‘THz Torch’ Technology: Secure Thermal Infrared Wireless Communications using Engineered Blackbody Radiation

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Declaration of Originality

I herewith certify that the research reported in this thesis is of the author’s own, conducted in the Optical and Semiconductor Device (OSD) Group, Department of Electrical and Electronic Engineering, Imperial College London from October 2010 to September 2014. Any other work mentioned in thesis has been properly referenced or acknowledged.

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To my family
Abstract

The thermal (emitted) infrared frequency bands, from 20 to 40 THz and 60 to 100 THz, are best known for applications in thermography. This underused and unregulated part of the spectral range offers opportunities for the development of secure communications. The ‘THz Torch’ concept, operating between the THz and mid-infrared ranges, was recently introduced. This technology fundamentally exploits engineered blackbody radiation, by partitioning thermally-generated spectral power into pre-defined frequency channels; the energy in each channel is then independently pulsed modulated to create a robust form of short-range secure communications in the far/mid-infrared.

In the thesis, the development of ‘THz Torch’ wireless communications systems will first be introduced. State-of-the-art THz technologies, infrared sources and detectors, as well as near-infrared and visible light communications technologies, will be reviewed in Chapter 2. Basic single-channel architecture of the ‘THz Torch’ technology will be presented in Chapter 3. Fundamental limits for the first single-channel proof-of-concept demonstrator will be discussed, and possible engineering solutions will be proposed and verified experimentally. With such improvements, to date, octave bandwidth (25 to 50 THz) single-channel wireless links have been demonstrated with >2 kbit/s data rate and >10 cm transmission distance. To further increase the overall end-to-end data rate and/or the level of security, multiplexing schemes for ‘THz Torch’ technologies are proposed in Chapter 4. Both frequency division multiplexing (FDM) and frequency-hopping spread-spectrum (FHSS) working demonstrators, operating between 10 and 100 THz spectral range, will be implemented. With such 4-channel multiplexing schemes, measured bit error rates (BERs) of $<10^{-6}$
have been achieved over a transmission distance of 2.5 cm. Moreover, the integrity of such 4-channel multiplexing system is evaluated by introducing four jamming, interception and channel crosstalk experiments. Chapter 5 gives a detailed power link budget analysis for the 4-channel multiplexing system. The design, simulation and measurement of scalable THz metal mesh filters, which have potential applications for multi-channel ‘THz Torch’ technology, will be presented in Chapter 6. The conclusions and further work are summarised in the last chapter.

It is expected that this thermodynamics-based approach represents a new paradigm in the sense that 19th century physics can be exploited with 20th century multiplexing concepts for low cost 21st century ubiquitous security and defence applications in the thermal infrared range.
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List of Symbols

Greek Symbols

\begin{itemize}
    \item $\lambda$ \hspace{1cm} wavelength \hspace{1cm} [m]
    \item $f$ \hspace{1cm} frequency \hspace{1cm} [Hz]
    \item $h$ \hspace{1cm} Planck constant \hspace{1cm} [J \cdot s]
    \item $c_0$ \hspace{1cm} speed of light in a vacuum \hspace{1cm} [m \cdot s^{-1}]
    \item $k_B$ \hspace{1cm} Boltzmann constant \hspace{1cm} [J \cdot K^{-1}]
    \item $b$ \hspace{1cm} Wien’s displacement constant \hspace{1cm} [m \cdot K]
    \item $\sigma_s$ \hspace{1cm} Stefan-Boltzmann constant \hspace{1cm} [W \cdot m^{-2} \cdot K^{-4}]
    \item $\varepsilon_0$ \hspace{1cm} free space permittivity \hspace{1cm} [F/m]
    \item $\mu_0$ \hspace{1cm} free space permeability \hspace{1cm} [N \cdot A^{-2}]
\end{itemize}

Acronyms and Abbreviations

\begin{itemize}
    \item ASK \hspace{1cm} amplitude-shift keying
    \item BER \hspace{1cm} bit error rate
    \item BIB \hspace{1cm} blocked-impurity-band
    \item BPSK \hspace{1cm} binary phase-shift keying
    \item BW \hspace{1cm} bandwidth
\end{itemize}
<table>
<thead>
<tr>
<th>Acronym</th>
<th>Description</th>
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<tbody>
<tr>
<td>CAD</td>
<td>computer aided design</td>
</tr>
<tr>
<td>CMOS</td>
<td>complementary metal-oxide-semiconductor</td>
</tr>
<tr>
<td>COTS</td>
<td>commercial-off-the-shelf</td>
</tr>
<tr>
<td>CSA</td>
<td>cross-sectional area</td>
</tr>
<tr>
<td>CST</td>
<td>computer simulation technology</td>
</tr>
<tr>
<td>CW</td>
<td>continuous wave</td>
</tr>
<tr>
<td>DC</td>
<td>direct current</td>
</tr>
<tr>
<td>DFG</td>
<td>difference-frequency generation</td>
</tr>
<tr>
<td>DSB</td>
<td>double-sideband</td>
</tr>
<tr>
<td>EBG</td>
<td>electromagnetic band gap</td>
</tr>
<tr>
<td>EO</td>
<td>electro-optic</td>
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<tr>
<td>FDM</td>
<td>frequency division multiplexing</td>
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<tr>
<td>FET</td>
<td>field-effect transistor</td>
</tr>
<tr>
<td>FHSS</td>
<td>frequency-hopping spread-spectrum</td>
</tr>
<tr>
<td>FIR</td>
<td>far-infrared</td>
</tr>
<tr>
<td>FPA</td>
<td>focal plane array</td>
</tr>
<tr>
<td>FSS</td>
<td>frequency selective surface</td>
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<tr>
<td>GDD</td>
<td>glow discharge detector</td>
</tr>
<tr>
<td>HBV</td>
<td>heterostructure barrier varactor</td>
</tr>
<tr>
<td>HEB</td>
<td>hot electron bolometer</td>
</tr>
<tr>
<td>HFSS™</td>
<td>high frequency structural simulator</td>
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<tr>
<td>IMPATT</td>
<td>impact ionisation transit-time</td>
</tr>
<tr>
<td>IR</td>
<td>infrared</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
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<tr>
<td>--------------</td>
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<tr>
<td>ITO</td>
<td>indium tin oxide</td>
</tr>
<tr>
<td>HBT</td>
<td>heterojunction bipolar transistor</td>
</tr>
<tr>
<td>HEMT</td>
<td>high-electron mobility transistor</td>
</tr>
<tr>
<td>LED</td>
<td>light-emitting diodes</td>
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<tr>
<td>LO</td>
<td>local oscillator</td>
</tr>
<tr>
<td>LTCC</td>
<td>low-temperature co-fired ceramic</td>
</tr>
<tr>
<td>LT-GaAs</td>
<td>low-temperature-grown gallium arsenide</td>
</tr>
<tr>
<td>MIR</td>
<td>mid-infrared</td>
</tr>
<tr>
<td>MMIC</td>
<td>monolithic microwave integrated circuit</td>
</tr>
<tr>
<td>MOMED</td>
<td>monolithic membrane-diode</td>
</tr>
<tr>
<td>NEP</td>
<td>noise-equivalent power</td>
</tr>
<tr>
<td>NIR</td>
<td>near-infrared</td>
</tr>
<tr>
<td>NLO</td>
<td>nonlinear optical</td>
</tr>
<tr>
<td>NRZ</td>
<td>non-return-to-zero</td>
</tr>
<tr>
<td>OOK</td>
<td>on-off keying</td>
</tr>
<tr>
<td>PC</td>
<td>photonic crystal</td>
</tr>
<tr>
<td>PCB</td>
<td>printed circuit board</td>
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<tr>
<td>PIR</td>
<td>pyroelectric infrared</td>
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<tr>
<td>QD</td>
<td>quantum dot</td>
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<tr>
<td>QCL</td>
<td>quantum cascade laser</td>
</tr>
<tr>
<td>RTD</td>
<td>resonant tunnelling diode</td>
</tr>
<tr>
<td>RZI</td>
<td>return-to-zero-inverted</td>
</tr>
<tr>
<td>SBD</td>
<td>Schottky barrier diode</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Full Form</td>
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<td>--------------</td>
<td>-----------</td>
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<tr>
<td>SIS</td>
<td>superconductor-insulator-superconductor</td>
</tr>
<tr>
<td>SLED</td>
<td>superlattice electronic device</td>
</tr>
<tr>
<td>SNR</td>
<td>signal-to-noise ratio</td>
</tr>
<tr>
<td>THz</td>
<td>terahertz</td>
</tr>
<tr>
<td>THz-TDS</td>
<td>terahertz time-domain spectroscopy</td>
</tr>
<tr>
<td>TUNNET</td>
<td>tunnel-injection transit-time</td>
</tr>
<tr>
<td>UTC-PD</td>
<td>uni-traveling-carrier photodiode</td>
</tr>
<tr>
<td>UV</td>
<td>ultraviolet</td>
</tr>
<tr>
<td>VEDs</td>
<td>vacuum electron devices</td>
</tr>
<tr>
<td>2D</td>
<td>two-dimensional</td>
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<td>3D</td>
<td>three-dimensional</td>
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1 Introduction

The terahertz (THz) spectral region is receiving increasing interest, due to its unique properties. For example, THz radiation is generally considered safe for humans because it is: (1) non-ionizing, due to the low photon energy; (2) sources of power are normally limited to micro/milliwatts; and (3) the depth of penetration does not extend beyond the thickness of human skin. Furthermore, electromagnetic waves having a frequency of up to ca. 0.6 THz are able to penetrate most non-metallic materials (e.g., dry clothing and plastics), making it useful for stand-off imaging systems. Additionally, THz “fingerprints” of materials, in the frequency-domain and time-domain, enable spectroscopy to be performed on material samples of interest. THz radiation also allows non-invasive and non-destructive testing and has already found its way into a number of civilian applications (e.g., packaging failure analysis, production quality control, food spoilage detection, historical art forensics, pollution monitoring, Tbit/s communications for backbone telecommunications networks and ‘instant’ high-definition video-on-demand optical links, etc.).

However, THz systems are normally notoriously large and very expensive, from complete systems down to individual front-end active devices and passive components. This is a major reason why there are currently no ubiquitous applications in the terahertz frequency spectrum (e.g., 0.3-10 THz). However, by moving into the high-THz part of the frequency spectrum (e.g., thermal infrared region, from 10 to 100 THz), for specific niche applications, it may be possible to create affordable systems for commercial exploitation. Secure wireless communications systems using pulsed-banded-noise sources are considered to have such potential.
1.1 Background

Wireless ship-to-ship communications, via optical signalling, was implemented by the British Royal Navy in 1867. This represents an early example of wireless communications using pulsed-banded sources, whereby data are transmitted intentionally within the visible spectral range. The signalling device, a lantern, was called the Aldis lamp (named after British inventor Arthur C. W. Aldis). It is also interesting to note that shortly after, in 1898, an English inventor, David Misell, filed Patent No. 617,592 for an electric device [1]. One year later, American Electrical Novelty and Manufacturing Co. manufactured the flashlight. Today, the commonwealth navies and NATO forces still use signal lamps when secure radio communications need to be silent or electronic “spoofing” is likely. Signalling at night is usually conducted in the near-infrared part of the spectrum, using filtered incandescent light bulbs and requiring night vision sights, minimising the risks of being intercepted or jammed.

Nowadays, coherent signal carrier sources are generally preferred for implementing modern wireless communications systems, as these enable higher levels of modulation and with improved signal-to-noise ratio (SNR) [2–6]. For example, cost-effective visible and near-infrared (NIR) laser diodes can be found in many commercial applications. On the other hand, incoherent signal carrier sources employing light-emitting diodes (LEDs) have also been widely used in many ubiquitous applications (e.g., NIR hand-held remote controls), for their low cost and high spectral efficiency. Unlike lasers, LEDs emit incoherent radiation and this represents an example of modern pulsed-banded-noise communications. Today, low cost LEDs and on-off keying (OOK) modulation have been established in many near-infrared [7] and visible [8,9] communications.

Banded-noise sources have also been utilized in both communications [10] and radar [11,12] systems in the microwave range, where the banded-noise carrier generators that are modulated by baseband signals. With noise radio, non-return-to-zero (NRZ) binary phase-shift keying (BPSK) for delayed-division multiple access communications was demonstrated at 2.4 GHz. Here, white noise generation spectrally appears as thermal noise to other radio systems – having a higher level of security,
through greater immunity to detection, interception and interference (both natural and man-made) – and is very low cost at microwave frequencies. Noise radar offers similar advantages to noise radio. Using wide-band noise in the microwave band, from 1-2 GHz, one can achieve high resolution and reduced ambiguities in range and Doppler estimations for applications including covert military surveillance and reconnaissance [11].

In contrast, little has been reported on wireless communications, using either coherent or banded-noise sources, in the thermal infrared spectral region. The reason is that, within this range, coherent laser sources are very expensive for most applications, while incoherent LED sources are not widely available due to the lack of appropriate material systems having sufficiently small bandgaps. Moreover, to achieve a minimum level of bit error rate (BER), increasing the level of transmitted power is through higher spectral radiance, within a fixed bandwidth; which may compromise its immunity to detection. The high atmospheric attenuation within this part of spectrum also prohibits its uses in wireless communications, especially for applications beyond a transmission distance of ~10 m. However, this will not be a hurdle for short-range communications. In fact, such high path losses can offer greater immunity to detection, interception and interference; in the same way unlicensed 60 GHz wireless networking (having the 802.11ad IEEE standard [13]) exploits the extremely high atmospheric oxygen (O₂) absorption band.

In addition to signalling, the infrared part of the electromagnetic spectrum has many other security and defence related applications. At one end of the cost spectrum, an amateur infrared security alarm system was reported in 1965, employing a domestic flashlight and an infrared filter made using exposed photographic film [14]. At the other end of the cost spectrum, BAE Systems incorporate their patented thermal/infrared/multi-spectral adaptive camouflage technology, known as ADAPTIV, on their CV90 armoured vehicles. Sheets of relatively large hexagonal ‘pixels’ can quickly change temperature. On-board cameras allow the vehicle to effectively display the background infrared image onto the vehicle, allowing even a moving tank to match its background infrared surroundings. This technology also facili-
tates identification friend or foe (IFF) capabilities. More ominously, in the recent armed conflict in Iraq, camouflaged passive infrared sensors, wired to the arming mechanisms of improvised explosive devices (IEDs) were discovered; which are triggered by the movement of personal and vehicles.

Having higher physical layer security in mind, the 19th century ideas of optical signalling and the flashlight are reconsidered here for the 21st century applications, with the emerging ‘THz Torch’ technology. This thesis will present the ‘THz Torch’ technology; introducing the strengths and weaknesses of using uncooled incoherent thermal infrared sources and detectors; showing proof-of-concept working demonstrators, intended for specific niche applications in security and defence. Unlike approaches normally associated with traditional THz systems, which bridge the gap between electronics and photonics, ‘THz Torch’ technologies can exploit ultra-low cost thermodynamic approaches. For example, simple miniature incandescent light bulbs are used as high-THz (thermal infrared) banded power source generators; while pyroelectric infrared sensors are used as thermal detectors.

1.2 Spectral Range Considerations

In general, for commercial applications, the infrared regions (which extend from 0.3 to 400 THz) remain one of the least explored parts of the electromagnetic spectrum. However, the technological ‘THz gap’, between conventional electronics and photonics, offers the real promise of creating new scientific discoveries and commercial exploitation. Depending on the particular community of origin, several definitions exist for the ‘terahertz’ (also known as “T-Ray”) part of the frequency spectrum. This is important, as several enabling technologists and/or applications may exist within the range specified by one definition, but not necessarily within another. The most loose definitions have their two-decade spectral ranges from 0.1 to 10 THz (corresponding to free space wavelengths of 3 mm to 30 µm) [15, 16] or 0.3 to 30 THz (corresponding to free space wavelengths of 1 mm to 10 µm) [16], and these encompass the vast majority of enabling technologies and applications. For example, at the lower end of the former, various electronics-based source and detector tech-
nologies can be employed, often in conjunction with a wide selection of guided-wave structures. However, with the latter, photonics-based source and detector technologies are generally employed, often in conjunction with free space (quasi-)optics. A more pragmatic view, adopted by the IEEE Transactions on Terahertz Science and Technology journal, has its range of interest from 0.3 and 10 THz [17]. This definition excludes the 0.1 to 0.3 THz frequency range that is freely open for commercial exploitation, including ultra-high speed telecommunications and security body scanners for the screening of illegal substances and concealed weapons (e.g., capable of detecting hidden metals, plastic, liquids, gels, ceramics and narcotics concealed beneath a person’s clothing) [18]. However, a more exclusive, single-decade, definition of the THz spectral range exists from 0.3 to 3 THz; widely cited in the open literature and having the least amount of commercial exploitation. This definition, previously referred to as the submillimetre wave region before ca. 1990, excludes the 3 to 30 THz part of the frequency spectrum that has a long history of commercial applications associated with the generation and detection of thermal radiation within the far-infrared region.

There are still no ubiquitous applications on the market operating from 0.3 to 10 THz, due to the lack of affordable enabling technologies (e.g., sources, detectors or guided-wave interconnects). Notable exceptions are the relatively basic ultra-low cost sensors for motion (with applications ranging from security to energy saving lighting systems) and fire detection systems, which operate between ca. 15 to 50 THz – i.e. at the upper half of the far-infrared band (0.3 to 30 THz) and lower part of the mid-infrared band (30 to 120 THz).

While the near-infrared band (120 to 400 THz) is well above the THz region, there are some interesting applications that can possibly be adopted for longer wavelength applications. For example, two decades ago, the Infrared Data Association (IrDA) published a set of protocols for wireless near-infrared communications that provide physically secure data transfer; having very low bit error rates (e.g., $<10^{-9}$), in environments where interference would make conventional radio-based wireless technologies unusable. Operating in the NIR, between 333 and 353 THz (i.e. 850 to 900
nm), the best operational range for IrDA communications is only between 5 and 60 cm. The slowest, serial infrared (SIR) physical layer protocol has specifications that include a 9.6 to 115.2 kbit/s data rate with asynchronous return-to-zero-inverted (RZI) pulses. For low-power mode to low-power mode of operation, the maximum range is just 20 cm. When compared to radio-based wireless technologies (e.g., Bluetooth), IrDA hardware is still a great deal less expensive and does not have the same security problems, which makes it ideal for line-of-sight data transfer between mobile phone handsets. However, IrDA links operate in the slow half-duplex mode, because when the transmitter is sending NIR pulses its receiver is effectively ‘blinded’ by them; making full-duplex unfeasible.

Other NIR applications of interest are ubiquitous remote controls and bi-directional voice & data communications systems used in secure environments; neither of which have specifications that conform to any international standards. Remote controls have been used since the early 1980s and today rely on ultra-low cost LED sources operating at approximately 319 THz (i.e. ~940 nm); using 33 to 60 kHz carriers, 100% amplitude-shift keying (ASK) modulation pulses transfer signals at data rates of between 4 and 120 bit/s. With secure communications, a mobile user’s voice is digitally encoded, having a bandwidth of only 3.4 kHz, and transmitted using Class 1 NIR LEDs across an area of 100 m² (indoors) and 25 m² (in direct sunlight) [19] or across a range of 3.2 km with a point-to-point link and with higher data rates [20]; offering a low probability of interception or jamming. For all these NIR applications, little research and development (R&D) has been reported in the open literature on similar systems operating at longer wavelengths in the high-THz (thermal infrared) part of the frequency spectrum. To this end, the ‘THz Torch’ concept was conceived.

1.3 Cost Driver

In general, wireless systems find countless applications in communications and remote sensing. Their front-end hardware can perform relatively complex signal processing functions directly at microwave (ca. 1 to 30 GHz) and optical (400 to 789 THz) frequencies. When combining ubiquitous wireless applications with advances
In enabling technologies, a positive spiral can be created that continually drives down manufacturing costs while still allowing performance and functionality to grow.

In contrast, there are two inherent reasons why millimetre-wave (30 to 300 GHz) and terahertz systems have yet to find ubiquitous applications; high data rate 60 GHz wireless communications systems [21] and vehicular 76.5 GHz autonomous or adaptive cruise control radar systems [22] are still considered luxury (non-ubiquitous) products on the domestic market. The first reason for this lack of progress is the increase in power loss (or its equivalent noise temperature) with frequency, within metal-based passive components and circuits. With coherent systems operating at ever higher frequencies, this problem is exacerbated by the increasing inability to generate sufficient spectrally-clean carrier power and to maintain power gain within an amplifying stage as frequency increases. As a result, the overall end-to-end systems performance degrades as frequency increases, requiring ever more expensive solutions (e.g., more exotic transistor technologies and/or the use of larger passive components – both of which may even have to be cryogenically cooled) that only high-end users (e.g., commercial, scientific or military) can afford and/or physically accommodate. The second reason is that wavelength is inversely proportional to frequency; thus, structures defined by a specific electrical length (e.g., resonators) generally become smaller in size as frequency increases. Therefore, fabrication tolerances become ever more important, requiring more expensive manufacturing technologies to be used for shorter wavelength applications. For example, to define a simple microstrip transmission line within a hybrid microwave integrated circuit, UV lithography can be used with simple organic-based printed circuit board (PCB) or low-temperature co-fired ceramic (LTCC) technologies. For millimetre-wave integrated circuits, a more expensive deep-UV lithographic system may be needed with the use of a microfabrication facility; while for terahertz monolithic integrated circuits it may be necessary to use very expensive E-beam/X-ray lithography, with more advanced microfabrication processing techniques.

For the above reasons, in terms of consumer-based systems intended for mass production, the cost of implementing a conventional radio-based wireless system
at upper-millimetre-wave frequencies (ca. 0.1 to 0.3 THz) becomes prohibitive for ubiquitous applications. As a result, manufacturers may be reluctant to invest in the kinds of R&D needed to make breakthroughs for the various enabling technologies that can bring prices down to affordable levels. This impasse is even more acute with operation at THz frequencies. The relatively recent emergence of THz technologies had been predominantly through the use of free space (quasi-)optical techniques, in preference to those found at longer wavelengths.

Historically, apart from purely scientific experiments, there was relatively little interest in THz frequencies. The main reasons for the fall-off in activity with reducing wavelength was the lack of demand for commercial exploitation, limited scope for realizing affordable front-end subsystems, high manufacturing costs and the lack of commercial equipment for undertaking accurate metrology. Today, the upper-millimetre-wave and terahertz parts of the frequency spectrum are slowly opening up to commercial exploitation. This is because of the introduction of new concepts in the development of low-cost front-end hardware and their associated manufacturing techniques, and the emerging market in commercial turnkey measurement systems operating up to 1.1 THz when implemented in the frequency-domain [23] and 4 THz using time-domain techniques [24].

Today, there is a sustained drive within both the scientific and engineering communities to promote the development of THz enabling technologies and their applications. With the former, new devices for THz generation and detection are being investigated, with ever increasing performances [25–30]. With the latter, there is real motivation for finding applications to commercially exploit the ‘THz gap’. When ubiquitous THz applications finally emerge, the costs associated with active devices and passive components (including their manufacture and metrology) will fall, creating a positive spiral of growth in all areas. To achieve and maintain this advancement, new engineering solutions are needed. Examples of such solutions, associated with: (1) material characterisation [31,32]; (2) analytical modelling of passive structures [33,34]; (3) stress testing of commercial numerical CAD software [35,36]; and (4) design of passive components and circuits [37,38], can be found online. More re-
cently, over the past three years, the ‘THz Torch’ technology has emerged [39–50], in an attempt to dramatically cut the cost for implementing specific niche applications in security and defence.

1.4 Outline of Thesis

The work presented in this thesis describes development of the ‘THz Torch’ concept and will be organized as follows:

- Chapter 1 has introduced the background, showed some wireless communications examples using pulsed-banded sources, the spectral range considerations and the cost driver for developing the thermal infrared ‘THz Torch’ technology using engineered blackbody radiation.

- Chapter 2 will provide a literature review of THz technologies and infrared sources and detectors. Furthermore, visible light and NIR wireless communications systems using LEDs will be reviewed, to compare the ‘THz Torch’ concept with these similar competing technologies.

- Chapter 3 will introduce the basic concept of the single-channel ‘THz Torch’ wireless communications system, and show the first proof-of-concept demonstrator. Fundamental limitations of this technology are also analysed. Engineering solutions are further proposed and verified experimentally.

- Chapter 4 discusses two multiplexing schemes for the ‘THz Torch’ concept. 4-channel frequency division multiplexing (FDM) and frequency-hopping spread-spectrum (FHSS) working demonstrators, within the 10-100 THz spectral range, will be demonstrated.

- Chapter 5 will give a detailed power link budget and noise performance analysis for the 4-channel multiplexing systems, in order to have a useful insight into the practical operation at component and systems levels, and to provide a valuable tool to investigate ways of improving the overall performances of both single and multi-channel systems.
• Chapter 6 shows the design, fabrication and preliminary measured results of scalable THz metal mesh filters based on 525 µm thick fused silica substrates. The effect of parameters in metal mesh filter design will be investigated. This technology is expected to be employed in the future multiplexing ‘THz Torch’ systems.

• Chapter 7 summarises the work described in this thesis, addresses the author’s original contributions and provides suggestions for future research.
2 Literature Review

In this chapter, a literature review of current THz and infrared (IR) technologies, and wireless communications using pulsed-banded sources will be presented. This review will be divided into four sections:

1. State-of-the-art THz technologies. In this section, THz sources and detectors, active devices and passive components, as well as THz systems and applications, will be reviewed. This aims to demonstrate the capability of current THz sources, detectors and other devices, showing the reasons why there are currently no ubiquitous applications in this part of frequency spectrum;

2. Infrared sources and detectors. In this section, thermal sources, LEDs and semiconductor-based quantum cascade lasers (QCLs), operating in the IR range, will be discussed. The performance of different types of IR detectors will also be investigated. This section tries to explain why banded thermal sources and pyroelectric sensors are suitable for the ‘THz Torch’ technology;

3. Visible and NIR LED-based wireless communications systems. Here, the wireless communications systems using LEDs will be introduced; This section tries to show the performance of such wireless systems which are similar to the ‘THz Torch’ concept, but working at much higher frequencies.

4. Conclusions. This section will summarise this chapter, and explain why the ‘THz Torch’ needs to be investigated.
2.1 State-of-the-art Terahertz Technology

Terahertz normally refers to electromagnetic waves at frequencies from 0.3 to 10 THz, as shown in Figure 2.1 [51]. Within this part of spectrum, wavelengths are of the same order of magnitude as molecular-level structures. This results in a unique set of behaviours that bring difficulties for THz generation and detection, and also, creating opportunities for a variety of valuable new applications in different fields.

![The electromagnetic spectrum indicating the THz range](image)

Figure 2.1: The electromagnetic spectrum indicating the THz range [51].

Despite attractive properties, different terahertz technologies are facing similar challenges. One significant issue is their uniformly low power conversion efficiency, which is typically much less than 1%. In order to obtain sufficient output power, more input power should be pumped into the sources. Low efficiency combined with the small size of the devices leads to another problem: extremely high power and current densities [52]. Therefore, cryogenic cooling systems are required to extend the performance of the devices, greatly limiting the use of THz technologies in everyday applications.

Moreover, with wavelengths in the same range as molecular structures, the THz waves can experience extreme atmospheric absorption, due primarily to water vapour and oxygen, as shown in Figure 2.2 [6, 53]. Therefore, considering the low output power (µW-mW) from most of existing THz sources, long-distance wireless communications or remote sensing using THz signals will rarely be practical. Fortunately, there still exist some transmission windows with reasonable bandwidths, as shown in Figure 2.3 [54,55], enabling potential applications for high-speed and short-range wireless data transfer.
2.1.1 Terahertz sources

As one of the most important devices in a THz system, terahertz sources have been intensively investigated. According to different physical mechanisms, technologies involved for THz generation fall into four general categories: electronics-based, thermal-based, photonics-based and using gas lasers or QCLs. Figure 2.4 gives an overview of the output power, as a function of frequency, for different types of THz and infrared sources [56, 57]. As seen from this figure, THz oscillators, amplifiers and multipliers can produce more than 1 W output power at 10 GHz. However, it will significantly decrease as the frequency increases. Uni-traveling-carrier photodiode (UTC-PD) photomixers, which are based on the photomixing of two lasers
with their frequency difference located in the THz range, faces a similar challenge: the output power is $>10$ mW at 0.1 THz and decreases to $<0.01$ mW at 1 THz. Higher output power can be obtained using difference-frequency generation (DFG) technique in nonlinear optical (NLO) materials. However, the complete system is not compact, as large laser systems have to be used to provide the optical pump. THz-QCLs can operate over a wide spectral range at the high-THz end with $>1$ mW output power, but unfortunately, up to this time they cannot work at room temperature. Therefore, the search for the ‘perfect’ THz source still continues.

Figure 2.4: Output power as a function of frequency for different types of THz and infrared sources [56,57].

**Electronics-based THz sources**

The electronics-based method utilizes solid-state electronic devices that have already been well established in the microwave and sub-mm wave regions. This type of THz source has advantages such as compactness, high-level integration and room temperature operation. In principle, it is possible to scale these readily available technologies for THz applications. However, at higher frequencies the power generated by solid-state electronic devices, such as transistors, Gunn oscillators and Schottky diode multipliers rolls off, owing to both the transit times and reactive
parasitic effects [58]. Therefore, faster operating speeds for these devices has to be achieved to attain sufficient output power.

Solid-state electronic fundamental and harmonic oscillators have been steadily improved by increasing the upper frequency limits and power efficiencies, and now these devices are able to be operated in the lower band of the THz gap. Typical output power is >100 µW in the 0.3-0.4 THz range and ~10 µW above 0.5 THz. Two-terminal devices such as InAs/AlSb and InGaAs/AlAs resonant tunnelling diodes (RTDs) [59–62], GaAs tunnel-injection transit-time (TUNNET) devices [63–65], Si, GaAs and GaN impact ionisation transit-time (IMPATT) diodes [66–70], InP Gunn diodes [63,71–73] and GaAs superlattice electronic devices (SLEDs) [74–76] are being developed towards low THz range. Transistor-based oscillators, e.g., InP and SiGe heterojunction bipolar transistors (HBTs) [77–82], InP-based high-electron mobility transistors (HEMTs) [83–86] have also been investigated. More recently, Si-based CMOS oscillators were demonstrated [87–93]. The signal sources are able to generate sub-mW power in the 0.2 to 0.5 THz range using standard 65 nm CMOS technology [93]. With further improvements on the fabrication techniques, e.g., using 45 nm or 32 nm CMOS manufacturing process, the operation speed of these devices is expected to be further increased. This will enable CMOS-based THz sources as lower cost alternatives for the realization of compact THz systems. Output power for selected state-of-the-art room temperature THz sources are shown in Figure 2.5 [94].

**Thermal-based THz sources**

The idea of using thermal methods to generate THz signal is straight-forward: THz radiation is naturally emitted by all objects above 0 K. According to Planck’s Law, the spectral radiance generated by a blackbody can be expressed as

\[
I(\lambda, T) = \frac{2\hbar c^2}{\lambda^5} \cdot \frac{1}{e^{\hbar c/\lambda k_B T} - 1} \quad [W/m^2/sr/\mu m]
\]

(2.1)

having a spectral peak at

\[
\lambda_{peak} = \frac{b}{T} \quad [m]
\]

(2.2)
where $I(\lambda, T)$ is the power radiated per unit area of emitting surface per unit solid angle per unit wavelength at absolute temperature $T$; $\lambda$ is the free space wavelength; $h$ is the Planck constant; $c$ is the speed of light in a vacuum; $k_B$ is the Boltzmann constant; and $b$ is Wien’s displacement constant.

As seen in (2.1) and (2.2), thermal-based THz sources are broadband, inherently low cost and have a good tunability. However, at room temperature, it is extremely challenging to use thermal-based sources in the THz regime, simply due to the very low output power level and high background noise within this spectral range. Therefore, cryogenic cooling and/or lock-in technique have to be used, in order to increase the SNR of the systems based on thermal sources.

Furthermore, since the output power from blackbody radiation is incoherent and unpolarized, only the intensity can be detected by using a broadband thermal detector. Other data such as frequency-domain or time-domain information cannot be provided by this type of source. However, for many imaging applications, plots of transmitted energy are sufficient. In exchange for the loss of such information, thermal-based sources can be applied to build compact, simple and ultra-low cost imaging systems.
Thermal sources have been employed as terahertz sources for many years in Fourier transform spectroscopy, but they can only generate about 1 nW power per wave number \[95\]. Nowadays, thermal-based THz sources are serving as measuring and calibrating THz detectors and other components such as filters \[96–99\]. Due to advances in microfabrication techniques, thermally powered terahertz radiation source using photonic crystals (PCs) were proposed \[100, 101\]. In the designs, different types of electromagnetic band gap (EBG) crystals were used to modify the thermal emission peak associated with the standard Planck blackbody spectral distribution, so that the output power in THz region can be dramatically enhanced. An interesting piece of research reveals that the simple act of unpeeling a roll of ‘off-the-shelf’ adhesive tape can generate terahertz radiation, due to triboelectric charging \[102\]. A broadband signal with its peak at approximately 2 THz was produced. The radiation was unpolarized and its power (\(<1 \mu W\)) rose with faster unwinding speeds.

**Photonics-based THz sources**

Here, photonics-based THz sources refer to those that generate THz signals by the optical-to-THz down conversion process. Direct THz wave generation using gas lasers or THz-QCLs will be discussed in the next section. With photonics-based methods, interaction media, such as photoconductors, photodiodes and nonlinear optical materials, are essential. Figure 2.6 gives a conceptual illustration of the optical-to-THz down conversion process for the generation of THz waves \[103\].

![Figure 2.6: Conceptual illustration of the optical-to-THz down conversion process for the generation of THz waves \[103\].](image-url)
Photoconductors can be used to generate either pulsed or continuous wave (CW) THz signals. With the former, an ultrashort laser pulse is applied to a DC-biased photoconductive switch. The generated photocarriers are then accelerated by the applied bias, producing a broadband pulsed THz beam via a photoconductive antenna [104–110]. Most commonly used materials for photoconductors are low-temperature-grown GaAs (LT-GaAs) for 700 to 900 nm lasers and LT-InGaAs for 1300 to 1600 nm ones. Pulsed photoconductive sources are capable of providing moderate THz powers in excess of 40 µW [108] and bandwidths as high as 20 THz [109]. This kind of THz sources have been commercialized [111,112] and employed in many THz time-domain measurement systems.

Without using bulky and expensive femtosecond lasers, photoconductors [27,113–115] and photodiodes [27,116,117] can be used to generate CW THz waves by means of photomixing. Two inexpensive, compact and tunable diode lasers with its difference frequency in the terahertz region are used, as shown in Figure 2.7 [29]. With this approach, commercial off-the-shelf (COTS) and cost-effective telecommunication-based components (e.g., laser diodes, fibres) can be employed, offering a reduction in size as well as cost for the complete THz system.

Among 1550-nm photodiodes, uni-traveling-carrier photodiodes (UTC-PD) and their modifications have exhibited the highest output power [103,118]. At frequencies below 1 THz, the optical-to-THz conversion efficiency of UTC-PDs is one order of magnitude higher than that of LT-GaAs photomixers [116]. Figure 2.8 illustrates the output power, as a function of frequency, for different types of photomixers [29]. Reported output power for UTC-PDs include: >1 mW at 0.3 THz [119], ~150 µW at
0.45 THz [120] and >10 µW at 1 THz [121]. This decrease is mainly due to intrinsic parasitic impedance of the device and the impedance mismatch between the device and the radiating antenna [122]. UTC-PD photomixer modules are now commercially available from several companies [123,124], while photoconductive photomixers have already been integrated into commercial THz systems [125].

Figure 2.8: Output power from different types of photomixers in the 0.1 to 1 THz frequency range [29].

THz generation schemes using NLO materials have also attracted a great deal of attention, due to high intensity and/or conversion efficiency [103]. Two approaches are commonly used. One is based on the optical rectification process, where a high-intensity ultrashort laser pulse is directly applied onto an electro-optic (EO) material (e.g., LiNbO$_3$, GaAs, GaSe, ZnTe and DAST) to produce a terahertz pulse [126–130]. This technique usually provides lower output powers than photoconductive antennas, but it has the advantage of providing very high bandwidths up to 50 THz [131]. The other is to use optical parametric process or difference frequency generation (DFG) in NLO materials, to generate “monochromatic” continuous or quasi-continuous THz waves [132–140]. In the optical parametric process, a laser pump beam is used to generate a second idler beam in a nonlinear crystal. Then the pump and idler signal beat to emit THz radiation, where the frequency of the THz wave is the difference of the pump and the idler signal. If the idler signal is given from the outside along with the pump beam, this process becomes DFG process. Single-frequency generation
based on DFG or optical parametric process can provide THz signals which are continuously tunable over a wide range and has higher signal-to-noise ratio and spectral resolution, making these techniques attractive for spectroscopy applications.

One of the main drawbacks for photonics-based THz systems is that they normally have discrete units, making the complete system rather complex and bulky. The realization of on-chip photonic devices is therefore highly desirable, as it could ultimately allow the production of convenient, compact and potentially low cost THz equipment [141]. Unfortunately, photonic devices are constrained in the dimensions by the fundamental laws of diffraction and tend to be at least one or two orders of magnitude larger than their nano-scale electronic counterparts [142]. This size limitation significantly affects the performance of many photonic devices and hampers large-scale integration of photonic devices and also the interfacing of these technologies to create new chip-scale devices [143]. Recent progress on plasmonics has the potential to overcome these limitations.

**THz semiconductor and gas lasers**

Semiconductor laser diodes have been widely used in visible and NIR ranges as compact and cost-effective sources. However, similar approaches to direct terahertz wave generation are limited, due to the lack of appropriate materials with sufficiently small bandgaps [57]. Before the invention of quantum cascade lasers, semiconductor THz lasers have relied on more exotic gain mechanisms, such as impurity-state-transition lasers [144], or the p-Ge hot-hole lasers [28,145,146]. QCLs [147] can emit waves by means of electron relaxation between subbands of quantum wells, allowing in principle the production of light at arbitrarily long wavelengths. The first THz-QCL at 4.4 THz and operating temperatures up to 40 K was demonstrated in 2002 [148]. One of the major challenges for THz-QCLs design is that higher operating temperature will cause the degradation of population inversion (and thus gain). However, with new materials, novel injection schemes and improved design strategies, significant progress has been made in this area. THz-QCLs now have covered a wide spectral range (0.84 to 5 THz) [149–152] and can operate at temperatures up to 200
K [153–157]. The output powers have also been increased; recently a pulsed-mode THz-QCL working at 10 K has demonstrated >1 W measured output power [158]. Figure 2.9 shows the values for some of the reported output power and operating temperature for different THz-QCL designs and operation modes [28].

Figure 2.9: Performance of some reported THz-QCLs: (a) output power; (b) operating temperature [28].

The far-infrared (FIR) gas laser is another common THz source. A carbon dioxide (CO$_2$) laser is applied to pump this low-pressure gas cavity, which lases at the gas molecule’s emission-line frequencies. Methanol is the most widely used gas, giving a powerful (∼1 W) emission line at 119 µm [159]. However, gas lasers are only line-tunable in the range of 0.25 to 7.5 THz, and require long cavities and kilowatt power supplies. This type of source is preferred for specific applications where continuous tunability is not required, such as heterodyne spectroscopy and plasma diagnostics [160]. THz gas laser are now commercially available from several companies [161, 162]. Note that there are also other types of THz sources such as vacuum electron devices (VEDs). However, this is beyond the scope this thesis. An excellent review of VEDs in THz region can be found in [163].

2.1.2 Terahertz detectors

THz detection methods are intimately linked to the THz generation methods. For example, if THz signals are generated by optical-to-THz down conversion schemes, the same principle can be used for THz detection [164]. However, it is not necessarily
the case. THz signal detection can be roughly divided into two categories: direct and heterodyne detection systems, as shown in Figure 2.10 [103]. For the former, electronics-based, thermal-based or photonics-based THz detectors can be used. For the latter, an electronic or photonic mixer can be employed, and driven by electrical or optical local oscillator (LO) signals, to improve the sensitivity of the complete detection system.

![Figure 2.10: Configurations of THz detection system: (a) direct detection; heterodyne detection with (b) electronic mixer and electrical LO; (c) electronic mixer and optical LO; (d) photonic mixer and optical LO [103].](image)

**Electronics-based THz detectors**

From the electronics side, Schottky barrier diodes (SBDs), as one of the basic components in THz technologies, are widely used as direct detectors, mixers, as well as frequency multipliers. One of the advantages is that the cut-off frequency of SBDs can exceed 10 THz with GaAs materials [165] and 1.5 THz with 130 nm CMOS technology [166], enabling direct signal detection from submillimetre wave bands up to 7 THz [167–172]. Zero-bias Schottky diodes are now commercially available, covering the frequency band from 50 GHz to 1.7 THz [173]. The fabrication cost of SBDs can be further reduced by being integrated into a CMOS process [166, 174]. Normal noise-equivalent power (NEP) values for SBD-based direct detectors range from a few to tens of pW/√Hz.

Nonlinear properties of plasma wave excitations in nano-scale field-effect tran-
sistor (FET) channels enable their response at frequencies appreciably higher than the device cut-off frequency. Both resonant and non-resonant (broadband) FET plasma wave detectors have been demonstrated [26, 175–180]. Different material systems, including Si, InGaAs, GaN/AlGaN and AlGaAs/GaAs have been used for the fabrication of different types of FETs [181–186]. Among these devices, Si-based FETs are more attractive as they can be fabricated using mature CMOS technology, and thus have the potential to offer cost-effective solutions to highly integrated detector arrays. Plasma wave detectors have a typical NEP value of \( \sim 100 \text{ pW/} \sqrt{\text{Hz}} \) for room temperature operation. More recently, a room temperature InAs nanowire FET, working at 0.3 THz, was fabricated, exhibiting a NEP in the order of \( 1 \text{ nW/} \sqrt{\text{Hz}} \) [187].

Glow discharge detectors (GDDs) are based on miniature neon indicator lamps. In detector operation, these lamps are biased with a direct current to break down the gas. The incoming THz signal is amplitude-modulated, and the GDD acts as an envelope detector [188]. Its detection mechanism is derived from enhanced ionizing collisions of electrons with neutral atoms generated by the incident electric field [189]. GDDs have been employed as both direct [190, 191] and heterodyne [192] THz detectors. The first system employing GDDs as direct detectors showed NEP values of 1-10 nW/\( \sqrt{\text{Hz}} \), while the improved system had a much lower NEP of \( \sim 100 \text{ pW/} \sqrt{\text{Hz}} \). GDDs have advantages such as fast response time (\( \sim 100 \text{ ns} \)), broadband and room temperature operation. Moreover, since each neon indicator lamp costs only $0.2-$0.5, GDDs can be further integrated to form focal plane arrays (FPAs) for THz imaging systems [193], while still maintaining its low cost advantage.

**Thermal-based THz detectors**

Thermal sensors can be used to detect both infrared and terahertz signals. Thermal-based THz detectors include Golay cells, bolometers, pyroelectric sensors and thermopiles. These types of detectors normally have flat spectral responses over a broad range and can operate at room temperature, although lower operating temperature can increase the sensitivity of these detectors. Golay cells have been widely used to
detect the radiation from sub-mm to mid-infrared (MIR) bands, due to their high sensitivity [194, 195], but they have limited modulation frequency (e.g., ∼20 Hz) and extremely fragile membranes. Bolometers measure the power of incident electromagnetic radiation via the heating of a material with a temperature-dependent electrical resistance. For different types of bolometers [196–204], cryogenically cooled hot electron bolometers (HEBs) [200–202] generally exhibit the highest sensitivity and fastest modulation speed. Transition edge sensor (TES) bolometers, which exploit the strongly temperature-dependent resistance of the superconducting phase transition, have the highest sensitivity. In principle, the TES bolometers are quite similar to the HEBs. In the case of HEB, high speed is achieved by allowing the radiation power to be directly absorbed by the electrons in superconductor. In TES bolometers, however, a separate radiation absorber is used that allows the energy to flow to the superconducting TES via phonons, as ordinary bolometers do [181]. TES bolometers are capable of reaching a NEP of $10^{-19}–10^{-20}$ W/$\sqrt{\text{Hz}}$ at hundreds of mK [196, 205, 206]. Pyroelectric sensors [207–209] and thermopiles [210] are broadband, room-temperature and more cost-effective alternatives. However, they have higher NEPs (e.g., ∼1 nW/$\sqrt{\text{Hz}}$) and the modulation speed is normally limited to <1 kHz. Table 2.1 lists the parameters of some common thermal type THz detectors, which are either commercially available or from open literature.

Table 2.1: Specifications for some common thermal-type THz detectors.

<table>
<thead>
<tr>
<th>Type</th>
<th>Frequency (THz)</th>
<th>NEP (pW/$\sqrt{\text{Hz}}$)</th>
<th>Modulation Frequency (Hz)</th>
<th>Operation Temperature (K)</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>Golay Cell</td>
<td>0.02-20</td>
<td>$\sim 10^2$</td>
<td>$\sim 20$ Hz</td>
<td>300</td>
<td>[194,195]</td>
</tr>
<tr>
<td>Bolometer</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Composite</td>
<td>0.15-20</td>
<td>$&lt; 0.1$</td>
<td>$&lt; 400$ Hz</td>
<td>4.2</td>
<td>[197]</td>
</tr>
<tr>
<td>Superconducting</td>
<td>0.1-20</td>
<td>$&lt; 1$</td>
<td>$&lt; 1$ kHz</td>
<td>8</td>
<td>[199]</td>
</tr>
<tr>
<td>HEB</td>
<td>0.06-2.5</td>
<td>$&lt; 0.5$</td>
<td>$\sim 1$ MHz</td>
<td>4.2</td>
<td>[202]</td>
</tr>
<tr>
<td>TES</td>
<td>THz-X-ray</td>
<td>$10^{-7}–10^{-3}$ MHz-MHz</td>
<td>$\sim 0.3$</td>
<td></td>
<td>[181]</td>
</tr>
<tr>
<td>Microbolometer</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Bi</td>
<td>0.3-30</td>
<td>$\sim 2 \times 10^2$</td>
<td>$&lt; 1$ MHz</td>
<td>300</td>
<td>[203]</td>
</tr>
<tr>
<td>VO$_x$</td>
<td>4.3-MIR</td>
<td>$\sim 3 \times 10^2$</td>
<td>$&lt; 100$</td>
<td>300</td>
<td>[204]</td>
</tr>
<tr>
<td>Pyroelectric</td>
<td>0.1-NIR</td>
<td>$\sim 10^3$</td>
<td>$&lt; 1$ kHz</td>
<td>300</td>
<td>[208,209]</td>
</tr>
<tr>
<td>Thermopile</td>
<td>3-MIR</td>
<td>$\sim 10^5$</td>
<td>$\sim 100$</td>
<td>300</td>
<td>[210]</td>
</tr>
</tbody>
</table>
Photonics-based THz detectors

As discussed above, photoconductors, photodiodes and EO materials can also be employed for the detection of THz signals, which are generated using the same concept. Another widely used photonics-based THz detector is the extrinsic photoconductor. This semiconductor photon detector can work in a wide spectral range, from NIR down to the THz regime, depending on the chosen materials. One of the major issues for this type of detector is that the cut-off frequency of many material systems is above the THz range, thus its application in this spectrum was limited in the past.

Photoconductive detectors made from doped Si and Ge have shown a good performance at wavelengths as long as 100 µm (i.e. 3 THz) [196]. The Ge:Ga combination can provide a photoconductive response at even longer wavelengths. In unstressed configurations, the cut-off wavelength is approximately 120 µm (i.e. 2.5 THz). This can be extended to beyond 200 µm (i.e. 1.5 THz) by using stressed Ge:Ga [211]. The performance characteristics of this detector depend critically on the operation conditions. In low radiation background (e.g., satellite astronomy), detector NEP can be as low as 1 fW/√Hz, but the modulation frequency is limited to ~20 Hz. In high radiation background the sensitivity will be significantly reduced, and the detector may operate at tens of kHz [212]. Blocked-impurity-band (BIB) structures can be applied to further extend the cut off wavelength, which is caused by the formation of the impurity band. Conventional Si:As BIB detectors have a cut-off wavelength of 28 µm (i.e. 10.7 THz) with a possible extension to ~50 µm (i.e. 6 THz) [213]. Ge-based BIB detectors have demonstrated cut-off wavelengths up to ~200 µm (i.e. 1.5 THz) [214,215]. GaAs BIB detectors have even longer cut-off wavelengths due to the smaller binding energy of shallow donors in this material, and have exceeded 500 µm (i.e. 0.6 THz) without applying uniaxial stress [216].

In general, semiconductor photon THz detectors have good SNR performance (with NEP values ranging from pW/√Hz to fW/√Hz or even less) and fast response time. However, they have restricted spectral responses depending on the material systems and thus are preferred to be applied at high-THz frequencies.
Other novel THz detectors

Due to advances in new materials and fabrication technologies, different kinds of novel THz detectors are being investigated, with ever improving performances. Carbon nanotubes are considered as an excellent alternative due to their small junction areas, high electron mobilities and low estimated capacitances. Schottky diodes using semiconducting single-walled nanotubes (s-SWNTs) were presented with a NEP of 0.1 pW/√Hz [217]. A THz detector based on the hot electron bolometric effect in metallic SWNTs (m-SWNTs) was demonstrated, with a detectable range from 0.694 to 2.54 THz and a thermal time constant of 1.5 ps at 77 K [218]. Carbon nanotube quantum dot (CNT-QD) transistors have also been used to detect THz signals from 1.4 to 4.2 THz, having an estimated NEP of 10 fW/√Hz [219].

Recent progress in fabricating graphene layer structures promises a significant enhancement of the performance for future optoelectronic devices [220]. Ryzhii et al. proposed a multiple graphene layer (GL) structure with lateral p-i-n junctions, for THz detection [221]. Room temperature operation at 0.3 THz was also demonstrated by single- and bi-layer graphene FETs [30]. More recently, two detection mechanisms of graphene FETs based on plasmonic and bolometric effects working from 1.63 to 3.11 THz were analysed [222]. Other graphene-based THz detectors include a plasma resonances detector utilizing a double-GL heterostructure [223] and an uncooled graphene bolometer based on n-type and p-type GLs [224].

Single-photon detectors have also been demonstrated in the THz range. This type of THz detector exhibits the highest sensitivity and, as a result, the lowest NEP, when compared to any other state-of-the-art THz detectors [225]. A single-electron transistor consisting of a semiconductor quantum dot (QD) in high magnetic field was used for the detection of single far-infrared photons between 1.43 and 1.71 THz (i.e. 175 to 210 µm), with its NEP reaching the order of $10^{-22}$ W/√Hz [25].

THz mixers

In order to obtain higher signal-to-noise ratio (SNR), heterodyne detection systems are preferred, in which a mixer and a local oscillator are essential components.
Instead of being used as direct THz detectors, Schottky diodes can also be used as mixers. SBD mixers can operate in temperature range from 4 to 300 K, and have been employed for heterodyne detection at frequencies up to 30 THz [167–170, 226–233]. Main disadvantages of SBD mixers are the poor sensitivity and high LO power requirement. Room temperature Schottky diode mixers typically have radiometric sensitivities near 0.05 K at 0.5 THz and 0.5 K at 2.5 THz [58]. LO power required by SBD mixers is ~0.5 mW of RF drive level at frequencies close to that of the observed signal, and even more (several mW) for multiple diode configurations.

Superconductor-insulator-superconductor (SIS) [234–242] and hot electron bolometer [241–253] mixers are two other popular THz mixers, having increased sensitivity when compared with SBDs. The upper operation frequency limit for SIS mixers is determined by the gap frequency of the superconducting material, which is 0.7 THz for niobium (Nb). Materials such as NbN or NbTiN have a higher gap frequency (~1.2 THz), and were used to increase the operation frequency of SIS mixers. A HEB can also be used as a heterodyne mixer, as long as it is fast enough to allow enough intermediate frequency (IF) bandwidths. In fact, only indium antimonide (InSb) HEBs [241,242] and superconducting HEBs [243–253] are considered to be suitable as practical THz mixers. Superconducting HEB mixers do not have an upper operation frequency limit, and require extremely low LO power (10 nW-1 µW) [103]. The latter feature enables HEBs be driven by QCLs [253], photomixers [254] and solid-state multiplier chains [255]. Figure 2.11 demonstrates the double-sideband (DSB) noise temperature of Schottky diode mixers, SIS mixers and HEB mixers operated in the terahertz spectral range [242].

Photonic mixers driven by optical LO signals have increased receiver bandwidths, due to the inherently wider frequency tunability in photonic signal generation techniques. This is quite useful for spectroscopic systems. Furthermore, optical fibres can be used in such heterodyne detection systems to deliver LO signals. Typical photomixers include photoconductors, photodiodes, and EO materials, which have been discussed in previous sections.
2.1.3 Terahertz active devices and passive components

The progression of terahertz technologies also depends on the realizations of efficient active devices and passive components. In this subsection, the development of frequency multipliers, amplifiers, modulators, filters and quasi-optical passive components for THz applications will be reviewed. These demonstrate the capabilities of state-of-the-art enabling technologies, showing the possibility of using such devices or components to build room temperature, compact and cost-effective THz systems.

Frequency multipliers

A frequency multiplier consists of a nonlinear circuit to distort the input signal and consequently generates its harmonics. They have been the ‘work horse’ for many decades as the sub-mm and THz sources. Traditionally, cascaded whisker-contacted Schottky diodes multipliers driven by phase-locked Gunn oscillators were used, due to low parasitic circuit elements and ease of diode fabrication [256–258]. Over the past two decades, major advancements have been made in the development of two fabrication techniques that have now fully superseded whisker-contacted diodes [94]. One is the planar technology with the use of air bridges, pioneered by the University of Virginia [259], and the other is the monolithic membrane-diode (MOMED) tech-
nology pioneered by the Jet Propulsion Laboratory [260]. Planar GaAs Schottky diode frequency multipliers now can produce tens or even hundreds of microwatts of power at frequencies up to 2.7 THz [261–266]. Schottky barrier diode circuits have also been fully integrated in a 130 nm SiGe BiCMOS process, demonstrating the capability of silicon microfabrication technologies for THz applications [267].

Another type of diode-based frequency multiplier is the GaAs or InP-based heterostructure barrier varactor (HBV). HBVs have attracted much interest for terahertz signal generation due to their symmetric capacitance-voltage and anti-symmetric current-voltage characteristics [268–273]. These devices produce only odd harmonics of an input signal and do not require any DC bias as the capacitance modulation region is centred at zero-bias; thus they are suitable for use in high-order odd harmonic multipliers such as triplers [269–272] or quintuplers [273]. Although promising, HBV diode-based frequency multipliers have not reached high efficiencies, and have mostly worked at lower end of the terahertz band [57]. Figure 2.12 shows the output power, at room temperature, from commercially available frequency multipliers between 0.05 and 3 THz [274].

![Figure 2.12: Output power at room temperature from commercially available frequency multipliers between 0.05 and 3 THz range [274].](image)

Since the operation speed of transistors has been steadily increasing, it is possible to design transistor-based frequency multipliers with integrated power amplifiers working at terahertz frequencies. In [275], a frequency tripler using 30 nm InP
HEMT process was demonstrated, giving a peak output power of 0.6 mW at 0.3 THz. Two active frequency doublers operating at 0.22 and 0.33 THz, and a 0.32 THz ×18 multiplier chain were fabricated using an SiGe HBT technology [276]. More recently, a passive 0.48 THz frequency doubler based on standard 65 nm CMOS technology was reported, with an unsaturated output power of 0.23 mW [277].

**THz power amplifiers**

In a terahertz frequency multiplier chain, power amplifiers are normally used to amplify a W-band (75-110 GHz) signal to generate enough power to drive the first stage multipliers. GaN power amplifiers are considered to be a good option due to their high power handling capability [278]. Modules for direct power amplification in the THz frequency range have also been demonstrated. A 50-mW module at 0.22 THz [279] and a 10-mW module at 0.34 THz [280], all using InP HEMT, have been utilized. With 250 nm InP HBT technology, a 0.32 THz amplifier with a peak power of 7 mW and measured small signal gain of 4.8 dB was reported [281]. More recently, a 0.3 THz amplifier module [282] and a 0.46 THz amplifier module [283] using metamorphic HEMTs (mHEMTs) were presented. InP HEMT power amplifiers now are able to operate at up to 0.65 THz, yielding a peak power of 3 mW and measured small signal gain of ≥10 dB [284]. With further improvements on HEMT and HBT devices, THz power amplifiers are expected to have higher operation frequencies, increased output power and efficiency, and less noise. Most recent result demonstrates a 10-stage common-source MMIC using 25 nm InP HEMT technology, achieving a 9 dB gain at 1.03 THz [285].

**THz modulators**

Modulators are very useful devices to control and manipulate electromagnetic waves. Although modulators have been well established in the optical regime, there is still a great demand for highly efficient, fast and versatile active THz modulators. Over the last decades, various types of THz modulators have been investigated. THz modulators can be categorized by the physical quantity they control, such as amplitude,
phase, polarization state, spectral, spatial and temporal properties [286]. They can also be classified by the material systems being used, where semiconductors, graphene, photonic crystals, liquid crystals and metamaterials are the most common materials. In this part, we divide THz modulators into three groups, depending on the modulation approach, i.e. optically-modulated, electrically-modulated, and thermally-modulated.

When optical radiation is incident upon a semiconductor, free carriers may be produced if the photon energy is greater than the band gap energy of the semiconductor. This will consequently change the complex conductivity of the semiconductor, which can be described by the Drude model [287], given by (2.3)

$$\tilde{\sigma}(\omega) = \varepsilon_0 \frac{\omega_p^2 \tau}{1 + j \omega \tau}$$

(2.3)

where $\omega_p = \sqrt{\frac{ne^2}{\varepsilon_0 m_e}}$ is the plasma frequency; $n$ is the free carrier density, $e$ is the electron charge; $\varepsilon_0$ is the free space permittivity; $m_e$ is the effective electron mass and $\tau$ is the recombination time.

The plasma frequency is a characteristic frequency, below which semiconductor materials act as pseudo-metals and above which they are transparent to light. Silicon and GaAs have been widely used as the semiconductor. Pulse widths as short as 10 ns for GaAs [288] and 5 ns for silicon have been shown at 1.4 THz [289]. More recently, spatially modulated light was used to generate tunable optical gratings in high-resistivity silicon to modulate terahertz waves [290, 291]. In addition to being used as large-aperture modulators, semiconductors can be employed as the substrate, on which metallic metamaterials are fabricated to actively control the transmission and/or reflection properties of the THz waves [292–296].

Similar to the principle of optically modulating THz waves using semiconductors, the carrier concentration in semiconductors can be changed by electron injection or depletion of charge carriers. A room temperature electrically controlled THz modulator based on electron density modulation was presented in [297]. These devices can be further integrated with metamaterials; demonstrators include electronic control
of amplitude [298–304], phase [305, 306], frequency [307] and spatial [308] modulation. The operation speed for these modulators ranges from 100 kHz to 10 MHz, and is expected to reach the order of 10 GHz [303]. More recently, graphene is found to be superior to semiconductors when used as an electrically driven modulator, due to its extremely high carrier mobilities [309–312].

It is also possible to actively manipulate the propagation of THz waves by thermally tuning the electrical conductivity and thus the optical response of materials, creating thermally-modulated THz modulators. These materials include semiconductors or metal oxides [313, 314], special insulator materials with metallic phase transition [315, 316] and superconductors [317]. Although they can provide a good modulation depth, the modulation frequency for these thermally controlled THz modulators is comparably low, due to large thermal time constant, which is normally ∼10 ms or longer.

Note that other modulation approaches, such as magnetic [318] and nonlinear [319] modulation, can also be employed for the tuning of the optical response of metamaterials, photonic crystals [320, 321] or liquid crystals [322], providing alternative ways of creating THz modulators. A complete review of different types of THz modulators and a comparison of their performance can be found in [286].

THz filters

THz filters are one of the most essential components in THz systems, for rejecting unwanted spectral bands of the incoming signals. Metal mesh filters, a type of frequency-selective surfaces (FSS) in a two (or three)-dimensional array [323], have been widely used in terahertz and infrared ranges, due to their properties such as compact, scalable and mature fabrication process [324]. Depending on the geometry and dimensions, they can exhibit high-pass, low-pass, band-pass, or band-rejection spectral responses. These THz filters can be either free-standing or on a substrate. With the former, the filter would be very fragile if the metal layer is not thick enough, while waveguide modes will be excited within the structure if the thickness of the metal is comparable to the wavelength. With the latter, as the frequency increases,
the effect from the substrate will be more significant and Fabry-Perot modes start to appear. Therefore, choosing a suitable substrate that has low losses in the spectrum of interest is important.

Free-standing cross-shaped THz band-pass filters are normally made up of metals (e.g., gold and copper) with tens of µm thickness, having high transmittance up to \( \sim 100\% \) [325–327]. These filters have been commercially available, from millimetre-wave to \( \sim 30 \) THz, as shown in Figure 2.13 [328]. At even higher frequencies, the fabrications of metal mesh filters are challenging and optical coating filters are generally preferred. Substrate-based metal mesh filters have lower power transmittance due to inevitable losses in the substrate materials. However, they are more rigid, so that the thickness of the metal can be decreased hundreds of nm to reduce the waveguide effect. Furthermore, with supporting materials, different types of structures, which cannot be made free-standing, can be utilized. In [329], a band-pass filter embedded in benzocyclobutene (BCB) was demonstrated at 1.2 THz. Trapped-mode excitation, which was created by adding an inner cross to the original cross-shaped slots, was employed to improve transmission at the resonant frequency. To further lower the fabrication cost, band-pass metal mesh filters were fabricated on a standard 525 µm thick fused silica wafer, using standard micromachining techniques [330–333]. Trapped-mode excitation was used to suppress the Fabry-Perot resonances caused by multiple reflections within the thick substrate, in order to further improve the out-of-band rejection.

Other THz filters include plasmonic-based high-pass [334] and band-pass filters [335] using engineered subwavelength microstructures, and narrow band-pass filters using photonic crystals [336].

**Quasi-optical passive components**

Quasi-optical components such as reflectors, beam splitters and lenses have been widely used in terahertz time-domain spectroscopy (THz-TDS) and fourier transform spectroscopy systems, communications and imaging systems for guiding and shaping THz waves. Fortunately, many quasi-optical components that have been established
in the visible or infrared ranges can be employed for THz applications.

Metal-coated plano and parabolic reflectors/mirrors can be employed in THz systems due to their high reflectivity (>99% for gold) over a broad frequency range (from visible to THz) [337] and are now commercially available [338]. Instead of metal coating, THz mirrors can also be utilized using dielectric materials. THz dielectric mirrors were demonstrated using different polymer films with a typical thickness of several tens of µm to build a THz reflector [339]. Commercially available glass plates coated with indium tin oxide (ITO) were used as far-infrared dichroic mirrors in [340], showing a high reflection from 0.1 to 2.8 THz. A ceramic mirror with reflection band from 0.3 to 0.38 THz was made by tape casting and sintering stacked ceramic layers of alumina and alumina-zirconia [341]. Dielectric mirror made from polypropylene (PP) and high-resistivity silicon [342], PP with rutile titania (TiO₂) and polyvinylidene fluoride (PVDF) with TiO₂ using a compounding process [343], have also been demonstrated. These dielectric type mirrors provide the means of producing high quality, flexible, omnidirectional terahertz mirrors at very low cost.

In the THz range, many optical beam splitter schemes cannot work efficiently, due
to the lack of material with desired properties at THz frequencies and/or difficulties in scaling device dimensions. Therefore, THz beam splitter should be carefully designed for THz applications [344]. For non-polarizing beam splitters that require a minimum polarization dependency, high resistivity float zone silicon (HRFZ-Si) with thickness up to several mm is normally used as the substrate [345] and is now commercially available [346]. Low cost THz beam splitters, fabricated using ultra-thin (∼μm) low-density polyethylene (LDPE) sheeting by spray coating [347], and other conductive polymers coated by inkjet printing, have also been demonstrated [348]. Polarizing beam splitters allow a beam with certain polarization transmitted, while beams with different polarizations are largely reflected. This can be utilized by using natural birefringence effects [349], polarization selectivity of multilayer dielectric coatings [350] or subwavelength metallic gratings on polymer substrates [344].

THz lenses need to have a high broadband transmission, low material cost, negligible aberration and small deviation in the THz range. Design schemes for THz lenses are similar to those for optical frequencies, but replacing glass by silicon or plastic. Hemispherical lenses made of high resistivity silicon are widely used by both THz emitters and detectors for beam collimating and focusing, and are readily available from several manufacturers. Plastic THz lenses, which are relatively cheap, can be made from different polymer with low loss and dispersion [351]. THz lenses made from polymethylpentene (TPX), high-density polyethylene (HDPE), and polytetrafluoroethylene (PTFE, Teflon), are already COTS components. Instead of using bulk polymers, THz lenses can also be fabricated by compression molding, employing micro-powders as base materials [352]. This fabrication process potentially enables the mass production of arbitrary lens shapes. Other types of THz lenses, including a variable-focus terahertz lens with its focal length adjustable by pumping a medical white oil in and out of the lens body (polymer foil) [353], Fresnel lenses using polyethylene [354] and silicon [355], diffractive lenses using metallic slab [356] and regular paper [357], Brewster lenses made from HDPE [358] and electrically controlled liquid crystal lenses [359] have also been reported.
Instead of using bulk materials, many quasi-optical components, such as beam splitters [360] and lenses [361,362], can also be made using periodic structures. This offers new opportunities for the design of bespoke and high performance THz quasi-optical devices.

2.1.4 Terahertz systems and applications

Front-end systems operating in the THz gap have generally existed for expensive scientific applications. They normally rely on precision free space (quasi-)optics and may need cryogenic cooling. However, in order to move away from less profitable high-end applications, engineering solutions are needed to create a positive spiral of technological growth with more profitable applications. In this section, three examples of low cost THz applications will be introduced, with a view to highlight an emerging trend in the development of low cost THz engineering demonstrators [363].

THz TDR tagging

There are countless tagging technologies, operating across the whole frequency spectrum, which have been commercially exploited; RFID tags and barcodes are the most common examples. However, their signals can be easily intercepted and reverse engineered. Simple THz engineering offers a new tagging technology, based on time-domain reflectometry (TDR), as shown in Figure 2.14 [364]. This approach relies on a stack of dielectric sheets having embedded defects. The defect could be a change in thickness or a sheet with a different dielectric constant, to create a unique TDR ‘key signature’ response. The main advantage of this technology, over current technologies, is that it is much harder to reverse engineer. Also, unlike RFID tags, they contain no metal and can be easily concealed.

Handheld THz-TDS

Time domain spectroscopy has been a driving application for THz technology. However, most systems are lab based and this limits their potential applications. Recent advances enable the development of a handheld battery powered system [365]. This
Figure 2.14: Illustration of the THz TDR Tag stack with spatial defects over the corresponding frequency response [364].

The system operates from 0.1 to 2 THz with a >50 dB dynamic range and a standoff detection of ∼10 cm. Figure 2.15 shows this handheld THz spectrometer and example chemical spectra. This demonstrates that THz systems are becoming mature enough to be deployed beyond the lab, while still performing accurate measurements.

Figure 2.15: Handheld THz spectrometer and example chemical spectra [365].

Real-time CMOS THz imagers

There is considerable interest in developing low cost security body scanners, for the screening of illegal substances and concealed weapons (e.g., capable of detecting
hidden metals, plastic, liquids, gels, ceramics and narcotics concealed beneath a person’s clothing). Most work in this area has shown that systems may take minutes or even longer to produce images. However, a recently reported THz video camera, which is fully integrated in a 65 nm CMOS bulk process technology, can produce real-time images [366]. This camera contains a $32 \times 32$ pixel array and consists of 1024 differential on-chip ring antennas coupled to NMOS direct detectors operating well beyond their cut-off frequency. In its video mode, the camera operates at up to 500 frames per second for the detection of THz signals between 0.7 and 1.1 THz. Figure 2.16 shows a picture of this terahertz camera module with and without its housing and a 1 Euro coin is placed in between for size comparison. As CMOS is a mature and relatively low cost technology, this type of imager can be expected to be easily integrated into existing systems for commercial exploitation.

Figure 2.16: 1k-pixel THz video camera with and without its housing [366].

### 2.2 Infrared Sources and Detectors

In the last section, THz sources, detectors and other devices and components have been reviewed. These enabling technologies and demonstrators show the capability of current-state-of-the-art THz technology. Since the ‘THz Torch’ technology is operating between the THz and infrared spectra, it is also necessary to have an overview on the infrared technology. This section contains two parts; the first one will focus on the IR sources, including thermal sources, LEDs and diode lasers; while the second part will look at thermal and photon-based IR detectors.
2.2.1 Infrared sources

The most significant difference between THz and IR sources is that only thermal and photonics-based methods can be used to generate IR signals, simply due to the fact that electronic oscillators cannot operate at such high frequencies.

IR thermal sources

By moving to the infrared spectrum, thermal-based sources can produce much more output power and, therefore, the effects from ambient thermal noise are less significant. As a result, thermal sources have been widely used in the IR range. There are many types of IR thermal sources, since all objects above 0 K can radiate power, according to Planck’s law. Room temperature objects have spectral peaks at $\sim$10 µm (i.e. 30 THz), enabling many applications such as motion detection [367] and thermal infrared remote sensing [368] in this part of spectrum.

The total output power from a non-ideal blackbody (i.e. greybody) can be calculated using Stefan-Boltzmann law

$$P = \varepsilon \sigma_s A_s T^4 \text{[W]}$$

(2.4)

where $\varepsilon$ is the emissivity; $\sigma_s$ is the Stefan-Boltzmann constant; $A_s$ is the radiating surface area; and $T$ is the absolute temperature of the object.

In order to generate more output power at a certain temperature, the radiator should have a high emissivity and large emitting area. Unfortunately, high-melting point, shiny metals such as platinum or tungsten have emissivity values in the order of 5% at room temperature. Therefore, black layers which have high absorptance (thus emissivity) are needed to achieve high power efficiencies. On the other hand, this black layer must be thin, in order not to increase the thermal time constant of the emitter significantly, which is directly proportional to the mass of the layer, given by

$$\tau \propto \frac{d V_{body} c_p'}{h_c A_s} \text{[s]}$$

(2.5)

where $d$ [kg/m$^3$] is the density of the material; $V_{body}$ [m$^3$] is the body volume; $c_p'$
\[ \frac{J}{\text{kg} \cdot \text{K}} \] is the specific heat capacity; and \( h_c \ [\frac{\text{W}}{\text{m}^2 \cdot \text{K}}] \) is the heat transfer coefficient. In order to have a low thermal time constant, materials with less mass, larger surface area and higher heat transfer coefficient are required.

The thermal evaporation technique is used widely for the fabrication of various thin metal black coatings, including gold, silver and bismuth [369]. Highly thermal conductive diamond-like dielectric materials, such as amorphous carbon nanocomposite (a-CNC), have also been used as the absorbing layer [370]. These thermal radiators can have a high emissivity (>80%) in the 1 to 20 µm spectral range and a modulation frequency up to \( \sim 100 \) Hz.

Due to advances in microfabrication techniques, thermal emitters can be integrated with metamaterials, and are capable of engineering the desired spectral emissivity profile over a large bandwidth. It is known that in thermal equilibrium, the absorptance is equal to the emissivity. With metamaterial-based periodic microstructures, it is possible to control the transmittance, absorptance and reflectance at a certain frequency by changing the structure parameters and periodicity. This idea can be used to control the in-band emissivity of thermal sources [371–376]. In order to further enhance the radiation efficiency, the directivity of the thermal emitters can also be increased [377–379]. All these demonstrators show the trends for further improving the performances of IR thermal sources. However, the long thermal time constant is still a fundamental limitation. Therefore, pulse-modulated thermal sources are more suitable for applications where high speed is not required.

**IR LEDs**

The light-emitting diode (LED) is a two-lead semiconductor light source, which resembles a basic pn-junction diode. When activated, it will emit infrared or visible light, with its frequency determined by the band gap of the chosen semiconductor material. Visible LEDs can be found in many electronic devices as indicator lamps, in automobiles as rear-window and brake lights, and on billboards. Infrared LEDs are employed in autofocus cameras, remote controls, and also as light sources in fibre-optic telecommunications systems.
Compared to thermal sources, LEDs can provide higher radiation efficiency, faster modulation rates and have similar or lower cost. However, LEDs are widely available only in the visible (i.e. 390 nm < λ < 760 nm) and near infrared (i.e. 760 nm < λ < 2500 nm) ranges. Furthermore, the central frequencies of LEDs are not continuously tunable over the whole visible and IR spectrum, due to lack of semiconductor materials with desired bandgaps. Different material systems have been used to expand the spectral coverage of LEDs. From 1.5 to 2.5 µm, heterostructures grown on GaSb are often used. InAs-based heterostructures can be employed to operate in the 2.8 to 4.6 µm range [380]. Optically immersed 7.0 µm LEDs with a CW power of 1 µW can be found in the market [381]. One of the major issues for LEDs is that the output power decreases as the central frequency decreases, from mW-W power levels for NIR LEDs to µW at 4.6 µm at room temperature. On the other hand, the unit price is increasing, from tens of pence up to hundreds of pounds.

LEDs operating at even longer wavelengths have also been reported. In [382], superlattice InAs/GaSb LEDs with peak emission wavelength of 8.6 µm and output power exceeding 300 µW at 77 K from a 120×120 µm² mesa were demonstrated. The output power in excess of 600 µW was further reported from a 520×520 µm² mesa [383]. However, to the author’s knowledge, room temperature LEDs at such long wavelengths have not been shown in open literature. Therefore, for room temperature operation, if high modulation speed is not required, thermal sources are preferred for many of the applications in the IR range.

**IR semiconductor lasers**

Near-infrared laser diodes have been widely used in many applications, including alignment, barcode scanning, machine vision and positioning. In the far/mid-infrared range, QCLs, both at room temperature and cryogenically cooled, are developing rapidly and becoming the leading semiconductor source, enabling a variety of applications in this part of spectrum. After the invention of the first QCL working at 4.2 µm [147], QCLs have expanded the achievable wavelength range to ~3 to 25 µm and the terahertz regime, with exemplary improvements in overall per-
formance [384]. In this spectral range, InGaAs/AlInAs alloys on InP are normally used, while GaAs/AlGaAs on GaAs are preferred for THz-QCLs.

However, even with ever increasing performance, in the past, most QCLs exhibited very low (<1%) conversion efficiencies – a metric known as the wall plug efficiency (WPE). This limits the use of QCLs in applications such as portable sensors and infrared counter-measures, where the power consumption is a major constraint. The WPE is a complicated parameter depending on several factors, including the temperature of the device, quantum-mechanical structure of the energy levels, and various device characteristics such as length and waveguide loss [385]. Significant advances in the WPE were achieved by exploring the design space and re-examining common design strategies. With carefully design considerations, WPEs have been increased, from 0.15% for the first QCL at 10 K, to more than 50% at 40 K [386]. The highest demonstrated WPE for room temperature CW operation in the mid-infrared range was 21% at 4.9 µm, and had a maximum output power of 5.1 W [387]. Figure 2.17 illustrates the total WPE for selected QCLs from literatures at various heat sink temperatures for both pulsed and CW operation [384]. From this figure, it is clearly seen that the power efficiency of QCLs is steadily improving. In the future, QCLs are expected to have better performance (e.g., higher output power and power efficiency), if new material systems, novel design schemes and advanced fabrication techniques can be employed.

![Figure 2.17: Power efficiencies as a function of heat sink temperatures for selected QCLs for pulsed and CW operation [384].](image-url)
2.2.2 Infrared detectors

Figure 2.18 shows the history of the development of IR detectors, for which both thermal and photon types have been used [388]. In fact, before the invention of IR detectors, the sensing of infrared radiation has been existing in nature throughout the animal kingdom. For example, snakes (e.g., vipers, pythons and boas) possess a unique sensory system for detecting infrared radiation, enabling them to generate a ‘thermal image’ of predators or prey [389]. As discussed in Section 2.1.2, all thermal-type THz detectors are readily applicable to infrared signal detection. This is because, although working at different frequency ranges, they share the same detection mechanism – by absorbing the energy of the incoming wave. Furthermore, at high temperatures, thermal emitters can generate more power in the IR range than that in the THz range. This makes thermal detectors favourable as cost-effective solution for detecting IR signals. Thermal detectors are broadband and enable room temperature operation (although working at lower temperature will further reduce the intrinsic and extrinsic noise), but their response time is limited to milliseconds. Therefore, they are suggested to be used for applications where fast modulation speed is not necessarily needed.

![Figure 2.18: History of the development of infrared detectors [388].](image)

Instead of using thermal detectors, photon detectors can be used for IR detection with improved sensitivity and SNR. Different detection principles, such as fundamental absorption (intrinsic photodetectors), impurity absorption (extrinsic photodetectors), low dimensional solids (e.g., superlattice (SL) and quantum well (QW) detectors), etc., have been exploited [390].
Intrinsic photon detector can be made from semiconductor materials include IV-VI (PbS, PbSe, PbSnTe), II-VI (HgCdTe) and III-V (InGaAs, InAs, InSb, InAsSb). One of the advantages of the intrinsic detector is that the operating temperature is higher than for any other type of semiconductor detectors. Among intrinsic photon detectors, HgCdTe is the most important semiconductor material system for IR detection, due to its higher performance in a wide spectral range from 1 to 25 µm [391]. But they have serious technological problems in mass production, including non-uniformity over large area, high cost in growth and processing, and lattice, surface and interface instabilities.

In the past, extrinsic photoconductors were widely used at wavelengths beyond 10 µm, when the intrinsic detector technology had not reached today’s maturity. Extrinsic detectors must be operated at lower temperatures to achieve similar sensitivity as of intrinsic detectors, and a sacrifice in quantum efficiency is required to avoid thick detectors [390]. Extrinsic detectors are based upon silicon and germanium materials, such as Si:In, Si:Ga, Si:As, Ge:Cu, Ge:Ga. To further improve the quantum efficiency, spectral range and detectivity of extrinsic photoconductors, BIB devices can be employed. BIB detectors have much higher primary doping, and are more resistant to the deleterious effects of radiation. The extended wavelength response is due to the formation of the impurity band. Ge:Ga BIB detectors can work at wavelengths between 50 and 220 µm [214], while the cut-off frequency is only ~120 µm for bulk Ge:Ga materials.

Among different types of quantum well infrared photodetectors (QWIP), technology of the type I GaAs/AlGaAs multiple quantum well detectors is the most mature. Rapid progress has been made during the last decades. Nowadays, QWIP focal plane arrays can be fabricated as large as 640×480 pixels, and the imaging performance is comparable to or even better than state-of-the-art HgCdTe detectors [392]. InAs/GaSb type II superlattice detectors working in the 3 to 5 µm window can operate at temperatures up to 180 K [393]. The success of QWIP has stimulated the development of quantum dot detectors. In general, quantum dot detectors are similar to QWIPs, only with the quantum wells replaced by quantum dots (QDs).
in order to apply size confinement in all spatial directions via these semiconductor nanostuctures. InAs/GaAs, InGaAs/InGaP and Ge/Si are the normally used material systems for these types of detectors. Currently QD photodetectors are not yet comparable to HgCdTe photodiodes; further optimizing the band structure and improving QD uniformity are key issues in improving the performance.

The specific detectivity $D^*$, which equals to the reciprocal of noise-equivalent power (NEP), normalized per square root of the sensor’s area and frequency bandwidth, is normally used to characterize performance of sensors. The values of $D^*$ of different types of IR detectors are compared, as shown in Figure 2.19 [390]. Each detector is assumed to view a hemispherical surrounding at a temperature of 300 K and operated at the indicated temperature. Theoretical curves for the background-limited $D^*$ (dashed lines) for ideal (e.g., only unavoidable intrinsic noise sources are considered) photovoltaic (PV), photoconductive (PC) detectors and thermal detectors are also shown.

Figure 2.19: Comparison of the specific detectivity ($D^*$, equals to the reciprocal of noise-equivalent power (NEP), normalized per square root of the sensor’s area and frequency bandwidth) of various commercially available infrared detectors [390].
2.3 Visible and NIR Wireless Communications using LEDs

Light-emitting diodes are considered to be more advantageous than incandescent light bulbs, for higher luminous efficiency and comparable cost. Furthermore, LEDs can be modulated at high-speed, offering the opportunity of using these devices for short-range wireless data communications. Currently available high-power and cost-effective LEDs are normally in the visible and NIR range, limited by available semiconductor materials. Therefore, in this section, LED-based wireless communications in these spectral ranges will be discussed. Note that laser diodes (LDs) can also be used as the light source in these ranges, but they are more expensive than LEDs and thus are not considered here.

2.3.1 LED-based visible light wireless communications

InGaN based highly efficient blue and green LED are now commercially available, making the generation of white light more convenient. This white LED technology is considered a strong candidate for future lighting technology. Visible light communications (VLC) is based on the idea that white LEDs can be simultaneously used as a light source as well as a communications device. These wireless communications systems are preferred for indoor deployment scenarios for lower background noise. The key challenges to achieving high-speed indoor VLC stem are from the path loss, ambient noise, multi-path effect, and interference. Induced signal degradation is greatly influenced by the system configuration. The most common link configurations, for indoor wireless systems, are as shown in Figure 2.20, including directed line-of-sight (LoS) links, non-directed LoS links, diffuse links and quasi-diffuse links, for which (a) and (b) are the most commonly used types for VLC systems [394].

On-off keying (OOK), pulse-position modulation (PPM), pulse-code modulation (PCM) and sub-carrier binary phase-shift keying (SC-BPSK) are some of the more popular modulation schemes used in conjunction with LED wireless systems [8]. In [395], a 40 Mbit/s data rate employing non-return-to-zero OOK modulation was
demonstrated and the speed was increased to 100 Mbit/s using a post equalized white LED [9]. With the deployments of new white LEDs and advanced modulation schemes, systems working at higher data rate have been reported. A white LED-based VLC system operating at data rates up to 400 Mbit/s was demonstrated in [396]. By employing quadrature-amplitude-modulation (QAM) on discrete multitone (DMT), and other system optimisations, a 513 Mbit/s gross transmission rate with BERs $<10^{-3}$ was achieved at a $\sim$0.3 m range [397]. Orthogonal frequency division multiplexing (OFDM) with QPSK for VLC systems was investigated in [398]. With the use of subcarrier multiplexing and wavelength division multiplexing and QAM-OFDM modulation schemes, a transmission link with 575 Mbit/s downstream and 225 Mbit/s upstream data rates was demonstrated [399]. The maximum data rate of 1.1 Gbit/s was shown in [400] by employing carrier-less amplitude and phase modulation (CAP) and a commercially available phosphorescent white LED, where BERs below the forward error correction (FEC) limit of $10^{-3}$ were obtained for a transmission distance of $>20$ cm. Furthermore, a 4-channel MIMO-OFDM system working at a speed of 1 Gbit/s (250 Mbit/s per channel) and 1 m transmission range was reported [401]. VLC systems have been commercialized for bi-directional, high
speed and fully networked wireless communications, and the term Li-Fi (similar to its counterpart Wi-Fi) has been used to describe this technology [402].

Although with ever increasing performance, there are still some fundamental limitations. For example, to obtain higher data rate and longer transmission distance, higher output power level is required from the transmitter. However, the safety regulations have to be applied to LEDs operating in the visible range. This will limit white LEDs to only short-range wireless communications applications, simply due to the reason that large spreading loss cannot be compensated by larger output power. However, further experimental and theoretical studies will provide enhanced foundations for important new developments in this very rapidly growing area, and more ubiquitous applications using VLC systems can be explored.

2.3.2 LED-based NIR wireless communications

NIR wireless communications systems have an even longer history for ubiquitous applications, before white LEDs were widely used in VLC systems. A simple example is the remote controls utilized for many consumer-based electronics. Wavelengths from 780 to 950 nm are currently the best choice for IR indoor wireless systems. In this range, low cost LED sources are readily available and this wavelength also matches the peak responsivity of the inexpensive silicon photodiodes [394]. The primary drawback of radiation in this band relates to eye safety: it can pass through the human cornea and be focused by the lens onto the retina, where it can potentially induce thermal damage [7]. The cornea is opaque to radiation at wavelengths beyond about 1400 nm, considerably reducing potential ocular hazards. Therefore, from a safety perspective, it is believed that the infrared C-band (1530-1565 nm) may be better suited. Unfortunately, the photodiodes available for this range, which are made of Ge or InGaAs, have much higher costs and capacitances per unit area than the silicon-based detectors.

Directed and non-directed LoS IR communications links are now able to achieve data rates higher than 100 Mbit/s, while still maintaining a very simple design [7]. An IR system based on the diffuse link operating at about 950 nm was first proposed
in 1979 [403], showing data rates of 125 kbit/s for PCM and 64 kbit/s for PSK. This was improved to 50 Mbit/s at a range of 2.9 m using OOK modulation [404]. A quasi-diffuse system demonstrated a 70 Mbit/s data rate at a BER of $10^{-9}$ over a 4 m range [405].

Over the past few decades, the Infrared Data Association (IrDa) has established a complete set of protocols for wireless infrared communications. For the Infrared Physical Layer (IrPHY) Specifications, LEDs with wavelength of 850 to 900 nm are employed, and the data rate starts from 2.4 kbit/s up to 1 Gbit/s (GigaIR). The ranges for these protocols are 1 m for standard applications, 0.2 m for low power to low power applications and several meters for GigaIR [406]. These standards will undoubtedly accelerate the development of IR wireless communications and encourage more real-life applications using this part of the electromagnetic spectrum.

2.4 Conclusions

In this chapter, THz and infrared technologies have been reviewed. As one of the most important devices in THz systems, THz sources have been extensively researched. Different types of devices for THz wave generation, either from the electronics or the photonics side, are facing the same challenge: the output power will significantly decrease as the frequency increases/decreases (depending on THz generation method). This gap severely restricts real-life applications of THz systems. Ubiquitous wireless communications systems using low cost LEDs can be found in both visible and NIR ranges, supporting data rates up to Gbit/s and transmission ranges up to several meters. However, at longer wavelengths, LEDs are not widely available and, therefore, not cost-effective. That is one of the reasons why the THz to the mid-infrared spectra are the least commercially exploited ranges, with only a few ubiquitous applications existing within this spectral range.

From the review it can be concluded that thermal-based sources and detectors may be the only options available for developing low cost systems within the thermal infrared band. This is the motivation for the research work presented in this thesis: a cost-effective wireless communications system in the far/mid infrared region, via
thermodynamic-based approaches. Due to fundamental physical limitations, thermal sources and detectors are not able to be operated at high data rates and support long transmission distances. Therefore, ‘THz Torch’ systems are expected to be employed for applications where high performance is not a necessity, but low cost is important. This is the starting point of the ‘THz Torch’ technology.
3 Single-channel ‘THz Torch’ Architecture

3.1 Introduction to Basic Components

3.1.1 Thermal sources

Various approaches, based on either electronic or photonic techniques, have been used to generate coherent THz radiation. However, for ultra-low cost applications, blackbody radiation is considered to be the best option, due to its simplicity, ease of tuning and affordability. The concept of using thermodynamic approaches to generate high-THz (i.e., 10 to 100 THz) thermal power is straightforward, as radiation is naturally emitted by all objects – as their temperature must be above zero degree Kelvin (as described by Planck’s Law). It should be noted that although thermal-based sources offer many benefits (such as simplicity, ease of tuning and affordability), the main drawback is that there is no carrier coherency, as with all unmodulated thermal sources. Thus, only the intensity of band-limited output power can be controlled. Other information (e.g., phasor representations and polarization) cannot be determined for the noise “carrier” and so for any system coherency a known modulating signal would have to be introduced.

In principle, many types of thermal radiators can be employed to generate and radiate electromagnetic energy at high-THz (thermal infrared) frequencies. As a simple proof-of-concept demonstrator, miniature incandescent light bulbs were used. At high-THz frequencies, such COTS bulbs do not contain perfect blackbody radiators (since their tungsten filaments have low emissivity), their windows have relatively
poor transmission characteristics (due to the high absorption and reflectance of the glass envelopes) and without an additional collimating lens and back reflector the spreading loss is high. Nevertheless, these COTS bulbs offer a very low cost means of converting input electrical signals into output thermal infrared radiation.

The thermal analysis of incandescent light bulbs involves all three fundamental methods of heat transfer: radiation, conduction and convection. This analysis is inherently complex, as it requires the study of a number of interacting mechanisms: (a) primary radiation from the filament (which appears mostly in the frequency spectrum between millimeter-wave and beyond visible); (b) absorption of primary radiation by the glass envelope, causing it to heat up; (c) thermal convection inside the glass envelope, causing the filament to heat up the glass envelope; (d) thermal conduction within the glass envelope and also within the two electrical leads to the outside world, which act as poor heat sinks; (e) secondary radiation from the glass envelope (which appears mostly in the thermal infrared spectral region, due to a much lower outside surface temperature); (f) thermal conduction from the glass envelope to the contacting environments on both sides; and (g) thermal convection outside the glass envelope. Due to the inherent complexity, it is not possible to individually quantify the effects of all the mechanisms. In Section 3.2.1, a detailed investigation into the radiation mechanisms associated with the basic transducer is given. It will be shown how both primary and secondary sources of radiation emitted from miniature incandescent light bulbs contribute to the total band-limited output power. The former is defined as the output power that is generated directly from the tungsten filament and passes through the glass envelope, while the latter is defined as the radiation generated by the glass envelope due to its increased temperature.

3.1.2 Thermal detectors

The specific detectivity $D^*$ is normally used to characterize the performance of a detector and is defined as:

$$D^* = \frac{\sqrt{A_x}}{NEP} = R_T \frac{\sqrt{A_x}}{S_n} \text{[cm}\sqrt{\text{Hz}}/\text{W}]$$

(3.1)
where $A_s$ is the surface area of the sensing element; NEP is the noise-equivalent power; $R_V$ is the responsivity of the detector and $S_n$ is the voltage-noise spectral density.

The NEP and voltage responsivity for a detector can be defined as

$$NEP = \frac{S_n}{R_V} \text{[W/}\sqrt{\text{Hz}}]\quad (3.2)$$

and

$$R_V = \frac{u}{\Phi_S} \text{[V/W]}\quad (3.3)$$

where $u$ is the output voltage of the detector and $\Phi_S$ is the input power or incident radiation flux.

For the detection of thermally-generated power radiation within thermal infrared range, two types of detector can be employed: high performance, expensive photonics-based detectors and low performance, ultra-low cost thermal-based detectors. Photon detectors have a high specific detectivity of $\sim 10^{10} \text{ cm} \sqrt{\text{Hz}}/\text{W}$ and fast response times of $< 1 \mu\text{s}$, but may require cryogenic cooling to minimise thermal noise. Furthermore, this type of detector has a relatively narrow instantaneous bandwidth, depending on the energy band gap of the specific semiconductor used.

The latter sensors include thermopile, thermistor bolometer and pyroelectric sensors. Thermopiles, comprising several radiation thermocouples connected in series or, less commonly, in parallel, generates an output voltage proportional to a local temperature difference. Typical specific detectivity and response times for thermopiles are $10^8 \text{ cm} \sqrt{\text{Hz}}/\text{W}$ and $\sim 10 \text{ ms}$, respectively. Thermistor bolometers measure the power of incident electromagnetic radiation via the heating of a material with a temperature-dependent electrical resistance, and have typical values of specific detectivity and response times of $2 \times 10^8 \text{ cm} \sqrt{\text{Hz}}/\text{W}$ and $\sim 1 \text{ ms}$, respectively. Pyroelectric infrared (PIR) sensors exploit pyroelectric materials (e.g., crystalline triglycine sulphate, TGS, or lithium tantalate, LiTaO$_3$; or the piezoceramic lead zirconate titanate, PZT) to generate a change in voltage due to a corresponding change in material temperature. Typical specific detectivity and response times are
Therefore, when compared to photon detectors, thermal detectors typically have a specific detectivity that is two orders of magnitude lower, while the response times are typically four orders of magnitude lower. However, important advantages of thermal-based sensors include operation across an ultra-wide instantaneous bandwidth and at room temperature. For most applications, these advantages may not be enough to justify employing thermal detectors. However, with cost-sensitive applications, there may well be a good argument for investigating engineering solutions with the use of thermal detectors.

The Golay cell should also be mentioned, as it is one of the most sensitive amongst the thermal-based detectors that can operate at room temperature. It consists of a gas-filled cavity. On one face of the cavity is an infrared window and on the other a flexible membrane; embedded within the cavity is an infrared absorbing film. Infrared radiation passes through the window and its energy is absorbed by the film. The absorbed energy is dissipated as heat and the surrounding gas expands, causing the membrane to flex outwards. On the other side of the membrane, an optical beam from a laser or LED is focused onto it, such that the reflected light can be measured by a photodiode. Therefore, when the membrane flexes, the amount of light measured by the photodiode will decrease and its calibrated readout will give an accurate measure of the incident infrared radiation flux. Golay cells typically have a high specific detectivity of $>10^9 \text{cm}\sqrt{\text{Hz}}/\text{W}$, over an ultra-wide instantaneous bandwidth. The main disadvantages of Golay cells are that they have linearity issues, relatively slow response times of $\sim 10 \text{ ms}$, are relatively large (i.e., not compatible with monolithic integration), have an extremely fragile membrane and are relatively expensive.

The pyroelectric sensor has been selected for the ‘$THz Torch$’ technology, due to its ultra-wide instantaneous bandwidth, room temperature operation and most importantly, its ultra-low cost; but this effectively represents the lowest benchmark performance. The pyroelectric sensor used in the first several experiments was the Murata IRA-E710ST1 (with a spectral range from 15 to 300 THz; the value of $D^*$
is unknown) [407], while later improved systems employed the InfraTec LME-553 detector, having a specific detectivity of $>1.1 \times 10^8 \text{ cm}\sqrt{\text{Hz}}/\text{W}$ at 100 Hz modulation frequency [408].

### 3.1.3 Optical coating filters

Radiation from conventional thermal emitters has a continuous frequency spectrum. Therefore, filters have to be employed in order to ensure that communications channels operate within their prescribed spectral ranges. For example, in the earliest single-channel experiments, identical commercially-available 5 µm long-pass filters [407], which came with the PIR detector IRA-E710ST1, were employed for both transmitter and receiver front ends. The spectral transmittance is $\gtrsim 70\%$ across the 25 to 50 THz octave range (which locates within the thermal infrared part), and the mean insertion loss is $\sim 2$ dB, as shown Figure 3.1.

![Figure 3.1: Spectral transmittance for the 5 µm long-pass filter [407].](image)

### 3.2 Introduction to Basic Subsystems

The basic architecture for a single-channel thermal infrared ‘THz Torch’ wireless communications link is shown in Figure 3.2. For the first proof-of-concept single-channel demonstrator, a transmitter consists of simple incandescent bulbs, a PIR
sensor and two identical filters, as introduced in Section 3.1, were employed.

Simple on-off keying is performed by either:

1. direct (internal, electronic) modulation: the transmitter is turned on and off electronically – used in the first experiment [39]

2. indirect (external, spatial) modulation: the transmitter generates CW signal and modulated by an external modulator – used in later experiments [40–48]

### 3.2.1 Basic transmitter subsystem

The transmitter consists of five Eiko 8666-40984 miniature incandescent light bulbs, having a length of 6.3 mm and diameter 2.6 mm. These bulbs are connected in series and assembled into a compact cylindrical package having an outer diameter of 8.2 mm, as shown in Figure 3.3.

Filament working temperature estimation

One of the most important parameters of the thermal source is the working temperature of the bulb’s filament radiator. The ideal spectral radiance, as a function of
wavelength, can be calculated using Planck’s law if the temperature of the radiator
is known. In theory, the working temperature of the filament cannot be measured
directly with the use of a calibrated thermal camera, due to the presence of the
glass envelope. An alternative, widely used indirect approach is to measure the
resistance of the filament, since this is a function of its working temperature. It
is not possible to ascertain the exact geometry of the filament within the commer-
cial bulbs. However, to a good degree of approximation, it can be assumed to be
perfectly cylindrical, having a uniform cross-sectional area $CSA$ and length $l$. The
electrical resistivity of tungsten $\rho(T)$, as a function of absolute temperature $T$, can
be expressed in terms of the filament resistance $R(T)$:

$$\rho(T) = R(t) \cdot (CSA/l)_{eff} \ [\Omega \cdot cm] \quad (3.4)$$

Measured data [409] for the resistivity of tungsten $\rho(T)$ against temperature,
shown in Figure 3.4, can be accurately modelled by the following empirical quadratic
expression:

$$\rho(T) = 2.228 \times 10^{-8} \times T^2 + 2.472 \times 10^{-4} \times T - 1.859 \times 10^{-2} \ [\Omega \cdot \mu m]$$

for $600 < T < 3,000 \quad (3.5)$

Figure 3.4: Measured resistivity of tungsten against temperature and curve fitting
(raw data were sourced from [409]).

At room temperature, $\rho(288 \text{ K}) = 5.14 \times 10^{-6} \ \Omega \cdot \text{cm}$ [409]. The combined resis-
tance of five Eiko 8666-40984 bulbs connected in series is directly measured to be
23.6 $\Omega$ at room temperature. Therefore, if the parasitic resistances of the short leads
are considered negligible (which is a good assumption), this gives an average value of \( R(288 \text{ K}) = 4.72 \ \Omega \) for each bulb filament. As a result, the effective ratio of cross-sectional area to length can be extracted using (3.4) and (3.5) to give a value of \((CSA/l)_{eff} = 1.09 \times 10^{-6} \) cm, which is assumed to be temperature independent. By indirectly measuring the resistance of a bulb, at a specific bias current, the working temperature now can be estimated to an acceptable degree of accuracy.

**Ideal spectral radiance estimation**

After estimating the working temperature of the bulb filament radiator, the ideal (i.e. blackbody) spectral radiance \( I(\lambda, T) \) can be obtained by applying Planck’s law, to give

\[
I_{BB}(\lambda, T) = \frac{2hc^2}{\lambda^5} \cdot \frac{1}{e^{hc/\lambda k_B T} - 1} \ [\text{W/m}^2/\text{sr}/\text{\mu m}]
\]  

(3.6)

where \( I_{BB}(\lambda, T) \) is the power radiated per unit area of emitting surface in the normal direction per unit solid angle per unit wavelength at absolute temperature \( T \) for a blackbody radiator; \( \lambda \) is the free space wavelength; \( h \) is the Planck constant; \( c \) is the speed of light in a vacuum; and \( k_B \) is the Boltzmann constant. Figure 3.5 illustrates the ideal spectral radiance against wavelength at different temperatures.

![Figure 3.5: Calculated ideal spectral radiance versus wavelength at: (left) various temperatures; (right) a working temperature of 772 K.](image)

With the earliest experiments, the Eiko 8666-40984 bulbs had a quiescent DC biasing current of 44 mA, which gives an estimated filament working temperature of 772 K, \( R(772 \text{ K}) \) and a corresponding spectral radiance peak at 80 THz, as shown.
in Figure 3.5. This peak frequency can be easily adjusted by changing the biasing current. With a larger biasing current, one can obtain higher spectral radiance levels, yielding an increased integrated output power for transmission. However, the penalty for this is a decrease in the band-limited output power to input DC power efficiency for this thermal source; which may be an issue where available DC supply power is at a premium (e.g., coin battery powered security key fob applications).

**Filament emissivity estimation**

Most radiator materials are far from being perfect blackbody radiators. As a result, they are not efficient at emitting blackbody radiation. Emissivity $\varepsilon(\lambda, T)$, defined as the ratio of energy radiated by a material to that radiated by an ideal blackbody at the same wavelength and temperature, is a parameter that characterises this efficiency. In practice, if the spectral range is not too large, the emissivity at the surface of a material is only a function of temperature (to a good approximation). Measured data [409] for the spectrum-average or total emissivity of tungsten filament $\varepsilon_{\text{filament}}(T)$, shown in Figure 3.6, can be modelled by the following linear expression:

$$\varepsilon_{\text{filament}}(T) \approx 1.343 \times 10^{-4} \cdot T - 2.019 \cdot 10^{-2}$$  \hspace{1cm} (3.7)

![Figure 3.6: Measured total emissivity of tungsten against temperature and its curve fitting (raw data were sourced from [409]).](image)
Calculated band-limited output radiated power

Having previously obtained the bandwidth of filters and estimated the working temperature of the filament radiators, taking the effective radiating area and emissivity into account and integrating over the spectral band of interest, the net band-limited output radiant intensity from a non-ideal blackbody can be expressed as

\[
I(T) = A_{eff} \cdot \int_{\lambda_1}^{\lambda_2} \varepsilon(\lambda, T) \cdot I_{net}(\lambda, T) d\lambda \quad [\text{W/sr}]
\]  

(3.8)

where \( A_{eff} \) is the effective radiating area of the thermal emitter; \( \varepsilon(\lambda, T) \) is the corresponding emissivity at \( \lambda \) and \( T \); \( \lambda_1 \) and \( \lambda_2 \) are the free space wavelengths associated with the upper and lower frequencies of interest, respectively; and \( I_{net}(\lambda, T) \) is the net spectral radiance emanated, which can be expressed as

\[
I_{net}(\lambda, T) = I(\lambda, T) - I(\lambda, T_0) \quad [\text{W/m}^2/\text{sr}/\mu\text{m}]
\]  

(3.9)

where first and second terms represent the absolute spectral radiance radiated from the source and absorbed by the source from ambient, respectively; and \( T_0 \) is the ambient background temperature.

In this particular case, the output radiant intensity can include both primary and secondary radiation. The former represents the power generated directly from the tungsten filaments and the latter is from the bulbs’ glass envelopes.

The radiant intensity from the primary radiation can be expressed as

\[
I_{primary}(T_{filament}) = I_{filament}(T_{filament}) \quad [\text{W/sr}]
\]  

(3.10)

with

\[
I_{filament}(T_{filament}) = A_{eff_{filament}} \cdot \int_{\lambda_1}^{\lambda_2} T_G(\lambda) \cdot \varepsilon_{filament}(\lambda, T_{filament}) \cdot I_{net}(\lambda, T_{filament}) d\lambda \quad [\text{W/sr}]
\]  

(3.11)

where \( T_{filament} \) represents the calculated working temperature for all identical fila-
ments, which has been estimated to be 772 K for a 44 mA bias current; $A_{\text{eff filament}} = A_{\text{filament}}/2$ is the estimated total effective radiating area (which assumes that only radiation in the outward direction is considered) for five filaments, if the total radiant intensity for the 5-bulb array is required; and $T_G(\lambda)$ is the power transmittance of the glass envelope in thermal equilibrium.

Since the filament can be assumed to be a grey surface, removing its wavelength dependency. $\varepsilon_{\text{filament}}(\lambda, T_{\text{filament}})$ in (3.11) can be replaced by $\varepsilon_{\text{filament}}(T_{\text{filament}})$, as given by (3.7). Therefore, (3.11) can be re-written as

$$I_{\text{filament}}(T_{\text{filament}}) = A_{\text{eff filament}} \cdot \varepsilon_{\text{filament}}(T_{\text{filament}}) \cdot \int_{\lambda_1}^{\lambda_2} T_G(\lambda) \cdot I_{\text{net}}(\lambda, T_{\text{filament}}) d\lambda \ [\text{W/sr}] \quad (3.12)$$

In order to calculate (3.12), $A_{\text{eff filament}}$ and $T_G(\lambda)$ first have to be obtained. With the former, to estimate $A_{\text{eff filament}}$, the filament is assumed to be a uniform cylinder. Therefore, if $CSA$ or $l$ can be physically measured then the other can be extracted using the previously estimated value of $(CSA/l)_{eff}$. Using a scanning-electron microscope (SEM), the average value for the diameter of the Eiko 8666-40984 bulb’s filament was measured to be 22.40 µm, as shown in Figure 3.7. Therefore, for the 5-bulb array configuration the total effective radiating area is $A_{\text{filament}}$ of 12.73 mm$^2$ and $A_{\text{eff filament}} \approx A_{\text{filament}}/2 = 6.37 \text{ mm}^2$.

![Figure 3.7: SEM images of the Eiko 8666-40984 bulb’s tungsten filament: (a) the complete filament within the circular perimeter of the broken glass envelope; (b) close-in view of the filament.](image-url)
With the latter, the inert gas within the class envelope is assumed to be air (which is a good assumption); resulting in air-glass and glass-air boundaries. This problem can be represented by an analogous 2-port network (adapted from [34]), as illustrated in Figure 3.8. Here, \( \hat{P}_i, \hat{P}_r, \hat{P}_a \) and \( \hat{P}_t \) represent the incident, reflected, propagated and transmitted power of the associated electromagnetic waves, respectively. Within the glass, the propagation constant \( \gamma = \alpha + j \beta \), where \( \alpha \) is the attenuation constant and \( \beta \) is the phase constant; \( T_h \) is the thickness of the glass.

\[ \gamma = \alpha + j \beta \]

From Figure 3.8, the voltage-wave reflection coefficients can be expressed as

\[ \rho_1 = \frac{Z_S - Z_T}{Z_S + Z_T} \] (3.13)

and

\[ \rho_2 = \frac{Z_T - Z_S}{Z_S + Z_T} = -\rho_1 \] (3.14)

where \( Z_S \rightarrow \eta_I = \sqrt{\frac{j \omega \mu}{\sigma_c + j \omega \varepsilon}} \) is the intrinsic impedance of glass; \( \mu \) is the intrinsic permeability; \( \sigma_c \) is the intrinsic conductivity; \( \varepsilon \) is the intrinsic permittivity and the termination impedance of this equivalent 2-port network \( Z_T \rightarrow \eta_0 = \sqrt{\frac{\mu_0}{\varepsilon_0}} \) is the intrinsic impedance of free space (i.e., air dielectric).

The associated voltage-wave transmission coefficients are given by

\[ \tau_1 = 1 + \rho_1 = \frac{2Z_S}{Z_S + Z_T} \] (3.15)

and

\[ \tau_2 = 1 + \rho_2 = \frac{2Z_T}{Z_S + Z_T} \] (3.16)
Taking into account the infinite number of reflections between the air-glass and glass-air boundaries, using scattering (or S-)parameters, the overall forward voltage-wave transmission coefficient $S_{21}$ and overall input voltage-wave reflection coefficient $S_{11}$ can be expressed as [34]

$$S_{21} = \tau_1 \cdot \left[ \frac{e^{-\gamma T_h}}{1 - (e^{-\gamma T_h} \rho_1)^2} \right] \cdot \tau_2$$

(3.17)

$$S_{11} = \rho_1 \cdot \left[ \frac{1 - e^{-2\gamma T_h}}{1 - (e^{-\gamma T_h} \rho_1)^2} \right]$$

(3.18)

To calculate $S_{21}$ and $S_{11}$, measured values for the complex index of refraction $	ilde{n}(\lambda) = n(\lambda) - jk(\lambda)$, where $n(\lambda)$ is the refractive index and $k(\lambda)$ is the extinction coefficient, for typical (soda lime silica) window glass at a room temperature was first needed [410]. The propagation constant through the glass $\gamma$ is expressed as:

$$\gamma = j\tilde{n}(\lambda) \cdot \frac{\omega}{c} = j2\pi \frac{\tilde{n}(\lambda)}{\lambda}$$

(3.19)

At normal incidence, the intrinsic impedance of the glass is given by

$$\eta_I = \frac{\eta_0}{\tilde{n}(\lambda)}$$

(3.20)

By applying (3.20) to (3.13)-(3.16), $\rho_1$, $\rho_2$, $\tau_1$ and $\tau_2$ can be obtained. As a result, $S_{21}$ and $S_{11}$ can be calculated using (3.17) and (3.18). The overall power transmittance and reflectance are determined from $S_{21}$ and $S_{11}$, respectively, while power absorptance is given by $1-|S_{11}|^2-|S_{21}|^2$.

The Eiko 8666-40984 bulb has a glass envelope thickness of $\sim 300$ µm at its sides and $\sim 400$ µm at the top end. If an average value of 350 µm is taken for the glass envelope, the overall power transmittance $T_G(\lambda)$, reflectance and absorptance from 1 to 100 THz (assuming typical window glass at room temperature) is shown in Figure 3.9 [44]. Here, $|\tau|^2 = |e^{-\alpha T_h}|^2$, $|\rho|^2 = |\rho_1|^2$ and $1-|\tau|^2-|\rho|^2$ correspond to the power transmittance, reflectance and absorptance, respectively, without considering multiple reflections; these parameters are often plotted in the open literature as
first-order approximations [410]. It can be seen in Figure 3.9 that these first-order approximations are very accurate below 65 THz, where absorptance dominates.

![Graph showing transmittance, reflectance, and absorptance](image)

Figure 3.9: Calculated power transmittance, reflectance and absorptance for 350 µm thick window glass at room temperature [44] (values of $n(\lambda)$ and $k(\lambda)$ were sourced from [410]).

It can also be seen from Figure 3.9 that typical window glass can be considered opaque below $\sim$60 THz. For most thermal infrared applications, this would only allow its use in its transparent region between ca. 70 to 100 THz band. However, for the ‘THz Torch’ applications, the high absorptance will contribute to the secondary source of radiation within the opaque spectral region of the glass envelope (due to the increased outer surface temperature of glass envelopes), while still utilizing the primary radiation from the bulb filaments within its transparent spectral region. It is also interesting to note that the 3 dB cut-off occurs at around 65 THz. Based on the calculated overall power transmittance $T_G(\lambda)$, the radiant intensity from the primary radiation can be calculated using (3.10).

Secondary radiation is defined as the radiation generated by the glass envelope due to its increased temperature. As seen in Figure 3.9, below 65 THz the power transmittance through the glass envelope is poor and the secondary source of radiation dominates. The band-limited net radiant intensity from secondary radiation can be expressed as

$$I_{\text{secondary}}(T_{\text{glass}}) = I_{\text{glass}}(T_{\text{glass}}) \text{ [W/sr]}$$

(3.21)
with

\[ I_{\text{glass}}(T_{\text{glass}}) = A_{\text{eff, glass}} \cdot \int_{\lambda_1}^{\lambda_2} \varepsilon_{\text{glass}}(\lambda, T_{\text{glass}}) \cdot I_{\text{net}}(\lambda, T_{\text{glass}}) d\lambda \ [W/\text{sr}] \quad (3.22) \]

where \( A_{\text{eff, glass}} \) is the total effective radiating area for all the glass envelopes and \( \varepsilon_{\text{glass}}(\lambda, T_{\text{glass}}) \) is the emissivity of the glass envelope with an outer surface temperature \( T_{\text{glass}} \). Here, we assume that \( \varepsilon_{\text{glass}}(\lambda, T_{\text{glass}}) \) is wavelength-dependent and does not change significantly as temperature increases from the room temperature to the highest elevated temperature of 366 K; this is a reasonable assumption, as stated in [409]. Furthermore, according to Kirchhoff’s law of thermal radiation, emissivity is equal to the power absorptance when in thermodynamic equilibrium. Therefore, the absorptance shown in Figure 3.9 is used as the emissivity of the glass, \( \varepsilon_{\text{glass}}(\lambda) \), which is only frequency-dependent.

With the particular 5-bulb array configuration shown in Figure 3.10a, the radiant intensity from secondary radiation can be further separated into two parts; the central higher temperature region and its surrounding lower temperature region, as shown in Figure 3.10b, which was measured using a FLIR E60 thermal camera [411].

![Figure 3.10: Five-bulb array: (a) transmitter configuration; (b) measured surface temperature distribution (THz band-pass filter and aperture removed). El1, Sp1 and Sp2 represents the average temperature within a circle and spot temperatures for two outer bulb centers, respectively.](image)

Therefore, (3.21) and (3.22) can be re-written as

\[ I_{\text{secondary}}(T_{\text{glass}}) = I_{\text{glass}}(T_{\text{high}}) + I_{\text{glass}}(T_{\text{low}}) \ [W/\text{sr}] \quad (3.23) \]
\[ I_{\text{glass}}(T_{\text{high}}) = A_{\text{eff, glass, high}} \cdot \int_{\lambda_1}^{\lambda_2} \varepsilon_{\text{glass}}(\lambda) \cdot I_{\text{net}}(\lambda, T_{\text{high}}) d\lambda \text{ [W/sr]} \tag{3.24} \]

\[ I_{\text{glass}}(T_{\text{low}}) = A_{\text{eff, glass, low}} \cdot \int_{\lambda_1}^{\lambda_2} \varepsilon_{\text{glass}}(\lambda) \cdot I_{\text{net}}(\lambda, T_{\text{low}}) d\lambda \text{ [W/sr]} \tag{3.25} \]

where \( T_{\text{high}} \) and \( T_{\text{low}} \) are the average temperatures for the high and low temperature regions, respectively; \( A_{\text{eff, glass, high}} \approx D^2(1 - \pi/4) = 1.45 \text{ mm}^2 \) is the total effective radiating area of the higher temperature region for the 5-bulb array; \( A_{\text{eff, glass, low}} \approx 8\pi D/2^2 = 42.47 \text{ mm}^2 \) is the total effective radiating area of the lower temperature region and \( D = 2.6 \text{ mm} \) is the diameter of the bulb’s glass envelope [44].

The surface temperature of the glass envelope depends on the emissivity, emitting area, temperature, position and shape of the tungsten filament. Instead of using complex thermodynamic modelling to simulate its surface temperature distribution, a more direct approach is to measure its temperature using a thermal camera. An experiment using the FLIR E60 thermal camera was performed. Note that the THz filter and associated aperture were removed, in order obtain the actual temperature distribution for the 5-bulb array.

Measured surface temperatures for 5-bulb array, without bias and with bias currents from 44 to 80 mA, are shown in Figure 3.11. For this experiment, room temperature was measured to be 21.5 °C (294.5 K). It can be seen that with this particular configuration, the central region of the source exhibits the highest temperatures.

The measured maximum and average temperatures for the 5-bulb array, as well as the temperature at the centre of the outer bulbs, for different bias currents, are given in Table 3.1. With reference to 3.10b, the average temperature is measured within a circle (denoted by El1) and spot temperatures are given for two outer bulb centers (denoted by Sp1 and Sp2). The outer bulb center values recorded in Table 3.1 represent the average values of Sp1 and Sp2, while the maximum temperature for the array is automatically captured by the camera within the selected area (located at the red triangle). After obtaining these temperature readings, the band-limited radiant intensity from secondary radiation then can be calculated using (3.23)-(3.25).

It is worth noting that the average temperature of the glass envelope is only slightly
Figure 3.11: Measured surface temperature (in °C) of the 5-bulb array as a function of bias current: (a) no current; (b) 44 mA; (c) 50 mA; (d) 60 mA; (e) 70 mA; (f) 80 mA.

higher than the low temperature, since $A_{\text{eff, glass, high}} \ll A_{\text{eff, glass, low}}$. By splitting the radiant intensity from secondary radiation into two parts, the temperature distribution and output radiant intensity estimation can be described more accurately.

The net spectral radiance for both radiation mechanisms can be obtained by considering the average emissivity of the filament $\bar{\varepsilon}_{\text{filament}}(T_{\text{filament}})$, and the power transmittance $T_G(\lambda)$ and emissivity $\varepsilon_{\text{glass}}(\lambda)$ of the glass envelope. Figure 3.12 shows the calculated net spectral radiance from these two radiation mechanisms at a bias current of 44 mA. It is interesting to note that below 50 THz, secondary radiation dominates the output power; while primary radiation dominates above 50 THz.

By inspection of Figure 3.12, the contribution of the secondary radiation may seem insignificant, when compared to the primary radiation, given the low temperature of operation. However, it will be shown that when the total effective radiating areas of $A_{\text{eff, glass}} = 43.92 \text{ mm}^2$ and $A_{\text{eff, filament}} = 6.37 \text{ mm}^2$ are taken into account for the glass envelope and filament, respectively, the resulting radiation intensities and
Table 3.1: Measured outer surface temperatures for the 5-bulb array at different bias currents.

<table>
<thead>
<tr>
<th>Bias Current (mA)</th>
<th>Measured Temperature (K)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Array Maximum</td>
</tr>
<tr>
<td>0</td>
<td>294.5</td>
</tr>
<tr>
<td>40</td>
<td>308.9</td>
</tr>
<tr>
<td>44</td>
<td>312.3</td>
</tr>
<tr>
<td>50</td>
<td>318.9</td>
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<tr>
<td>55</td>
<td>325.3</td>
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<td>60</td>
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</tr>
<tr>
<td>70</td>
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</tr>
<tr>
<td>75</td>
<td>353.7</td>
</tr>
<tr>
<td>80</td>
<td>366.1</td>
</tr>
</tbody>
</table>

Figure 3.12: Calculated spectral radiance from both primary and secondary radiation for a bias current of 44 mA.

Band-limited output radiated powers are now of the same order of magnitude for a bias current of 44 mA. Figure 3.13 shows the calculated band-limited output power between 1 to 100 THz. It can be seen that, as the bias current increases, the peak in spectral radiance increases in frequency, as dictated by Planck’s law. However, when compared to the secondary radiation mechanism, the output power from the primary radiation mechanism is more strongly coupled to changes in bias current.

The overall total radiant intensity, with contributions from both radiation mech-
In order to verify this analysis, an experiment was performed to measure the total output power from the ‘THz Torch’ transmitter. The Ophir 10A thermal power sensor was used; this commercial power detector has a spectral range of 15 to 1,580 THz (i.e., from 20 \( \mu \text{m} \) to 0.19 \( \mu \text{m} \)) [412]. The source and detector were positioned as close as possible to each other, so that all the generated power is captured by the large sensing element, having a diameter of 16 mm. In this experiment, there was no THz band-pass filter.

With this experiment, the spectral range used for the calculation was chosen to be 15 to 300 THz (i.e., 20 \( \mu \text{m} \) to 1 \( \mu \text{m} \)); the upper frequency limit is dictated by the

\[
I_{TX}(T) = I_{primary}(T_{filament}) + I_{secondary}(T_{glass}) \ [\text{W}/\text{sr}] (3.26)
\]

By considering a projected solid angle of \( \pi \), the total band-limited output radiated power from the ‘THz Torch’ transmitter can be expressed as

\[
P_{TX}(T) = \pi \cdot [I_{primary}(T_{filament}) + I_{secondary}(T_{glass})] \ [\text{W}] (3.27)
\]

**Measured band-limited output radiated power**

In order to verify this analysis, an experiment was performed to measure the total output power from the ‘THz Torch’ transmitter. The Ophir 10A thermal power sensor was used; this commercial power detector has a spectral range of 15 to 1,580 THz (i.e., from 20 \( \mu \text{m} \) to 0.19 \( \mu \text{m} \)) [412]. The source and detector were positioned as close as possible to each other, so that all the generated power is captured by the large sensing element, having a diameter of 16 mm. In this experiment, there was no THz band-pass filter.

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\[
P_{TX}(T) = \pi \cdot [I_{primary}(T_{filament}) + I_{secondary}(T_{glass})] \ [\text{W}] (3.27)
\]
available data for the complex index of refraction for glass [410]. However, this will not introduce any significant discrepancy, as most of the output radiated power (e.g., >99%) is between 15 and 300 THz, for all the temperatures considered here. The measured and calculated values for band-limited output radiated power are shown in Figure 3.14. It can be seen that the measured results are in excellent agreement with the calculated values, even though we have made many assumptions throughout, confirming the validity of the primary and secondary radiation modelling approaches.

![Figure 3.14: Measured and calculated band-limited (15-300 THz) output radiated power against bias current for the 5-bulb array.](image_url)

**Band-limited output radiated power to input DC power conversion**

**Efficiency Calculations**

The band-limited output radiated power and conversion efficiency for both single and multi-channel ‘THz Torch’ transmitters, employing the same 5-bulb array configuration described previously, can be calculated. The spectral range for the first proof-of-concept single-channel ‘THz Torch’ system was defined over the 25 to 50 THz octave bandwidth [39,42,44], while four non-overlapping spectral ranges for the 4-channel multiplexing systems are: 15 to 34 THz for Channel A; 42 to 57 THz for Channel B; 60 to 72 THz for Channel C; and 75 to 89 THz for Channel D [40,44,45].

Figure 3.15 shows the calculated band-limited output radiated power levels. It can be seen that, for low bias currents, the single-channel and Channel A thermal sources can generate more band-limited output radiated power. The reason for this...
is that the spectral peak for the secondary radiation is at \( \sim 32 \text{ THz} \), which is within the operation bands of these two sources. While the spectral peak for the primary radiation is at \( \sim 80 \text{ THz} \) (for a bias current of 44 mA), which lies within Channel D, the output power for Channel D is lower due to the smaller total effective radiating area. As the bias current increases, much more band-limited output radiated power will be generated within Channel D, surpassing the values from any other channel beyond a bias current of 50 mA. In the other extreme, Channel B is far from both primary and secondary radiation spectral peaks, and so its band limited output radiated power is less than all the other channel sources.

![Figure 3.15: Calculated band-limited output radiated power against bias current for single and multi-channel 'THz Torch' transmitters.](image)

The power conversion (or transducer) efficiency is defined as the ratio of band-limited output radiated power to input DC power. This efficiency has been calculated for different channels, as shown in Figure 3.16. It has been previously estimated that for a 44 mA bias current, the filament has a working temperature of 772 K and that for a single bulb the corresponding resistance \( R(772 \text{ K}) = 17.02 \text{ \Omega} \). Therefore, the input DC power is \( P_{DC} = I^2R = 33 \text{ mW} \). As a result, the conversion efficiency can be estimated to be \( \sim 1\% \) for the single channel and Channel A thermal sources and approximately half this value for the other channel sources at this bias point. It should be noted that the insertion loss of the THz band-pass filter and also spreading losses are not taken into account. As the bias current increases, the efficiency for the
single channel and Channel A and B sources will decrease, as the spectral peak for the primary radiation moves higher in frequency, and more power is radiated above the spectral ranges of these channels. For this reason, Channel C and D sources show an increase in power conversion efficiency at higher bias currents.

![Figure 3.16](image)

**Figure 3.16:** Calculated power conversion efficiency against bias current for single and multi-channel ‘THz Torch’ transmitters.

For specific niche applications, high power conversion efficiency can represent important attributes for thermal sources implemented using incandescent light bulbs. Indeed, the band-limited output radiated power and power conversion efficiency can be optimized with appropriate levels of quiescent DC bias current. Although low cost NIR LEDs can provide higher efficiency and switching speeds, their output spectral frequency cannot be tuned. In contrast, the spectral peak of thermal sources can be continuously tuned over a vast spectral range, simply by changing the quiescent DC bias current. Furthermore, with multi-channel systems, different channels can have different quiescent DC bias currents, so that the output radiated power can be balanced between channels, optimizing the performance of complete frequency multiplexed systems.

### 3.2.2 Basic receiver subsystem

The pyroelectric sensor is able to detect objects by measuring the difference in temperature between a target and its background. The pyroelectric materials create
a spontaneous polarization charge, with a magnitude depending on the material’s change of temperature. The equivalent circuit model for typical PIR sensors is as shown in Figure 3.17 [44].

![Equivalent circuit model for typical PIR sensors](image)

Figure 3.17: Equivalent circuit model for typical PIR sensors [44].

When incident radiation flux \( \Phi_S \) is absorbed by the sensing element’s material, there is a corresponding change in its physical temperature. This temperature change can be expressed as [414]:

\[
\Delta T = \frac{\alpha \Phi_S}{G_T \sqrt{1 + (\omega_m \tau_T)^2}} \text{[K]}
\]

(3.28)

where \( \alpha \) is the absorption efficiency; \( \Phi_S \) is the incident power, \( \omega_m \) is the angular modulation frequency of the incoming signal; \( \tau_T = H_P / G_T \) is the thermal time constant of the detector; \( H_P \) and \( G_T \) are the heat capacity and thermal conductance of the sensing element’s material, respectively.

A change in temperature will result in a small polarization current, \( i_P \), due to the pyroelectric effect, which is proportional to the temperature difference and the sensing element’s surface area \( A_s \):

\[
i_P = \omega_m p A_s \Delta T \text{[A]}
\]

(3.29)

where \( p \) is the pyroelectric coefficient of the pyroelectric sensing element.

Therefore, the output voltage from the PIR sensor is given by:

\[
u = \omega_m p A_s \alpha \Phi_S \frac{R}{G_T} \frac{1}{\sqrt{1 + (\omega_m \tau_T)^2}} \frac{1}{\sqrt{1 + (\omega_m \tau_E)^2}} \text{[V]}
\]

(3.30)
where \( \tau_E \) is the electrical time constant of the PIR sensor.

Depending on the preamplifier type, there are two types of PIR sensors: voltage mode and current mode. The voltage mode can be implemented using a voltage follower, while the current mode using a transimpedance amplifier (TIA), as seen in Figure 3.18 [414].

![Figure 3.18: Voltage and current mode preamplifier for PIR sensors [414].](image)

In (3.30), \( R = R_G \) and \( \tau_E = R_G \cdot C_P \) for voltage mode detectors, where \( C_P = \frac{\varepsilon_0 \varepsilon_r A_s}{d_P} \) is the capacitance of the pyroelectric sensing element; \( d_P \) is the thickness of the sensing element. For the current mode detectors, \( R = R_{fb} \) and \( \tau_E = R_{fb} \cdot C_{fb} \), where \( R_{fb} \) and \( C_{fb} \) are the feedback resistance and capacitance of the integrated preamplifier, respectively.

From (3.3), the voltage responsivity for a PIR sensor can be expressed as [414]

\[
R_V = \omega_m \alpha A_s d \frac{R}{G_T} \frac{1}{\sqrt{1 + (\omega_m \tau_T)^2}} \frac{1}{\sqrt{1 + (\omega_m \tau_E)^2}} [V/W]
\]

(3.31)

It should be noted that thermal time constant \( \tau_T \) and the electrical time constant \( \tau_E \) determine the corner frequencies of voltage responsivity of PIR sensors [414]. For voltage mode sensors, the lower corner frequency is determined by \( \tau_E \), while the upper corner frequency is determined by \( \tau_T \). For the current mode type, \( \tau_E \) is significantly lower due to a much smaller feedback capacitance \( C_{fb} \), determining the upper corner frequency of the voltage responsivity. Furthermore, current mode detectors have a much higher voltage responsivity above the electrical corner frequency than that of the voltage mode. Figure 3.19 shows a comparison of frequency dependencies of voltage responsivity for these two types of PIR sensors [414].

99
Figure 3.19: Frequency dependencies of voltage responsivity for voltage and current mode PIR sensors [414].

The Murata IRA-E710ST1 sensor used in the first experiments has two $2 \times 1 \text{ mm}^2$ sensing elements. Such dual-element sensors are normally used for detecting objects in motion, as the background environment can be subtracted from the moving target. As a result, it was necessary to remove this function by introducing a short circuit to one of the elements.

Even with such integrated preamplifiers, the output voltages are still small, in the mV range. As a result, back-end electronics is needed. In the first experiments, analogue circuitry is introduced, consisting of a 2-stage high gain low-noise amplifier (LNA), employed for signal amplification and DC blocking; this is followed by a Schmitt trigger for simple analogue-to-digital conversion. This simple back-end electronics represents an ultra-low cost solution for on-off keying (OOK) digital modulation applications.

### 3.2.3 First proof-of-concept single-channel demonstrator

The basic single-channel architecture, shown in Figure 3.2, was implemented. The transmitter and receiver were positioned 0.5 cm apart, to create a very short LoS wireless communications link. A rectangular pulse generator was used to digitally modulate the bulbs, via an intermediate driver circuit. The quiescent DC bias current through five bulbs connected in series was set to 44 mA, giving the desired peak spectral radiance at $\sim 80 \text{ THz}$ [39].
Experimental result shows a maximum effective bit rate of 10 bit/s over a 0.5 cm transmission distance for non-return-to-zero (NRZ) pulses, as illustrated in Figure 3.20. From Figure 3.20a, it can be seen that the bit rate decreases dramatically as the transmission distance increases. This is mainly due to losses from beam spreading, since no back reflector or collimating lenses were employed. These contributions will be discussed further in Section 3.3. While this attenuation characteristic would otherwise be viewed as undesirable, for specific niche security and defence applications this offers a low probability of interception or jamming. It can also be seen, in Figure 3.20b, that as more bulbs are introduced the maximum effective bit rate increases for a fixed transmission distance. However, the law of diminishing returns applied, where a saturation effect can be seen as the number of bulbs increases beyond three, for this particular experimental setup.

![Figure 3.20: Maximum effective bit rate: (a) versus transmission distance (with 5 bulbs); (b) versus number of bulbs at a fixed distance of 0.5 cm [39].](image)

3.3 Fundamental Limits and Engineering Solutions

Ultra-low cost short-range wireless communications links based on the ‘THz Torch’ technology have been demonstrated experimentally in Section 3.2.3. While the measured performances of these first proof-of-concept demonstrators have been limited, in terms of bit rate and transmission range, it is important to recognise that this emerging technology is still in its infancy. By understanding the physical limits of the individual technologies, depending on the particular application, it can be justified whether it is worth pushing the performance of an individual component to its
limit, replace it with a more expensive higher performance component or adopt an alternative solution [44].

### 3.3.1 Operating spectrum

The recommended operating spectrum for the ‘THz Torch’ technology has been identified as lying in the high-THz (thermal infrared) region of 10 to 100 THz, which is above the frequencies normally associated with terahertz technologies. This recommendation is based on the levels of spectral radiance, which are relatively low below 10 THz: lying in the long wavelength tail creating by applying Planck’s law. Increasing the bulb’s quiescent DC bias current will increase the working temperature of the filament radiator and, therefore, increase both the amplitude and frequency of the spectral radiance peak. However, the levels of spectral radiance in the long-wavelength tail will not increase by anywhere near the same amount as that at the peak. As a result, of the high levels of spectral power generated near the peak will be wasted (since the peak will be above 100 THz) and the DC-to-THz power conversion efficiency will reduce dramatically.

The operating spectrum is also limited by the background thermal noise associated with the transmitter hardware or target, ambient channel environment and receiver hardware. The equivalent photon energy is 25 meV at room temperature, corresponding to a frequency of ~6 THz. Therefore, for room-temperature operation, the ‘THz Torch’ concept has a lower frequency limit of ~6 THz. The thermodynamics of the ‘THz Torch’ concept relies on temperature changes; temporal (in the case of secure RFID, smart key fobs, secure data transfer, covert communications, night signalling and IED trigger applications) or spatial (in the case of IED triggers, high ‘pixel’ resolution FIR counter-measures for adaptive thermal camouflage and sentry decoys and identification friend or foe applications). Introducing an active cooling subsystem (e.g., Peltier effect cooler, PEC) to the transmitter/target can increase its temperature change ΔT, improving the end-to-end systems performance, but this is only when the transmitter/target dominates the receiver’s field of view (FOV). Introducing a PEC to the receiver’s back-end electronics only will help to improve
its output signal-to-noise ratio. When employing a PEC, for either the transmitter or receiver, there are the inevitable increases in associated DC power consumption, complexity, size, mass and cost. More advanced equipment, e.g., lock-in amplifiers, can be used to further improve the SNR performance.

3.3.2 Transmitter thermal time constant

The response times associated with incandescent light bulbs’ filaments are not of importance if the source radiation is to be modulated indirectly (e.g., via spatial modulators) or if applications do not require fast switching speeds when modulated directly (e.g., via electronic modulation). However, for applications that require faster data rate and that cannot employ indirect modulation, the physical limitations associated with direct modulation response times need investigating.

There are two thermal time constants that need to be considered, corresponding to the heating and cooling processes. The heating (turn-ON) thermal time constant $\tau_H$ is defined as the time it takes to go from 10% to 90% of the temperature difference between the initial and final steady-state temperatures. The cooling (turn-OFF) thermal time constant $\tau_C$ is defined in a similar way to $\tau_H$; being the time taken to decrease from 90% to 10% of the temperature difference between the initial and final steady-state temperatures. Figure 3.21 shows the instantaneous filament temperature, which was extracted from the measured transient filament resistance, for a square wave having a mark(M):space(S) ratio of 1:1 and modulating frequency of 0.16667 Hz (representing a very slow bit rate).

**Filament turn-ON thermal time constant**

When the bulb is electrically turned ON with a step response function, at its initial temperature $T(0)$, there is a large injection of current and the temperature of the filament increases. Since the resistivity of tungsten has a positive temperature coefficient, the instantaneous bulb resistance also increases; from its initial value $R(T(0))$ until a steady-state value is reached, at thermal equilibrium, where the input power is exactly balanced out by all of the dissipative (i.e., heat transfer) loss mechanisms.
Figure 3.21: Instantaneous filament temperature for a square wave having M:S = 1:1 and modulating frequency of 0.16667 Hz. \((t_m > \tau_H \text{ and } t_s > \tau_C)\).

Without knowing all the parameters of the filament and glass envelope, it is difficult to have an accurate model to describe the complex thermodynamic processes. A simple alternative is to assume that heat transfer occurs without any energy leaving the whole system (i.e., adiabatically). It is then possible to calculate the ideal heating or turn-ON thermal time constant \(\tau_{H,\text{ideal}}\) of the filament. This represents the lower bound for the turn-ON time for the thermal source.

If the input power is only used to increase the temperature of the filament, the energy balance equation can be expressed as

\[
M \cdot C(T) \cdot \frac{dT}{dt} = P_{IN}(T) \quad \text{where} \quad P_{IN}(T) = \frac{V_{DC}^2}{R(T)}
\]

(3.32)

where \(M \text{ [kg]}\) is the mass of the filament; \(C(T) \text{ [J/kg/K]}\) is the specific heat capacity; \(t\) is the time dependency; \(P_{IN}(T)\) is the input power, which is a function of temperature; and \(V_{DC} = 0.749\text{ V}\) (for a bias current of 44 mA) is the DC voltage across a single filament.

Since the filament is assumed to be a uniform cylinder with a diameter of 22.40 \(\mu\text{m}\) and \((CSA/l)_{eff} = 1.09 \times 10^{-6}\text{ cm}\), then \(M = CSA \cdot l \cdot d_{\text{tungsten}} = 275\ \mu\text{g},\) where \(d_{\text{tungsten}} = 19.25\ \text{g/cm}^3\) is the density of tungsten. In practice, the specific heat capacity of tungsten is temperature dependent. Within the temperature range of 300 to 3,000 K, with constant pressure, this can be modelled using an empirical
curve fit [415]

\[ C(T) = \frac{1}{0.18385} \sum_{n=0}^{4} A_n \left( \frac{T}{1000} \right)^{n-1} \text{[J/kg/K]} \]  
(3.33)

where \( A_0 = -0.20869 \), \( A_1 = 23.70345 \), \( A_2 = 5.13206 \), \( A_3 = -1.99922 \) and \( A_4 = 0.73417 \) are the best fit coefficients. Figure 3.22 shows the modelled (3.33) and measured [415] specific heat capacity for tungsten, showing excellent agreement.

Figure 3.22: Specific heat capacity against temperature for tungsten (raw data were sourced from [415]).

Substituting (3.33) into (3.32) and integrating over the temperature range of interest, we have

\[ \tau_{H_{ideal}} = M \cdot \int_{T_{H_{initial}}}^{T_{H_{final}}} \frac{C(T)}{P_{IN}(T)} dT \text{[s]} \]  
(3.34)

Here, the heating thermal time constant is defined as the time taken to go from 10% to 90% of the temperature difference between the initial and final steady-state temperatures. For a 44 mA bias current, \( T_{H_{initial}} = 347.2 \text{ K} \) and \( T_{H_{final}} = 724.8 \text{ K} \). Therefore, \( \tau_{H_{ideal}} \) is calculated to be 207 ms.

Figure 3.23a shows the measured transient turn-ON current profile for a bias current of 44 mA. The transient temperature of the filament can also be extracted, according to the measured bulb resistance, as shown in Figure 3.23b.

A simple empirical curve fit can be applied to the instantaneous turn-ON filament temperature shown in Figure 3.23b, as given by the following expression with less than 1.5% error:

\[ T(t) = T(\infty) - \Delta T_{MAX} \cdot e^{-1.9995t/\tau_H} \]  
(3.35)
Figure 3.23: Turn-ON responses giving the experimental values for instantaneous:
(a) current; (b) extracted filament temperature [44].

where \( t \) is the instantaneous heating time starting from the initial working temperature \( T(0) \); \( \Delta T_{MAX} = [T(\infty) - T(0)] \) is the maximum change in filament temperature; and \( T(\infty) \) is the final steady-state temperature in thermal equilibrium. With this example, \( T(0) = 300 \text{ K} \) and \( T(\infty) = 772 \text{ K} \). The turn-ON thermal time constant was found experimentally to be \( \tau_H = 645 \text{ ms} \). By comparing the calculated and measured thermal time constants, it is found that \( \tau_{H,\text{ideal}} / \tau_H \approx 0.32 \), showing that approximately \( \frac{2}{3} \)rd of the input power is dissipated through heat transfer during the turn-ON process.

**Filament turn-OFF thermal time constant**

When the bulb is electrically turned off with a step response function, at its initial temperature \( T(0) \), the filament starts cooling down and its instantaneous resistance decreases with time from its initial steady-state value of \( R(772 \text{ K}) \) to its final steady-state value of \( R(300 \text{ K}) \). However, the current will instantly drop to zero once the bulb is turned off. Therefore, the technique used to determine the turn-ON thermal time constant cannot be used here. To measure the turn-OFF thermal time constant, a new technique was proposed [44]. The instantaneous turn-ON currents are measured at different delayed intervals of cooling after the turn-OFF from the initial steady-state condition.

Cooling of the hot tungsten filament is a relaxation process [416]. Therefore, \( \tau_C \) is expected to be larger than \( \tau_H \). Measured turn-OFF responses for instantaneous
current and extracted filament temperature are shown in Figure 3.24. A simple empirical curve fit can be applied to the instantaneous turn-OFF filament temperature, shown in Figure 3.24b, as given by the following expression with less than 3% error:

\[ T(t) = T(\infty) + \Delta T_{\text{MAX}} \cdot e^{-2.415t/\tau_C} \]  \hspace{1cm} (3.36)

where \( t \) is the instantaneous cooling time from the initial working temperature \( T(0) \); \( \Delta T_{\text{MAX}} = [T(0) - T(\infty)] \) is the maximum change in filament temperature; and \( T(\infty) \) is the final steady-state temperature. With this example, \( T(0) = 772 \) K and \( T(\infty) = 300 \) K. The turn-OFF thermal time constant was found to be \( \tau_C = 2.415 \) ms, which is much larger than \( \tau_H \). This severely limits the switching speed for the bulb and set a practical limit on the data rate for OOK digital modulation in a communications channel – although this may not be an issue with applications that do not required high data rates.

![Figure 3.24: Turn-OFF responses giving the experimental values for instantaneous: (a) current; (b) extracted filament temperature [44].](image)

**Glass envelope thermal time constants**

The instantaneous outer surface temperature of the glass envelopes can also be measured directly, giving the thermal time constants for the secondary radiation. Measured transient maximum and average temperatures for the 5-bulb array, and temperature at the centre of the outer bulbs, for a 44 mA bias current, are shown in Figure 3.25. It can be seen that the thermal time constants for the average temperature of the 5-bulb array is similar to that of the centre of the outer bulbs.
Using the same definitions as those for the filament, the heating and cooling thermal time constants for the glass envelopes are measured to be 195 and 385 seconds, respectively, for a bias current of 44 mA. Therefore, the thermal time constants for the glass envelopes are between two and three orders of magnitude larger than those for the filament. This would create a fundamental limitation on the signalling rate if direct modulation is applied. However, this would not present a fundamental limit for ‘THz Torch’ applications if external modulators (e.g., programmable mechanical shutters or spatial light modulators) are employed.

**Effect of thermal time constants on the data rate**

If the incandescent light bulbs are directly modulated, both the heating and cooling thermal time constants of the filament, as well as glass envelopes impose fundamental limitations on their signalling speed; this was observed in the first experiment, where the maximum bit rate was only 10 bit/s. Since pyroelectric sensors can only detect the change in temperature $\Delta T$ of the source, increasing the bit rate will result in a smaller $\Delta T$, as shown in Figure 3.26, giving a lower output SNR.

One solution is to employ indirect modulation, whereby a constant level of spectral power (generated from a bulb having a fixed quiescent DC bias current) is externally pulsed using an optical switch. As a result, this avoids any issues of thermal time constant limitations associated with the transmitter. If a simple optical chopper is employed, the thermal time constant limitation associated with the PIR sensor can
Figure 3.26: Measured instantaneous filament temperature, showing the $\Delta T$ of the source as bit rate increases: (a) 2 bit/s gives larger $\Delta T$; (b) 4 bit/s gives smaller $\Delta T$ [44].

be eliminated.

An experiment was conducted with the use of the Murata IRA-E710ST1 sensor, across a single channel covering the 25 to 50 THz spectral range. Here, the receiver was positioned 1.0 cm from the transmitter and three values of quiescent DC bias current were considered: with $I = 44, 50$ and 60 mA, having estimated spectral radiance peaks at 80, 93 and 108 THz, respectively. The performances with the lowest quiescent DC bias current of 44 mA can be compared with the very first experiment having direct modulation. As shown in Figure 3.27a, with a fixed bit rate of 30 bit/s, the maximum transmission distance was 2 cm. This represents almost a five-fold increase in the transmission distance with three times the bit rate. Moreover, as shown in Figure 3.27b, with a fixed transmission distance of 1.0 cm, the maximum bit rate was 100 bit/s. This represents a ten-fold increase in bit rate with double the distance, when compared with the same conditions using direct modulation.

For most practical applications, an optical chopper is of little use. However, there are other indirect modulation technologies that can be employed. For example, depending on the speed of operation, mechanical shutters (realized using miniature mechanical or even microelectromechanical systems, MEMS [417], technologies) can provide perfect transmission and extinction, although this may present an expensive solution. High-contrast terahertz modulators, based on extraordinary transmission through ring aperture arrays, can be ultra-fast (with expected switching speeds >10
Figure 3.27: Experimental results for different bias currents, with the use of an optical chopper, showing the output voltage from the LNA against: (a) transmission distance with a fixed bit rate of 30 bit/s; (b) bit rate for a fixed distance of 1.0 cm [44].

Gbit/s) and potentially low cost [303], although being based on resonant arrays they will have a narrow instantaneous bandwidth.

3.3.3 Detector thermal time constant

As seen from (3.31), the output voltage from PIR sensors is the function of the modulation frequency. Furthermore, its modulation frequency dependant characteristics are dictated by the thermal ($\tau_T$) and electrical ($\tau_E$) time constants. The modulation frequency dependency of detector responsivity for the Murata IRA-E710ST1 and InfraTec LME-553 devices can be seen in Figure 3.28a and Figure 3.28b, respectively.

Figure 3.28: PIR sensor relative responsivity versus chopping frequency: (a) Murata IRA-E710ST1 [407]; (b) InfraTec LME-553 [408].

As seen from Figure 3.28, the Murata IRA-E710ST1 has a very low 70.7% cut-off modulation frequency at $\sim$1.3 Hz. Indeed, the Murata IRA-E710ST1 sensor was
originally developed for detecting relatively slow movements of warm bodies (i.e., for human motion sensing). As a result, the low modulation frequencies of operation (normally between 1 to 10 Hz) can be realised with a basic technology that is very cheap to manufacture in large volumes. The InfraTec LME-553 has a much larger 70.7% cut-off modulation frequency at ~600 Hz and a relative responsivity that remains almost constant between 10 and 200 Hz, thus more suitable for obtaining higher data rates.

The InfraTec LME-553 sensor element is made from the crystalline pyroelectric material lithium-tantalate (LiTaO₃), which also exhibits piezoelectric properties. As a result, the performance of this sensor will be severely degraded by microphonic effects, from external mechanical vibrations and acoustic pressure waves. Besides the careful mounting for the PIR sensors, one simple engineering solution is to connect two sensors (one blind) directly to the differential inputs of a low-noise operational amplifier. The output from this LNA is then band-pass filtered before the analogue signal reaches the Schmitt trigger. The back-end circuit is shown in Figure 3.29.

Figure 3.29: Back-end electronics for the improved ‘THz Torch’ wireless communications link.

This circuit consists of an instrumentation amplifier INA163 with a voltage gain of
100, which is used to amplify the differential signal from two LME-553 detectors. The amplified signal then passes through a unity-gain band-pass filter with a designed central frequency of 1 kHz. The filtered signal is further threshold detected by the Schmitt trigger, to convert the received signal into the digital form. Two voltage regulators TPS7A4091 and TPS7A3001 are applied to provide ±5 V voltages for the sensors and back-end circuit. Figure 3.30 illustrates the PCB layout of the circuit and the complete receiver module.

Figure 3.30: PCB layout for back-end electronics and picture of the complete receiver module.

With the improved sensors and back-end electronics, much faster data rates can be achieved. The same experimental setup was used, where the transmitter was positioned 1.0 cm from the receiver, to create a LoS wireless link, and the optical chopper was applied to modulate the signal. With the same 44 mA bias current and 2 kbit/s data rate, the output voltage from the optical chopper, LNA and Schmitt trigger are shown in Figure 3.31.

To further demonstrate the performance of the improved single-channel ‘THz Torch’ wireless communications link, the bit error rate (BER) as a function of bit rate and transmission distance is shown in Figure 3.32. In the experiments, an end-to-end binary data stream having $2 \times 10^6$ bits was used. The output from the channel receiver’s Schmitt trigger was recorded by the PicoScope 2205 MSO. This data series was then threshold detected from one centrally positioned sample point per bit (i.e.,
values greater than zero representing logic 1 and less than zero representing logic 0) using the developed MATLAB code, and compared with the originally transmitted data set, to obtain the BER.

The overall BER performance was evaluated for distances from 1 to 5 cm and data rates from 2 to 4 kbit/s. In the case where there was no recorded transmission errors for the $2 \times 10^6$ bit data set, the corresponding BER was considered to be $< 1 \times 10^{-5}$ and, therefore, was not shown in Figure 3.32. In Figure 3.32a it can be seen that, as the transmission range increases, the BER will increase dramatically due to the
spreading loss and misalignment at longer distances. The BER performance will also degrade as the bit rate increases, due to decreased voltage responsivity from the PIR sensor and increased acoustic noise from the optical chopper.

### 3.3.4 Free space attenuation and spreading loss

One of the inherent limitations of operating the high-THz spectral range of 10 to 100 THz is the high free space attenuation. Figure 3.33 demonstrates various spectral windows within the infrared range from 1 to 28 µm (i.e., 300 THz to 10.7 THz) [418], including a low attenuation band from 21 to 40 THz (i.e., 14 µm to 7.6 µm), followed by a high attenuation band from 40 to 56 THz (i.e., 7.6 µm to 5.4 µm), followed by mixed low and high attenuation bands in the rest part of this spectral range. In principle, it may be possible implement a multi-channel ‘THz Torch’ architecture that can exploit the spectral locations of the low attenuation windows.

![Figure 3.33: Atmospheric attenuation in the IR range from 1-28 µm [418].](image)

Since the transmission distances of the proof-of-concept experimental demonstrators are of the order of centimetres, the dominant loss mechanism is beam spreading. For a compact design (whereby five miniature bulbs are assembled into a compact cylindrical package, having an outer diameter of only 8.2 mm), it is not appropriate to employ a much large parabolic mirror. A simple engineering solution is to use a conformal metal reflecting film (e.g., foil), placed behind the five-bulb array, to reduce the spreading loss. Moreover, by including collimating lenses made from low absorption materials, located at both the transmitter and receiver, multi-kbit/s links having transmission distances of the order of metres can be expected.
Collimating lenses made from potassium bromide (KBr) can be used to increase the data rate as well as the transmission range. KBr material has excellent transmission properties, with a minimum transmission level of 90% from 15 to 750 THz (i.e., 20 µm to 0.4 µm), as shown in Figure 3.34 [419].

![Figure 3.34: Transmittance for KBr from 0.2 to 30 µm [419].](image)

In the experiments, two plano-convex lenses with a diameter of 1.5 cm were applied for both the transmitter and receiver. Since the focal length \( f = 1.6 \text{ cm at } 5 \mu\text{m} \), the transmission range in the measurements varies from 5 to 17 cm and the bit rate was adjusted from 2 to 4 kbit/s. Figure 3.35 shows the measured output voltage from the optical chopper, LNA and Schmitt trigger for a 5 cm distance, when the transmitter was biased by a 44 mA current and with 2 kbit/s data rate.

![Figure 3.35: Output voltages from the chopper, LNA and Schmitt trigger for 44 mA bias current, 2 kbit/s data rate and 5 cm distance with the use of two KBr lenses.](image)
The BER performance, as a function of both the transmission distance and the bit rate, was further demonstrated, as shown in Figure 3.36. The BER experiments were performed using the same methodology as described in the last subsection. As seen from Figure 3.36a, with the use of KBr collimating lenses, the transmission range is significantly improved. Wireless communications links with BERs $< 10^{-5}$ were obtained for distances $> 10$ cm at a 80 mA bias current. With the same bias current and 5 cm distance, the maximum bit rate was $> 3$ kbit/s for which no bit error was found for the $2 \times 10^6$ bits transmitted data. As discussed before, data rates are further limited by the responsivity of the PIR sensor and the rotating speed of the optical chopper.

![Figure 3.36: BER performance for the improved single-channel 'THz Torch' wireless link with two KBr lenses: (a) BER against distance with a fixed data rate of 2 kbit/s; (b) BER against data rate for a fixed distance of 5 cm.](image)

Finally, the maximum bit rates (no bit error is found for the $2 \times 10^6$ bits data) for the improved single-channel 'THz Torch' wireless link, both with and without KBr collimating lenses, are demonstrated. Figure 3.37 shows the maximum bit rate, as a function of transmission distance, at different bias currents from 44 to 80 mA. It is observed that by applying KBr collimating lenses, both the maximum bit rate and transmission range have been improved. Note that KBr lenses were used here, rather as a proof-of-concept solution (as it is hygroscopic). In future, more suitable materials (e.g., plastics), which have good transmittance, are stable and low cost should be employed for beam collimating and focusing.
Figure 3.37: Maximum bit rate for the improved single-channel ‘THz Torch’ wireless link, both with and without two KBr lenses.

3.3.5 Bulb glass envelope absorption

COTS incandescent light bulbs have a hermetically-sealed envelope, to prevent the hot tungsten filament from oxidizing in air. The envelope material (e.g., fused silica or soda lime glass) has a high spectral transmittance in the optical and NIR spectral ranges, but this is far from true at longer wavelengths, as shown in Figure 3.9. As discussed before, the Eiko 8666-40984 bulb has an average glass envelope thickness of \( \sim 350 \) \( \mu m \). Therefore, the bulb's glass envelope is expected to contribute significantly to the overall transmission path loss, especially in 10 to 60 THz range. In this part of spectrum, a secondary radiation mechanism dominates, due to increased temperature of the glass envelope.

One method of minimizing the glass envelope loss is to selectively remove its end section and replace it with a more transparent window material. For example, Zinc selenide (ZnSe) has a relatively low absorption coefficient across a wide spectral range, from 16 to 600 THz (i.e., 19 \( \mu m \) to 0.5 \( \mu m \)). Furthermore, unlike KBr materials, ZnSe is also non-hygroscopic and stable for practically all user environments. As a result, it has been used in optical component (e.g., windows, lenses and beamsplitters) for pyrometry applications. A Russian company, TYDEX\textsuperscript{®}, supply such materials with broadband anti-reflection (BBAR) coatings. Figure 3.38a gives the spectral transmittance of a CVD-ZnSe window material with BBAR coating for
the 25 to 100 THz range [420]. Synthetic (CVD-grown) diamond also has a high transmittance (\(\sim 70\%\)) from UV to far-IR ranges, as shown in Figure 3.38b [421]. Although not having a flat and high transmittance across the entire 10 to 100 THz region, some plastics have transmission windows within part of the spectrum of interest. For example, polytetrafluoroethylene (PTFE, Teflon) can offer several transmission windows, including 1 to 7 \(\mu\)m, 8.5 to 12.5 \(\mu\)m and 22 to 45 \(\mu\)m, as shown in Figure 3.38c [422]. Other plastics such as polypropylene (PP) and polyethylene (PE) also have wide transmission windows in the IR bands. More importantly, they are more cost-effective when compared to KBr lenses. By applying materials with high transmittances within the 10-100 THz spectral range, the output power, thus the bit rate and transmission distance, are expected to be further increased.

Figure 3.38: Spectral transmittance of: (a) CVD-ZnSe with BBAR coating for the 25-100 THz range [420]; (b) CVD-diamond for the UV to IR ranges [421]; (c) 100 \(\mu\)m thick PTFE film [422]; (d) Polyethylene in the IR region [423].
3.4 Conclusions

In this chapter, the basic single-channel ‘THz Torch’ architecture was presented. After introducing the basic components and subsystems, the first proof-of-concept single-channel demonstrator was implemented. However, only very low data rates (∼10 bit/s) and short transmission distance (∼1 cm) have been achieved. Fundamental limits of this technology were thus analysed, in order to further improve the performance of this system. By understanding the physical limitations, such as thermal time constants of sources and detectors, spreading loss, and high absorption from the glass envelopes, engineering solutions were proposed and verified experimentally. With these improvements, the maximum bit rate and the transmission range were increased to >2 kbit/s and >10 cm, respectively.

Single-channel ‘THz Torch’ concept can be enhanced by implementing multi-channel multiplexing schemes, for further increasing data rate and/or the level of security. The multi-channel systems use pre-defined channels within the 10 to 100 THz; thermal power in each channel will be independently pulse-modulated, transmitted and detected. Chapter 4 will introduce the multi-channel ‘THz Torch’ concept, demonstrate the proof-of-concept working demonstrators and evaluate the system’s integrity.
4 Multi-channel ‘THz Torch’

Architectures

4.1 Introduction

In communications systems, multiplexing schemes can offer important benefits, including increased overall end-to-end data rates (with band-limited channels), increased robustness to interference and jamming (both natural and manmade) and enhanced protection from interception, which are essential physical layer attributes for secure communications [40].

The ‘THz Torch’ concept can be easily extended using multiplexing schemes, by partitioning thermally-generated spectral power into pre-defined frequency channels; the energy in each channel is then independently pulse-modulated. For multi-channel operation, a number of standard or bespoke filters can be used to create filter banks for implementing different multiplexing schemes. The bandwidth, selectivity and transmittance of the associated channel filters should be carefully chosen, so as to allow roughly equal levels of energy transfer through each channel.

In this chapter, the multi-channel ‘THz Torch’ concept will be presented first, following by the frequency division multiplexing (FDM) and frequency-hopping spread spectrum (FHSS) multiplexing schemes implementation details. Both FDM and FHSS demonstrators will be verified experimentally, and their BER performance will be shown. Finally, the integrity of this multiplexing system will be evaluated by introducing four different jamming, interception and channel crosstalk experiments.
4.2 ‘THz Torch’ Frequency Division Multiplexing System

4.2.1 Basic architecture for ‘THz Torch’ FDM scheme

To increase the end-to-end bit rates beyond those of individual transmitter-receiver pair channels, the ‘THz Torch’ concept can be enhanced by implementing a FDM scheme, while still maintaining its low cost advantage [40].

Here, the frequency spectrum is divided up into $N$ non-overlapping frequency bands within the 10 to 100 THz region. The end-to-end serial data stream is converted into $N$ parallel streams, where data is simultaneously sent over the associated free space channels at the maximum bit rate that the existing enabling technologies can support. After detection, amplification and digitization at the channel receivers, the $N$ output parallel data streams are multiplexed to reconstruct the original transmitted serial data stream; having an overall data rate that is $N$ times that supported by each channel. A simple 4-channel ‘THz Torch’ FDM scheme is illustrated in Figure 4.1 [45]. In this multiplexing system, each individual channel has the same architecture as the single-channel ‘THz Torch’ technology.

Figure 4.1: Basic architecture for ‘THz Torch’ FDM scheme [45].
4.2.2 First proof-of-concept FDM demonstrator

The first proof-of-concept ‘THz Torch’ FDM system demonstrator was reported in [40]. Here, four channels were defined within the thermal infrared region from 15 (assumed) to 90 THz. In this work, four 1 mm thick optical coating filters, sourced from Northumbria Optical Coatings Ltd. [424], are employed to define the non-overlapping frequency bands within the thermal infrared. The bandwidth, selectivity and transmittance for each channel filter are chosen to create approximately equal levels of band-limited blackbody radiation within each transmission channel. Measured transmittances for each channel filter are given in Figure 4.2, from 20 to 100 THz, and associated specifications for each of these four channel filters are listed in Table 4.1 [424].

Figure 4.2: Measured transmittances for transmission channel (A, B, C and D) filters from Northumbria Optical Coatings Ltd. [45].

<table>
<thead>
<tr>
<th>Specifications</th>
<th>Channel</th>
<th>Northumbria Optical Coatings Ltd. Stock Codes</th>
<th>Cut-off (THz)</th>
<th>Cut-on (THz)</th>
<th>Transmittance (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>B</td>
<td>SWBP-6177-000111 42 [7.059]</td>
<td>57 [5.295]</td>
<td>~84.2</td>
<td></td>
</tr>
<tr>
<td></td>
<td>C</td>
<td>SWBP-4596-000070 60 [5.004]</td>
<td>72 [4.188]</td>
<td>~75.7</td>
<td></td>
</tr>
</tbody>
</table>

Table 4.1: Optical properties of COTS channel filters [424].
The first proof-of-concept ‘THz Torch’ FDM system employed transmitter (with Eiko 8666-40984) and receiver (with Murata IRA-E710ST1) similar to those used for the first single-channel demonstrator shown in Section 3.2.3. Four channel transmitter-receiver pairs were mounted face-to-face to create line-of-sight transmission links. The separation distance between the channel transmitter-receiver pairs was fixed at a distance of 1.0 cm (i.e., double the distance of the first single-channel experiment). Unlike the first proof-of-concept single-channel ‘THz Torch’ demonstrator, where the bulbs were directly modulated electronically, indirect modulation using an optical chopper was adopted for simplicity, as seen in Figure 4.3. The DC bias current was 44 mA, to give the same peak spectral radiance at ~80 THz.

![Assembled channel transmitter array behind the optical chopper (left) and receiver array (right) [40].](image)

It is worth mentioning that the proximity of adjacent channel transmitter-receiver pairs may have to be considered; for example, in order to avoid adjacent channel interference. This is not important in this particular application, as there are a number of simple solutions that can be considered (e.g., employing collimating lenses, ensuring band separation and increasing filter selectivity). However, for this particular experimental setup, the adjacent channel transmitter-receiver pairs were spatially separated in order to allow them to be completely shielded by the blades of the 12 cm diameter optical chopper, as shown in Figure 4.3.

The maximum bit rate for each individual channel was measured to be 20.6 bit/s, as shown in Figure 4.4. It was found that the recovered signal from the channel
transmitter-receiver pair using filter SWBP-3685-000091 was distorted if the chopping frequency is increased further. This is because the bandwidth and/or transmittance for this particular channel filter were smaller than desirable. The maximum effective bit rate for this ‘THz Torch’ FDM system demonstrator was achieved after combining the parallel outputs from the four channel transmitter-receiver pairs into one serial output bit stream. In this experiment, as expected, the maximum end-to-end bit rate was measured to be 82.4 bit/s. This represents an eight-fold increase in bit rate, and over twice the distance, when compared to first single-channel experiment shown in Section 3.2.3.

Figure 4.4: Output bit steams from the four channel transmitter-receiver pairs, each having a bit rate of 20.6 bit/s [40].

4.2.3 Improved FDM demonstrator

An improved FDM demonstrator was implemented by using the faster PIR sensor InfraTec LME-553 and new back-end electronics shown in Section 3.3.3. To perform the OOK modulation of the incoherent band-limited blackbody radiation from each channel individually, unlike in the first FDM demonstrator where an optical chopper was applied, here, for convenience, four customized choppers were employed – each independently driven by their stepper motor (SY35ST36-1004A). A single microcontroller (LPC1768) was used to control all four stepping drivers (DM320C). The end-to-end binary data stream was generated in software (MATLAB) and then loaded into the microcontroller. Band-limited output power from each transmitter was transmitted through a slot in the chopper if the data bit was at logic 1 and blocked by a blade if the data bit was at logic 0. The four transmitter-receiver pairs are spatially aligned to create non-interfering (low channel crosstalk) short-range
LoS wireless links. In all instances, the distance between the chopper blades and the transmitter’s filter was fixed at 5 mm, while the distance between the chopper blades and the channel receiver was allowed to vary [45].

A maximum channel data rate of 640 bit/s was achieved with the existing non-optimal demonstrator setup; the speed is limited by the bespoke mechanical chopper, operating at 192 rpm. This corresponds to an effective end-to-end data rate of 2,560 bit/s for the complete multiplexing system. When compared to the first proof-of-concept FDM demonstrator [40], reasons for the ×32 improvement in data rate include the use of (1) faster PIR sensors, (2) noise cancelling using dual-PIR sensors and (3) improved low noise back-end electronics. It is easy to see that by implementing FDM scheme, each channel receiver’s detector is operating at 320 Hz modulation frequency, which is lower than its 70.7% cut-off modulation frequency at ∼600 Hz, as shown in Figure 3.28b; while the effective overall data rate is >2 kbit/s. This is another advantage with the FDM scheme: each channel receiver can operate at or close to its optimal modulation frequency, giving a larger voltage responsivity of the sensor.

To compare channel quality with such multiplexing schemes, BER analysis was performed. In these experiments, an end-to-end binary data stream having $2 \times 10^6$ bits was used. With the FDM scheme, $0.5 \times 10^6$ bits were sent over each channel. The analogue output from the channel receiver’s Schmitt trigger (having voltage levels of almost ±5 V) was input to the microcontroller. The microcontroller has an internal analogue-to-digital converter, to sample and digitally record the quantized levels - for later threshold detection from one centrally positioned sample point per bit (i.e., values greater than zero representing logic 1 and less than zero representing logic 0). The received data set was then compared with the originally transmitted data set, to obtain the BER. The overall BER performance was evaluated for channel transmitter bias currents from 44 to 80 mA and distances from 1 to 5 cm.

Figure 4.5 shows the measured overall BER for the improved 4-channel ‘THz Torch’ FDM system [45]. In the case where there was no recorded transmission errors for the $2 \times 10^6$ bit data set, the corresponding BER was considered to be $<10^{-6}$ and,
therefore, are not shown in Figure 4.5. For a channel transmitter (i.e. incandescent bulb array) bias current of 44 mA, the BER was measured to be $\sim 2 \times 10^{-4}$ over a distance of 1 cm. As the range increases, the BER becomes progressively worse, due to the spreading loss of the radiated power, as no collimating lenses are employed in this experiment. For an 80 mA bias current, no errors were detected in transmission for distances up to 2.5 cm; and the characteristic BER curve can then be seen at distances beyond 3 cm.

![Figure 4.5: Measured overall BER for the improved 4-channel FDM system versus free space transmission distance for different channel transmitter bias currents [45].](image)

The inset of Figure 4.5 shows the measured eye diagrams for each of the four channels, labelled in accordance with the channel filter responses given in Figure 4.2, having a bias current of 80 mA and a range of 3 cm. It can be seen that the eye is wide open with channels A and D, with the former being slightly noisier due to exposure to more ambient noise seen within this channel. Channel B has an almost closed eye, which dominates the overall BER performance of the complete system, due to having the lowest band-limited output power level. Fortunately, these eye diagrams can be adaptively controlled, by changing the bias currents for each channel, to equalise the output voltage between each channel receiver; thus, optimising the BER performance of the complete multiplexing system.
4.3 ‘THz Torch’ Frequency-hopping Spread Spectrum System

4.3.1 Basic architecture for ‘THz Torch’ FHSS scheme

In addition to FDM, a frequency-hopping spread spectrum (FHSS) scheme can also be implemented to further enhance immunity to detection, interception and interference, for secure applications. A simple 4-channel ‘THz Torch’ FHSS architecture is illustrated in Figure 4.6 [45]. With this FHSS scheme, the end-to-end serial data stream is transmitted into the same channels as for the FDM system, but only within one channel at any time – dictated by pseudo-random channel allocation. Although there is no advantage in the overall end-to-end data rate for the FHSS scheme, there can be a significant enhancement to security.

Figure 4.6: Basic architecture for ‘THz Torch’ FHSS scheme [45].

4.3.2 First proof-of-concept FHSS demonstrator

For the first proof-of-concept FHSS demonstrator, each channel has the same pre-defined frequency bands and hardware, as used in the improved FDM experiments. With this FHSS demonstrator, slow frequency hopping (SFH) was employed to transmit a 1 kbit data packet through individual channels, within a single hop.
For convenience, a predetermined pseudo-random hopping pattern was applied to both the transmitter and synchronized receiver, to establish the secure end-to-end communications link. In the FHSS experiments, 2,000 packets were transmitted and the received packets were then analysed, resulting in a total $2 \times 10^6$ bit data. On average the same number of bits ($\sim 0.5 \times 10^6$) was sent over each channel.

Having a 640 bit/s channel data rate which is limited by the chopper speed, the measured overall BER for the first proof-of-concept FHSS working demonstrator is shown in Figure 4.7. As expected, the overall BER performance is similar to that for the improved FDM scheme, as the same transmitter and receiver hardware are used in both experiments. In general, however, a slightly better BER performance was observed with FHSS, as only one of four channel links is operating at any time and, therefore, there is no interference due to crosstalk between channels (e.g., through direct electromagnetic or mechanical coupling). The results for an experiment to investigate channel crosstalk, due to direct electromagnetic coupling, will be given in Section 4.4.

![Figure 4.7: Measured overall BER for the first proof-of-concept 4-channel FHSS system versus free space transmission distance for different channel transmitter bias currents [45].](image)
4.4 Physical Layer Multi-channel System Integrity Evaluation

In the previous sections, both FDM and FHSS schemes have been implemented for the 4-channel ‘THz Torch’ systems. To test the integrity of such multi-channel thermal infrared communications system using engineered blackbody radiation, four different experiments, including two jamming, one interception and one channel crosstalk measurements, will be introduced in this section [45].

4.4.1 Jamming experiment #1

In the first jamming experiment, an additional pulse-modulated thermal source (i.e. jamming signal) was introduced to each of the four channels in turn, to act as an unwanted jammer to one of the channel receivers. Here, the thermal source for the jammer employs the same five-bulb array as that for the channel transmitter. However, while having an identical incandescent bulb array source to that of a channel transmitter, the optical filter is removed for it to have a spectral bandwidth that covers all channels. Moreover, since the filter at the jammer is removed, it will have a 5 dB increase in the level of output power for the same bias current within any of the channels; the 40% minimum aperture blockage and 72-84% transmittances from the filter are avoided. However, at the channel receiver, the received signal power from the jammer will be attenuated, due to the off-axis angle of incidence.

In this experiment,

1. The channel transmitter is placed at a fixed position (0° from normal incidence and 3 cm distance to the PIR detector) and has a fixed bias current of 80 mA and voltage of 2.484 V (giving a DC power of 199 mW for each bulb, corresponding to a total channel transmitter input DC power of 994 mW).

2. With the large bespoke mechanical choppers, it is difficult to introduce an extra chopper to change the modulation frequency of the jamming signal. Therefore, the same chopper is used for both the channel transmitter and jammer. As a result, the jammer operates with its maximum effectiveness; having the same
320 Hz modulation frequency and 50% duty cycle, while being approximately 180° out-of-phase with the channel transmitter.

3. The jammer is at a fixed location (45° from normal incidence and 3 cm distance to the channel receiver’s detector), and the bias is adjusted from 0 to 80 mA.

The effect of the jammer is investigated, one channel at a time, and the resulting overall BERs for both FDM and FHSS systems were recorded for different interfering noise amplitudes (i.e. jammer bias currents), as shown in Figure 4.8. As expected, the overall BER integrity is similar for both FDM and FHSS schemes. Furthermore, it can be seen that Channel D is the most robust to the effects of jamming, while Channel B is the most sensitive. The reason is that Channel D is the nearest to and Channel B is further away from the spectral radiance peak of the primary source of radiation. As a result, the jamming signal has a greater influence on the weaker signal received from the Channel B transmitter. It should be noted that Channel A is more robust than Channel B, even though its spectral band is furthest away from the spectral radiance peak, because it benefits from a secondary source of radiation from the glass envelope of the five-bulb array. One simple solution is to increase the channel transmitter bias currents, by different amounts for the various channels, to effectively balance the overall performance of the complete multiplexing system.

![Figure 4.8](image)

Figure 4.8: Experiment for the jamming of different channels for the worst-case condition of the jammer modulation frequency being equal to the channel modulation frequency of 320 Hz: (a) measured overall BER for the 4-channel FDM scheme versus jammer bias current; (b) measured overall BER for the 4-channel FHSS scheme versus jammer bias current [45].
4.4.2 Jamming experiment #2

For the second experiment, instead of using the same five-bulb array, the jammer is made from a single, more expensive, miniature thermal source (commercially available blackbody infrared radiation emitter INTX 17-0900 [425]), which can be pulsed modulated up to a frequency of 100 Hz (i.e. without the need for the jammer to use a mechanical chopper).

In this experiment,

1. The channel transmitter is placed at the same position with the same 80 mA bias current as the jamming experiment #1.

2. The jammer (INTX 17-0900) is at the same location as with the previous jamming experiment, with the same duty cycle of 50%, and has a bias current of 117 mA and voltage of 5.9 V (corresponding to a total input DC power of 690 mW).

3. The jammer’s modulation frequency is investigated from 0 to the maximum modulation frequency of 100 Hz.

With a single INTX 17-0900 thermal emitter, having 70% of the maximum input DC power when compared to the previous jamming experiment, the overall output spectral intensity will still be higher than that from the five-bulb array, due to a much higher emissivity (∼0.8) of the emitter. However, as can be seen from the overall BER results shown in Figure 4.9, both multiplexing schemes are less sensitive to the jamming signal; the BER increases as the modulation frequency approaches the 320 Hz centre frequency of the baseband (BB) BPF. In the existing channel receiver design, a 4th order Sallen-Key Butterworth filter has been implemented. With either a Chebyshev or elliptical filter of the same order or a higher order filter implementation, the integrity of the multiplexing schemes can be further improved; such that the jammer would need to operate much closer to the operating modulation frequency of the channel transmitter to have a significant effect. This sensitivity to modulation frequency can be exploited by a channel receiver that can be reconfigured
for more than one modulation frequency, introducing more resilience to mitigate the effects of a jamming signal.

![Figure 4.9: Experiment for the jamming of different channels with a commercially available infrared emitter: (a) measured overall BER for the 4-channel FDM scheme versus jammer modulation frequency; (b) measured overall BER for the 4-channel FHSS scheme versus jammer modulation frequency [45].](image)

4.4.3 Interception experiment

With the third experiment, the possibility of interception by an intruding receiver will be investigated. Here, for convenience, an intruding receiver is located in the horizontal plane of the established communications link. The channel transmitter has the same fixed location and bias current as with the previous jamming experiments, and the intended channel receiver has been removed in this experiment. The output RMS voltage is recorded at angles of interception from -90° to 90° in the horizontal plane.

The measured output RMS voltage $V_{rms}$ from the intruding receiver positioned at various angles from normal incidence is shown in Figure 4.10, for each of the four channels. It can be seen that the output voltages, at normal incidence, correspond in amplitude to the BER performances for each channel. As the off-axis angle increases the output voltage rolls off. With the existing demonstrator setups, having angles of interception within ±45°, it is theoretically possible to intercept the transmitted channel data. However, in practice, since the recommended communications link is confined to a free space transmission distance $R \lesssim 1$ m, the size of both the
intended channel and intruding receivers limit how close the interceptor can be positioned to normal incidence. Furthermore, with the use of collimating lenses at the channel transmitters, the beamwidth of the transmitted power would be confined and this would dramatically reduce the likelihood of interception. To get closer to normal incidence, the interceptor would have to be located at a further distance from the intended channel receiver; however, the detected power advantage would be offset because of the $1/R^2$ spreading loss dependency and by the possibility of signal blockage from the intended channel receiver.

Figure 4.10: Experiment for the interception of different channels showing the measured output RMS voltage from the intruding receiver positioned at various angles [45].

4.4.4 Channel crosstalk experiment

With the last experiment, the effects of channel crosstalk are quantified. Each channel receiver is separately aligned, at normal incidence, to the four channel transmitters; this represents the worst-case condition, as there is no spatial displacement introduced between the different channels. In this experiment, the channel transmitter has the same fixed location and bias current as with the jamming and interception experiments. The measured output RMS voltage $V_{rms}$ for each channel receiver can be seen in Figure 4.11a for different transmit channels. It can be seen that the output voltages, along the main diagonal, correspond in amplitude to the BER performances for each channel. The off-diagonal values represent worst-case chan-
nel crosstalk leakage. This can be represented by the calculated channel crosstalk rejection, shown in Figure 4.11b.

![Experiment for worst-case channel crosstalk](image)

**Figure 4.11:** Experiment for worst-case channel crosstalk: (a) output RMS voltage $V_{rms}$ for each receive (Rx) channel for different transmit (Tx) channels; (b) channel crosstalk rejection [45].

In the existing multiplexing scheme implementations, the high levels of out-of-band rejection seen in Figure 4.2 are sufficient to provide high levels of crosstalk rejection. However, since crosstalk rejection is defined by the ratio of received power from the wanted channel $P_{RX-wanted}$ to that from an unwanted channel $P_{RX-unwanted}$, if $P_{RX-wanted}$ is low for a fixed value of $P_{RX-unwanted}$, then the calculated crosstalk rejection will be artificially low. With this experiment, the worst-case crosstalk rejection of 6.7 dB exists when the channel B receiver detects the weak $P_{RX-wanted}$ and the significant $P_{RX-unwanted}$ from channel A. While the output power from the channel D transmitter is higher than that for channel A, the level of out-of-band rejection is greater at channel D because of its larger spectral separation from channel B. Calculated values for crosstalk rejection will increase by increasing the source bias current for channel B.

### 4.5 Conclusions

This chapter has introduced an ultra-low cost multi-channel system for implementing a short-range wireless communications link using engineered blackbody radiation within the thermal infrared. Both FDM and FHSS multiplexing systems have been demonstrated, and the BER performance of these systems have been evaluated. At a
1 cm distance and a 44 mA bias current, the end-to-end data rate was increased from only 80 bit/s for the first-of-concept FDM system to \( \sim 2.6 \text{ kbit/s} \), where BER was still \( <10^{-3} \) for the improved FDM working demonstrator. The BER performance can be further improved by applying higher channel transmitter bias currents. For example, for a 80 mA bias, the BER was \( <10^{-6} \) for the same distance and increased to \( \sim 10^{-3} \) over a 3 cm transmission range. The first FHSS working demonstrator, having a 640 bit/s data rate and a slightly improved overall BER performance, was also shown experimentally. The performance of these system demonstrators can be easily enhanced through the independent control of the channel transmitter bias currents. Moreover, adopting forward error correction (FEC) coding algorithms, which has a BER limit of \( 10^{-3} \) [400], can further reduce the BER for these systems.

It also has been shown that, in order to undermine the inherent immunity to interference and interception, both the jammer and intruding receiver, respectively, must be designed to meet the same channel spectral bandwidth and operating modulation frequency specifications. With the largely unregulated part of thermal infrared band, a communications link can be designed to a bespoke set of specifications that may not be known, \textit{a priori} to the jammer/interceptor; while reconfigurable techniques can be introduced to further enhance the integrity of \textit{THz Torch} systems.

With both FDM and FHSS demonstrators, the data rate was ultimately limited by the rotational speed of the mechanical choppers, while the transmission distance is limited by spreading loss and channel transmitter bias currents. With the former, faster external modulators can be employed (e.g., spatial light modulator). With the latter, collimating lenses can be used to increase transmission distances by an order of magnitude. As shown in section 3.3.4, single-channel \textit{THz Torch} wireless communications links have demonstrated data rates of \( >2 \text{ kbit/s} \) over a distance of \( >10 \text{ cm} \), by employing two KBr plano-convex lenses. By exploring a diverse range of methods, significant enhancements to both data rate and free space transmission distance can be expected.
5 Link Budget Analysis for ‘THz Torch’ Systems

5.1 Introduction

Until very recently, there has been little in the way of enabling technologies within thermal infrared (ca. 10 to 100 THz) part of the frequency spectrum to support wireless communications, offering opportunities for developing secure communications within this largely unregulated part of the electromagnetic spectrum. In the previous chapters, both single and multi-channel thermal infrared wireless communications systems have been demonstrated for short-range applications.

To assess the performance of thermodynamics-based links, accurate signal and noise power link budget analysis is required. In this chapter, a detailed power link budget analysis for the ‘THz Torch’ thermal infrared wireless communications system, using engineered blackbody radiation, will be presented. Here, a generic 4-channel system, with a 640 bit/s data rate per channel and a transmission range of 1 cm is investigated as a function of channel transmitter bias currents. The calculated output signal and noise voltages from each of the uncorrelated channel receivers are compared with measured values. Furthermore, the signal to noise ratio (SNR) of each channel will be measured, showing good agreement with calculated results.

The overall signal power link budget breaks down into three blocks (transmitter, free space channel and receiver), as shown in Figure 5.1. Here, $I_{\text{filament}}$ is the radiant intensity of the tungsten filament; $T_G$ is the power transmittance of the bulb’s glass envelope; $I_{\text{primary}}$ and $I_{\text{secondary}}$ are the radiant intensity from primary and secondary source of radiation, respectively; $T_{THz,BPF}$ is the average power
transmittance of the optical coating band-pass filter (BPF); $I_{TX}$ is the total radiant intensity from the channel transmitter’s source; $L_{FS}$ is the free space channel loss; $P_{RX}$ is the power received by the sensor; $R_V$ is the voltage responsivity of the PIR sensor; $u$ is the output RMS voltage from the detector; $A_{BB\text{-}LNA}$ is the voltage gain of the baseband (BB) LNA; $A_{BB\text{-}BPF}$ is the voltage gain of the baseband BPF; and $V_{out}$ is the output RMS voltage at the channel receiver.

![Diagram](image)

Figure 5.1: Overall signal power link budget representation for the thermodynamics-based ‘THz Torch’ wireless link.

### 5.2 Transmitter Output Radiant Intensity

In Section 3.2.1, the basic transmitter subsystem has been characterised. The radiant intensity from primary and secondary source of radiation mechanisms are given by (3.10) and (3.23), respectively.

The overall total radiant intensity from the transmitter, with contributions from both primary and secondary radiation mechanisms, can be expressed by

$$I_{TX}(T) = \left[I_{primary}(T_{filament}) + I_{secondary}(T_{glass})\right] \cdot \left(\frac{A_{THz\text{-}BPF}}{A_{eff_{glass}}}\right) \cdot T_{THz\text{-}BPF} \ [\text{W/sr}] \ (5.1)$$

where $A_{THz\text{-}BPF} = 3.7 \times 4.7 \ \text{mm}^2$ is the aperture size for the THz BPFs; $A_{eff_{glass}} = A_{eff_{glass\text{-}high}} + A_{eff_{glass\text{-}low}} \approx 43.92 \ \text{mm}^2$, giving a transmitter aperture efficiency of $A_{THz\text{-}BPF}/A_{eff_{glass}} \approx 40\%$. The four channel filters are the same as used in previous multi-channel experiments. Note that the transmittances of these filters are wavelength dependent, and should be calculated within the integration process. However, using the spectral ranges and average power transmittances shown in Table 4.1, similar results can be obtained. Therefore, the average transmittance for each
filter is used throughout the calculation.

### 5.3 Free Space Loss

Free space losses include both spreading losses and atmospheric attenuation. With the latter, clear conditions are assumed throughout this chapter. With the former, if one considers the transmitter to be a point source Lambertian radiator, which radiates uniformly in all directions, the free space loss \( L_{FS} \) can be calculated by applying Lambert’s cosine law to give

\[
L_{FS} \approx T_{ATMOSPHERIC} \cdot \frac{A_s}{R^2} \cos \theta \text{ [sr] \quad (5.2)}
\]

where \( T_{ATMOSPHERIC} \) the average atmospheric power transmittance across the channel bandwidth for a specific propagation distance (1 cm in this case); \( A_s = 3 \text{ mm}^2 \) is the surface area of the detector’s sensing element; \( R = 13.8 \text{ mm} \) is the total transmission distance in free space between the point source and detecting element of the PIR sensor (i.e., 1 cm distance between transmitter and receiver, plus the 2 mm space between the bulbs and the optical coating filter, plus the 1.8 mm distance between the receiver’s optical coating filter and sensing element); \( \theta \) is the angle between the central line of sight and the offset from the surface normal of the detector (zero in this case).

For short-range communications, \( T_{ATMOSPHERIC} \) is expected to be unity. To confirm this, Line-By-Line Radiative Transfer Model (LBLRTM) simulation software is used [426, 427]. This is an accurate, efficient and highly flexible model for calculating spectral transmittance and radiance; representing the best approach for calculating the atmospheric attenuation for each of the four channels. At the shorter wavelengths (\( \gtrsim 75 \text{ THz}, \text{i.e.,} \lesssim 4 \mu\text{m} \)) considered in this study, the influence of scattering from atmospheric aerosol (dust, smoke, etc.) might be expected to increase and this process is not accounted for in LBLRTM. However, over the shortest paths considered here, this effect should not be significant.

LBLRTM extracts absorption line parameters from the HITRAN line database,
as well as additional line parameters from other sources. It incorporates the water vapour continuum absorption model, as well as continuum extinctions for carbon dioxide, oxygen, nitrogen and ozone [428]. Since only generic atmospheric attenuation is being investigated in this section, a horizontally homogeneous atmospheric profile based on the values given at sea-level for the US Standard 1976 Model [429] is assumed, where temperature and atmospheric pressure are 288.15 K and 101.325 kPa, respectively. The 1976 US Standard atmosphere was defined to be representative of annual average conditions experienced at mid-latitudes. Within this particular standard model, the percentage concentrations of several gas species are listed in Table 5.1. Note that in Table 5.1 Other represents radiatively inactive gases (i.e., they do not absorb infrared radiation, since they do not possess an electric or magnetic dipole moment), dominated by nitrogen (N$_2$) and with contributions from argon (Ar), etc.. Moreover, since the effects of atmospheric ionization from solar radiation is not an issue at the wavelengths in this applications, they are thus not considered in the atmospheric modelling.

<table>
<thead>
<tr>
<th>Gas Species</th>
<th>Concentrations (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Other</td>
<td>78.29228034 (assumed)</td>
</tr>
<tr>
<td>O$_2$</td>
<td>20.9</td>
</tr>
<tr>
<td>H$_2$O</td>
<td>0.7745</td>
</tr>
<tr>
<td>CO$_2$</td>
<td>0.033</td>
</tr>
<tr>
<td>CH$_4$</td>
<td>0.00017</td>
</tr>
<tr>
<td>N$_2$O</td>
<td>0.000032</td>
</tr>
<tr>
<td>CO</td>
<td>0.000015</td>
</tr>
<tr>
<td>O$_3$</td>
<td>0.00000266</td>
</tr>
</tbody>
</table>

The calculated banded-average atmospheric transmittance, as a function of the propagation distance from 1 mm to 1 km, for our 4-channel ‘THz Torch’ system is given in Figure 5.2 [50]. Figure 5.3 shows the atmospheric transmittances across the spectrum from 10 to 100 THz, with a resolution of 300 MHz, at various distances. It can be confirmed that in principle, a single-channel ‘THz Torch’ wireless link
operating in Channel D (75-89 THz) can operate up to a range of ~1 km; although in order to minimize the significant spreading losses, collimating lenses would be needed at both the transmitter and receiver. In practice, a realistic transmission range of a few meters can be expected. The transmittances for all channels exceed 83%, with Channel A and D greater than 98%, which is acceptable and can be easily compensated for by increasing the source bias current.

Conversely, Channel B (42-57 THz) has the worst transmittance performance, due to the very high water absorption, limiting the effective range to approximately 1 m with the use of collimating lenses. Between these extremes, Channel C (60-72 THz) is partially affected by carbon dioxide, as is Channel A (15-34 THz) that also suffers from atmospheric absorption from ozone over longer distances. It is interesting to note that banded-average atmospheric transmittance does not follow any simple scaling law with distance. As a result, accurate signal power link budget calculations require the use of atmospheric attenuation modelling software simulations to be performed for a specific bandwidth, time of day, precipitation, site locations (i.e., taking into account transmission path inhomogeneity), etc.. Nevertheless, for a range of only 1 cm, the simulated banded-average atmospheric transmittances for each channel are given in Table 5.2.
Figure 5.3: Transmittance across the spectrum from 10 to 100 THz for a transmission distance of: (a) 1 cm; (b) 1 m; (c) 1 km. [50].

Table 5.2: Simulated banded-average atmospheric transmittance for each channel (A, B, C, D).

<table>
<thead>
<tr>
<th>Channel</th>
<th>A: 15-34 (THz)</th>
<th>B: 42-57 (THz)</th>
<th>C: 60-72 (THz)</th>
<th>D: 75-89 (THz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$T_{ATMOSPHERIC}$ (%)</td>
<td>99.98435</td>
<td>99.58088</td>
<td>99.79324</td>
<td>99.99992</td>
</tr>
</tbody>
</table>

5.4 Receiver Output Signal Voltage

At the receiver, the incident power $I_{TX}(T) \cdot L_{FS}$ passes through an identical THz BPF and the resulting received power $P_{RX}(T)$ at each PIR sensor can be estimated using the following

$$P_{RX}(T) \approx I_{TX}(T) \cdot L_{FS} \cdot T_{THz-BPF} \ [W] \quad (5.3)$$

Since mechanical choppers are employed, with the prototype hardware implementation, power is transmitted sinusoidally and, therefore, $P_{RX}(T)$ is representative of peak values. The LME-553 PIR sensor from InfraTec GmbH was chosen for its
ability to detect incoherent band-limited thermal power, having also an ultra-wide bandwidth, room temperature operation, being ultra-low cost and having a relatively fast response time (~1 ms) [408]. Its output RMS voltage \( u(T) \) is proportional to the incident radiation power \( P_{RX}(T) \), representing a square-law detector operating in its linear region:

\[
u(T) = R_V \cdot \frac{P_{RX}(T)}{2\sqrt{2}} \quad [V]
\]  

(5.4)

where \( R_V \) [V/W] is the voltage responsivity of the PIR detector. In this case, the employed LME-553 PIR sensor is made from LiTaO\(_3\), with an additional black absorption layer. Here, without any optical filtering, the maximum RMS voltage responsivity is quoted as 6,500 V/W (centred at a 100 Hz chopping frequency) [408]. For a speed of 320 Hz, the roll-off value of RMS voltage responsivity is approximately 90% of its maximum value, giving an estimated responsivity of 5,850 V/W.

Considering the voltage gains from the back-end electronics, the output RMS voltage can be expressed as

\[
V_{out}(T) = u(T) \cdot A_{BB,LNA} \cdot A_{BB,BPF} \quad [V]
\]

(5.5)

where \( A_{BB,LNA} = 99.4 \) and \( A_{BB,BPF} = 1.06 \) with the demonstrator.

The calculated output RMS voltages \( V_{out}(T) \) for each channel receiver are shown in Figure 5.4a, as a function of source bias current. As expected, the output voltage increases as the transmitter’s source bias current increases. It was seen in Section 5.3 that there is negligible atmospheric attenuation with any of the channels over this short distance. However, from Figure 5.4a, it can be seen that Channel D exhibits the most significant voltage increase with biasing current. The reason for this is that primary radiation dominates (due to the high power transmittance and low power absorptance of the glass envelope), increasing from 80 to 131 THz, and Channel D is the nearest to the spectral radiance peak, as seen in Figure 5.4b.

Similarly, Channel A benefits from the spectral radiance peak of the secondary radiation source (due to the low power transmittance and high power absorptance of the glass envelope) increasing from 31.7 to 35.6 THz and from 32.3 to 37.9 THz.
Figure 5.4: Source bias current dependency: (a) calculated channel receiver output voltages; (b) calculated primary and secondary radiation spectral radiance peaks; (c) measured channel receiver output voltages.

for low and high temperature regions, respectively. However, Channel B is the least responsive, because it is the furthest away from any spectral radiance peaks and/or associated tails. Again, it can also be seen in Figure 5.4a that primary radiation is more sensitive to low source bias currents, when compared to secondary radiation.

The measured output RMS voltages are given in Figure 5.4c. When compared to predictions, it can be seen that there is good agreement in both trends and values. Note that only measured RMS voltages up to $\sim 2.83$ V were recorded, as the back-end circuit saturates when the peak-to-peak voltage reaches $\sim 8$ V.

There are additional loss mechanisms that have not yet been taken into consideration. For example, the LiTaO$_3$ pyroelectric material with black absorption layer does not have a flat absorptive spectral response across the entire thermal infrared. As a result, the voltage responsivity will be wavelength dependent, with a decreasing value below its cut-off frequency of $\sim 15$ THz [430]. Also, in practice, the transmitter aperture efficiency of 40% does not take into account diffraction effects or any sig-
nificant mechanical misalignments. Finally, specular and molecular reflections have not been included in the atmospheric attenuation calculations. These additional loss mechanisms require detailed numerical CAD modelling, which is beyond the scope of this study. Nevertheless, a good agreement has been achieved between the predicted and measured receiver output signal voltages, for all channels, using the proposed end-to-end signal power link budget analysis.

5.5 Receiver Output Noise Voltage

To determine the signal-to-noise (SNR) performance of the 'THz Torch' wireless communications system, the noise analysis for each channel receiver must also be undertaken. The additive intrinsic noise for each channel can be separated out into two sources: noise from the font-end sensor and noise from the back-end electronics. The block diagram for the complete channel receiver is illustrated in Figure 5.5a.

![Block Diagram of Channel Receiver](image)

**Figure 5.5:** Complete channel receiver: (a) detailed block diagram; (b) noise contributions.

Within each channel receiver, a dual-sensor configuration is employed, where two identical PIR sensors LME-553 are used; one is a dummy sensor that is opaque to the environment, to minimize unwanted microphonic effects caused by mechanical/acoustic vibrations incident to the pyroelectric material. The output from each
PIR sensor will pass through the buffering stage and is then amplified and filtered, before it is threshold detected by the Schmitt trigger (ST). Intrinsic noise will be generated at each stage, and combined with noise contributions from all previous stages. The total noise of the channel receiver is evaluated at the output of the baseband BPF, as shown in Figure 5.5b. Note that without a THz BPF the intrinsic noise performance would not be channel specific. Also, noise contribution from the ST is not included here as signal and noise power are measured at the output of the baseband BPFs.

5.5.1 Pyroelectric infrared sensor noise sources

The pyroelectric sensor transduces not only the useful incident thermal power but also ambient background noise power. Here, the received noise pulse is detected by the resulting surface temperature differences, giving rise to surface charges (due to the pyroelectric effect), which in turn generates a short circuit current. This extremely low current, supplied by the high impedance of the pyroelectric material, is then converted to the required output voltage by an integrated TIA.

With the current mode TIA-based PIR sensor (LME-553) there are a number of intrinsic noise sources, which include: temperature noise $V_{NT}$, due to temperature fluctuation; dielectric noise $V_{ND}$, due to the dielectric loss associated with the pyroelectric material; noise from the input resistance $V_{NR}$; noise from the large feedback resistor $V_{NFB}$; current noise from the TIA’s operational amplifier (op-amp) $V_{NI}$; and voltage noise from the op-amp $V_{NU}$. Each of these noise contributions will now be considered in turn, for completeness, as there is no single reference that covers all possible contributions.

**Temperature noise $V_{NT}$**

When the sensing element is in thermal equilibrium with its surroundings, there will be no output voltage. However, temperature fluctuation will cause a response from
the sensing element, resulting in the following voltage-noise spectral density [414]

\[ V_{NT} = \frac{R_V}{\alpha} \sqrt{4k_B T^2 G_T} \frac{1}{\sqrt{\text{Hz}}} \]  

(5.6)

where \( \alpha \) is the absorptance of the pyroelectric sensing element, which quantifies how much incident thermal power will be absorbed by the material; \( G_T = \frac{H_P}{\tau_T} \) [W/K] is the thermal conductance of the pyroelectric sensing element; \( H_P = c'_p d_P A_s \) [J/K] is the heat capacity; \( c'_p \) [Jm\(^{-3}\)K\(^{-1}\)] is the volume-specific heat capacity; \( d_P \) [m] is the thickness of the pyroelectric sensing element; \( A_s \) [m\(^2\)] is the surface area of the pyroelectric sensing element; and \( \tau_T \) [s] is the thermal time constant of the PIR detector. For the current mode TIA-based PIR sensors, the voltage responsivity \( R_V \) is given by [414]

\[ R_V = \omega_m \alpha A_s p R_{fb} G_T \frac{1}{\sqrt{1 + (\omega_m \tau_T)^2}} \frac{1}{\sqrt{1 + (\omega_m \tau_E)^2}} \frac{1}{\text{[V/W]}} \]  

(5.7)

where \( \omega_m = 2\pi f_m \) [rad/s] is the angular frequency of the modulation signal; \( f_m \) is the modulation frequency; \( p \) [Cm\(^{-2}\)K\(^{-1}\)] is the pyroelectric coefficient of the sensing element; \( R_{fb} \) [\( \Omega \)] is the feedback resistance of the integrated op-amp; \( \tau_E = R_{fb} C_{fb} \) is the electrical time constant for the integrated op-amp; and \( C_{fb} \) [F] is the feedback capacitance.

It can be seen in (5.6) that temperature noise is frequency independent at a constant temperature, and is considered as the only noise source for an idea PIR detector. Therefore, \( V_{NT} \) represents the background limit to the detector’s noise equivalent power NEP, which is defined by the value of incident power for which the SNR of the detector equals unity with a noise bandwidth of 1 Hz. At an ambient room temperature of 300 K, \( G_T = 4\alpha \sigma_s T^3 A_s \) [431], where \( \sigma_s \) [Wm\(^{-2}\)K\(^{-4}\)] is the Stefan-Boltzmann constant and, if one considers a 100% absorptance across all the detectable wavelengths, then

\[ V_{NT} = R_V \sqrt{16k_B \sigma T^5 A_s} = 5.5 \times 10^{-11} \sqrt{A_s} \cdot R_V \frac{\text{[V/Hz]}}{\sqrt{\text{Hz}}} \]  

(5.8)
The NEP and specific detectivity \( D^* \) for an ideal pyroelectric sensor can be obtained from \[432\]

\[
NEP = \frac{V_{NT}}{R_{V}} = 5.5 \times 10^{-11} \sqrt{A_s[cm^2]} [W/\sqrt{Hz}]
\]

(5.9)

\[
D^* = \frac{\sqrt{A_s}}{NEP} = 1.8 \times 10^{10} [cm\sqrt{Hz}/W]
\]

(5.10)

These values represent the upper theoretical bounds for the performance of pyroelectric detectors.

**Dielectric noise \( V_{ND} \)**

A pyroelectric sensing element acts as a dielectric, with an associated Johnson-Nyquist voltage-noise spectral density expressed as \[431\]

\[
V_{ND} = \frac{R_{fb} \cdot \sqrt{4k_B T \cdot \omega_m C_P \tan \delta_P}}{1 + (\omega_m \tau_E)^2} [V/\sqrt{Hz}]
\]

(5.11)

where \( C_P = \frac{\varepsilon_0 \varepsilon_r A_s}{d_p} [F] \) is the electrical capacitance of the pyroelectric sensing element; \( \varepsilon_0 [F/m] \) and \( \varepsilon_r \) are the free space permittivity and dielectric constant of the pyroelectric material, respectively; and \( \tan \delta_P \) is the dielectric loss tangent. It can be seen that this noise contribution is modulation frequency dependent, and proportional to the electrical capacitance, as well as the feedback resistance of the integrated op-amp. This noise contribution is more significant at higher modulation frequencies.

**Input resistance noise \( V_{NR} \) and feedback resistor noise \( V_{NFB} \)**

Both contributions are Johnson-Nyquist noise, generated by the thermal agitation of the charge carriers within the large integrated op-amp feedback \( R_{fb} [\Omega] \) and input \( R_{input} [\Omega] \) resistors at thermal equilibrium. The respective voltage-noise spectral densities are given by \[433\]

\[
V_{NR} = R_{fb} \cdot \sqrt{\frac{4k_B T}{R_{input} \cdot [1 + (\omega_m \tau_E)^2]}} [V/\sqrt{Hz}]
\]

(5.12)
and

\[ V_{NFB} = \sqrt{\frac{4k_B T \cdot R_{fb}}{1 + (\omega_m \tau_E)^2}} \text{[V/√Hz]} \]  

(5.13)

**Op-amp current noise \( V_{NI} \)**

The integrated op-amp also generates noise from within the PIR detector. The voltage-noise spectral density introduced by the associated current noise can be expressed as [433]

\[ V_{NI} = \frac{i_{opamp} \cdot R_{fb}}{\sqrt{1 + (\omega_m \tau_E)^2}} \text{[V/√Hz]} \]  

(5.14)

where \( i_{opamp} \) [A/√Hz] is the equivalent input current-noise spectral density.

**Op-amp voltage noise \( V_{NU} \)**

The voltage-noise spectral density associated with \( V_{NU} \) is due to the equivalent input noise voltage of the integrated op-amp, which can be expressed as [433]

\[ V_{NU} = \frac{u_{opamp} \cdot R_{fb}}{R_{eq}} \sqrt{\frac{1 + (\omega_m \tau_E)^2}{1 + (\omega_m \tau_E)^2}} \text{[V/√Hz]} \]  

(5.15)

where \( u_{opamp} \) [V/√Hz] is the equivalent input voltage-noise spectral density; \( R_{eq} = R_{input} || R_{fb} || R_P \) [Ω] is the equivalent resistance of the circuit; \( R_P = \frac{1}{\omega_m C_P \tan \delta_P} \) is the equivalent resistance of the pyroelectric sensing element; \( \tau_E = R_{eq} \cdot C_{eq} \) [s] is the electrical time constant of the equivalent circuit; \( C_{eq} = (C_P + C_{input} + C_{fb}) \) [F] is the equivalent capacitance of the circuit; and \( C_{input} \) [F] is the input capacitance of the integrated op-amp.

**Total voltage-noise spectral density \( V_{N-LME-553} \) for a single LME-553 detector**

Figure 5.6 shows the calculated voltage-noise spectral densities for each individual noise source for the LME-553 PIR detector. Table 5.3 lists the associated values for all parameters used in the calculations. Note that for parameters which are not specified in LME-553 datasheets, typical values for similar LiTaO₃ type PIR detectors have been given (indicated by *).
Figure 5.6: Calculated voltage-noise spectral density for the LME-553 detector and associated parameters.

Table 5.3: Associated parameters for the calculation of voltage-noise spectral density for the LME-553.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Unit</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\alpha$ (*)</td>
<td>0.7</td>
<td>NA</td>
<td>[433]</td>
</tr>
<tr>
<td>$d_p$ (*)</td>
<td>3.2</td>
<td>J/cm$^2$/K</td>
<td>[432]</td>
</tr>
<tr>
<td>$d_p$ (*)</td>
<td>30</td>
<td>$\mu$m</td>
<td>[434]</td>
</tr>
<tr>
<td>$A_p$</td>
<td>9</td>
<td>mm$^2$</td>
<td>[408]</td>
</tr>
<tr>
<td>$H_p$ (*)</td>
<td>864</td>
<td>$\mu$J/K</td>
<td>N/A</td>
</tr>
<tr>
<td>$\tau_T$</td>
<td>200</td>
<td>ms</td>
<td>[408]</td>
</tr>
<tr>
<td>$G_T$</td>
<td>4.32</td>
<td>mW/K</td>
<td>N/A</td>
</tr>
<tr>
<td>$\varepsilon_r$ (*)</td>
<td>47</td>
<td>N/A</td>
<td>[432]</td>
</tr>
<tr>
<td>$p$ (*)</td>
<td>23</td>
<td>nC/cm$^2$/K</td>
<td>[432]</td>
</tr>
<tr>
<td>$C_p$ (*)</td>
<td>125</td>
<td>pF</td>
<td>N/A</td>
</tr>
<tr>
<td>$\tan \delta_p$ (*)</td>
<td>0.003</td>
<td>N/A</td>
<td>[432]</td>
</tr>
<tr>
<td>$R_{fb}$</td>
<td>4</td>
<td>$\Omega$</td>
<td>[408]</td>
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<tr>
<td>$C_{fb}$</td>
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<td>[408]</td>
</tr>
<tr>
<td>$\tau_E$</td>
<td>0.24</td>
<td>ms</td>
<td>N/A</td>
</tr>
<tr>
<td>$R_{input}$ (*)</td>
<td>100</td>
<td>$\Omega$</td>
<td>[433]</td>
</tr>
<tr>
<td>$i_{opamp}$</td>
<td>0.6</td>
<td>$fA/\sqrt{Hz}$</td>
<td>[435]</td>
</tr>
<tr>
<td>$u_{opamp}$</td>
<td>19</td>
<td>nV/\sqrt{Hz}</td>
<td>[435]</td>
</tr>
<tr>
<td>$C_{input}$ (*)</td>
<td>2</td>
<td>pF</td>
<td>[433]</td>
</tr>
</tbody>
</table>

From these calculations, it can be seen that $V_{NFB}$ dominates at lower modulation frequencies, due to the large feedback resistor. At modulation frequencies above 200 Hz, $V_{NU}$ dominates, because of the large equivalent input voltage-noise spectral density at the input of the op-amp. Furthermore, it is shown that $V_{ND}$ surpasses all
but $V_{NU}$ above 200 Hz. Noise from temperature fluctuation, input resistance and op-amp current noise, are less significant.

The overall voltage-noise spectral density $V_{\text{LME-553}}$ can be calculated from the summation of each individual noise source contribution:

$$V_{\text{LME-553}} = \sqrt{V_{NT}^2 + V_{ND}^2 + V_{NR}^2 + V_{NF}^2 + V_{NI}^2 + V_{NU}^2} \text{ [V/\sqrt{Hz}]} \quad (5.16)$$

Figure 5.7 shows the calculated overall voltage-noise spectral density and the measured results given by the LME-553 datasheet [408]. It can be seen that there is excellent agreement at low frequencies; discrepancy above this frequency is mainly due to not having exact values for all parameters. For example, in order to obtain $C_P$, which will significantly affect the value of $C_{eq}$ and $\tau'_E$, the thickness of the pyroelectric material $d_P$ had to be assumed to be a typical value of 30 µm [434]. This will introduce inaccuracy in calculating $V_{NU}$ and $V_{ND}$, especially at high frequencies where these noise sources dominate.

By integrating over the modulation bandwidth, assuming a 1 Ω reference load resistance, the noise power for a single LME-553 PIR sensor can be calculated as

$$N_{\text{LME-553}} = \int_{f_{m1}}^{f_{m2}} V_{\text{LME-553}}^2 df_m \text{ [W]} \quad (5.17)$$

where $f_{m1} = 1$ Hz and $f_{m2} = 1$ kHz are the lower and upper modulation frequency, respectively. The noise power is calculated to be 0.84 µW. Note that there are a num-
ber of additional sources of unwanted signals at the PIR sensor (mostly associated with the environment), such as atmospheric noise, stray electromagnetic interference and microphonics. These extrinsic effects can only be modelled once specific ambient deployment conditions are known and thus are not considered further here.

After identifying the intrinsic noise sources associated with the LME-553, its NEP and specific detectivity $D^*$ can be calculated using (5.9) and (5.10), as shown in Figure 5.8a and 5.8b, respectively. These results are also compared with associated measured data [408]. It can be seen that the predicted and measured results have a good fit over three orders of magnitude in the modulation frequency.

![Figure 5.8: Calculated and measured [408] (a) noise equivalent power and (b) specific directivity for LME-553 detector.](image)

### 5.5.2 Back-end electronics noise sources

The back-end electronics is used to further amplify the output signal from the PIR sensor and also filter-out unwanted noise from the PIR sensor. With the particular circuit, two identical PIR sensors were used for noise reduction. The output from each detector passes through a unity gain buffer amplifier. The small-signal output voltage will be amplified by a common low-noise instrumentation amplifier (INA), having a designed voltage gain of 100. The signal from the output of the INA is then filtered by a 4$^{th}$-order Sallen-Key Butterworth baseband BPF, having a designed centre frequency of 320 Hz (corresponding to the modulation frequency of 320 Hz). Therefore, the noise from the back-end electronics include: voltage and current noise from the external op-amps (OPA227) used in buffer and filter stages; voltage and current noise from the instrumentation amplifier (INA163); and noise from all the
resistors. It should be noted that burst and avalanche noise sources associated with op-amps are too small to be considered further [436].

Noise analysis for the back-end electronics can be performed using the SPICE-based analog simulation program TINA-TI [437]. Within this simulator, resistors having resistance \( R \) are represented by Johnson-Nyquist noise by default (having voltage-noise spectral density \( \sqrt{4k_B T R} \)). However, not all SPICE macros contain correct noise information for op-amps and INAs. Therefore, within TINA-TI, the uncorrelated voltage and current noise sources were added to the input of the op-amps and the INA. The current and voltage noise sources have two different noise contributions: Johnson-Nyquist noise, which has a flat spectral density; and flicker noise, which dominates at low frequencies. Figure 5.9a and 5.9b show the noise models for the OPA227 and the INA163, respectively. In the noise model, a current-controlled voltage source (CCVS) with a transresistance of 1 \( \Omega \) is applied to convert current-noise spectral density to voltage-noise spectral density, as the noise analysis in TINA-TI only considers voltage noise. The simulated voltage and current noise spectral densities for OPA227 and INA163 are shown in Figure 5.9c.

![Figure 5.9: Back-end noise modelling: (a) model for the op-amp OPA227; (b) model for the low-noise instrumentation amplifier INA163; (c) associated simulated voltage-noise spectral densities.](image)
Since the noise sources for all the components have been modelled, the output voltage-noise spectral density for the back-end electronics can be calculated using

\[ V_{N-BE} = \sqrt{(V_{N-buffer}\cdot A_{INA}\cdot A_{BB\cdot BPF})^2 + (V_{N-INA}\cdot A_{BB\cdot BPF})^2 + V_{N-BPF}^2} \text{[V/\sqrt{Hz}]} \]

(5.18)

where \( A_{INA} \) is the voltage gain of the INA. The output noise power from the complete back-end electronics \( N_{BE} \) is predicted to be 3.52 nW, over a modulation bandwidth from 1 Hz to 1 kHz.

If noise from both LME-553 PIR detectors are considered uncorrelated, the overall voltage-noise spectral density for the complete channel receiver is expressed as

\[ V_{N-receiver} = \sqrt{2V_{N-LME-553-output}^2 + V_{N-BE}^2} \text{[V/\sqrt{Hz}]} \]

(5.19)

where \( V_{N-LME-553-output} = V_{N-LME-553}\cdot A_{BB\cdot LNA}\cdot A_{BB\cdot BPF} \) is the voltage-noise spectral density of a single LME-553 at the output of the back-end electronics; \( A_{BB\cdot LNA} = A_{buffer}\cdot A_{INA} \) and \( A_{buffer} = 1 \) is the voltage gain of the buffer stage.

Figure 5.10 shows the simulated voltage-noise spectral density for a single LME-553 PIR detector and back-end noise sources for the channel receiver. As expected, the former predominantly dominates the channel receiver, while the latter can be ignored at all but very low modulation frequency (where flicker noise dominates). The intrinsic noise power from a single LME-553 is \( N_{LME-553} = 0.84 \mu W \). At the output of the channel receiver, this is \( N_{LME-553-output} = 3.26 \text{ mW} \). The intrinsic noise power from the back-end electronics is only \( N_{BE} = 3.52 \text{ nW} \), while the calculated intrinsic noise power for the complete channel receiver is \( N_{receiver} = 6.52 \text{ mW} \) – all values have been determined over a modulation bandwidth from 1 Hz to 1 kHz, and assume a 1 \( \Omega \) reference load resistance.

### 5.5.3 Channel receiver noise measurements

The noise performance for each channel receiver is measured directly using a PicoScope 2205 MSO digital oscilloscope. A noise floor at 65 \( \mu V_{RMS} \) was first recorded by short-circuiting the high-impedance input of the oscilloscope. The duration of the
measurements was 5 minutes for each channel. Since RMS values of noise voltage can be measured directly, the corresponding total noise power can be calculated, as shown in Table 5.4, where *Opaque* and *Filter* denote the PIR detector having either a blocked aperture or filled with its assigned channel THz BPF, respectively.

Table 5.4: Measured noise power for each PIR detector and channel receiver.

<table>
<thead>
<tr>
<th>Channel</th>
<th>(N_{\text{LME-553}}) ((\mu\text{W}))</th>
<th>(N_{\text{receiver}}) (mW)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Calculated</td>
<td>Measured</td>
</tr>
<tr>
<td>Opaque</td>
<td>Filter</td>
<td></td>
</tr>
<tr>
<td>A</td>
<td>0.84</td>
<td>2.50</td>
</tr>
<tr>
<td>B</td>
<td>0.84</td>
<td>2.22</td>
</tr>
<tr>
<td>C</td>
<td>0.84</td>
<td>2.40</td>
</tr>
<tr>
<td>D</td>
<td>0.84</td>
<td>1.96</td>
</tr>
</tbody>
</table>

When compared with the calculated value of \(N_{\text{LME-553}} = 0.84 \mu\text{W}\), the measured values are 2 to 3 times higher. This discrepancy is due to two main practical reasons: first, the calculated values were obtained by integrating over a modulation bandwidth from 1 Hz to 1 kHz, while the noise detected by the PIR sensor is much wider in the measurement. Second, in practice, the PIR sensor was exposed to background ambient conditions; extrinsic noise sources, which were not included in previous simulations, have also been measured by the sensor. It should also be noted that the calculations are based on values from the data sheet, which does not give information
on measurement conditions (e.g., temperature).

5.5.4 Measured output signal-to-noise ratio

Since both the signal and noise properties of this wireless communications system have been measured, assuming a 1 Ω reference load resistance, the measured output SNR can be calculated as

$$SNR_{dB} = 10 \log_{10} \left( \frac{V_{out}^2}{N_{receiver}} \right) \text{ [dB]} \quad (5.20)$$

To validate the results, measured end-to-end SNRs are compared against calculated values, resulting in the scatter plot shown in Figure 5.11. In Figure 5.11, SNR values were obtained by varying the source bias current from 44 to 80 mA and the transmission range from 1 to 3 cm, to obtain different receiver output signal voltages. In general, the measured SNR values agree well with the predicted data. In low SNR conditions, the measured results tend to be higher than predicted. This is because the calculated noise power is higher than measured, for Channels B, C and D – resulting in lower calculated SNRs.

Figure 5.11: Measured SNR against calculated SNR performance for each channel receiver (A, B, C, D). Dashed line represents the “ideal” results that measured and calculated SNR values are the same.
Figure 5.12 shows the measured BER against measured SNR. The measured BER was obtained using the methodology described in Section 4.2.3, and an end-to-end binary data stream having $2 \times 10^6$ bits was used. When no bit errors were observed, actual BER values are considered to be smaller than $10^{-6}$, and, therefore, not shown. A simple empirical curve fit is also shown in Figure 5.12, given by

$$BER \approx 0.5 e^{-SNR/10}$$ (5.21)

Figure 5.12: Measured SNR against measured BER performance for each channel receiver (A, B, C, D).

It can be seen that (5.21), with our current implementations, resembles the classical relationship $BER = 0.5 e^{-E_b/N_0}$ obtained for the optimum differential binary phase-shift keying (DBPSK) [438]. When compared to DBPSK, the existing hardware prototype demonstrator requires much higher values of SNR for the same levels of BER. Therefore, Figure 5.12 and (5.21) provide a useful tool for predicting the performance of this system. By applying other techniques, such as forward error correction (FEC) algorithms, the BER and overall performance for the complete system is expected to be improved.

5.6 Conclusions

In this chapter, a detailed end-to-end power link budget analysis for the ‘THz Torch’ thermal infrared banded-noise wireless communications link has been investigated.
Here, a number of assumptions have had to be made, in order to limit the scope and depth of this work (e.g., filament emissivity and detector's spectral responsivity, both being frequency independent; effective radiating area of the filament being half the calculated physical area; bulbs having a hemispherical shape; the bulb's glass envelope having a constant thickness; glass envelope emissivity being temperature independent; simplistic atmospheric modelling; ignoring diffraction effects with the THz BPFs; ignoring any mechanical misalignment between the transmitter and receiver THz BPFs; adopting parameter values for the PIR sensor and a modulation bandwidth from 1 Hz to 1 kHz; and ignoring extrinsic noise contributions). Nonetheless, the predicted output RMS voltages from each receiver of the multi-channel ‘THz Torch’ system agree well with independent measurements. It has been shown that, with this thermodynamics-based technology, the range of the wireless link has the potential to reach \( \sim 1 \) km. Moreover, the overall noise performance is dominated by the front-end PIR sensor, as found with conventional electronic/photonic systems.

The results from this detailed power link budget analysis can create a useful insight into the practical operation at both component and systems levels. With further refinement, it will prove to be an invaluable tool for engineering optimal performances with single and multi-channel systems. For example, with the former, channel frequency allocation can be optimized to promote longer ranges. With the latter, signal levels can be adaptively controlled and equalized; simply by changing the source bias currents for each channel, thus, balancing out the performance (e.g., having the same minimum output SNR across the channels) of the complete multi-channel system. With both, spreading loss is an important factor that limits the transmission range; this issue can be improved considerably by employing collimating lenses, as with optical systems. In addition, more channels can be introduced to overcome limited response times for a single PIR sensor, to further increase data rates, while also improving the resilience of the system to interception and jamming at the physical layer.
6 THz Metal Mesh Filters for ‘THz Torch’ Applications

6.1 Introduction

Metal mesh filters have been used in far-infrared, THz and submillimetre astronomical instruments since the first publication by Ulrich [323] on the transmission properties of metal grids [326]. They have advantages such as compact, excellent scalability, and an easier fabrication process in comparison with other THz filters. Metal mesh band-pass filters exhibit high transmission (normally higher than 80%) at the resonant frequency, have adjustable pass band characteristics, and with good out-of-band rejection [327].

In this chapter, the electromagnetic theory of metal mesh filters will be introduced first, where both free-standing and substrate-based types will be discussed. Moreover, conventional cross-shaped THz metal mesh filters on standard 525 µm fused silica substrates will be designed, simulated, fabricated and measured. Trapped-mode excitation is employed to further improve the out-of-band rejection. This is validated by preliminary experimental results. This metal mesh filter work and its potential applications for ‘THz Torch’ technology are finally discussed in the conclusion part.
6.2 Metal Mesh Filters Basics

6.2.1 Theory

Metal mesh filters can be described as two-dimensional (2D) metamaterials, where metallic unit cells with certain pattern are arranged periodically, forming a 2D array. In microwave/RF engineering, they are normally referred to as frequency-selective surfaces (FSSs). In Ulrich’s early work, the properties of two types of structure: a simple grid with a regular array of square openings; and its complementary structure of an array of metallic squares supported on a thin dielectric substrate were investigated [323]. Key parameters for these structures are the width/spacing ($K$) and the periodicity ($G$) of the grids.

Using the transmission line method, the behaviour of these structures was modelled as either lumped inductance (square openings) or a lumped capacitance (free-standing squares). The former exhibits a high-pass frequency response characteristic while the latter has a low-pass spectral property. These two types of meshes are commonly referred to as inductive or capacitive grids. The corresponding equivalent circuits are as shown in Figure 6.1 [439], where $R = 0$ if the grids are assumed to be lossless (i.e., no absorption).

![Inductive and capacitive grids and corresponding equivalent circuits](adapted from [439]).

In [323], it was shown that at resonant frequency $\omega_0$, the normalized characteristics impedance of the inductor and capacitor for these two equivalent circuits can be
expressed as

\[ Z_0 = \frac{\omega_0 L}{\omega_0 C} \]  (6.1)

Therefore, corresponding admittances can be described by

\[ Y_{\text{ind}}(\omega) = \frac{1}{R - jZ_0\Omega} \]  (6.2)

and

\[ Y_{\text{cap}}(\omega) = \frac{1}{R + jZ_0\Omega} \]  (6.3)

where \( \Omega = \frac{\omega - \omega_0}{\omega} \) is the generalized frequency and \( \omega = G/\lambda \).

Using the transmission line theory, the power transmittance \(|\tau(\omega)|^2\) and reflectance \(|\rho(\omega)|^2\) for inductive and capacitive grids are

\[ |\tau(\omega)|^2_{\text{ind}} = \frac{R^2 + Z_0^2/\Omega^2}{(1 + R)^2 + Z_0^2/\Omega^2} \]  (6.4)

\[ |\tau(\omega)|^2_{\text{cap}} = \frac{R^2 + Z_0^2\Omega^2}{(1 + R)^2 + Z_0^2\Omega^2} \]  (6.5)

\[ |\rho(\omega)|^2_{\text{ind}} = \frac{1}{(1 + R)^2 + Z_0^2/\Omega^2} \]  (6.6)

\[ |\rho(\omega)|^2_{\text{cap}} = \frac{1}{(1 + R)^2 + Z_0^2\Omega^2} \]  (6.7)

The power absorptance \(|A(\omega)|^2\) then can be calculated to be [323]

\[ |A(\omega)|^2 = 1 - |\tau(\omega)|^2 - |\rho(\omega)|^2 = 2R|\rho(\omega)|^2 = |\rho(\omega)|^2 \eta \left( \frac{c}{\lambda \sigma_c} \right)^{1/2} \]  (6.8)

where \( \eta \) is the ratio of the area of the grid to that of a flat sheet, which can be calculated by \( \eta = G/K \) for capacitive grids and \( \eta = 1/(1 - K/G) \) for inductive grids; \( \sigma_c \) is the conductivity of the metal [323]. Since the values of \( Z_0 \) and \( \omega_0 \) can be calculated numerically, the power reflectance, transmittance and absorptance at various frequencies can be obtained analytically.

Although many unit cell geometries have been used to provide desired band-pass spectral responses, the most known element is the cross-shaped structure, which
is obtained by superimposing the two patterns shown in Figure 6.1. Figure 6.2a illustrates the structure of conventional cross-shaped metal mesh filters. It is similar to the inductive grid, except that the repeated unit cell is a cross instead of a square. This type of filter has both inductive and capacitive properties, yielding a self-resonance. As a result, a band-pass characteristic will appear. The corresponding equivalent circuit is shown in Figure 6.2b. Similarly, its complementary structure, called capacitive cross-shaped metal mesh filter, has a band-stop spectral property, but it requires a substrate for the realization.

![Figure 6.2: Inductive cross-shaped metal mesh filters: (a) structure and parameters [327]; (b) equivalent circuit (adapted from [326]).](image)

The spectral characteristics of cross-shaped metal mesh filters are controlled by their periodicity \( G \), length \( L \) and width \( K \) of crosses. The effect of each parameter on the frequency response will be discussed in Section 6.2.2. In order to avoid the waveguide effect, the thickness of the metal film \( h \) is normally made very thin (e.g., \(<\lambda_r/100\), where \( \lambda_r \) is the resonant wavelength). Literatures have shown cross-shaped metal mesh filters operating within the range from 0.1 to 14 THz, exhibiting high transmission peaks and having bandwidths between 13% to 50% of the resonant frequency [324]. Metal mesh band-pass filters have been commercially available, covering the spectral range from 0.5 to 30 THz [328].

At even higher frequencies, one of the challenges for the fabrication of metal mesh filters is the thickness of metal layer: it has to be made very thin when compared to the resonant wavelength; otherwise the waveguide mode will be introduced. Therefore, free-standing filters for high frequency applications are mechanically fragile.
Supporting materials, which are transparent or semi-transparent within the frequency range of interest, have to be employed. Substrates such as Mylar (polyester film), polyamide, polytetrafluoroethylene (Teflon) and benzoctyloclobutene (BCB) are normally used. Recently, other interesting works have been published using different materials as substrates, including polymethylpentene (TPX), TydexBlack and high-density polyethylene (HDPE) [324]. These cross-shaped metal mesh filters, either free-standing or substrate-based, can be fabricated using photolithography, Direct Electron Beam Litography (DEBL), nanoimprint, laser ablation or lithography electroplating and molding (LIGA).

6.2.2 Free-standing metal mesh filters

Due to the advances of computational electromagnetics and computing platforms, the capabilities and efficiencies of numerical techniques have been greatly improved. Commercially available simulation softwares, e.g., CST and HFSS™, are much more powerful than before, making the simulations and optimisations of 2D periodic structures much easier. However, in order to further decrease the time consumption on the design stage, it is still necessary to investigate the effect of each geometrical parameter and the scalability of metal mesh filters.

Referring to Figure 6.2a, the resonant frequency, peak transmission and bandwidth are determined by $G$, $L$ and $K$. To start with, a test model with $G = 600 \mu m$, $L = 400 \mu m$ and $K = 100 \mu m$ are used to evaluate the effect of each parameter, with an estimated resonant wavelength of $G/2 < \lambda_r < G$. The thickness of the thin gold layer was set to 1 \(\mu\)m so that the waveguide effect can be avoided. The CAD modelling and simulated transmission and reflection curves are as shown in Figure 6.3. The simulated results show a resonant frequency ($f_r$) of 0.375 THz, a peak transmission ($\tau$) of 98.5% and a -3 dB bandwidth (BW) of 20.6%. The transmission zero ($f_{\text{zero}}$) occurs at $\lambda_d = c_0/G = 0.5$ THz, due to the diffraction of the infinite array, and is only dependent on its periodicity.
Effect of the periodicity $G$

The effect of the periodicity (lattice constant) of inductive metal mesh filters will be investigated in this section. The value of $G$ is adjusted from 420 to 780 $\mu$m, which is $\pm 30\%$ of its original value, while the other geometrical parameters are kept the same. Simulation results shown in Table 6.1 were obtained using CST.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>$G$ ($\mu$m)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>420</td>
</tr>
<tr>
<td>$f_r$ (THz)</td>
<td>0.386</td>
</tr>
<tr>
<td>$f_{zero}$ (THz)</td>
<td>0.699</td>
</tr>
<tr>
<td>Peak $\tau$ (%)</td>
<td>99.2</td>
</tr>
<tr>
<td>BW (%)</td>
<td>54.3</td>
</tr>
</tbody>
</table>

From Table 6.1, it can be seen that within $\pm 30\%$ parameter variations, the resonant frequency will increase when the periodicity decreases. The limit is that when $G \to L$, and the structure becomes the capacitive grids, which has a low-pass spectral response. This will decrease the lower -3 dB cut-off frequency and result in a wider bandwidth. On the other hand, the frequency of the transmission zero, at which the diffraction occurs, determines the upper -3 dB cut-off frequency and thus the bandwidth. As the value of $G$ decreases, $f_{zero}$ will move towards higher
frequency, giving a wider BW. Due to these two reasons, the BW of the metal mesh filters will increase as $G$ decreases, from 8.4% to 54.3% in this case. In general, the peak transmission $\tau$ will slightly increase as the periodicity decreases, because of a larger aperture to total area ratio, which permits more power of the incident wave transmitting through this structure.

**Effect of the cross length $L$**

The effect of the length of the crosses will be investigated in this section. To keep the same increment ratio, the value of $L$ is adjusted from 280 to 520 $\mu$m, which is $\pm 30\%$ of its original value, while the other geometrical parameters are kept the same. Simulation results are shown in Table 6.2.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>$L$ ($\mu$m)</th>
<th>280</th>
<th>320</th>
<th>360</th>
<th>400</th>
<th>440</th>
<th>480</th>
<th>520</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_r$ (THz)</td>
<td></td>
<td>0.461</td>
<td>0.433</td>
<td>0.404</td>
<td>0.375</td>
<td>0.346</td>
<td>0.320</td>
<td>0.294</td>
</tr>
<tr>
<td>$f_{zero}$ (THz)</td>
<td></td>
<td>0.501</td>
<td>0.500</td>
<td>0.500</td>
<td>0.500</td>
<td>0.499</td>
<td>0.498</td>
<td></td>
</tr>
<tr>
<td>Peak $\tau$ (%)</td>
<td></td>
<td>97.0</td>
<td>97.8</td>
<td>98.3</td>
<td>98.5</td>
<td>98.8</td>
<td>99.0</td>
<td>99.1</td>
</tr>
<tr>
<td>-3 dB BW (%)</td>
<td></td>
<td>5.8</td>
<td>10.5</td>
<td>15.5</td>
<td>20.6</td>
<td>26.8</td>
<td>32.2</td>
<td>39.5</td>
</tr>
</tbody>
</table>

It is shown that as the cross length increases, the resonant frequency will be significantly decreased. This relationship is straightforward: smaller lengths correspond to higher frequencies. A simple equation $\lambda_r \approx 2L$ can be used to estimate the resonant wavelength. The frequency of the transmission zero will not change when the cross length varies, as it is only dependent on the periodicity of this two-dimensional structure. Furthermore, increasing the value of $L$ has the same effect as decreasing the value of $G$, resulting in a slightly higher peak transmission. Finally, the -3 dB bandwidth will increase as the cross length increases. This is because: (1) for a larger value of $L$, the separations between the unit cells decrease, which is effectively the same as decreasing the periodicity. This will decrease the lower 50% cut-off frequency, giving a larger bandwidth; (2) with the same transmission zero frequency
which determines the upper 50% cut-off frequency of this band-pass filter, smaller $f_r$ will result in a wider bandwidth due to a less sharp roll-off slope characteristic.

**Effect of the cross width $K$**

The effect of the width of the crosses will be shown in this section. The value of $K$ ranges from 70 to 130 µm, which is ±30% of its original value, while the other geometrical parameters are kept the same. Simulation results are shown in Table 6.3.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>$K$ (µm)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>70</td>
</tr>
<tr>
<td>$f_r$ (THz)</td>
<td>0.365</td>
</tr>
<tr>
<td>$f_{zero}$ (THz)</td>
<td>0.502</td>
</tr>
<tr>
<td>Peak $\tau$ (%)</td>
<td>97.9</td>
</tr>
<tr>
<td>-3 dB BW (%)</td>
<td>17.1</td>
</tr>
</tbody>
</table>

As seen from Table 6.3, increasing the value of $K$ will slightly move $f_r$ towards higher frequencies. The peak transmission and bandwidth will also increase, as the cross width increases. Again, the transmission zero point is not a function of $K$. In general, the spectral responses are less dependent on the cross width. Nevertheless, without careful design of this parameter, one will be able to obtain the desired band-pass characteristics.

**Effect of the metal layer thickness $h$**

With original parameters, where $G = 600$ µm, $L = 400$ µm, $K = 100$ µm and $h = 1$ µm, the resonant frequency of the band-pass filter was simulated to be 0.375 THz, as seen in Figure 6.3. This corresponds to a resonant wavelength $\lambda_r$ of ~800 µm. Since $h \ll \lambda_r$, the waveguide mode effect from the metal layer can be ignored in these simulations. However, it has to be taken into account if its thickness is comparable to $\lambda_r$. In this section, the effect of the metal layer thickness will be investigated, by
increasing the value of $h$ from $\lambda_r/1000$ to $\lambda_r$, while the other parameters are kept the same. Simulation results are shown in Figure 6.4.

![Simulation results](Figure 6.4: Simulated spectral responses for different metal layer thickness $h$ with $G = 600 \mu m$, $L = 400 \mu m$ and $K = 100 \mu m$.)

It can be seen that, for $h \ll \lambda_r/10$, the effect from the metal layer thickness can be ignored. From $h \approx \lambda_r/10$, a new transmission peak appears. This peak corresponds to a waveguide mode of the openings and the energy from the surface waves on the front is transferred to the back side. The wavelength of the transmission peak is determined by the vector addition of wavevector $k$ of the surface waves and the waveguide modes [440]. By increasing $h$, this peak will move towards lower frequencies until combined with the original transmission peak.

Actually, by carefully choosing the metal layer thickness, it is able to obtain desirable spectral band-pass responses, for which the $1^{st}$ waveguide mode transmission peak has moved near to the original transmission peak, while the $2^{nd}$ peak have not appeared within the spectrum of interest. Figure 6.5 gives the simulated transmission for $h = 400, 500$ and $600 \mu m$. It is observed that for $h = 500 \mu m$, a band-pass filter from 0.390 to 0.455 THz with a bandwidth of 24.6% is obtained, although there is a transmission dip of 63.4% occurs at 0.420 THz. This is quite similar to the operation principle of some THz dichroic filters, for which the cross-shaped resonant structures are normally replaced by circular elements.
Effect of the incident angle $\theta_{\text{inc}}$

All the simulation results shown above are based on plane-wave excitation at normal incidence. However, the spectral responses of metal mesh filters also depend on the incident angle. In this subsection, the performance of these filters will be investigated at different incident angles. To be more specific, $\theta_{\text{inc}}$ will range from $\theta = 0^\circ$ (normal incident) to $\theta = 45^\circ$, while $\varphi$ is kept as $0^\circ$ in these simulations. Figure 6.6 demonstrates the corresponding simulated spectral responses.

Figure 6.6 shows that if we no longer assume normal incidence, more peaks/dips will appear within the spectrum of interest due to diffraction [441,442]. It is known that normal component of the E-field cannot induce surface currents since the elec-
trons are constrained to move in only two dimensions. For normal incident condition, the E-field is always in the filter plane (in-plane polarisation) and always with the same amplitude at a certain time. In this case, the effectiveness of the incoming wave is at its maximum. However, the E-field has components normal and tangent to the surface (out-of-plane polarisation) for non-normal incidence, and the E-field is not with a uniform distribution in the filter plane. The corresponding impedance of the metal mesh filter will be changed for different polarisation states. Therefore, the overall transmission and reflection coefficients are the linear combinations of E-field vectors at different states [443].

Scalability of free-standing metal mesh filters

The geometrical parameters for the original design are scaled down by a maximum factor of 50, and the scalability of free-standing metal mesh filters will be demonstrated. In order to describe the scalabilities of different parameters, the relative scaling error is used, which is defined as

$$e_s = \frac{\lambda_c - \text{simulated} - \lambda_c - \text{predicted}}{\lambda_c - \text{predicted}} \times 100\%$$

where $\lambda_c - \text{predicted}$ is the predicted value of the parameter (e.g., resonant frequency) using the scaling factor, while $\lambda_c - \text{simulated}$ is the simulated value based on the scaled structure.

The predicted and simulated resonant frequency $f_r$, peak transmission $\tau$ and bandwidth are shown in Figure 6.7. In the simulations, the metal thickness is kept as 1 $\mu$m to avoid waveguide effect. Other parameters are scaled according to the scaling-down factor.

It is found that the resonant frequency has a good scaling property, with a relative scaling error of $|e_s| < 1\%$. However, both the peak transmission and -3 dB bandwidth will decrease as the structure scales down. The former has $|e_s| < 8\%$ while the latter is with $|e_s| < 42\%$. This is mainly due to the increased absorption at higher frequencies. Nevertheless, one can still say that these thin layer free-standing metal
mesh filters are scalable, and have the potential to be applied to the ‘THz Torch’ technology, especially within the lower part of the spectrum of interest.

**Multi-layer free-standing metal mesh filters**

Many applications require sharp roll-off characteristics. In this case, two or more metal mesh filters can be stacked together to form multi-layer band-pass filters for better out-of-band rejection. For free-standing metal mesh filters, the only variable is the separation distance of different layers. By minimizing the interaction between layers, a desired spectral response can be obtained. Air-gap separations can be utilized by using metal or dielectric spacers.

Figure 6.8 shows the simulated spectral responses for different separation distances $d_s$. Here, $G = 600 \, \mu m$, $L = 400 \, \mu m$ and $K = 100 \, \mu m$ are used, giving a resonant frequency of 0.375 THz. It is found that the desired result, e.g., only has a single transmission peak, was obtained at $d_s \approx \lambda_r/4 = 200 \, \mu m$. This means that these two layers of filters have a minimum interaction when $d_s$ is equal to a quarter of the resonant wavelength. As $d_s$ increases further, more transmission peaks will appear. For $d_s = \lambda_r/2$, three transmission peaks were observed within 0.25 to 0.5 THz range. Another important separation distance is $d_s \approx G/2$, at which a flat and wide pass band can be obtained. Therefore, depending on the specific requirement, it is recommended to choose a separation distance between $\lambda_r/4$ and $\lambda_r/2$.

The above results assume a perfect alignment for these multi-layer band-pass
filters. However, in practice, the performance of stacked metal mesh filters will always be affected due to non-aligned elements. A mismatch distance can be used to describe the misalignment, as shown in Figure 6.9, where $d_m = 0$ (left) means a perfect alignment, and $d_m = G/2$ (right) represents a total misalignment.

The simulated spectral responses for these two conditions (perfect alignment and total misalignment) at different separation distances are shown in Figure 6.10, with $G = 600 \, \mu m$, $L = 400 \, \mu m$ and $K = 100 \, \mu m$. It is seen that for small $d_s$, the misalignment will severely degrade the performance of the two-layer metal mesh filters, due to the near-field interaction. As this value increases, this near-field effect becomes less significant. For $d_s = 400 \, \mu m$, both conditions can produce similar spectral response. In this case, the contribution from misaligned elements disappears.
Figure 6.10: Effect of the misalignment for different separation distances $d_s$: (a) 100 µm; (b) 200 µm; (c) 300 µm; (d) 400 µm. Here, $G = 600$ µm, $L = 400$ µm and $K = 100$ µm are used in these simulations.

and the band-pass characteristics only depend on the separation distance.

6.2.3 Substrate-based metal mesh filters

Effect of the substrate thickness $h_s$

The realization of capacitive cross-shaped metal mesh filters has to rely on substrates. Furthermore, at high frequencies, to avoid the waveguide effect while still keeping the filters mechanically robust, substrate materials are normally used to support these thin metal layers. As a result, the air-metal-air boundaries for free-standing mesh filters are altered to air-metal-dielectric-air boundaries for the substrate-based type. This will cause a resonant frequency shift, whose value depends on both the refractive index $n$ and thickness of the supporting material. For thick substrate, the shifted resonant wavelength can be predicted using [440]

$$\lambda_{r_0} \rightarrow \lambda_{r_d} \approx \lambda_{r_0} \sqrt{\frac{n_1^2 + n_2^2}{2}}$$

(6.10)
where $\lambda_{r_0}$ is the resonant wavelength for free-standing metal mesh filters; $\lambda_{rd}$ is the shifted resonant wavelength for substrate-based filters; and $n_1$ and $n_2$ are the refractive indices of the materials on either side of the metal layer. If fused silica is used as the substrate, $n_1 = 1$ for the free space and $n_2 = 1.94$ for the substrate, a shift factor of 1.546 can be obtained.

Note that (6.10) can only give good estimations for thick substrate (e.g., $h_s > \lambda_{rd}/20$). The shift of resonant frequency is less significant if the substrate is thin, as shown in Figure 6.11; the solid line represents the estimated values of $\lambda_{rd}$ obtained by (6.10), while the markers are the simulated values of $\lambda_{rd}$.

![Figure 6.11: Estimated (solid line) and simulated (square marker) shifted resonant frequency for different substrate thickness $h_s$.](image)

Another challenge for the design of substrate-based metal mesh filter is the Fabry-Perot effect: thick dielectrics will behave as a Fabry-Perot etalon and multiple reflections within the supporting material can produce transmission peaks within the spectrum of interest. For a substrate with a thickness of $h_s$ and refractive index of $n$, the phase difference $\delta$ can be expressed as

$$
\delta = \left( \frac{2\pi}{\lambda_0} \right) 2nh_s \cos \theta
$$

(6.11)

where $\lambda_0$ is the free space wavelength; and $\theta$ is the incident angle.

Assume surfaces on both sides of the substrate have a reflectance $R$, the trans-
mittance is given by [444]

\[
T = \frac{(1 - R)^2}{1 + R^2 - 2R \cos \delta} = \frac{1}{1 + \frac{4R}{(1-R)^2} \sin^2(\delta/2)},
\]  

(6.12)

The transmission peaks occur when \(\sin^2(\delta/2) = 0\). Therefore, \(\delta = 2\pi \cdot m\), where \(m = 1, 2, \cdots\). By substituting this into (6.11) and considering the normal incidence condition, we have

\[
\lambda_0 = \frac{2nh_s}{m}
\]

(6.13)

From (6.13) it can be seen that when the value of \(h_s\) increases, the transmission peak wavelength will also increase. As shown in Figure 6.12, for \(h_s/\lambda_{r_0} = 0.01\), only one resonance, which is associated with the metal mesh structure, can be excited in 0.15 to 0.5 THz range. When the thickness increases to \(h_s/\lambda_{r_0} = 0.1\), a transmission peak caused by Fabry-Perot mode appears at 0.445 THz. For \(h_s = \lambda_{r_0}/4\), the first Fabry-Perot peak moves to \(~0.178\) THz, which is not far away from the designed resonant frequency of 0.243 THz. As a result, the expected peak transmission is reduced. More Fabry-Perot modes will also appear within the same spectral range. This will, in turn, degrade the performance of such band-pass metal mesh filters. In practical designs, the substrate thickness and the desired spectral response have to be carefully considered, so that the effect from thick substrates is kept to a minimum.

![Figure 6.12: Simulated spectral responses of the filter with \(G = 600\) µm, \(L = 400\) µm and \(K = 100\) µm and different substrate thickness \(h_s\).](image)
Multi-layer substrate-based metal mesh filters

For multi-layer free-standing metal mesh filters, the only variable is the separation distance. Having a quarter-wavelength distance, the interaction between the two layers is kept to a minimum. Substrate-based metal meshes have more options: substrate thickness and configuration type, although the former sometimes is not adjustable. In this part, different configurations for two-layer band-pass metal mesh filters on thick substrates will be discussed.

The first type employs only one substrate, and each side has a metal layer on top. Such configuration can effectively decrease the absorption from thick substrates. However, the separation distance cannot be adjusted in this case, and the fabrication process would be more challenging, due to the alignment of the top and bottom metal layers. Multi-layer metal mesh filters can also be realized by combining two conventional two single-layer substrate-based filters and separated by an air-gap. The advantage is that the separation distance can be adjusted based on the resonant frequency, to have a minimum interaction between the layers. But with an additional substrate compared to the first configuration, the absorption from supporting materials will increase.

Figure 6.13 shows the simulated spectral responses for these two configuration types. The geometrical parameters of the metal mesh filters are kept the same as previous examples. With Type A configuration, it is clearly seen that the first Fabry-Perot resonance appears at 0.257 THz for \( h_s = 100 \, \mu m \). As the substrate thickness increases, this transmission peak moves towards the desired resonant frequency of 0.220 THz for \( h_s = 200 \, \mu m \), and then to 0.198 THz for \( h_s = 300 \, \mu m \). While for Type B configuration, by changing the width of the air-gap, Fabry-Perot resonant frequencies can be adjusted. For example, for \( d_s = 510 \, \mu m \), a transmission peak was observed at 0.299 THz. By increasing/decreasing the separation distance, this peak can be moved towards/away from the required resonant frequency. Figure 6.14 shows the change in frequency for the first Fabry-Perot mode. By carefully choosing the separation distance, better pass band characteristics can be obtained to meet specific design requirements, especially in the 0.1 to 0.3 THz range.
Figure 6.13: Different multi-layer configurations and simulated spectral responses: (a) two configuration types; (b) transmission curves for Type A; (c) transmission curves for Type B.

6.3 THz Metal Mesh Filters on Electrically Thick Substrates

6.3.1 Conventional cross-shaped THz metal mesh filters

Previous examples have shown the effects from geometrical parameters and substrates. In the practical designs, all these factors must be taken into consideration to obtain the desired band-pass characteristics. To develop mechanically robust and
Figure 6.14: Change in frequency for the first Fabry-Perot mode for different separation distance $d_s$ ($h_s = 200 \, \mu m$).

Low cost band-pass metal mesh filters, supporting materials are needed. There are other techniques, such as applying electrically thin substrates or encasing within polymer layers to fabricate substrated-based metal mesh filters. However, thin substrates are fragile and expensive, while encasing the film within a polymer increases the cost due to the extra fabrication steps. Therefore, a standard 525 $\mu m$ fused silica wafer is used here as the substrate, due to its low losses at THz frequencies, low dielectric constant and its compatibility with standard surface micromachining.

The design process for THz metal mesh filters on electrically thick substrates can be divided into four steps:

1. Calculate the shift factor using (6.10). In this case, $n_1 = 1$ and $n_2 = 1.94$, giving a shift factor of 1.546;

2. Estimate the values for parameters $G$, $L$ and $K$. Here, the desired resonant frequency $\lambda_{rd}$ should be first converted to the resonant frequency $\lambda_{r0}$ of free-standing metal meshes. Then $G$, $L$ and $K$ can be estimated using $G/\lambda_{r0} \approx 0.8$, $L/\lambda_{r0} \approx 0.5$ and $K/\lambda_{r0} \approx 0.15$ [324];

3. Simulate the initial model and obtain its spectral response;

4. Based on the initial result, considering the effect of each parameter and desired pass band characteristics, e.g., resonant frequency, bandwidth, adjust the parameters and continue the optimisation until a satisfactory result is obtained.
Table 6.4 shows the optimised parameters to achieve resonances at 0.1 THz intervals over the 0.1 to 0.5 THz band. Simulation results show that these conventional cross-shaped metal mesh filters can produce the expected filter responses at desired frequencies. However, the out-of-band rejection for these cross-shaped filters is poor due to multiple reflections within the thick substrates, as seen in Figure 6.15.

Table 6.4: Optimised parameters for conventional cross-shaped metal mesh filters.

<table>
<thead>
<tr>
<th>( f_r ) (THz)</th>
<th>( G ) (µm)</th>
<th>( L ) (µm)</th>
<th>( K ) (µm)</th>
<th>Peak ( \tau ) (%)</th>
<th>-3 dB BW (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.1</td>
<td>1150</td>
<td>975</td>
<td>100</td>
<td>91.9</td>
<td>19.0</td>
</tr>
<tr>
<td>0.2</td>
<td>680</td>
<td>425</td>
<td>50</td>
<td>86.1</td>
<td>17.7</td>
</tr>
<tr>
<td>0.3</td>
<td>440</td>
<td>293</td>
<td>30</td>
<td>92.0</td>
<td>5.6</td>
</tr>
<tr>
<td>0.4</td>
<td>360</td>
<td>210</td>
<td>20</td>
<td>87.1</td>
<td>3.5</td>
</tr>
<tr>
<td>0.5</td>
<td>290</td>
<td>160</td>
<td>15</td>
<td>79.2</td>
<td>3.3</td>
</tr>
</tbody>
</table>

Figure 6.15: Simulated spectral responses for conventional cross-shaped metal mesh filters on electrically thick substrate at 0.1, 0.3 and 0.5 THz.

6.3.2 Trapped-mode THz metal mesh filters

It has been shown that THz metal mesh filters on electrically thick substrates suffer from Fabry-Perot resonances caused by multiple reflections within the substrate. As discussed in Section 6.2.1, a complementary spectral response can be obtained by
using a complementary structure. Since inductive cross-shaped metal meshes show band-pass characteristics, its complementary structure, i.e., capacitive cross-shaped meshes, will have band-stop properties. With substrates, it is possible to add an inner cross to the original design and create the trapped-mode excitation [445], in order to improve the out-of-band rejection.

Figure 6.16a shows the structure of trapped-mode metal mesh filters. The excitations of trapped modes can be verified by the simulation of the surface current distribution within the structures, as illustrated in Figure 6.16b. At resonance frequencies, the induced currents are counter propagating at distinct sections of the structure. The resulting dipole moment and, therefore, the dipolar coupling to external electromagnetic fields is strongly reduced, which results in a high transmission at or near the resonant frequencies [329].

![Figure 6.16: Trapped-mode cross-shaped metal mesh filters: (a) structure; (b) simulated surface current distribution.](image)

Table 6.5 shows the optimised filter parameters for two cross-shaped band-pass metal mesh filters based on trapped-mode excitation. Compared to the corresponding simulations with conventional cross-shaped filters, there is a greatly improved out-of-band performance, as seen in Figure 6.17.

These trapped-mode metal mesh filters can also be stacked to achieve a narrower bandwidth and better out-of-band rejection. Type B configuration is used here to have adjustable separation distance. Corresponding simulated spectral responses for different $d_s$ are shown in Figure 6.18. As seen from this figure, for $d_s \approx \lambda_{rd}/4$,
Table 6.5: Optimised parameters for cross-shaped metal mesh filters based on trapped mode excitation.

<table>
<thead>
<tr>
<th>$f_r$(THz)</th>
<th>$G$ ($\mu$m)</th>
<th>$L$ ($\mu$m)</th>
<th>$L_{inner}$ ($\mu$m)</th>
<th>$K$ ($\mu$m)</th>
<th>$K_{inner}$ ($\mu$m)</th>
<th>Peak $\tau$ (%)</th>
<th>-3 dB BW (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.1</td>
<td>978</td>
<td>744</td>
<td>540</td>
<td>92</td>
<td>51</td>
<td>89.3</td>
<td>12.3</td>
</tr>
<tr>
<td>0.3</td>
<td>365</td>
<td>234</td>
<td>193</td>
<td>36</td>
<td>16</td>
<td>87.2</td>
<td>5.3</td>
</tr>
</tbody>
</table>

Figure 6.17: Simulated spectral responses for cross-shaped metal mesh filters based on trapped mode excitation at 0.1 and 0.3 THz.

the resonant frequency occurs at $\sim$0.1 THz with a lower peak transmission and narrower bandwidth. By further increasing $d_s$ to $\lambda_{rd}/2$, a wider pass band with a sharper roll-off can be achieved, if a transmission dip is acceptable. This shows that different transmission characteristics can be obtained by carefully choosing the separation distance of multi-layer cross-shaped metal mesh filters based on trapped-mode excitation.

Figure 6.18: Simulated spectral responses for the two-layer 0.1 THz trapped-mode metal mesh filter at different separation distances.
6.3.3 Fabrication and measurement

Both conventional and trapped-mode THz metal mesh filters have been fabricated using surface micromachining [331–333]. Figure 6.19a shows the fabrication steps for these filters [333]. First, the substrate is sputter-coated with 35 nm of chrome, acting as a seed layer for the 150 nm gold layer. A 1 µm thick S1813 photoresist layer is then spun on the wafer, soft baked, exposed and developed using standard photolithographic techniques. The two metal layers are etched separately, using selective gold and chrome wet etchants, respectively. The photoresist is stripped using acetone and the sample is finally cleaned with IPA. Microscope images of the fabricated filters are shown in Figure 6.19b [332].

![Fabrication process diagram](image)

**Figure 6.19**: Fabrication of cross-shaped metal mesh filters: (a) surface micromachined processing steps; (b) microscope images [332].

The filters were measured using the turnkey TeraView 3000 terahertz time-domain
spectroscopy (THz-TDS) system. Broadband THz signal can be produced by illuminating an ultrafast laser onto a DC-biased photoconductive switch. A frequency range from 0.01 to 3.5 THz or even wider can be covered. Figure 6.20 gives the time and frequency-domain THz signal generated by TeraView 3000. Here, “scanner position” represents the position of the motorized stage, which is used to control the optical delay line for varying the timing of the detection pulse. The temporal shape of the THz wave is then measured in femtosecond time-resolution. The terahertz spectra can be obtained using Fourier transform of the time-domain data.

![Figure 6.20](image)

Figure 6.20: THz signal from THz-TDS: (a) time-domain; (b) frequency-domain.

A preliminary broadband measurement result from 0.01 to 3 THz is as shown in Figure 6.21 [333]. As expected, the conventional 0.4 THz cross-shaped filter shows a resonant frequency at 0.4 THz; however, a second mode at 0.5 THz is also observed. The improved 0.3 THz trapped-mode metal mesh filter shows better out-of-band rejection, as well as an increase in pass band transmittance.

### 6.4 Conclusions

In this chapter, cross-shaped THz metal mesh filters were introduced, designed, fabricated and measured. The effect of each geometrical parameter on the spectral responses was investigated first. Numerical results showed that the periodicity $G$ of the two-dimensional array will significantly affect the transmission zero frequency
and thus the bandwidth of these filters; by increasing $G$ the BW will be decreased. The resonant frequency $f_r$ and peak transmission $\tau$ will decrease slightly when $G$ increases. Furthermore, $f_r$ is heavily dependent on the cross length $L$: a larger value of $L$ corresponds to a lower $f_r$ and a wider bandwidth.

The effect of thickness of metal layer was also discussed, verifying that its effect can be ignored if it is $h_s \lesssim \lambda_{0}/100$. As $h_s$ increases to $\sim \lambda_{0}/10$, a new peak associated with the waveguide mode of the opening crosses appears. Furthermore, the thickness of the substrate will de-tune the filter towards longer wavelengths. The shift factor depends on the refractive index and thickness of the supporting material. As the thickness of substrate increases, Fabry-Perot modes start moving towards lower frequencies and more transmission peaks will appear within the spectrum of interest. The dependency of transmission on the incident angle was also presented, showing that non-normal incidence will not have a good spectral band-pass response,

Figure 6.21: Preliminary measured results for conventional and trapped mode THz metal mesh filters: (a) time-domain; (b) frequency-domain [333].
due to phase differences of the incident waves on the metal layer.

Metal mesh filters can be stacked to achieve a sharper roll-off slope characteristic. The separation distance and misalignment of two-layer free-standing metal mesh filters were investigated. It was shown that at a $\lambda_{r0}/4$ separation distance, the interaction between filter layers is kept to a minimum, and a good spectral band-pass response with a single transmission peak and narrower bandwidth can be obtained. The substrate-based multi-layer metal mesh filters can be realized using a single substrate, one metal layer on each side. This configuration can effectively decrease the absorption of the substrates. However, the design is challenging, due to alignment issues and the separation distance cannot be easily adjusted. Using another configuration for which two substrate-based metal mesh filters are separated by the air-gap, the resonant frequencies of Fabry-Perot modes can be adjusted, yielding better pass band characteristics within the spectral range of interest.

To develop low cost and mechanically robust THz band-pass filters, a standard 525 $\mu$m thick fused silica wafer was used as the substrate for the design. Trapped-mode excitation, which is formed by adding an inner cross to the original structure, was applied to reduce the effects of Fabry-Perot resonances and improve the out-of-band rejection. Both conventional and trapped-mode metal mesh filters were designed, fabricated and measured within the 0.1 to 0.5 THz range. Preliminary experimental results confirmed that a better out-of-band rejection and higher transmission can be obtained using the trapped-mode excitation. This low cost fabrication process using standard surface micromachining are suitable for mass production.

Band-pass filters employed in current ‘THz Torch’ technology are based on optical coating techniques, whose spectral response depend only on the bulk substrate and its coating material. These filters are not scalable and may not be available on the market if more channels need to be defined for multiplexing systems. With the cross-shaped metal mesh filters, both the resonant frequency and bandwidth can be designed to meet required specifications. Therefore, they are expected to be employed in the future ‘THz Torch’ multi-channel systems, especially in the lower part of the operating spectral range.
7 Conclusions and Further Work

7.1 Summary of Work

The work reported in this thesis aims to develop the recently proposed ‘THz Torch’ concept for secure wireless communications applications. This technology fundamentally exploits engineered blackbody radiation, by partitioning thermally-generated spectral power into pre-defined frequency channels. Filtered band-limited output power in each channel is then independently modulated, transmitted and detected by a thermal detector, creating a robust form of short-range secure communications in the thermal infrared bands.

This concept has been verified experimentally, with its performance increasing from $\sim 10$ bit/s data rate and 0.5 cm transmission distance for the first single-channel proof-of-concept demonstrator, to $>2$ kbit/s and $>10$ cm for the improved system, within an octave bandwidth (25 to 50 THz). Fundamental limitations for such thermodynamic-based communications systems were analysed. It was shown that the thermal time constants of both the transmitter and the pyroelectric detector were the hurdles for further improving the data rate, while the transmission range was limited by the spreading loss of the transmitter.

The ‘THz Torch’ concept can be enhanced by utilizing multiplexing schemes. Frequency division multiplexing (FDM) and frequency-hopping spread-spectrum (FHSS) schemes have been proposed and experimentally demonstrated. The 4-channel FDM working demonstrator can operate at a data rate of $\sim 2.6$ kbit/s over 1 cm with BERs of $< 10^{-4}$, for a 44 mA channel transmitter bias current. Using the same pre-defined channels and hardware, FHSS multiplexing system was implemented, with a maximum data rate of 640 bit/s. Although there is no advantage in
the overall end-to-end data rate, FHSS systems have an enhanced immunity to de-
tection, interception and interference. The integrity of these multi-channel systems
was further evaluated by introducing four experiments.

Moreover, a detailed end-to-end power and noise link budget analysis for the 4-
channel thermal infrared banded-noise wireless communications link was presented
to predict the channel behaviour and channel-to-channel variations. This work will
serve as a valuable engineering tool, enabling system optimisation or fault diagnosis
to be performed.

Scalable THz metal mesh filters were investigated to evaluate the possibility of
employing such band-pass filters in future ‘THz Torch’ systems. To keep a low
fabrication cost and complexity, a standard 525 µm thick fused silica wafer was
used as the substrate. Both conventional cross-shaped and trapped-mode structures
were designed, fabricated and measured within the 0.1 to 0.5 THz frequency range.
However, there are still challenges to overcome before metal mesh filters can be suc-
cessfully used by the ‘THz Torch’ technology. As frequency increases, the structures
will be decreased and this will result in extra fabrication costs. Furthermore, the
substrate material needs to be carefully chosen to have a low loss in the spectral
range of interest.

7.2 Original Contributions

The major contribution and, to a certain extent, the primary novelty of the work
has been the development of the ‘THz Torch’ concept for secure wireless commu-
nications. The author did most of the fundamental theoretical and experimental
work for this newly proposed technology. First of all, both single and multi-channel
proof-of-concept demonstrators were designed, implemented and verified experimen-
tally. Fundamental limits were analysed, and engineering solutions were proposed for
further improving the data rate and transmission distance of such thermodynamic-
based systems. Mathematical models have also been developed for analysing the
two sources of radiation of the transmitter for this particular application, as well as
power link budget of the multiplexing system. With extensive analysis and exper-
ments, the noise and bit error performances of these multi-channel systems were characterised.

The second contribution is the designs and simulations of THz metal mesh filters on electrically thick substrates. It is a challenge to design conventional cross-shaped metal mesh filters on standard 525 µm thick materials, as Fabry-Perot resonances occurs within thick substrates due to multiple reflections. By adding an inner cross to the conventional design, the effect of Fabry-Perot modes can be reduced, making it possible to have a single peak transmission in the desired spectral range. The effects of geometrical parameters and substrates on the spectral responses of these filters were also investigated, and this will serve as a valuable guide for further THz metal mesh filter designs.

7.3 Future Work

7.3.1 Increased density of channels

To date, ‘THz Torch’ wireless communications systems can work at low kbit/s data rates and over a range of tens of centimetres. The former is ultimately limited by the thermal time constants of the receiver, while the latter is due to the spreading loss. To further improve the performance of both single and multi-channel ‘THz Torch’ systems, these limitations have to be overcome.

As shown in (2.5), the thermal time constant of a thermal emitter is directly proportional to its mass and inversely proportional to its surface area (and thus the thickness). Tungsten filaments are not considered good thermal emitters, because of their large thermal time constant and low emissivity. With the glass envelopes, the band-limited output power is further reduced, especially below ~70 THz, where the power transmittance for glass is low. This can be improved by employing commercially available thermal emitters. Such infrared sources have high emissivity (>80%), with similar emitting area (~mm²), can be modulated up to 100 Hz [425], and more importantly, are not significantly more expensive (~£10). If employed in future ‘THz Torch’ systems, more output power can be expected. Furthermore, since they
can be modulated at higher frequencies, mechanical choppers are no longer needed, making the complete system more compact and lightweight.

Another benefit is the increase in overall output power and, therefore, the bandwidth of each pre-defined channel can be reduced while still keeping the same band-limited output power level. As a consequence, more channels can be defined within the same spectral limits (10 to 100 THz in our case). For example, if an overall end-to-end data rate of 8 kbit/s is required, a 4-channel system has to operate at 2 kbit/s per channel, which is not within the optimal operation range for most of the pyroelectric sensors. As a result, this degrades the signal-to-noise ratio of the complete system, due to a lower specific detectivity from the detectors. Introducing more channels, for the same end-to-end data rate, each channel should operate at a lower modulation speed, within the optimal operation range of the PIR detectors, improving the SNR performance.

7.3.2 System with higher level of integration

For practical applications, ‘THz Torch’ systems need to be compact. Without employing bespoke mechanical choppers, the size, system complexity, power consumption, mass and cost of the complete systems will be reduced. However, a higher level of integration is still essential for producing even better working demonstrators.

One possible improvement is to integrate quasi-optical components (e.g., lenses) at both the transmitter and receiver ends. In previous experiments, two KBr collimating lenses were used in the single-channel system and the transmission distance was increased to >10 cm. Although the KBr lenses have high transmittance across the entire spectrum of interest, the main drawbacks are that this material is hygroscopic and not cost-effective. As an low-cost alternative, plastic-based lenses can be employed. Although not having a flat and high transmittance across the entire 10 to 100 THz region, some plastics have transmission windows within part of the spectrum of interest. For example, polytetrafluoroethylene (PTFE, Teflon) can offer several transmission windows, including 1 to 7 µm, 8.5 to 12.5 µm and 22 to 45 µm, as shown in Figure 3.38c [422]. Other plastics such as polypropylene (PP) and
polyethylene (PE) also have wide transmission windows within the IR bands. More importantly, they are much more cost-effective, when compared to KBr lenses, and can be fabricated fairly easily using 3D printing [446] and other techniques. Fresnel lenses can also be designed to further decrease the mass, volume and power loss of lenses. By applying materials with high transmittances, the output power and, thus, the bit rate and transmission distance are expected to be further increased.

### 7.3.3 Integrated frequency-selective thermal emitter

It has been shown in Chapter 6 that inductive cross-shaped metal mesh filters exhibit band-pass characteristics at/near the resonant frequencies. Its complementary structure (capacitive metal mesh filters), as a result, will have a band-stop spectral response. Figure 7.1 shows the structures and corresponding simulated transmission curves for these two types of filters with the same geometrical parameters. As expected, the complementary structures can provide complementary transmission properties. For such resonant structures, a low transmittance can be obtained at \( \sim 0.375 \) THz in this case.

![Figure 7.1: Structures and transmission curves for inductive and capacitive cross-shaped metal mesh filters.](image)

According to Kirchhoff’s law of thermal radiation, the absorptivity of a material is equal to its emissivity at thermal equilibrium. At its resonant frequency, since capacitive metal mesh filters have low transmission, then high absorptivity and emissivity can be expected from this structure when it is used as a thermal emitter, if the reflection is minimized. In [375], capacitive cross-shaped metal mesh structures
were used to produce near-blackbody radiators with frequency-selective properties in the infrared bands. The designed single and dual-band structures, as well as the measured absorptivity and emissivity, are shown in Figure 7.2 [375]. This verifies that such cross-shaped periodical structures are capable of engineering the desired spectral emissivity profile over a large bandwidth; in principle, this technique can be employed in the ‘THz Torch’ systems. As a result, no optical coating or metal mesh filters need to be further applied. Moreover, for these thin metal layers with small mass, smaller thermal time constants can be expected. This enables one to pulse-modulate such thermal emitters by an external fast laser diode. However, the performance (e.g., modulation speed) for implementing these thermal sources still require further investigation.

![Figure 7.2: Structures and measured absorptivity/emissivity for two frequency selective thermal emitters (adapted from [375]).](image)

### 7.3.4 Potential applications for ‘THz Torch’ technologies

In this thesis, the performances of both single and multi-channel ‘THz Torch’ systems for short-range wireless communications have been demonstrated. In the future, potential applications using this thermodynamics-based ‘THz Torch’ technologies need to be explored. One of the proposed applications is the implementation of low cost absorption spectroscopy in the thermal infrared range. Absorption spectroscopy measures the absorption of radiation, as a function of frequency or wavelength, due to
its interaction with a sample. Having increased channel density within 10 to 100 THz frequency range, more channels can be defined, by either metal mesh filters or optical coating filters. This, in turn, will provide more spectral information of a certain material under test. With further advanced signal processing techniques, it may be possible to extract the absorption spectrum of the sample, compare it with a built-in database and determine the material of the sample. The ‘THz Torch’ spectroscopy is believed to be a low cost alternative to existing expensive infrared spectrometers, especially for applications where accurate and high-resolution spectral information of materials are not a priority. Other applications to be investigated include secure RFID, key fobs and secure contactless payment.

7.4 Publications Arising From This Work

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Book Chapter

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