different dynamics. Initially, the traces exhibit similarity, but their dynamics quickly diverge from one another.

\[
\begin{align*}
\dot{x} &= \sigma(y - x) \\
\dot{y} &= r x - y - xz \\
\dot{z} &= xy - bz
\end{align*}
\]

(Equation 4.1)

The parameters \(\sigma\), \(r\), and \(b\) are system parameters that hold great significance. Even a slight fluctuation in their values can result in entirely different dynamics within the system. In Figure 4.1, the chaotic attractor of the Lorenz equation is depicted, exhibiting the distinctive “butterfly” shape. However, it is important to note that chaos had been observed in the dynamics of other systems prior to this discovery. A notable example is the ruby solid-state laser, which exhibited variations in the output pulses that were described as "erratic" in nature, as depicted in Figure 4.2.a. Although this peculiar behavior was initially attributed to noise, it marked an early encounter with chaotic phenomena in physical systems. The unpredictability arises from the extreme sensitivity to initial conditions in chaotic dynamics. As a result, it becomes nearly impossible to accurately predict the future evolution of a chaotic system. This inherent unpredictability has found applications in various fields, including random bit generation (RGB), communication, and lidar, where chaos-based techniques can be harnessed for their unique properties.

0.16, \(r = 45.6\), and \(b = 4\).

### 4.2.1 Chaos characterization

Chaos can be identified by several properties, such as sensitivity to initial conditions, irregular behavior, and the presence of a strange attractor. Irregular behavior is characterized by the lack of periodicity or predictability in the system’s behavior, making it difficult to
Figure 4.1: The strange attractor associated with the Lorenz equation system, as presented in Equation 4.1, illustrates the average spectral power projection onto two distinct planes concerning the three variables of the equation. In this representation, \( u(t) = \frac{x(t)}{10} \), \( v(t) = \frac{y(t)}{10} \), and \( w(t) = \frac{z(t)}{20} \) are utilized for visualization purposes. For this system, the parameters are \( \sigma = 16 \), \( r = 45.6 \) and \( b = 4 \). The distinct butterfly shape of the attractor is what led to it being named after Edward Lorenz, who first discovered this chaotic behavior. Courtesy of Cuomo\(^7\).

forecast its future state accurately.

4.2.1.1 Strangem attractor

The presence of a strange attractor as shown Figure 4.2.c) implies that the system’s motion does not converge to a fixed point or a periodic orbit, but rather settles onto a complex geometrical structure that defines the long-term dynamics of the system. It is a set of points in phase space that an initially nearby set of points evolves into, as time goes on. The fractal structure of the attractors in a chaotic system is a notable characteristic, as they exhibit self-similarity at different scales. This self-similarity feature implies that if one focuses on a specific region of the attractor and zooms in, the details would look similar to the entire attractor. Moreover, strange attractors are known for their sensitivity dependence on initial conditions, which means that minor differences in the initial conditions can result in significantly different trajectories over time. This sensitivity is a distinct feature of chaotic systems, which leads to the unpredictability of their behavior. The reconstructed trajectory is useful for illustrating the complex geometrical and topological structure of chaos, showing the local instability, and global stability\(^1\). The Figure 4.1 and Figure 4.2

4.2. "Does the flap of a butterfly’s wings in Brazil set off a tornado in Texas?"
illustrates the representation of a strange attractor obtained from a modified version of the Lorenz equation, which was simulated using an experimental electrical circuit. The plot displays the average spectral power across two distinct planes.

**Figure 4.2:** a) It displays the irregular pulsing dynamics of a ruby solid-state laser, taken from\(^8\). b) It presents a numerical simulation of Lorenz chaos, which is described by three nonlinear differential equations depicted in Equation 4.1 with parameters \(r\), \(b\), and \(\sigma\). The values of these parameters have been fixed at \(r=28\), \(b=8/3\), and \(\sigma=10\). The representation presents two time-traces (red and black) for two slightly different initial conditions. c) It displays the trajectories in the three-dimensional phase space (black) with projections in the two-dimensional phase planes (grey). d) It shows the computation of the three largest LEs, \(\lambda_i\), using the Wolf algorithm, which reveals a single positive Lyapunov exponent, \(\lambda_1\). e) It displays the computation of the correlation dimension, \(D\), using the Grassber–Procaccia algorithm, which estimates the lower bound of the fractal dimension by measuring the dimensionality of the space occupied by a set of points. The red dashed line corresponds to \(C_D=2.05\). Courtesy of Sciamanna\(^{12}\).

### 4.2.1.2 Lyapunov exponent

The Lyapunov exponent (LE) has been introduced to study the complexity of chaos. It represents the rate at which two closely situated trajectories deviate from each other. It is used to quantify the sensitivity to initial conditions in chaotic systems. A positive Lyapunov exponent \(\lambda > 0\) indicates that the nearby trajectories will diverge exponentially over time, which is a characteristic of chaos. In contrast, a negative Lyapunov exponent \(\lambda < 0\) indicates that the nearby trajectories will converge, which is characteristic of stable systems. The magnitude and the number of positive Lyapunov exponent is a measure of the complexity of chaos in a system. Larger exponents indicate stronger chaotic behavior. Interestingly, when multiple large positive LEs are present, it guarantees the presence of
complex chaos in the system. This relationship can be expressed as follows:

$\lambda(x_0) = \lim_{n \to +\infty} \frac{1}{n} \sum_{i=0}^{n-1} \ln(|f'(x_i)|)$

(4.2)

where $f$ is a non-linear function verifying $x_{n+1} = f(x_n)$ and $x_0$ is a starting orbit. To gain further insight into the meaning of the Lyapunov exponent, we can express it as $|\delta Z(t)| = e^{\lambda t} |\delta Z_0|$, where $|\delta Z_0|$ represents the initial separation vector, illustrated in the Figure 4.3. It should be noted that the Lyapunov exponent can vary depending on the orientation of the initial vector, resulting in a different spectrum of Lyapunov exponents, however the number of positive LE remain constant. There are various methods available to evaluate these exponents. The definition provided here is applicable to discrete models, but the evaluation of Lyapunov exponents can also be extended to continuous media. As well, the precise evaluation of LE has been a very important challenge through the years$^{13-15}$. In cases where both the dynamical equation and the time series are well known, several algorithms have been developed to assess the LE spectrum. Examples include the algorithms proposed by Sano and Sawada in$^{16}$, Shimada and Nagashima in$^{17}$. The Wolf algorithm is a commonly used method for calculating the maximal LE exponent of a time trace.$^{13}$ The first step of the algorithm involves converting time-domain data into phase space through the application of the time-delay embedding theorem. Next, a point in the trajectory is chosen, and its nearest neighbor is identified by computing the distance between the two points (using a metric such as Euclidean distance), which is denoted as $L_0$. The algorithm then proceeds to track the trajectory of the chosen point over time, continually calculating the distance between it and its nearest neighbor. This process continues until the distance exceeds a predefined threshold, at which point the current distance is stored as $L'_0$ and a new nearest neighbor trajectory is identified, with a starting point at $z(t_1)$ in the Figure 4.3. This process is repeated until the desired level of accuracy is achieved.

4.2.1.3 Dimensions of a chaotic dynamics

The correlation dimension $C$ is a way to measure the dimensionality of a set of random points, which is often considered a type of fractal dimension. The correlation dimension is derived from the correlation integral, which measures the probability that two points in phase space are within a certain distance of each other. By analyzing the scaling behavior of the correlation integral as a function of the distance, the correlation dimension can be estimated. A higher correlation dimension indicates a more complex and intricate attractor.
Figure 4.3: The wolf algorithm is depicted in the schematic representation. Initially, the evaluation of the first Lyapunov exponent (LE) is carried out by assessing the distances between the black and blue curves until a specified threshold is reached. Subsequently, the second Lyapunov exponent is computed using the nearest neighbor approach. This calculation is then repeated for the red curve.

with a greater number of independent degrees of freedom. In comparison to other methods for evaluating dimension, such as those discussed in\textsuperscript{18} and\textsuperscript{11}, the correlation dimension offers several advantages. It can be quickly and easily computed, offers reduced noise for specific data points, and often coincides with calculations in other dimensions. A conservative estimate of correlation dimension can be implemented with the Grassberger-Procaccia (G-P) algorithm\textsuperscript{19,20} that we will describe hereafter.

4.2.1.4 Classification of semiconductor laser: Interest in chaos generation

In addition to steady and pulsing dynamics, semiconductor laser can exhibit complex dynamics such as chaos, which occurs in high-dimensional nonlinear physical systems. For example, laser diodes possess a unique property that makes them highly sensitive to optical perturbations: their emission frequency is detuned from the gain spectrum peak, which creates an anomalous dispersion effect at the lasing frequency.\textsuperscript{21} This property results in a refractive index variation with carrier density, leading to a $\alpha$-factor that explains laser chirp and linewidth broadening. Moreover, this effect facilitates laser instabilities. There are three distinct categories for the laser; class A, class B and class C laser. The distinction lies in the disparity between the gain recovery time, which is compared to the cavity round trip time, and the photon lifetime. Class C laser has a much longer cavity recovery time compare to the photon lifetime; this laser is intrinsically easy to destabilize. They can even destabilize in continuous wave operation. For instance, gas lasers such as NH$_3$ lasers and Ne-Xe lasers are Class C lasers. Lorenz-Haken chaos refers to the chaotic behavior
observed in a free-running laser operating at 81.5µm using NH₃ gas. This chaotic regime is achieved by utilizing low pressure and long wavelength, which collectively contribute to the establishment of chaotic conditions.²² Class B lasers operating without external perturbations do not exhibit deterministic chaos, as they require an additional degree of freedom. However, external perturbations can easily destabilize them. Class B lasers, such as CO₂ lasers and laser diodes, have been extensively studied for their chaotic behavior. Experimental and numerical studies have demonstrated chaotic trajectories in CO₂ lasers by introducing loss modulation as an additional degree of freedom.²³ In this scenario, the laser becomes a chaotic system and shows instabilities. Finally, Class A laser are very hard to destabilize since they need two extra degree of freedom to be destabilized.

4.2.2 Generation of chaos in semiconductor laser

The focus of this section is on the generation of non-linear chaotic dynamics in semiconductor lasers. As depicted in Figure 4.4, various methods can be employed for generating temporal chaos, such as optoelectronics feedback,²⁴ optical injection,²⁵,²⁶ and electrical modulation of the laser’s bias.²⁷,²⁸ External optical feedback in Figure 4.4.a) consist of returning a small fraction of the emitted laser output in order to create a competition between the relaxation oscillation frequency and the external cavity frequency. Optical feedback can either be directly redirected into the laser cavity with the assistance of a mirror, or can undergo modification. For example, we can put in the path diffraction grating, or incorporating polarizers or retarding plates, are methods for manipulating the characteristic of the returning field. The availability of these techniques for controlling chaotic dynamics has made them widely used for generating chaos, given their ease of installation and the ability to choose the desired dynamics. Another way of generating chaos is optical injection as depicted in Figure 4.4.b). When there is significant frequency detuning and/or strong injection, the injected laser can undergo destabilization to chaos through various bifurcation mechanisms. The third methods is using low current modulation to destabilize the laser, leading to a chaotic behavior as depicted in the Figure 4.4.c). Finally, optoelectronic feedback involves converting the laser diode’s output into an electrical current using a photodiode, which is then amplified and reintroduced via the laser driving current. This process results in a time-delayed contribution to the laser diode’s dynamics. The setup is depicted in Figure 4.4.d) In our work, we concentrated on external optical feedback to generate non-linear dynamics. The pioneering experimental investigation of a semiconductor laser subjected to optical feedback was conducted by TKach and Chraplyvy, as presented in²⁹.
Various setups have been devised for generating chaos in semiconductor lasers. Figure 4.4: Various setups have been devised for generating chaos in semiconductor lasers.

This chaos could be used for example to build chaotic LiDARs at near-infrared wavelength. The detection and ranging in CLIDAR are achieved by correlating the waveform of the signal reflected from the target with a reference waveform that has been delayed. The use of reliable detection systems with a range of several meters and precision below one centimeter has been made possible by the high-speed complex dynamics and the fast-decaying autocorrelation function.

### 4.2.3 Rate equation of semiconductor laser subjected to optical feedback

The formalism of a semiconductor laser under optical feedback was initially proposed by Lang and Kobayashi, considering only one roundtrip in the external cavity, which implies a relatively low level of feedback. We inspired from the Olivier Spitz’s thesis for our research. The rate equations describing the laser subjected to optical feedback can be expressed as follows, where \( E \) represents the slowly varying envelope of the complex electric field, and \( N \) is the carrier density of the upper laser state:

\[
\begin{align*}
\frac{dN}{dt} &= \frac{I}{e} - \frac{N}{\tau_c} - G(N)|E|^2 \\
\frac{dE}{dt} &= \frac{1+i\alpha}{2} \left( G(N) - \frac{1}{\tau_p} \right) E + kE(t - \tau_{ext})e^{-i\omega_{ext}t}
\end{align*}
\]  

(4.3)

In the Equation 4.3, various parameters contribute to the overall rate of the system. These parameters include the bias current \( I \), the electron charge \( e \), the carrier lifetime \( \tau_c \), the linewidth enhancement factor \( \alpha \), the round-trip time of light in the external cav-
ity $\tau_{\text{ext}} = \frac{2L}{c}$ where $L$ is the cavity length and $c$ is the speed of light, the gain per unit time $G(N)$, the photon lifetime $\tau_p$, and the free-running laser angular frequency $\omega_0$. The term $\omega \tau$ represents the feedback phase, which corresponds to the angular shift introduced by the feedback compared to the laser’s own phase. Notably, the only modification arising from the feedback loop is given by the term $kE(t - \tau_{\text{ext}})e^{-i\omega_0\tau_{\text{ext}}}$. This additional term accounts for the delay time $\tau_{\text{ext}}$, and the complex field is described by a delay differential equation. It is this delay and feedback mechanism that underlie the instability and chaotic dynamics observed in semiconductor lasers. The feedback coefficient resulting from multiple reflections in the external cavity is denoted as $k$.

$$k = \frac{1}{\tau_{\text{in}}} 2C_l \sqrt{f_{\text{ext}}} \quad (4.4)$$

where $\tau_{\text{in}}$ is the internal cavity roundtrip time and $C_l$ is an external coupling coefficient. It can be expressed for a Fabry-Perot laser as:

$$C_l = \frac{1 - R_2}{2\sqrt{R_2}} \quad (4.5)$$

with $R_2$ the reflection coefficient of the laser front facet subjected to the reinjection. In the case of a DFB laser, the expression of $C_l$ becomes much more complex and depends on the complex reflectivity at both laser facets:

$$C_l = \frac{2(1 - |\rho|^2)e^{-i\phi_0}(q^2 + \kappa^2)L^2}{i\kappa L(1 + \rho_L^2) - 2\rho_L q L} - \frac{1}{2qL - \sum_{k=l,r}^2} \frac{(1 - \rho_k^2)\kappa L}{2iqL\rho_k + \kappa L(1 + \rho_k^2)} \quad (4.6)$$

The given equation represents the complex reflectivity at either the left or right facet of the laser, denoted by $\rho_k = |\rho_k|e^{i\phi_k}$, where $k = l$ or $r$, respectively. Here, the right facet faces the external cavity. $L$ refers to the length of the laser, $\kappa$ is the coupling coefficient of the DFB grating, and $q$ can be expressed as $q = \alpha_{\text{tot}} + i\delta_0$. Here, $\alpha_{\text{tot}}$ refers to the total internal losses, and $\delta_0$ represents the Bragg detuning, which is defined as the deviation between the lasing and the Bragg wavenumbers. Starting from Equation 4.3, if we write $E$ as $\sqrt{S}e^{i\phi}$, we can deduce two rate equations for the amplitude and phase of the electric field.

4.2. "Does the flap of a butterfly’s wings in Brazil set off a tornado in Texas?"
Using Equation 4.7, we can model the behavior of a DFB semiconductor laser under the influence of optical feedback. The simulations can capture the nonlinear and chaotic dynamics, but accurately predicting the actual output of the laser is impossible due to the strong dependence on initial conditions. Through numerical simulations, we can observe that semiconductor lasers can exhibit various behaviors, including instability, bistability, self-pulsation, and coherence collapse states. These findings highlight the rich and diverse range of dynamics that can be observed in semiconductor lasers under optical feedback.

4.2.4 Chaotic dynamics in semiconductor laser

In this section, we will explore various chaotic behaviors observed in semiconductor lasers exposed to optical external feedback. By adjusting the level of feedback \( f \) that is reflected back to the diode, we can demonstrate diverse chaotic phenomena.

- **Regime I:** This regime is characterized by very small feedback effects, where the feedback strength of the electric field amplitude is less than 0.01 %. Depending on the feedback fraction, the laser oscillation linewidth can become broad or narrow.

- **Regime II:** In this regime, the feedback effects are small but not negligible, with a feedback strength of less than 0.1 %. The generation of external modes leads to mode-hopping among internal and external modes.

- **Regime III:** This is a narrow region where the feedback strength is around 0.1 %. In this regime, the mode-hopping noise is suppressed, and the laser restabilizes on the mode with the narrowest linewidth.

- **Regime IV:** This regime corresponds to a moderate feedback strength of around 1 %. Here, the relaxation oscillation becomes undamped, resulting in a greatly broadened laser linewidth. The laser exhibits chaotic behavior and sometimes evolves into unstable oscillations in a coherence collapse state, with a significant increase in the fluctuation level.

- **Regime V:** This is the strong feedback regime where the feedback strength is higher than 10 %, which can only be achieved if the emitting facet of the laser is adequately
coated. In this regime, the internal and external cavities behave like a single cavity, and the laser oscillates in a single mode, with a significantly narrowed laser linewidth.

**Figure 4.5:** Various forms of nonlinear dynamics have been observed in semiconductor lasers, and these patterns have also been documented in mid-infrared QCLs. Courtesy of Spitz\textsuperscript{33}.

### 4.2.4.1 Special focus on coherence collapse regime

In this section, we will focus on the coherence collapse regime, which is the most interesting regime for the following study. The first demonstration of this regime was in 1985 by Lenstra\textsuperscript{35}. The coherence collapse regime arises due to the complex interplay between the laser’s gain, feedback, and coherence properties. In a semiconductor laser, the coherence of the laser output can be described in terms of its spectral linewidth. When the laser operates in a coherent regime, its linewidth is narrow and stable. However, under certain conditions and if the laser is destabilized, the linewidth can begin to broaden and fluctuate irregularly. This is known as coherence collapse and can lead to chaotic behavior in the laser output. This operational condition can present challenges, as it can manifest even without very high feedback in the system, resulting in substantial performance deterioration for telecom diodes. Additionally, incorporating an integrated isolator can be a complex task and may introduce substantial losses to the system.

Nevertheless, the fascinating characteristic of this type of chaos lies in its remarkably wide spectral footprint, as depicted in Figure 4.6, accompanied by intricate chaotic patterns. This intriguing property presents an opportunity to conceal a message within the chaotic dynamics.

### 4.2.4.2 Chaos in cascaded devices lasers

In 2016, researchers demonstrated that quantum cascade photonic can exhibit chaotic behavior. They exhibited low frequency chaotic fluctuation at the power output of the QCL under external optical feedback\textsuperscript{36}. This finding was significant as it was not initially evident.
Figure 4.6: Example of Coherence Collapse chaos in a continuous wave Interband Cascade Laser (ICL) at 15°C and 150 mA. a) Times series of the output of the ICL in a chaos configuration mode. b) Related frequency spectrum of the ICL response, exhibiting a very broad GHz chaos bandwidth. The drop of power around 2.6 GHz is due to the maximum bandwidth of the oscilloscope.

due to the unique structure of the original QCL. The absence of a relaxation oscillation frequency, attributed to the intersubband property and the low $\alpha$-factor, raised doubts about the potential for QCLs to exhibit chaotic dynamics. Indeed, those dynamics only appear for near-threshold pump current and high feedback strength. Initially, oscillations emerged at the frequency of the external cavity. As the feedback strength increased, the observed dynamics evolved to include both the frequency of the external cavity and the distinct dropout signature, indicative of low-frequency fluctuations (LFFs). In addition to experimentally demonstrated temporal LFF chaotic waveforms, the researchers conducted extensive numerical simulations using the Lang and Kobayashi equations, which were previously mentioned. By considering the unique characteristics of QCL, such as a low $\alpha$-factor and an exceptionally low carrier-to-photon lifetime, the simulations revealed a Hopf bifurcation followed by a chaotic bubble as illustrated in Figure 4.7.a)

In the absence of feedback, the laser remains stable and lacks the nonlinear dynamics depicted in Figure 4.7. The initial study demonstrated the potential for triggering chaotic dynamics, but it primarily occurred at current levels very close to the threshold. However, O.Spitz extensively explored the generation of chaotic dynamics in QCLs in his thesis. The research revealed the possibility of exhibiting diverse chaotic behaviors such as Low-Frequency Fluctuation (LFF), Period 1 oscillation, Period 2 oscillation, and Coherence Collapse. Intersubband technology seems to be an elective choice because it does not exhibit a relaxation oscillation frequency, which is a common limitation for non-linear dynamics bandwidths. Moreover, simulations have shown the possibility to generate broadband chaos with QCL. Nonetheless, the maximum demonstrated chaos bandwidth
Figure 4.7: a) Numerical bifurcation diagram for a bias $P = I/I_{th} - 1 = 0.02$ and an external cavity length $L_{ext} = 35\text{cm}$ and associated time series. b) Stable regime of the QCL for a feedback ratio of 0.11 %. c) Period 1 oscillation for a feedback ratio of 2.14 %. d) Low-Frequency-Fluctuation (LFF) for a feedback ratio of 2.59 %. Courtesy of Jumpertz\textsuperscript{36}

still remains confined to the range of 100 MHz. The underlying factors contributing to this surprisingly limited bandwidth are not yet fully comprehended.\textsuperscript{40,41} Interestingly, we were able to study a different cascaded laser structure as so called ICL. These lasers exhibit relaxation oscillations in the GHz range, as confirmed through both numerical simulations\textsuperscript{42} and experimental\textsuperscript{43} investigations, which is particularly intriguing for broad band chaos. Recent experimental findings\textsuperscript{43} have revealed that ICLs can generate hyperchaos, characterized by multiple positive Lyapunov exponents\textsuperscript{44}. Simulation results have shown promising dynamics bandwidth\textsuperscript{42}, although further experimental investigation is required to explore the influence of the number of stages on the dynamic behavior. Notably, the typical bandwidth of hyperchaos in ICLs is one or two orders of magnitude larger than that in QCLs, reaching frequencies above 1.2 GHz, as illustrated in Figure 4.6.

4.2.5 Chaos synchronization

The synchronization of chaotic systems, wherein two chaotic systems can align their outputs, is a captivating phenomenon. It is remarkable due to the inherent dependence of chaotic systems on initial conditions. Initially, many researchers believed that synchronizing two chaotic systems was highly improbable due to the complexity involved. The
exponential divergence of nearby trajectories made it seem paradoxical that chaotic systems could converge. Chaos and synchronization were considered incompatible. However, this perspective was challenged and disproven in the late 1980s, establishing synchronization of chaos as a natural phenomenon. Interestingly, synchronizing chaos had to be rediscovered multiple times before it gained widespread acceptance. The breakthrough came with the pioneering work of Pecora and Carroll, who introduced the concept of chaos synchronization. They demonstrated the synchronization of chaotic systems using a straightforward coupling scheme known as the "Pecora-Carroll method" or "identical systems method". Subsequently, several experimental demonstrations of chaos synchronization were conducted in various systems, including electronic circuits, lasers, and mechanical systems. These experiments validated the theoretical concepts and opened up new possibilities for practical applications. The first demonstration of electronic circuit synchronization occurred in the early 1990s by Ott, Grebogi, and Yorke, who published the influential paper titled "Controlling Chaos". They achieved chaos synchronization by constructing two nearly identical electronic circuits, known as "chaos generators," and introducing a weak coupling between them. The chaotic signals from both circuits synchronized, with their states becoming correlated and following similar trajectories. In 1994, Roy and Thornburg achieved a significant breakthrough by demonstrating the synchronization of two mutually coupled solid-state Nd:YAG lasers. Sugawara also demonstrated chaos synchronization with two CO2 lasers. Later, synchronization of the phase of two lasers was demonstrated as well. These early demonstrations of chaos synchronization paved the way for further research, leading to a deeper understanding of synchronization phenomena and their applications in various fields.

**Figure 4.8:** Schematic diagram of chaos synchronization systems in semiconductor lasers with optical feedback.

As shown in the Figure 4.8, there are various methods to achieve complete chaos synchronization between two lasers. Chaos synchronization is achieved by injecting the master...
signal from a laser into the cavity of a second laser that satisfies the injection locking condition. However, in an open-loop configuration, the receiver laser is only driven into a chaotic state if it is injected with the transmitted signal. Due to the practicality and simplicity of this setup, the ability to ignore phase-matching parameters, and the effectiveness of the chaos synchronization, the open-loop configuration has become the preferred method.

4.3 Laser chaos communication

Chaos synchronization, as discussed before, plays a crucial role in chaos communications. The theoretical studies in the 1990s played a crucial role in establishing the foundation for chaos communication. Researchers recognized the potential of chaotic systems for secure communication due to their sensitivity to initial conditions and the inherent unpredictability of their future behavior. Different methods could be used to use chaos to conceal an information.

4.3.1 Chaos Modulation (CMO)

Chaos modulation involves utilizing the chaotic signal generated by a chaotic system as the carrier signal. The information signal, whether it is in analog or digital form, is then combined with this chaotic carrier signal through a modulation process. The modulation techniques employed can include amplitude modulation (AM), frequency modulation (FM), or phase modulation (PM), depending on the specific requirements of the implementation. We encode the desired message into the carrier chaotic signal. Experimental demonstrations using analog electronic circuits arrived just after the demonstration of chaotic synchronized system and confirmed the feasibility of chaos modulation, and subsequent developments focused on chaos demodulation techniques to extract the original message from the chaotic carrier. The initial demonstration of chaotic modulation and demodulation (CMO) was conducted by Colet and Roy in their work published in 1994. Through numerical simulations using modulated solid-state lasers, they explored the potential of chaotic communications using nonlinear systems. Their research highlighted the possibility of transmitting chaotic data, specifically a binary bit-sequence, at a rate of 100 kbit/s within the system. VanWiggeren and Roy conducted an influential study on data transmission for secure communications using chaotic modulation and demodulation (CMO). Their research involved laser systems, specifically proposing a ring fiber laser system equipped with an optical feedback loop, also known as a delay loop, to generate
Figure 4.9: Different method for achieving private chaos based communication. a) Chaos Masking. b) Chaos Modulation. c) On-Off Chaos Sift Keying. Courtesy of Locquet\textsuperscript{54}

chaos. By introducing the message to be transmitted as a modulation within the feedback loop, a new chaotic oscillation was generated in the system. The researchers successfully demonstrated data transmission at a rate exceeding 100 Mbit/s, showcasing the potential of CMO for high-speed communication.
4.3.2 Chaos Masking (CMa)

In this communication scheme, a chaotic laser carrier is utilized, and a small message is embedded within it. You can retrieve a schematic diagram in the Figure 4.9.a. The entire signal, comprising the chaotic oscillation and the embedded message, is transmitted to the receiver. Through chaos synchronization and a chaos-pass filtering effect within the receiver laser, only the chaotic oscillation is reproduced. By subtracting the output of the receiver from the transmitted signal, the message can be successfully decoded. The first experimental attempts in chaotic communications involved analog circuits employing different dynamical systems, such as the Chua system and the Lorenz system. In the case of the Lorenz system, a setup with two unidirectionally coupled Lorenz circuits was used, where one circuit served as the emitter and the other as the receiver. The primary objective was to conceal a small speech signal within the fluctuations of the chaotic signal generated by the master circuit. The slave circuit generated its own synchronized chaotic signal, allowing for the retrieval of the hidden speech signal by subtracting the two signals.

4.3.3 Chaos Shift Keying (CSK)

In the chaos-shift keying (CSK) technique, as depicted in Figure 4.9, the emitter system’s parameter, denoted as \( p \), can have two possible values, \( p_0 \) or \( p_1 \), based on a binary information-bearing signal. This parameter modulation leads to a synchronization error, \( e(t) = x_e(t) - x_r(t) \), at the receiver, where \( x_e(t) \) and \( x_r(t) \) represent the emitter and receiver signals, respectively. The amplitude of this error depends on the parameter modulation and reveals the information encoded in the binary signal. Alternatively, two receivers can be utilized, one set to \( p_0 \) and the other to \( p_1 \), to recover the message. The alternating synchronization of the two receivers reveals the value of the binary message. The idea is to implement an array of dual-paired chaotic lasers as is implemented in those papers.

4.3.3.1 Communication principle for laser private transmission

In the case of on-off chaos-shift keying (OOC SK), which is a specific variant of CSK, the receiver parameter is set to one of the emitter parameter values. Depending on the binary message value, the receiver can either synchronize or fail to synchronize with the emitter. However, it is also possible for the receiver to synchronize in both configurations, but with an error that carries the message information. This configuration is commonly used for private laser transmission, as it allows for easy manipulation of the parameter values (\( p_0 \) and \( p_1 \)) by modulating electrically the laser. A schematic representation is depicted in the
Figure 4.10: On-Off Chaos Sift keying schematic scheme. The red color shows the possibility of the laser to synchronize their chaotic output. Courtesy of Locquet.\textsuperscript{54}

Figure 4.10. In a real-world application, a 2.4 Gbit/s chaos-based private transmission was successfully demonstrated in the near-infrared domain at a wavelength of 1.55\,\textmu m over a 120-km fiber link in the metropolitan area of Athens, Greece\textsuperscript{50}. However, the use of both the fiber medium and commercial laser diodes imposes limitations on the application’s relevance and data integrity.

### 4.4 Private free space communication in the long wave infrared domain

Free-space communication, on the other hand, is inherently suitable for the cryptographic method since there is no physical layer protecting the message. A chaos-based communication between unidirectionally-coupled 5.6\,\textmu m QCLs has been achieved at a data rate of 0.5 Mbit/s\textsuperscript{60}. Unfortunately, the emission wavelength is not within the thermal atmospheric window, making long-distance implementation impractical. Also, the generated chaos bandwidth hindered the maximum data rate accessible. In this study, we achieve an 8 Mbit/s chaos-based private communication between two coupled DFB QCLs emitting at 9.3\,\textmu m.

#### 4.4.1 Experimental setup

The experimental setup, depicted in Figure 4.11, involves a continuous wave room-temperature distributed feedback QCL emitting at 9.3\,\textmu m as the transmitter, referred to as master laser (ML). The QCL is biased with a low-noise generator, and modulated with a OOK PRBS through an arbitrary waveform generator. The setup includes an isolator to avoid back-
Figure 4.11: The experimental setup is presented here. A Continuous-Wave Distributed Feedback Quantum Cascade Laser (DFB QCL) emitting at $9.3 \mu m$ is subjected to an external optical feedback configuration which makes him chaotic. An Arbitrary Waveform Generator (AWG) modulates the QCL using a Pseudo-Random-Binary-Sequence (PRBS) On-Off-Keying (OOK) modulation with an amplitude of less than $5 \text{ mV}$. The transmitted beam (message+chaos) is sent to the receiver through an isolator to prevent back-coupling reflections. The path length can be adjusted between a 2-meter and a 31-meter configuration using a Herriott cell. The transmitted signal is injected into a second QCL, which is synchronized with the chaos of the first QCL. The emitted signals are collected by two Mercury-Cadmium-Telluride (MCT) detectors with $700\text{MHz} 3\text{dB}$ bandwidth, one for the master signal and the other for the slave signal. A custom MATLAB program processes the master and the slave signals to extract the concealed message.

coupling reflections, and a Herriott cell to adjust the path length. The transmitted signal is synchronized with a second QCL referred to as Slave Laser (SL), which receives the transmitted signal after passing through an isolator. The master and slave signals are collected using two $700\text{MHz} 3\text{dB}$-bandwidth Mercury-Cadmium-Telluride detectors and processed by a custom MATLAB program which will be described below.

4.4.1.1 Laser characterization

The QCLs are designed using molecular-beam-epitaxy with a $6\mu m \times 2 \text{ mm}$ buried structure, with AlInAs electron barriers and GaInAs quantum wells. The lasers are designed for monomode emission and have a high reflective back facet to minimize losses. The feedback intensity is controlled using a mid-infrared polarizer, and the cavity length is set at approximately 20 cm. To protect the master QCL from back reflections, a mid-infrared optical isolator is used. The receiver SL presents the same design geometry as the master, with a matched emitting wavelength. The data is captured by a Tektronix oscilloscope.
Figure 4.12: a) Light-Intensity-Voltage (LIV) of the two DFB-QCLs: Master laser (ML) and the Slave Laser (SL) at 290 K. The two LIV responses are very similar, showing the matching geometry. b) Optical response of the ML and the SL at a current of 280 mA and a temperature around 283 K. The bandwidth of the master and the slave laser are 10 MHz and 7 MHz, respectively. c) Optical spectrum of the ML and the SL. The two QCLs are DFB up to 380 mA and show a near-perfectly-matched emission wavelength. After that, the two QCLs become highly multimode because of geometry defects.

set at a sample rate of 250 MS/s, and we restrain the electrical bandwidth to 20 MHz to only account for the main features in the QCLs’ signals. The identical physical parameters required for synchronization are difficult to achieve due to process variations during fabrication, leading to deviations in the Light-Voltage-Intensity curves. The characteristics of the two DFB lasers are highlighted in Figure 4.12.

Figure 4.13 illustrates the evolution of the optical spectrum of the slave laser as the bias current increases, while maintaining a temperature of 293 K. The dominant lasing mode occurs at approximately 9.3 µm, exhibiting a single mode characteristic for bias currents below 380 mA. However, beyond this threshold, additional side modes emerge, growing in intensity with increasing current. Notably, the spectrum in Figure 4.13 demonstrates a red shift in the lasing mode as the bias current is raised. Observations of mode hopping can be made at specific current values (e.g., at 340 mA in Figure 4.13.a)). This phenomenon arises due to competition between modes and non-uniform broadening of the gain, particularly associated with Fabry-Perot modes. The weak coupling coefficient (estimated at
Figure 4.13: Optical spectrum of the free-running slave laser when it is biased above threshold from 300 mA to 400 mA. The QCL is monomode when biased with currents below 380 mA; above, side modes appear. The amplitude of the side modes increases with the injection current. Optical red shift of the lasing mode as the current bias increases. A wavelength shift approximately $7 \text{ cm}^{-1}$) indicates a limited interaction of the field with the active region of the device, making mode jumping more likely. Furthermore, it is important to note that the threshold of the slave laser decreases to approximately 291 mA when it is optically injected with the master laser beam.

4.4.2 Chaos synchronization

To achieve chaos synchronization, the master signal is injected into the slave cavity under the injection locking condition. The open-loop configuration is preferred as it allows for easier setup, neglecting the phase-matching parameters of the feedback and ensuring efficient chaos synchronization as already discussed. For precise optical injection of the master laser beam into the slave laser cavity, we employ a method that involves modulating the injection voltage of the master laser using a sinusoidal waveform with a frequency of 100 kHz and an amplitude of 200 mV. The resulting modulation of the master laser’s output emission can be observed in Figure 4.14.a). The slave laser, in response to the injection, exhibits periodic intensity peaks when the wavelengths of the two lasers overlap. This behavior enables us to accurately align the optics between the master and slave lasers. Moreover, precise setting of the operating current and temperature of both sources is crucial to complete effective chaos synchronization. Next, we evaluate the optimal current value that enables the synchronization of the two lasers. The continuous wave operation is maintained for both lasers, with the slave laser bias gradually varied from the laser threshold up to 410 mA, while the master laser bias remains fixed at 370 mA. The choice of the master current value is based on the observation of relevant chaotic dynamics in...
the laser output at that particular current spot. By sweeping the current of the slave laser, it leads to three equidistant injection locking ranges, as shown in Figure 4.14.b). Each injection locking range displays both synchronization and anti-synchronization behavior, with a drop in the cross-correlation coefficient at the turning point. The dependence of synchronization quality on the coupling with the slave side-modes is still under analysis, with the maximum cross-correlation coefficient at 97% around 290 mA, and slightly smaller values reached around 350 mA.

Figure 4.14: a) The time traces of ML (shown in blue) and SL (shown in red) are displayed during injection locking. The ML is biased at 371 mA and a temperature of 276 K, while the SL is biased at 297 mA and a temperature of 287 K. The effectiveness of the injection of the master QCL is indicated by the peaks in the slave laser response. b) The black dots in the experimental data represent the maximum of the master-slave cross correlation (CC) as a function of the slave bias current. The synchronization intervals are marked in blue, while the anti-synchronization intervals are marked in orange. The master operating point is at a bias of 370 mA and a temperature of 276 K. The slave operating point is at a temperature of 287 K, with a bias sweeping from 290 mA to 410 mA.

4.4.3 Chaos characterization

The complexity of chaos plays a crucial role in our system’s ability to resist attacks from eavesdroppers. Unpredictability, which is a characteristic of chaos, makes it impossible for an illegitimate receiver to distinguish a low-amplitude message from high-amplitude chaos. In this experiment, chaos is generated using optical feedback and the message is encrypted using chaos on-off-shift-keying (OOSK) scheme.\textsuperscript{61,62} The optical self feedback setup consists of a beam splitter and a tilted feedback mirror, and the feedback strength is adjusted with a wire-grid mid-infrared polarizer. The master laser is maintained at 276 K to observe more complex chaos dynamics, and the low-amplitude message is used to conceal
it within the chaotic trace. The chaos pattern is extremely sensitive to initial conditions, and a small-amplitude modulation of the master current helps to stabilize it. The objective of this experiment is to achieve the most complex chaotic dynamics, which is represented by coherence-collapse dynamic, since the chaos bandwidth determines the maximum data rate of the encrypted message. Figure 4.15 shows the temporal-frequency characteristic of complex dynamics in QCL, and statistical analysis using Lyapunov exponents (LEs) was performed to uncover the features of chaotic dynamics. This chaos characterization has been performed by Pr. Wu from Southwest University. The number of positive LE values obtained indicates that the optical feedback QCL transmitter emits a hyperchaotic waveform, which is suitable for secure communication.

![Fig 4.15](image)

**Figure 4.15:** a) The Fourier transform power spectrum of the chaotic temporal waveform. b) Calculated Lyapunov exponents (LEs) spectrum. The curves converge to the LE values as $\text{LE}_1 \approx 23 \mu s^{-1}$, $\text{LE}_2 \approx 14 \mu s^{-1}$, $\text{LE}_3 \approx 7 \mu s^{-1}$, $\text{LE}_4 \approx 0.7 \mu s^{-1}$ and $\text{LE}_5 \approx -6.2 \mu s^{-1}$. c) Measured three-dimensional portrait in phase space. Blue dots are the measured raw data, while the solid red curve is the reconstructed trajectory. d) The correlation integral $C_D(r)$, where the black curves are the Logarithmic plots of $C_D(r)$ versus sphere radius r based on the Grassberger-Proccacia algorithm, and the red curves are the slope of the $C_D(r)$ versus sphere radius r. A clear plateau on the slope of the correlation integral is observed, and marked with the correlation dimension $D_2$ is estimated at $\approx 6$.

In Figure 4.15.c), the phase portrait of chaos is shown in a 3D space defined by the amplitude of the waveform, the first and second derivatives (vertical axis). The blue...
phase portraits depict projections of the 3D portrait onto the corresponding phase planes, revealing the complex geometrical and topological structure of chaos, as well as local instability and global stability\textsuperscript{11}. In order to obtain accurate measurements from recorded timetraces, it is crucial to mitigate the influence of noise in the measured waveform. To achieve this, a state-space averaging method can be applied, which has demonstrated its effectiveness in suppressing noise in chaotic waveforms.\textsuperscript{11,63} The noise-removed waveform is denoted as \( \tilde{X}_n \), where \( X_n \) represents the n-th data point of the original waveform. The specific process is described by Eq. (4.8).

\[
\tilde{X}_n = \frac{\sum_{k=m}^{N-m} X_k \omega_n(k)}{\sum_{k=m}^{N-m} \omega_n(k)} , \quad \text{where } \omega_n(k) = \exp\left(- \frac{\sum_{j=-m}^{m} (X_{k-j} - X_{n-j})^2}{\sigma_{\text{noise}}^2}\right) \quad \text{(4.8)}
\]

where \( N \) is the time-series vector size, \( m \) is the embedded dimension, and \( \omega_n(k) \) is the corresponding weight. \( \tilde{X}_n \) averages the dataset \( X_n \) in a sliding \( \pm m \) points window. Next, the Lyapunov exponents (LEs) provide a quantitative characterization of the average rate at which neighboring trajectories in phase space diverge from each other. In the reconstructed multidimensional phase space, two trajectories in phase space with initial separation \( \delta Z_0 \) diverge (being treated with the linear approximation) at a rate given by: \textsuperscript{11} \[ |\delta Z(t)| \approx e^{\lambda t} |\delta Z_0| , \]

where \( \lambda \) is the LE. The rate of separation is usually different for different orientations. Thus, there is a spectrum of Lyapunov exponents. Two (or more) positive LEs system is usually called a hyperchaotic system, meaning complex divergent dynamics. Zero LE is characteristic of the periodic trajectory component. For the estimation of correlation dimension \( D_2 \), we implement the Grassberger-Proccacia (G-P) algorithm.\textsuperscript{20} The correlation dimension offers the advantage of being computationally efficient and requiring only a small number of data points. When calculating the correlation dimension, the value of \( D_2 \) can be obtained by calculating \( C(r) \):

\[
C(r) = \frac{1}{N^2} \left( \sum_{n=1}^{N} \sum_{m=1}^{N} (d(X_n, X_m) < r) \right) \quad \text{(4.9)}
\]

This formula is a numerical computation of the average number of vectors that could be found within a sphere of radius \( r \) around a given vector. Distance \( d \) is the Euclidean norm. Then we obtain the formula for \( D_2 \):

\[
D_2 = \lim_{r \to 0} \frac{\partial \log_2(C(r))}{\partial \log_2(r)} \quad \text{(4.10)}
\]
4.4.4 Private communication performance evaluation

4.4.4.1 Algorithm

In Figure 4.16, a sequence of digital post-processing operations is illustrated, which aims to recover the message, enhance transmission quality, and evaluate system performance. MCT detectors are employed to distinguish the master and slave signals, which are then analyzed using the oscilloscope and a MATLAB program. The program applies a finite-impulse response (FIR) bandpass filter (BPF) to eliminate low-frequency noise, as MCT detectors have a minimum detectable frequency of 10 kHz. The passband and stopband high frequency are adjusted based on the data rate of the encrypted message to optimize the signal-to-noise ratio while preserving information transmission. Moreover, this filtering process removes the high-frequency chaos contribution, which is not taken into consideration during the difference computation. The message is reconstructed by evaluating the intensity difference between the chaos transmitter \( I_{ct} = |E_{master} + M(t)|^2 \) and the chaos-synchronized slave \( I_{cs} = |E_{slave}|^2 \approx |E_{master}|^2 \), resulting in \( I_{ct} - I_{cs} = 2 \text{Re}(E_{master}M(t)) + |M(t)|^2 \). To compensate for the amplitude disparity between the master and slave time traces, a multiplication factor.
factor called $\beta$ is introduced, ranging from 0.7 to 1.3 in steps of 0.015. Digital equalization plays a vital role in compensating for systematic channel distortions and minimizing inter-symbol interference (ISI) to enhance the maximum data rate. For this purpose, a fractional-spaced FFE is utilized on the down-sampled difference signal, and the coefficients of the FFE filter are determined using a gradient descent algorithm with a convergence parameter $\mu$, which controls the learning step’s speed and accuracy. The received signal is equalized using the learned filter, and the number of filter coefficients, defined as $n_{\text{tap}}$, is set to maintain low latency, typically below 400 samples (or 100 symbols). The values of $\mu$ and $n_{\text{tap}}$ are optimized to achieve efficient transmission quality. A smaller value of $\mu$ yields a more precise but slower estimation of channel distortions, while a higher value leads to a faster but less accurate estimation. Finally, pulse-amplitude demodulation through threshold detection is performed, followed by the visualization of the eye diagram for the master signal and the difference signal. The associated BER is calculated to quantify and visualize the transmission privacy and system performance.

4.4.4.2 Evaluation of synchronization quality

Figure 4.17.a) and Figure 4.17.d) present the master signal (in blue), the slave signal (in red), the original message (in green), and the difference signal (in purple), respectively, after suitable filtering and normalization. The message is a non-return-to-zero (NRZ) pseudo-random binary sequence (PRBS) with a length of $2^7 - 1$ and a modulation amplitude ranging from 2 to 10 mV. Figure 4.17.b) and Figure 4.17.e) show the 1D auto-correlation and cross-correlation, respectively, for two different data rates. With the use of a 127-bit sequence repetition, the auto-correlation diagram reveals the presence of this repetition through side peaks in the master signal. Specifically, at a transmission rate of 5 Mbit/s, only two side peaks with low amplitude levels (<15 %) are visible. However, at an 8 Mbit/s transmission rate, the presence of a single narrow main lobe at nearly zero time lag indicates a very low signal-to-chaos ratio, indicating effective encryption of the PRBS signal. In a real transmission scenario, the bit patterns may not be as clearly repetitive. To evaluate the synchronization quality between the master and slave signals, the cross-correlation is computed as a function of the time lag between the time traces. A higher maximum value in the cross-correlation indicates better synchronization. Figure 4.17.c) and Figure 4.17.f) illustrate the 2D correlation heatmap, where the master intensity is plotted against the slave intensity. In the case of synchronization, a good synchronization is indicated by a straight line with a positive slope of 45°. To account for anti-synchronization, the intensity of the slave laser is flipped, causing the dashed white line indicating the trend
Figure 4.17: Experimental timetraces for 5 Mbit/s (a,b,c) and 8 Mbit/s (d,e,f) data-rate transmission. a) (blue) master signal, (red) inverted slave signal, (green) original message, (purple) equalized difference. b) cross-correlation and auto-correlation diagrams for the filtered intensity of the slave and the master QCLs. c) 2D Correlation diagram for the intensity of the ML (y axis) and SL (x axis). The signal of the slave is flipped to have a positive correlation. The d), e), and f) has the same purpose as the a), b) and c) respectively but for the 8 Mbit/s transmission.

to have a positive slope.

4.4.4.3 Performance of the private transmission

The quality of the signal transmission is verified using the eye diagram presented in Figure 4.18. The eye diagram is created by overlaying all the transmitted bits (approximately 50 kbits for a data rate of 8 Mbit/s) within a time interval equal to twice the inverse of the data rate. The eye diagram and BER are visualized for both the master signal and the difference signal at two different data rates: 5 Mbit/s and 8 Mbit/s. In this analysis, it is assumed that the eavesdropper is in an unlikely configuration where they know the exact data rate and coding format (NRZ) of the encrypted message. From Figure 4.18.a) and c), it is clear that the signal retrieved at the receiver end, after digital processing and equalization, is completely indiscernible. However, the legitimate receiver (Figure 4.18.b) and d)) benefits from a very low BER (<4%), enabling error-free transmission after FEC algorithm. The insets highlight the possibility of obtaining a decipherable picture without
error code correction. Higher data rates are not considered due to laser bandwidth limitations, which would result in higher difference BER. For visual purposes, insets 1-4 of Figure 4.18.a)-d) display the results of the attempted private transmission, where a colored image represents the encrypted message. Each pixel’s information is converted into a bit stream for transmission and then transformed back into an image format at the receiver end. The image remains well concealed in both scenarios. Analyzing the difference signal (Figure 4.18.b) and d)), the 1.1% difference in BER between the two cases results in a greater number of correctly deciphered pixels for the 5 Mbit/s transmission. Overall, the meaningful content of the image is recovered regardless of the data rate.

4.4.4.4 Herriott cell transmission.

We employed an Herriott cell to increase the range of the free-space transmission path. To ensure a high-quality beam output, we added an extra telescope before the multi-pass cell to shape the beam. With this setup, we achieved a private transmission rate of 1Mbit/s, while maintaining a low BER of approximately 4% for the intended receiver and a high
BER of around 20% for potential eavesdroppers, as shown in Figure 4.19. The outcomes of utilizing higher amplitude electrical signal modulation for transmission can be observed in appendix D. In this case, the BER for both the legitimate user and the eavesdropper decreases simultaneously.

![Eyediagram for 1 Mbit/s through the Herriott cell which is equivalent to 31 meters propagation: a) Master signal exhibiting an error rate of 25%. b) difference signal exhibiting an error rate of 4.1%.

**Figure 4.19:** Eyediagram for 1 Mbit/s through the Herriott cell which is equivalent to 31 meters propagation: a) Master signal exhibiting an error rate of 25%. b) difference signal exhibiting an error rate of 4.1%.

### 4.4.4.5 Attack on the communication system

Finally, let’s consider the challenges faced by an eavesdropper in a ground-to-ground communication application. Firstly, localizing the transmitted beam in free space poses numerous obstacles. The divergence at MIR wavelengths is lower than that at radio frequencies, making it more difficult to precisely locate the beam. Additionally, the 9.3 µm wavelength falls within a range of the electromagnetic spectrum where thermal atmospheric radiation is at its highest, further complicating interception. Assuming that the eavesdropper can still detect the signal, there are two potential approaches they could take to decipher the hidden message. The first possibility is to directly process the chaotic signal in an attempt to recover the encrypted message. As depicted in Figure 4.15, the high complexity of the generated photonic chaos ensures a high level of unpredictability, making the task extremely laborious. Consequently, the signal-to-chaos ratio is very small (< 1/50), rendering traditional signal processing methods ineffective for decryption. Advanced chaos analysis techniques may be explored, although their analysis falls outside the scope of this thesis. The second option for the eavesdropper is to fabricate and process their own QCL without having information about the transmitter source, relying solely on the emitted and transmitted signals. To achieve synchronization with the master laser, the eavesdropper’s laser would need to have a very close emission wavelength and an identical geometry. If this synchronization cannot be guaranteed, only partial synchronization may be achieved, making effective message recovery impossible. Even in our optimized configuration, the
BER is already at a few percent, which is close to the maximum error value that a 27% overhead FEC can tolerate. Fine-tuning of temperature and current parameters is crucial to achieve good correlation between the master and slave lasers. Finally, if the private transmission is intercepted, the power experienced by the slave laser would undergo a significant drop, resulting in an immediate interruption of communication. This serves as an additional barrier for eavesdroppers, as any attempt to intercept the signal would disrupt the communication process.

**Figure 4.20**: Evaluation of the quality of the transmission for the legitimate user and the eavesdropper. The summary encapsulates BER outcomes for private FSO transmission, considering different data rates and two transmission scenarios, including a 2-meter B2B configuration and a 31-meter Herriott cell transmission. The dashed line represents extraction of the message from the signal transmitted by the master, i.e. the signal intercepted by an eavesdropper. The straight line corresponds to the deciphered message after performing the difference.

We studied the BER as a function of the amplitude of the modulated PRBS message embedded within the chaotic signals. Our investigation considered three different data rates: 1 Mbit/s (with HC), 5 Mbit/s, and 8 Mbit/s. The results are presented in Figure 4.20. Increasing the modulation amplitude enhances the signal-to-chaos ratio, which simultaneously reduces privacy and increases the likelihood of an eavesdropper recovering the message. However, it also improves the ability of the intended receiver to decode the signal. To achieve a BER greater than 25% for an eavesdropper and less than 4% for the intended receiver, it is crucial to carefully adjust the voltage amplitude of the modulating signal. The amplitude of the modulating signal exhibits a strong correlation with the BER observed by an eavesdropper in the 1 Mbit/s transmission through the Herriott cell. This suggests that the chaos amplitude was only marginally sufficient to correctly mask the message. A lower efficiency in message masking leads to a lower BER for an eavesdropper.
However, for transmissions at 5 and 8 Mbit/s without the Herriott cell, the increase in message amplitude has minimal impact on the eavesdropper’s BER, which remains above 40%. In these cases, the chaos amplitude is sufficiently high to maintain a low signal-to-chaos ratio. For higher data rate transmissions, larger message amplitudes are required to achieve an acceptable BER for the intended receiver. Hence, the measurements reported at 8 Mbit/s were limited to the range of 7-10 mV. This is specific to our configuration, as the electrical modulation bandwidth of the master QCL is approximately 8 MHz (as shown in Figure 4.12.b)), and frequencies above a few MHz are already attenuated in our system. Therefore, inputs with larger amplitudes are needed to compensate for this attenuation.

4.4.4.6 Summary of the results

A private mid-infrared free-space communication at 9.3 µm with improved message recovery through digital signal processing techniques was successfully demonstrated. The utilization of the Herriott cell in the experimental setup indicates the potential for achieving long-distance secure communication in the mid-infrared region, particularly for applications involving ground stations or vehicles and aircrafts with lower data rate requirements.

Table 4.1: Summary table reporting the main private transmission experimental results. The term 'HC' designates a transmission distance of 31 meters.

<table>
<thead>
<tr>
<th>Data rate</th>
<th>CC</th>
<th>Master BER</th>
<th>Difference BER</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 Mbit/s (HC)</td>
<td>92%</td>
<td>25%</td>
<td>4%</td>
</tr>
<tr>
<td>5 Mbit/s</td>
<td>84%</td>
<td>42.9%</td>
<td>1.4%</td>
</tr>
<tr>
<td>8 Mbit/s</td>
<td>85%</td>
<td>41.3%</td>
<td>2.5%</td>
</tr>
</tbody>
</table>


Chapter 4 • Free-space private communications at long-wave infrared wavelength


Chapter 4 • Free-space private communications at long-wave infrared wavelength
The increasing number of broadband internet users and the emergence of the Internet of Things (IoT) have driven the development of Free-Space-Optical (FSO) communication systems as a supplement to the existing fiber network. FSO systems provide a fast, reliable, and cost-effective solution for ground-to-ground communication and transmissions between the ground and low-orbit satellites. As depicted in a previous chapter, recent breakthroughs in transmission rates have been achieved using QCLs and ICLs. Despite the benefits of mid-infrared FSO systems, their use is currently limited to line-of-sight operations, which can hinder communication through obstacles. Furthermore, the low divergence of the beam makes impossible the end-user communication.

5.1 Current RF system and needs for higher frequency

An RF signal refers to an electrical signal that carries information or data through the air using radio waves. It is a type of electromagnetic radiation that fall within the radio frequency portion of the electromagnetic spectrum. They have longer wavelengths and lower frequencies compared to other forms of electromagnetic radiation, such as visible light, X-rays, and gamma rays. RF waves are used extensively in various applications, including wireless communication, broadcasting, remote sensing, and scientific research. The current RF system has served as the fundamental framework for wireless communication, offering a diverse array of applications and services. These existing RF systems predominantly operate within the S band, typically ranging from 2 to 4 GHz, with some extending to higher frequencies in the tens of gigahertz.

5.1.1 Low frequency: < 1 GHz

The lower frequencies in the RF spectrum (0-1 GHz) find extensive applications in wireless communication, broadcasting (television, radio). These frequencies are chosen due to their ease of generation and detection. Moreover, they exhibit good propagation characteristics,
allowing signals to travel long distances and penetrate building walls and foliage with ease. The low frequency carrier hindered the possible data rate as well as necessitating important size of antennas. Development has moved to higher carrier frequency to increase the potential data rate.

5.1.2 S band: 1-6 GHz

The shift towards higher frequencies arises from the demand for increased data rates and smaller antenna dimensions. In this context, antenna size is closely linked to the wavelength, and higher carrier frequencies lead to shorter wavelengths, enabling the design of compact antennas. This attribute is especially beneficial for space-limited settings and portable devices, facilitating the development of mobile internet connectivity technologies such as EDGE, 1G, 2G, 3G, and 4G, as well as enabling functionalities like Wi-Fi and Bluetooth.

5.1.3 Ka band: 26-40 GHz

The need for Ka band stay the same as the move to S band like increase data rates and smaller antenna size. Actual application of the Ka band is satellite intercommunication or point to point links to backhauling network traffic between cellular towers, connecting remote locations, and establishing communication links between data centers. In our re-
search endeavors, we initially focused on the Ka band (26-40 GHz) due to its favorable cost-effectiveness in terms of electronics and detector bandwidth. We explore the potential of mid-infrared intersubband technology and heterodyne mixing as a promising solution for achieving high-speed photonic-wireless links in the RF band. While these frequency ranges have effectively supported numerous applications, they are now facing challenges due to the growing number of wireless devices and the limited availability of spectrum.

5.1.4 W band: 75-110 GHz

However, the W band (75-110 GHz) is the ultimate target frequency for several compelling reasons. Firstly, it offers exceptional data transmission capabilities, enabling high-speed communication with the use of compact antennas. The wider frequency range allows for larger bandwidth, which translates to increased data rates and enhanced system performance. Additionally, the W band exhibits reduced interference compared to lower frequency bands, resulting in improved signal quality and reliability. The utilization of the W band is currently limited due to the very expensive cost of such systems. Nonetheless, commercialized systems already exist. LightPoint, for example, has introduced an RF system operating at 80 GHz with a capacity of up to 20 Gbit/s.²

5.1.5 Reaching the THz range...

The congestion of RF spectrum has prompted the need for higher frequency solutions to meet the escalating demands for wireless connectivity such as the terahertz (THz) and sub-terahertz frequencies. These higher frequency ranges, with their larger available bandwidth and reduced interference, hold great potential for meeting the escalating requirements of modern wireless communication systems. Furthermore, within those frequency bands, there persists a limitation in generating significant THz sources. Terahertz quantum cascade lasers (THz-QCLs) are a compelling avenue, but their use remains contingent on cooling operations, as visualized in Figure 5.2. Undoubtedly, numerous challenges must be surmounted.

The first one is thermal management of the QCL device. At room temperature, the thermal energy is significant compared to the low-energy THz photons. This can lead to a significant level of thermal population in the upper laser states, causing rapid non-radiative recombination and limiting the laser’s efficiency and performance. Moreover, THz QCLs often require cryogenic cooling to achieve efficient operation due to the higher energy levels involved. However, cryogenic cooling adds complexity, cost, and limitations to practical

5.1. Current RF system and needs for higher frequency
applications.\textsuperscript{3}

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{figure5_2.png}
\caption{a) Summary of THz-QCL results, revealing the hurdles associated with the development of room-temperature THz-QCL. Courtesy of Vitiello\textsuperscript{4} b) State-of-the-art result of a THz-QCL. It shows the rapid decrease of the output power while increasing the temperature. Courtesy of Khalatpour\textsuperscript{5}.}
\end{figure}

\section{5.2 Radio-Frequency antenna}

An antenna is a device that converts a guided electromagnetic wave into a radiated electromagnetic wave, and vice versa. It is physically connected to a waveguiding device, which can be a coaxial cable, a printed line (such as a coplanar line or microstrip line), or a waveguide. In our experiment, we utilized horn antennas to generate our RF signal at 26 GHz. These antennas are metallic structures that enable a gradual transition from a waveguide to free space. The use of horn antennas offers several advantages, including high gain and a wide bandwidth. To estimate the received power at the receiving end, the Friis formula can be applied.\textsuperscript{1}

\begin{equation}
P_r = P_e G_e G_r \cdot \left(\frac{\lambda}{4\pi d}\right)^2.
\end{equation}

The received power $P_r$ in a wireless communication system is determined by factors such as the gain of the emitter antenna ($G_e$), the gain of the receiver antenna ($G_r$), and the distance ($d$) between the two antennas. It is important to note that the received power decreases proportionally to the square of the distance $d^2$ due to the spreading of the electromagnetic wave. This means that for very long-distance RF links, the received power becomes significantly weaker, making it challenging to maintain reliable communication. It is very directional while benefiting from a quite broad spectral range as depicted in the
Figure 5.3. The frequency range goes from 23 GHz to 40 GHz, the upper range was limited by the bandwidth of the VNA used.

Figure 5.3: The characteristics of the 26 GHz horn antenna were measured using a Vector Network Analyzer (VNA). a) The graph illustrates the response of two antennas arranged in a back-to-back configuration. A 30 dB attenuation was observed when the antennas were positioned in this configuration, by comparing the modulated signal emitted with the received signal at the receiver. The frequency range of operation for the horn antenna was approximately 23 GHz to 40 GHz, with the upper limit determined by the capabilities of the VNA. b) The impact of the emitter antenna’s orientation on the reception quality is examined. The two antennas are separated by an approximate distance of 20 cm.

5.3 Photonic microwave generation

Photonic microwave generation is the utilization of photonics to generate signals spanning the RF to THz frequency range. By converting optical signals into microwave frequencies, it becomes possible to transmit, process, and manipulate microwave signals using light-based technologies. Over the past three decades, microwave photonics has made significant progress in the development and demonstration of various devices, technologies, and applications. Table 5.1 illustrates various techniques that can be employed for this purpose, each with its own advantages and disadvantages. In the field of semiconductor lasers, there have been successful demonstrations of InGaAsP quantum-well lasers operating at frequencies exceeding 30 GHz, specifically at a wavelength of 1.55 µm. One approach to improving the modulation bandwidth of lasers is to enhance their resonant frequency response. This can be achieved through the use of external cavity lasers or monolithic multisection lasers. Additionally, membrane lasers based on photon-photon resonance have been developed, with demonstrated bandwidths of up to 108 GHz. Another development is the use of...
Table 5.1: Comparison between different types of photonic microwave techniques. Reproduced from Qi.

<table>
<thead>
<tr>
<th>Techniques</th>
<th>Electronics</th>
<th>Tunability</th>
<th>Maximum Frequency</th>
<th>Microwave Linewidth</th>
<th>Single Sideband</th>
<th>Modulation type</th>
<th>Stability</th>
<th>Optical loss</th>
</tr>
</thead>
<tbody>
<tr>
<td>Direct modulation</td>
<td>Moderate</td>
<td>Fair</td>
<td>108 GHz</td>
<td>Determined by source</td>
<td>Special require design</td>
<td>AM</td>
<td>Good</td>
<td>None</td>
</tr>
<tr>
<td>OPLL</td>
<td>Complicated</td>
<td>Fair</td>
<td>330 GHz</td>
<td>Narrow (~mHz)</td>
<td>Yes / /</td>
<td>Poor</td>
<td>None</td>
<td>None</td>
</tr>
<tr>
<td>Dual-mode laser</td>
<td>Complicated</td>
<td>Good</td>
<td>42 GHz</td>
<td>Moderate (~MHz)</td>
<td>Yes / /</td>
<td>Good</td>
<td>None</td>
<td>None</td>
</tr>
<tr>
<td>External modulation</td>
<td>Moderate</td>
<td>Fair</td>
<td>hundreds of GHz</td>
<td>Determined by source</td>
<td>Special design required</td>
<td>AM/PM</td>
<td>Poor</td>
<td>Large</td>
</tr>
<tr>
<td>Mode locking</td>
<td>Moderate</td>
<td>Poor</td>
<td>30 GHz</td>
<td>Narrow (~Hz-mHz)</td>
<td>No / /</td>
<td>Good</td>
<td>Large</td>
<td>None</td>
</tr>
<tr>
<td>OEO</td>
<td>Complicated</td>
<td>Poor</td>
<td>75 GHz</td>
<td>Narrow (~mHz)</td>
<td>No / AM/PM</td>
<td>Good</td>
<td>Large</td>
<td>None</td>
</tr>
<tr>
<td>Period-one dynamics</td>
<td>Complicated</td>
<td>Poor</td>
<td>100 GHz</td>
<td>Moderate (~MHz-mHz)</td>
<td>Yes / FM</td>
<td>Good</td>
<td>None</td>
<td>None</td>
</tr>
<tr>
<td>Optical heterodyne</td>
<td>Simple</td>
<td>Good</td>
<td>~THz</td>
<td>Broad</td>
<td>Yes / AM</td>
<td>Poor</td>
<td>None</td>
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distributed Bragg reflector lasers with a coupled-cavity injection-grating design, operating at a wavelength of 1.55\(\mu\)m. These lasers have shown an intrinsic 3-dB modulation bandwidth of 37 GHz. Optical injection locking is a technique that can further enhance the resonance frequency of semiconductor lasers, and researchers have achieved a 60 GHz 3-dB bandwidth using a directly modulated optically injection-locked VCSEL. External modulators have also been extensively developed for photonic microwave generation. Lithium niobate interferometric modulators, for example, have showcased bandwidths exceeding 100 GHz and relatively low drive voltages. These advancements in semiconductor lasers and modulators contribute to the continuous improvement of photonic microwave generation techniques. Another method can generate high-frequency signals including leveraging the non-linear dynamics of semiconductor lasers (period-one dynamics). In our work, we choose to consider heterodyne technique. It offers numerous of advantages in the generation of high-frequency signals, making it a compelling choice for a wide range of applications. Its key strength lies in its exceptional tunability, enabling precise control and adjustment of frequencies according to specific requirements. This versatility makes it well-suited for various frequency ranges, including the gigahertz and terahertz domains. One of the notable benefits of heterodyne systems is their compatibility with simple electronics. This
simplicity enhances their practicality and efficiency, particularly in high-frequency applications where complex circuitry can be challenging to implement. A significant challenge in heterodyne frequency generation is achieving stable frequencies due to the dependence on the stability of the two lasers involved. However, by leveraging the advantages of the heterodyne technique, we have focused our efforts on the generation of high-frequency signals, enabling us to explore novel applications.

5.3.1 Heterodyne principle for microwave transmission

![Figure 5.4: Principle of photonics wireless system using heterodyne detection. Adapted from Pang](image)

Our system utilizes heterodyne mixing for frequency conversion, where we generate two signals with frequencies $\omega_{RF}$ and $\omega_{OPT}$ by combining two distinct signals with frequencies $\omega_s$ and $\omega_c$. A schematic representation is presented in the Figure 5.4. The resulting heterodyne frequencies can be either the sum $\omega_{OPT} = \omega_s + \omega_c$ or the difference $\omega_{RF} = \omega_s - \omega_c$. To achieve this, we illuminate a detector with two signals of different frequencies. Specifically, in our system, we mix two signals on the detector: $E_s$, which carries the modulated (possibly amplitude $I(t)$ and phase-modulated $Q(t)$) information and has a frequency of $\omega_s$, and $E_c$, the carrier signal with a frequency of $\omega_c$. In the following equations, we will focus only on the frequency carried by $\omega_{RF}$ since higher frequency like $\omega_{OPT}$ can be easily cut by filtering. The expressions for their electrical fields can be given as $^{15,16}$:

\[
\begin{align*}
E_s(t) &= \sqrt{P_s} \left( I(t) + jQ(t) \right) \exp \left( -j(\omega_s t + \phi_s(t)) \right) \hat{e}_s^s \\
E_c(t) &= \sqrt{P_c} \exp \left( -j(\omega_c t + \phi_c(t)) \right) \hat{e}_c^c.
\end{align*}
\]

The optical power of the signal laser and carrier frequency laser are denoted by $P_s$ and $P_c$. 

5.3. Photonic microwave generation
respectively. The initial phase of the signal is represented by $\phi_s$, while the initial phase of the carrier is represented by $\phi_c$. The polarization unit vectors for the signal and carrier are denoted by $\vec{e}_s$ and $\vec{e}_c$, respectively. When these two electrical field vectors interact in the heterodyne mixer, one component of the resulting signal corresponds to an electrical signal with a frequency of $\omega_{RF} = |\omega_s - \omega_c|$ and a phase $\phi_{RF}(t)$. The expression for this component of the electrical signal can be retrieved:

$$E_{RF}(t) = 2\sqrt{P_sP_c} \left( I(t) \cos(\omega_{RF}t + \phi_{RF}(t)) + Q(t) \sin(\omega_{RF}t + \phi_{RF}(t)) \right) \vec{e}_s \cdot \vec{e}_c. \quad (5.3)$$

After performing heterodyne beating on the received signal, down-conversion is necessary to accommodate the limited bandwidth of the oscilloscope. It is important to note that losses incurred during the down-conversion process are not considered here. The local oscillator signal has a frequency $\omega_{LO}$, a phase $\phi_{LO}$, and a power $P_{LO}$. The electrical signal can be expressed as $E_{LO}(t) = \sqrt{P_{LO}} \cos(\omega_{LO}t + \phi_{LO}(t))$. In the following expression, only the component at the frequency $\omega_{IF} = \omega_{RF} - \omega_{LO}$ is considered, as the component at $\omega_{RF} + \omega_{LO}$ will be filtered out using a low-pass filter:

$$E_{IF}(t) = \langle E_{RF}(t) \cdot E_{LO}(t) \rangle = \sqrt{P_sP_cP_{LO}} \left( I(t) \cos(\omega_{IF}t + \phi_{IF}(t)) + Q(t) \sin(\omega_{IF}t + \phi_{IF}(t)) \right) \vec{e}_s \cdot \vec{e}_c. \quad (5.4)$$

The use of angle brackets indicates the application of low-pass filtering to reject components at $\omega_{RF} + \omega_{LO}$. Subsequently, the IF signal is converted into the digital domain for the purpose of digital down-conversion and demodulation. The resulting signal after the digital down-conversion can be represented as:

$$E_{Rx}(t) = \langle E_{IF}(t) \exp(j\omega_{IF}t) \rangle = \frac{1}{2} \sqrt{P_sP_cP_{LO}} \left( I(t) + jQ(t) \right) \exp(-j\phi_{IF}(t)) \vec{e}_s \cdot \vec{e}_c. \quad (5.5)$$

The expressions do not take into account the system loss. The accumulated phase offset and phase noise during transmission are encompassed within the term $\phi_{IF}(t)$, which can be subsequently corrected with equalization algorithm. The highest level of RF signal power is attained when the polarization states, $\vec{e}_s$ and $\vec{e}_c$ are aligned.
5.3.2 Interest of intersubband device for heterodyne generation

In our research, we extensively investigated the potential of intersubband technology for heterodyne generation, and we found several compelling reasons why this technology stands out. One significant advantage is the absence of relaxation oscillations, as mentioned earlier in this paper. This characteristic enables the development of ultrafast devices such as lasers, detectors, and external modulators. The bandwidth capabilities of intersubband technology are estimated to reach up to hundreds of GHz, opening up exciting possibilities for terahertz generation. For instance, recent advancements in QWIPs have demonstrated bandwidths of up to 70 GHz, as shown in Figure 5.5.

![Figure 5.5](image)

**Figure 5.5**: Demonstration of a 70 GHz heterodyne 3dB-bandwidth QWIP detector at 10.3µm. Courtesy from Hakl.

Furthermore, intersubband-based devices offer other intriguing features. Emission utilizing intersubband transitions exhibits high spectral purity, making it highly desirable. QCLs can achieve kHz linewidth through optical feedback and even sub-Hz linewidth using an ultra-stable near-infrared laser to stabilize a QCL with an optical frequency comb. Additionally, detectors based on intersubband transitions are noteworthy due to their high saturation current (between 300kW cm\(^{-2}\) and MW cm\(^{-2}\)), enabling the generation of strong heterodyne signals using high-output lasers.

5.3.3 Heterodyne detection for bandwidth evaluation

Currently, heterodyne techniques are predominantly employed for the characterization of mid-infrared devices. This method enables high-resolution spectral measurements, facilitating a detailed analysis of detector performance at high speeds, although limited by the
source’s bandwidth. Heterodyne detection provides crucial data, including linewidth, frequency stability, and phase noise, which are vital for evaluating the quality and reliability of mid-infrared devices. In our study, the QWIP detector’s bandwidth, assessed by Djamaal Gacemi from ENS Paris using heterodyne beating, reached tens of GHz, as illustrated in Figure 5.6.

Figure 5.6: Optical response of the QWIP detector. The QWIP detector demonstrates optical response of 30 GHz 3 dB bandwidth using heterodyne beating between two 9 µm DFB laser.

5.4 Ka band transmission

The objective is to build a FSOoRF (Free Space Optics on Radio Frequency) system using mid-infrared technology. It consists of the direct transfer from an FSO link to an RF link. While this technique has been extensively developed in the near-infrared, its application in the mid-infrared range is of great interest. Research has demonstrated successful transmission using 1.5 µm laser and detector technologies, along with advancements in modulation techniques. Notably, different research teams have achieved impressive transmission rates of 100 Gbit/s over fiber in the RF Q band and W band, respectively. In this work, we used the combination of an external modulator and heterodyne mixing to directly convert information from a mid-infrared carrier to an RF carrier. This experiment takes advantages of the very high bandwidth and saturation current of mid-infrared device. We demonstrated the possibility to convert a mid-infrared modulated signal to an RF band located around 26 GHz which is a commonly used communication band. These works were a collaboration with Livia Del-Balzo in ENS Paris, who processed the external modulator and assisted me in setting up the experimental setup.
Two continuous-wave DFB-QCLs operating at 9 µm detuned by 26 GHz have been used. To modulate the signal of the second QCL, we employed an external modulator based on the Stark effect. The up-conversion of the modulated signal to the Ka-band was accomplished through direct heterodyning using a QWIP. This electrical signal was transmitted via a first Ka-band horn antenna through free space. Then, a second antenna intercepted the wireless signal which was down-converted to be captured by the oscilloscope, it could be further processed using a custom MATLAB algorithm.

5.4.1 Experimental setup

The experimental configuration utilized in this investigation is depicted in Figure 5.7. The setup involved two commercial DFB-QCLs. They emit both maximum 100 mW at room-temperature and at 9 µm. An external Stark-effect modulator with a 3dB-bandwidth of 5 GHz was employed to modulate one of the two lasers. The characteristics of the modulator can be observed in Figure 5.8.a), demonstrating its suitability for QCL emission around 9 µm, which is located on the linear region. Additionally, Figure 5.8.b) reveals a modulation depth of approximately 25%, ensuring the production of distinct sidebands at the output of the detector corresponding to the modulation of the heterodyne signal. To optimize the overlapping of the two beams and then the heterodyne signal, mirrors have been used. The signal was heterodyned on the detector, resulting in its up-conversion to the Ka-band, employing a nitrogen-cooled QWIP with a 25GHz 3dB-bandwidth and a responsivity of approximately 10mA/W as depicted in the Figure 5.6
Figure 5.8: Parameters of the stark effect based external modulator. a) Represents the absorption in function of the wavelength. We can see that the modulator is well adapted for the DFB QCL. b) Absorption in function of the voltage bias. We can reach a maximum modulation depth of approximately 25%. Courtesy of Livia Del Balzo from ENS Paris.

5.4.2 Heterodyne spectrum signal

By maximizing the amplitude of the modulated signal and carefully aligning the two QCL beams to overlap on the detector, we have optimized the system for direct injection into the Ka-band horn antenna, enabling free space transmission. The two horn antennas have been characterized with a Vector Network Analyzer previously presented in the manuscript. The operating frequency range of the antenna was estimated to be from 23 GHz to 40 GHz, as shown in Figure 5.3. In a back-to-back configuration, there was an observed attenuation of 30 dB between the electrical signal at the emitter and the signal received at the receiver. In Figure 5.9, we present an example of the resulting signal on the QWIP for a 24 GHz signal with a 1 GHz sine modulation. The related spectrum has been captured with a 26 GHz spectrum analyser. Despite a maximum carrier frequency amplitude of -20 dBm, the signal achieved a signal-to-noise ratio exceeding 40 dB, thanks to a 30 dB amplification. It is worth noting that the high bias modulation resulted in non-linearity, causing multiple peaks corresponding to a multiple of the modulation frequency $f$ in the spectrum.

The generated signal is then injected into the first horn antenna. Following the transmission through free space, the RF signal is received by a second antenna, capturing the wireless signal within the 23-40 GHz range. To enable measurement using a 16 GHz oscilloscope, we perform down-conversion of the signal to the 0-10 GHz range by utilizing RF frequency mixing with a local-oscillator signal around 23 GHz. Subsequently, the signal is recorded and subjected to post-processing using a MATLAB algorithm. It is worth mentioning that, for the transmission experiment, the carrier frequency was set to approximately 28 GHz. This frequency selection was a compromise, aiming to reside in the middle of the antenna’s frequency range while ensuring a frequency below 30 GHz to avoid attenuation caused by
Figure 5.9: The output signal of the QWIP exhibits a beat between a 9\(\mu\)m signal QCL and a 1 GHz modulated signal shifted by 25 GHz. Observing the modulated band centered at 1 GHz, we notice its multiple frequencies stemming from the modulator’s non-linearity. Moreover, we observe the modulated signal in the vicinity of the carrier frequency, as shown in the Figure 5.6.

5.4.3 Transmission results

To assess the feasibility of message transmission, we use a OOK PRBS message. The transmission was conducted at different rates, and the resulting down-converted message is illustrated at the top of Figure 5.10.a). The signal envelope from the black curve at the top is extracted using a Matlab Hilbert filtering function (red curve). It is a signal processing technique used to obtain the analytic representation of a real-valued signal. The analytic representation is a complex-valued signal that enables various operations, including envelope detection for amplitude demodulation. Then, the message can be extracted from the signal envelope (in red at the bottom), which is subsequently equalized \((\mu_1=10^{-4} \text{ and } n_{\text{tap}}=391)\) to enhance signal quality. The eye diagram is presented in Figure 5.10.b). We achieved a Bit Error Rate of 0.2% for a 5 Gbit/s transmission, which is deemed error-free following the application of a 7% HD-FEC algorithm.

We successfully converted a 5 Gbit/s transmission from the mid-infrared range to an RF signal at 26 GHz. To extend the RF distance link, we conducted additional experiments with the inclusion of a wood plate in the signal path, showing that the electrical modulated signal from the QWIP was powerful enough. This experiment paves the way for extensive research on the monolithic integration of heterodyne emitters and detectors in mid-infrared platforms. In particular, the development of coherent detectors will be a sig-
The QWIP output signal exhibits a heterodyne beat phenomenon at 26 GHz, resulting from the frequency difference of the 9 µm DFB-QCLs. One of the two laser beams is modulated with a 1 GHz signal, leading to sidebands around 26 GHz. Observing the modulated band centered at 1 GHz, we notice its multiple frequencies stemming from the modulator’s non-linearity. Moreover, we observe the modulated signal in the vicinity of the carrier frequency.

Significant milestone for coherent transmission, building upon recent breakthroughs in phase modulation with external modulators. The upcoming experiment aims to establish an experimental setup for transmission in the W band, utilizing heterodyne techniques between two mid-infrared lasers.


Conclusion

This PhD thesis investigates high-speed free-space optical transmission (FSO) in the mid-infrared region, exploring new sources and detectors based on devices with cascade structures. Compared with near-infrared wavelengths, mid-infrared wavelengths have the advantage of undergoing minimal attenuation in the atmosphere, and being intrinsically resistant to adverse atmospheric conditions. Recent research has developed a variety of optical sources operating at mid-infrared wavelengths, particularly in the thermal atmospheric windows of 3 to 5 µm and 8 to 14 µm. These sources include optical parametric oscillators, supercontinuum generation in fibers, and cascade technologies. Our research encompasses three key aspects: high-speed mid-IR FSO transmission, private communication via chaotic synchronization, and the integration of mid-IR Free Space Optics on Radio Frequency (FSOoRF) communication systems via heterodyne beat transfer. This part of the spectrum represents a promising technology for effective transmissions that can withstand degraded atmospheric conditions. The first two chapters serve as an introduction to free-space optics systems and mid-infrared devices.

In the Chapter 3, the research on high-speed free-space optical transmission (FSO) presented in the thesis focuses on exploring the possibilities of fast data transmission in the mid-infrared spectrum using different types of infrared sources. We worked on optimizing various system parameters such as modulation schemes, coding and filtering techniques. In addition, the implementation of signal processing techniques such as finite impulse equalization (FFE) and decision feedback (DFE) algorithms enabled to significantly improve the data rates accessible at this wavelength. Sources based on cascaded devices provide access to two transmission windows, namely 3-5 µm and 8-14 µm. For the mid-infrared, interband cascade lasers (ICLs) were considered as a relevant source for low-power free-space communications. We began by studying the intensity noise of ICLs produced at the Université de Montpellier, revealing a relaxation oscillation around GHz. This discovery confirmed the possibility of directly modulating IC lasers at frequencies of several GHz. This research was used to set up an FSO transmission system using a detector based on interband cascade transitions (ICIP) to detect the mid-infrared spectrum. By doing so, we achieve data rates of up to 12 Gbit/s and 14 Gbit/s for the OOK and PAM4 modula-
tion formats respectively, while guaranteeing an error rate of less than 0.4 % imposed by the error-correcting codes. The appeal of these devices for FSO applications lies in their low current consumption and low current threshold. Many applications require optimized power consumption, such as satellite or short-range transmissions. For transmissions in the long-infrared range (8-14 µm), we designed a transmission system using a QCL optimized for fast modulation, i.e. operating with a low threshold current and a short cavity length for such a device (0.5 mm). We achieved data rates of around 5 Gbit/s using several types of modulation and guaranteeing low error rates. To prove the effectiveness of these systems, we set up real-time transmission of video streams of up to HD1080p quality. In addition, a transmission system based on external modulators was set up in collaboration with ENS Paris. These modulators utilizing the Stark effect for electrical modulation of absorption. The principle is based on the modulation of the electric field applied to two asymmetrical quantum wells, which results in the modification of the energy levels of the electronic transitions. This modulation, in turn, leads to the absorption or non-absorption of photons. By controlling the electric field, the device can selectively induce changes in the absorption properties, allowing for precise manipulation of the transmitted or received photons based on the desired modulation scheme. An initial study using this external modulator demonstrated data rates of up to 30 Gbit/s over a distance of more than 30 meters using cooled QWIPs with a BER <0.4 %. To demonstrate the ease of implementation of such a system, we also achieved data rates well above the state of the art using a QCD, which does not need to be cooled or polarized. This was followed by a new external modulator and QCD detector based on metamaterial patch antennas. These new devices enabled to achieve data rates of up to 68 Gbit/s with an error rate of less than 0.4 %. Our results exceed by far the current state of the art in this wavelength range. In addition to the very high data rates achievable, the main advantage of these devices is their ability to generate high modulated power for long-distance transmissions. The results obtained provide valuable information for the design and implementation of high-speed FSO systems. The results show significant advances in improving the data rate and reliability of FSO systems in the mid-infrared. However, despite the relative stealth of FSO links, there is a need to secure the links for certain applications.

In the chapter 4, the focus of the thesis turns to the study and applications of chaos synchronization techniques for secure and private communications over FSO links. The recently demonstrated chaotic properties of light sources in the mid-infrared range are exploited, offering the possibility of hiding messages in chaotic signals for FSO transmission. The aim of this research is to improve data security and confidentiality, opening up a promising avenue for private communication in FSO systems. To this end, experiments...
were carried out to demonstrate the feasibility of chaotic synchronization in FSO systems. Correlation coefficients of the order of 90% were obtained between the signal emitted by the master laser and that emitted by the slave laser. The retrieval of a message by an individual depends on the complexity of the generated chaos. The more complex and unpredictable the chaotic behavior, the more challenging it becomes to decipher and extract the original message. A study was therefore carried out to assess this complexity. The results confirmed the generation of hyperchaos corresponding to three positive Lyapunov exponents, opening up the possibility of establishing a private transmission. A 2-level pseudo-random-binary-signal modulation format was used to test our system’s ability to transmit a message at different data rates. At the same time, it was possible to guarantee that a non-legitimate user would obtain a BER of around 40%, while a legitimate user would have a BER of less than 4%. To confirm the use of this system over long distances, the signal propagation distance was increased to 30 meters using a Herriott cell. The importance of the signal’s electrical modulation amplitude was highlighted, as it is necessary to strike a balance between preventing the message from being too obvious, while enabling the legitimate user to find it. This study demonstrates the feasibility of such a system for private FSO communications.

In the final chapter, we look at the possibility of integrating FSO systems into existing communications systems. The interest in creating a conversion between a high-speed FSO link and an RF link is of paramount importance. The first step in this experiment was to use the Ka band of the atmosphere between 26 GHz and 40 GHz. An FSOoRF system was set up using the heterodyne technique on a QWIP detector with a bandwidth of 26 GHz. To produce the beat, we employed distributed feedback quantum cascade lasers with a wavelength of 9 µm, each operating at a frequency separation of 26 GHz. One of the beams is modulated by an external modulator. We then modulate a heterodyne beat with a frequency corresponding to the frequency difference between the two lasers. So subsequently, we injected the signal into a horn antenna specifically designed for the desired frequency band. Additionally, we incorporated a second antenna to receive the RF signal and subsequently process it using a Matlab program.

The research findings presented in this thesis hold significant promise across multiple domains, encompassing telecommunications, satellite communication, and secure data transmission. It represents a comprehensive investigation into high-speed FSO transmission, private chaotic synchronization communication, and the integration of FSO and RF communication links.
"C'est un voyage. Au début de ce voyage, dit un poème zen, la montagne au loin a l'air
d'une montagne. Au fil du voyage, la montagne ne cesse de changer d'aspect. On ne la
reconnaît plus, c'est toute une fantasmagorie qui remplace la montagne, on ne sait plus
du tout vers quoi on s'achemine. À la fin du voyage, c'est de nouveau la montagne, mais
cà n'a rien à voir avec ce qu'on apercevait de loin il y a longtemps, quand on s'est mis en
Emmanuel Carrère
I  Mid-infrared outdoor transmission

Currently, there have been limited outdoor demonstrations comparing the performance of near-infrared and mid-infrared wavelengths in free-space optical (FSO) links. However, it is crucial to empirically showcase the advantages of mid-infrared over near-infrared in FSO communications in a real field communication. To address this, a proposed approach is to conduct a comparative study over several hundred meters, directly comparing the near-infrared and mid-infrared wavelengths in challenging atmospheric condition. Through this comparison, important insights can be gained regarding the performance and advantages of mid-infrared FSO communications in real-world scenarios and then consider mid-infrared as very viable candidate for the future of FSO communication. Adaptive Optics will be an essential component for achieving precise pointing, turbulence compensation, and wave-front shaping.

II  Monolithic integration on photonic platform

The perspectives of this work are extensive, with future research focusing on the integration of all components onto a single photonics platform, further enhancing the technology’s attractiveness for industrial applications. Integrated design offers significant advantages in terms of size, weight, and easy integration into larger systems. Furthermore, it facilitates improved coupling between components, thereby enhancing the overall system performance. Although mid-infrared integration is still at its early stages, recent studies have demonstrated promising potential for various materials to be utilized for the development of functional monolithic platforms. As research progresses, the realization of a fully functional and efficient monolithic integration platform in the mid-infrared becomes increasingly feasible.
II.I Silicon platform

Silicon is a standout material for mid-infrared integrated platforms. It is a highly versatile material widely used in integrated photonics due to its excellent optical properties. Moreover, the utilization of Si-based platforms confines the spectral range of mid-IR photonic integrated circuits to wavelengths below 7 µm, owing to silicon’s optical absorption. The heterogeneous integration process is enabled through techniques like direct wafer bonding, transfer-printing, direct growth of III-V materials on silicon, or flip-chip bonding. Although alloying silicon with germanium (SiGe) have been shown to attain reasonable transparency in the wavelength range of 8 – 14 µm. The mid-infrared silicon platform offers several advantages. First, it is cost-effective and scalable manufacturing, relying on the already well-established silicon photonics. Second, it allows for the integration of mid-infrared functionalities with existing silicon photonic components, such as waveguides, modulators, and detectors. Third, the mid-infrared silicon platform presents opportunities for integrating mid-infrared devices with complementary metal-oxide-semiconductor (CMOS) electronics. Several research groups have already successfully implemented mid-infrared devices on those platform including ICL, QCL, QCD or modulator and SiGe waveguide. One major challenge arises from the lattice mismatch between the materials used for mid-infrared devices and silicon. This lattice mismatch leads to dislocation problems, significantly degrading the performance of the devices. To overcome this issue, it requires careful engineering and optimization of the heterostructures to mitigate the impact of dislocations on the device performance.

II.II Indium Phosphide (InP) platform

The development of a mid-infrared Indium Phosphide (InP) platform is another significant area of research in the field of mid-infrared integrated photonics. InP is a highly suitable material for mid-infrared applications due to its unique optical properties, which is a direct bandgap in the mid-infrared wavelength range. In recent years, there have been notable advancements in extending the capabilities of InP-based devices into the mid-infrared range. Moreover, InP is compatible with well-established fabrication processes, allowing for efficient and scalable production. Finally, InP-based devices can be integrated with other InP-based components, such as waveguides, modulators, and detectors, enabling the realization of complex integrated systems. Recent works show the possibility to implement all required devices on InP. These materials have the advantage of achieving low background doping levels, below $10^{15}$ cm$^{-3}$, either by maintaining low impurity levels or through the use of doping compensation techniques. As a result, absorption losses in bulk structures
can be effectively minimized, staying well below $1 \text{ dB/cm}$. It can enable the development of efficient and high-performance devices such as lasers, modulators, detectors, and integrated photonic circuits operating in the mid-IR range. Further research and optimization efforts are underway to improve the performance and reduce losses in waveguide structures based on these materials. As an example, devices such as optical ring resonators with an high Q-factor of approximately 120,000 have been successfully demonstrated.

II.III Monolithic integration platform application

II.III.1 Transmission platform

To enable industrial applications, it is crucial to achieve full integration on a photonics platform. The implementation of integrated components offers several advantages, including miniaturization, improved reliability, and scalability. By consolidating all the necessary elements on a single photonic platform, better control over system parameters can be achieved, leading to enhanced efficiency. For instance, in the case of the platform shown in Figure 5.11, the size of the waveguide plays a critical role in optimizing the coupling efficiency into the external modulator. Consequently, by realizing a complete integration, significant improvements can be achieved in terms of performance and overall system functionality.

Figure 5.11: Photonic Integrated Platform on InP. The objective is to integrate a transmission platform using a QCL and a Stark effect external modulator connected with an InP waveguide.

Previous studies have demonstrated the successful use of lasers and waveguides on InP, showcasing impressive performance. The next logical step would be to integrate all these components monolithically on the same chip, allowing for a more compact and efficient system.

An interesting possibility is to leverage the ease of generating a frequency comb in QCLs.
to achieve multiwavelength modulation, as illustrated in Figure 5.12. This approach involves separately modulating the teeth of the comb, which can drastically increase the data rate of the FSO link. The stability and the optimization of teeth power will have to be optimized. By independently controlling the modulation of each wavelength, it becomes feasible to transmit multiple data streams simultaneously, significantly enhancing the overall capacity of the FSO communication system. Compared to the near-infrared spectrum, mid-infrared FSO offers a distinct advantage in terms of window transparency. It enables the exploitation of extensive transmission bands: a 2 \( \mu \text{m} \) range from 3-5 \( \mu \text{m} \) and a 6 \( \mu \text{m} \) range from 8-14 \( \mu \text{m} \), allowing for the simultaneous utilization of hundreds of wavelengths, as shown in right part of the Figure 5.12. In contrast, the near-infrared range provides a significantly narrower transmission band of approximately 0.3 \( \mu \text{m} \). The mid-infrared multiplexed FSO system accommodates an impressive 20-fold increase in available wavelengths compared to its NIR counterpart. This technique holds great promise for achieving future high-speed and efficient data transmission in FSO networks.

**Figure 5.12:** Multi-wavelength channel using QCL frequency comb associated with stark effect external modulator. OFC: Optical Frequency Comb.

**II.IV Advanced modulation scheme for transmission**

Building upon the discovery of high phase modulation capabilities, we can further utilize this advancement to develop QAM modulation techniques as described in the Figure 5.13. The approach involves splitting the beam of a single QCL into two branches using an Multi-Mode Interferometer (MMI) which is an optical device used in integrated photonics for splitting or combining light signals into multiple paths. It operates on the principle of interference to achieve efficient signal distribution or recombination. In the first branch, the beam undergoes direct amplitude modulation using an external modulator. In the second branch, the beam passes through two external modulators. The first modulator introduces
phase shifting to achieve a phase offset of 90°. The beam is then modulated in amplitude and then recombined with the first branch. This process results in the generation of a QAM signal, where we have two modulated signal with a phase shift offset of 90°. To prepare the demodulation, it is necessary to also transmit the direct output of the QCL to serves as a local oscillator. The generated QAM signal is therefore coupled and transmitted into free space for communication purposes. This approach enables efficient and high-capacity data transmission by leveraging the combined encoding capabilities of both amplitude and phase variations.

**Figure 5.13:** The objective is to implement a QAM (Quadrature Amplitude Modulation) transmission platform utilizing a Stark effect external modulator. The schematic diagram illustrates both the QAM a) transmitter and b) receiver components. For practical purpose, heterodyne coherent detection has been preferred. EAM: Electrical Amplitude Modulator. MMI: Multi Mode Interferometer. LO: Local Oscillator.

In the demodulation process, differentiation between the two signals with different phases is crucial. To achieve this, a local oscillator (LO) signal that is also transmitted through free space has to be considered. However, due to uncertain power recovery of the LO signal in free space, another QCL laser with a shifted emission wavelength. This QCL laser is locked onto the QCD detector, which detects the LO signal. By combining the LO beam with the QAM transmitted beam, a beating effect is created. This allows us to independently detect the amplitude modulation associated with the two different phases, facilitating the subsequent demodulation process. Implementing a homodyne method would require 2 pairs of detectors with balanced detection, which is more complex for initial demonstrations. The use of the QCD is particularly meaningful due to its passive nature and its room temperature operation, which simplifies its integration into the system. Its passive behavior eliminates the need for additional power sources, streamlining the overall design and integration process. Phase modulation is an alternative to consider in the context of our research. However, the currently available phase external modulators have limitations. Some
have demonstrated high bandwidth but low phase shift,\textsuperscript{21,22} while others have unknown bandwidth with a limited 60° phase shift.\textsuperscript{23}

II.V Heterodyne platform

As highlighted previously, the creation of a monolithic platform for heterodyne emitters and receivers holds significant value. The objective is to integrate the setup presented in the last chapter to enhance usability and convenience. The development of this platform opens up two powerful applications.

The first application, in the Figure 5.14, involves building a Free-Space Optical (FSO) to RF system, enabling the direct conversion of a signal transmitted over FSO into an RF signal. This conversion allows for the transmission of signals to end-users or for applications that require non-line-of-sight communication. By leveraging the integrated FSO and RF platform, this application facilitates efficient and seamless communication in situations where visibility or line-of-sight access is limited or restricted.

![Figure 5.14: Heterodyne monolithic integration. This platform can be employed for FSO-to-RF signal conversion, facilitating the direct transfer of messages sent in the MIR-FSO domain to RF W-band signals. EOM: Electrical Optical Modulator. MMI: Multi Mode Interferometer.](image)

The second application, in the Figure 5.15 (light grey), involves the utilization of a microwave photonics platform that incorporates two DFB QCLs, QWIP and an external modulator for high-speed RF microwave generation in the band W. This setup holds significant relevance in the mid-infrared region due to the intersubband nature of the devices. The unique characteristic of these devices is their ability to handle high saturation currents while offering a potential for very high bandwidth.
Figure 5.15: Heterodyne monolithic integration. This platform can be used to directly generate a high frequency W band modulated signal. AOM: Amplitude Optical Modulator. MMI: Multi Mode Interferometer.


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List of publications

Journals Articles


† These authors have contributed equally to this work and share first authorship
**National Conferences**


**International Conferences**


temperature semiconductor lasers emitting in the transparency window around 4 \( \mu m \),
Optical Fiber Communications Conference, 2021

Post deadline


Invited papers


A.1 Transfert function

In this section, we will provide a detailed calculation of the transfer function formula for the ICL. By substituting the expressions of I, N, and S from Equation 2.13 into Equation 2.9 and Equation 2.10, we can obtain the system of equation:

\[
\begin{bmatrix}
\gamma_1 + j\omega & \gamma_2 \\
-\gamma_1 & \gamma_2 + j\omega
\end{bmatrix}
\begin{bmatrix}
\delta N(\omega) \\
\delta S(\omega)
\end{bmatrix}
= \begin{bmatrix}
\eta \frac{\delta I}{\delta q} \\
0
\end{bmatrix}.
\] (A.1)

where

\[
\begin{align*}
\gamma_1 &= G_0 S + \frac{1}{\tau_p} + \frac{1}{\tau_{aw}} \\
\gamma_2 &= G_0 (N - N_{tr}) \\
\gamma_22 &= mG_0 S + m \frac{\beta}{\tau_p} \\
\gamma_21 &= \frac{1}{\tau_p} - mG_0 (N - N_{tr})
\end{align*}
\] (A.2)

Using the Cramer’s rule, we can obtain:

\[
\begin{align*}
\frac{\delta S(\omega)}{\delta I} &= \frac{\eta \frac{\delta I}{\delta q}}{(\gamma_1 + j\omega)(\gamma_2 + j\omega) + \gamma_21} \\
\frac{\delta S(0)}{\delta I} &= \frac{\eta \frac{\delta I}{\delta q}}{\gamma_1 \gamma_2 - \gamma_1 \gamma_21}
\end{align*}
\] (A.3)

This leads to the transfer function, which is calculated using the following formula: \[7\]
\[ H(\omega) = \frac{\delta S(\omega)}{\delta I} / \frac{\delta S(0)}{\delta I} = \frac{\omega^2}{\gamma_1 \gamma_2 + \gamma_2 + \gamma_1} \cdot (A.4) \]

A.2 Relative Intensity Noise (RIN)

To evaluate the relation of the RIN of the ICL, we use the two Equation 3.8 and Equation 3.9:

\[
\begin{bmatrix}
\gamma_1 + j\omega & \gamma_2 & 0 \\
-\gamma_1 & \gamma_2 + j\omega & 0 \\
-\gamma_1 & 0 & j\omega
\end{bmatrix}
\begin{bmatrix}
\delta N(\omega) \\
\delta S(\omega) \\
\delta \phi(\omega)
\end{bmatrix} =
\begin{bmatrix}
F_N(\omega) \\
F_S(\omega) \\
F_\phi(\omega)
\end{bmatrix}.
\]

(A.5)

We can deduce then the following expression of \( \delta S \) using again the Cramer rules on this matrix system:

\[ \delta S(\omega) = \frac{(\gamma_1 + j\omega)F_S(\omega) + \gamma_2 F_N(\omega)}{\omega_R^2 + j\gamma \omega - \omega^2}. \quad (A.6) \]

Next, we can express the RIN formula using the following equation:

\[ RIN(\omega) = \frac{|\delta S(\omega)|^2}{S} \]

\[ = \frac{|(\gamma_1 + j\omega)F_S(\omega) + \gamma_2 F_N(\omega)|^2}{S^2 \omega_R^4} \left| \frac{\omega_R^2}{\omega_R^2 + j\gamma \omega - \omega^2} \right|^2. \quad (A.7) \]

We can expand the numerator of the second term as follows:

\[ |(\gamma_1 + j\omega)F_S(\omega) + \gamma_2 F_N(\omega)|^2 = (\gamma_1^2 + \omega^2)F_S^2(\omega)F_N^2(\omega) + \gamma_2(\gamma_1 + j\omega)F_S(\omega)F_S^*(\omega)F_N(\omega) + \gamma_2(\gamma_1 - j\omega)F_S(\omega)F_S^*(\omega)F_N(\omega) + \gamma_2^2 F_N(\omega)F_N^*(\omega) \quad (A.8) \]
Using the formulas Equation 3.5 from the description of the Langevin noise in semiconductor laser, we have finally:

\[ RIN(\omega) = \frac{|\delta S(\omega)|^2}{S} \]

\[ = \frac{a_1 + a_2 \omega}{S^2 a^2 R} |H(\omega)|^2. \]  

(A.9)

where \( a_1 \) and \( a_2 \) are already described in the equation Equation 3.11 from the corpus of the document.
In the context of the mid-infrared wavelength range, it can be difficult to find detectors that possess the necessary bandwidth for precise demonstrations. In this appendix, we will discuss the implementation of the electrical rectification technique, as employed in our study. The fundamental idea behind this technique is to leverage the non-linear current component of the device to enhance its electrical bandwidth and enable accurate measurements. This approach is particularly relevant for intersubband devices, where the bandwidth is primarily limited by electrical factors. The current-voltage (I-V) characteristic of the device can be described for a voltage $V_0$ and a microwave electrical excitation equivalent to $V_1 \exp(i\omega t)$. This results in a microwave excitation that is dependent on the electrical bandwidth of the system.

$$\text{Re}(V_{MW}) = \text{Re}(H(\omega)V_1 \exp(i\omega t)) = V_1 |H(\omega)| \cos(\omega t + \phi(\omega)). \quad (B.1)$$

We have the non-linear (I-V) response:

$$I(V_0 + V_{MW}) \approx I(V_0) + \left(\frac{\partial I}{\partial V}\right)_{V_0} V_{MW} + \frac{1}{2} \left(\frac{\partial^2 I}{\partial V^2}\right)_{V_0} V_{MW}^2 + o(V_{MW}^2). \quad (B.2)$$

The second term, which is relatively small, can be expanded as follows:

$$\frac{1}{2} \left(\frac{\partial^2 I}{\partial V^2}\right)_{V_0} V_{MW}^2 = \frac{1}{4} \left(\frac{\partial^2 I}{\partial V^2}\right)_{V_0} V_1^2 (1 + \cos(2\omega t + 2\phi(\omega))). \quad (B.3)$$

Unfortunately, the resulting microwave signal is too weak to be directly detected. Therefore, the approach is to use a lock-in amplifier to amplify the desired signal. However, lock-in amplifiers have a limited bandwidth, typically up to several MHz, for optimal performance. To overcome this limitation, an amplitude modulation technique is employed. This involves creating a small frequency envelope with an angular frequency $\omega_{LF}$, which is modulated onto the high-speed frequency $\omega_{RF}$. The resulting modulated signal $V_{MW \times LF}$ at the $\omega_{LF}$ frequency can be expressed as:
\[ \text{Re}(V_{MW \times LF}) = V_1|H(\omega_{RF})|(1 + m \cos(\omega_{LF})) \cos(\omega t + \phi). \quad (B.4) \]

The expression for the detected non-linear component of the current-voltage (I-V) relationship at the frequency \( \omega_{LF} \) is:

\[ I_{\text{Lock-in}} \propto V_1^2|H(\omega_{RF})|^2m \cos(\omega_{LF}). \quad (B.5) \]

The current value at a frequency of \( \omega_{LF} \) is directly proportional to the magnitude of \( |\omega_{RF}| \). As a result, by sweeping the frequency of \( \omega_{RF} \) while keeping \( \omega_{LF} \) constant, we can assess the magnitude of \( |H(\omega_{RF})| \) using a lock-in amplifier. The signal-to-noise ratio (SNR) of the measurements is highest just above the threshold, where the (I-V) curve displays nonlinear behavior. However, as the current increases beyond the threshold, the SNR decreases.
Annexe 3 : Optimization of the FFE Equalizer

C.1 Stark effect external modulator

As mentioned in Section 3.4.2.1, this annex provides in-depth details on optimizing the equalizer parameters to achieve a low error rate. We present the results using both a QWIP and a QCD in both B2B and 31 meters configurations. The goal is to identify a parameter combination that meets one of the two hard-decision forward error correction (HD-FEC) conditions. Ideally, the 7% HD-FEC case is preferred, as it only incurs a 7% data rate penalty. When multiple options are available for $n_{\text{tap}}$ and $\mu$, it is beneficial to choose the combination with the lowest value of $n_{\text{tap}}$. This simplifies the implementation process, as larger values facilitate faster estimations. For instance, in Figure C.3, specific parameters significantly improve transmission quality through equalization, while in Figure C.1, the equalization result is less dependent on the parameters.

C.1.1 Transmission with QCD in B2B

![Image](image_url)

**Figure C.1:** a) Eye diagram of the QCD transmission after a free-space propagation of approximately 31 meters for an OOK format at 14 Gbit/s, exhibiting an error rate of 0.035%. b) Optimization of the FFE equalization parameters in order to achieve a BER compatible with the 7% HD-FEC case.
C.1.2 Transmission with QCD with HC

Figure C.2: a) Eye diagram of the QCD transmission after a free-space propagation of approximately 31 meters for an OOK format at 14 Gbit/s, exhibiting an error rate of 0.35 %. b) Optimization of the FFE equalization parameters in order to achieve a BER compatible with the 7% HD-FEC case.

C.1.3 Transmission with QWIP with HC

Figure C.3: a) Eye diagram of the QWIP transmission after a free-space propagation of approximately 2 meters for an OOK format at 30 Gbit/s, exhibiting an error rate of 0.3 %. b) Optimization of the FFE equalization parameters in order to achieve a BER compatible with the 7% HD-FEC case.
Annexe 4 : Private transmission

D.1 Evaluation of the complexity of the generated chaos

In this section, we provide a detailed analysis of the chaos characterization at two different data rates, namely 1 Mbit/s and 5 Mbit/s.

D.1.1 1 Mbit/s B2B transmission

![Figure D.1](image)

**Figure D.1:** a) The Fourier transform power spectrum of the chaotic temporal waveform. b) Calculated Lyapunov exponents (LEs) spectrum. The curves converge to the LE values as $\text{LE}_1 \approx 19.7 \mu s^{-1}$, $\text{LE}_2 \approx 11 \mu s^{-1}$, $\text{LE}_3 \approx 5 \mu s^{-1}$, and $\text{LE}_4 \approx -1.2 \mu s^{-1}$. c) Measured three-dimensional portrait in phase space. Blue dots are the measured raw data, while the solid red curve is the reconstructed trajectory. d) The correlation integral $C_D(r)$ is calculated. The correlation dimension $D_2$ is estimated at $\approx 5$. 
The same method has been employed as in the main manuscript. Figure D.1.a) illustrates the chaotic nature of the 1 Mbit/s back-to-back transmission, evident from the presence of a unique and distinct strange attractor. The phase portrait of the attractor displays the characteristic features typically observed in chaotic systems. Additionally, Figure D.1.b) provides further confirmation of the system’s hyperchaotic behavior through the calculation and display of three positive Lyapunov exponents. These results reinforce our findings of the system’s chaotic dynamics.

### D.1.2 5 Mbit/s B2B transmission

The same graph, which was shown just before, illustrates for a 5 Mbit/s transmission.

![Figure D.2:](image)

- **Figure D.2:** a) The Fourier transform power spectrum of the chaotic temporal waveform. b) Calculated Lyapunov exponents (LEs) spectrum. The curves converge to the LE values as $LE_1 \approx 24 \mu s^{-1}$, $LE_2 \approx 14 \mu s^{-1}$, $LE_3 \approx 7.5 \mu s^{-1}$, and $LE_4 \approx -0.7 \mu s^{-1}$. c) Measured three-dimensional portrait in phase space. Blue dots are the measured raw data, while the solid red curve is the reconstructed trajectory. d) The correlation integral $C_D(r)$ is calculated. A clear plateau on the slope of the correlation integral is observed, and marked with the correlation dimension $D_2$ is estimated at $\approx 75$. 

D.2 1 Mbit/s private transition over 30 meters using an Herriott Cell

The results for the 1 Mbit/s transmission message through the Herriott Cell can be found in Figure D.3. We used a higher message modulated voltage, which resulted in a higher Bit Error Rate (BER) for the legitimate user, meeting the requirements for an FEC decoding algorithm. However, this increased the risk of message security, as the eavesdropper would only have 19.7 % errors in decoding the message.

**Figure D.3:** 31 meters 1 Mbit/s private transmission. a) Eyediagram for 1 Mbit/s through the Herriott cell which is equivalent to 31 meters propagation: a) Master signal exhibiting an error rate of 19.7 %. b) difference signal exhibiting an error rate of 1.7 %.
Titre : Nouveaux dispositifs quantiques en cascade dans le moyen infrarouge pour les applications en optique espace libre, sécurité des données et photonique micro-ondes.

Mots clés : photonique dans le moyen infrarouge, optique en espace libre, théorie du chaos

Résumé : Cette thèse de doctorat porte sur la transmission optique en espace libre (FSO) dans la région de l'infrarouge moyen couvrant trois aspects principaux : la transmission FSO à grande vitesse, la communication privée par synchronisation chaotique, et l'intégration de systèmes de communication FSO-RF. Dans le domaine de la transmission FSO à grande vitesse, cette recherche s’intéresse à différents schémas de modulation, de filtrage et d’égalisation. Des dispositifs en cascade tels que les lasers en cascade interbande (ICL) et les lasers en cascade quantique (QCL) sont utilisés. Des débits de données allant jusqu’à 14 Gbps sont obtenus en utilisant des ICL et des photodétecteurs interbandes en cascade (ICIP) dans la fenêtre 3-5 µm. Des QCL et des modulateurs externes basés sur l’effet stark sont également mis en œuvre, permettant d’atteindre des débits de données allant jusqu’à 68 Gbps. Ces dispositifs offrent une puissance modulée élevée, ce qui les rend adaptés aux transmissions sur de longues distances. La thèse explore également l’application des techniques de synchronisation chaotique pour la communication privée sur les liaisons FSO. Les propriétés chaotiques des sources lumineuses dans l’infrarouge moyen sont exploitées, ce qui permet de dissimuler des messages dans des signaux chaotiques. La complexité du chaos généré permet une transmission privée ; les utilisateurs légitimes obtenant de faibles taux d’erreur alors que ceux des utilisateurs non légitimes sont important. En outre, la thèse étudie l’intégration des systèmes de communication FSO et RF. L’objectif est de créer une conversion entre les liaisons FSO à grande vitesse et les liaisons RF. Des techniques de battement hétérodyne sont utilisées, combinant des lasers à cascade quantique pour générer des signaux de battement. Cette approche permet la transmission de signaux FSO sur la bande Ka par l’intermédiaire d’un détecteur QWIP. La recherche démontre la faisabilité de l’intégration FSO-RF, ouvrant des possibilités de combiner les avantages des deux systèmes de communication. Dans l’ensemble, cette thèse présente des avancées en matière de transmission FSO à grande vitesse, de communication sécurisée utilisant la synchronisation chaotique, et d’intégration des systèmes FSO et RF.

Title : Novel Mid-Infrared Quantum Cascade Devices for applications in free-space optics, data security and microwave photonics

Keywords : mid-infrared photonics, free-space optics, chaos theory

Abstract : This doctoral thesis focuses on free-space optical (FSO) transmission in the mid-infrared region covering three main aspects: high-speed FSO transmission, private communication through chaotic synchronization, and integration of FSO-to-RF communication systems. In the field of high-speed FSO transmission, the research optimizes modulation schemes, equalization for high speed data transmission in the mid-infrared spectrum. Cascade devices such as interband cascade lasers (ICLs) and quantum cascade lasers (QCLs) are utilized. Data rates of up to 14 Gbps are achieved using ICL and interband cascade interband photodetectors (ICIPs). QCLs and external modulators based on electrically modulated absorption are also implemented, achieving data rates up to 68 Gbps. These devices high modulated power, making them suitable for long-distance transmissions. The thesis also explores the application of chaos synchronization techniques for private communication over FSO links. Chaotic properties of light sources in the mid-infrared are leveraged, enabling the concealment of messages within chaotic signals. The complexity of the generated chaos allows for private transmission, with legitimate users achieving low error rates while non-legitimate users experience a higher error rate. Furthermore, the thesis investigates the integration of FSO and RF communication systems. The aim is to create a conversion between high-speed FSO links and RF links. Heterodyne beating techniques are utilized, combining quantum cascade lasers to generate beat signals. This approach enables the transmission of FSO signals over the Ka band through a QWIP detector. The research demonstrates the feasibility of FSO-to-RF integration, opening possibilities for combining the advantages of both communication systems. Overall, this thesis presents advancements in high-speed FSO transmission, secure communication using chaotic synchronization, and integration of FSO and RF systems. The research findings have implications for various fields, including telecommunications, satellite communication, and secure data transmission.