



Heterogeneous Brillouin microwave photonic filters: Towards complete photonic integration

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Declaration

I, *Matthew Garrett*, declare that this thesis is submitted in partial fulfilment of the requirements for the conferral of the degree *Doctor of Philosophy (PhD)*, from the University of Sydney, is wholly my own work unless otherwise referenced or acknowledged. This document has not been submitted for qualifications at any other academic institution.

Matthew Garrett

2023-12-05

Supervisor Statement

I, *Professor Benjamin J. Eggleton*, declare that, in my opinion, this thesis is sufficiently well presented to be examined, and does not exceed the prescribed word limit or any extended word limit for which prior approval has been granted.

Professor Benjamin J. Eggleton

2023-12-05

Abstract

Next-generation radio-frequency (RF) systems are required to operate at higher frequencies, over wider frequency ranges and support broader signal bandwidths due to increased utilisation of the RF spectrum. Although traditional electronic approaches can meet these challenges individually, it is difficult for current RF electronic technology to meet all requirements simultaneously while maintaining a compact footprint, which is essential for Size, Weight and Power (SWaP) constrained applications. Microwave photonics (MWP) is a paradigm-shifting field that combines RF electronics with photonics to create RF devices with unprecedented frequency agility and tunability by leveraging the inherent broadband nature of photonic components. When implemented in compact photonic integrated circuits (PICs), MWP systems enable widely tunable RF devices that support broadband signals in compact chip-scale footprints, thereby overcoming many future challenges that traditional RF electronic systems face.

In particular, MWP devices that use on-chip stimulated Brillouin scattering (SBS), a nonlinear acousto-optic interaction, offer flexibility, reconfigurability and fine spectral resolution unmatched by traditional electronic or optical processing techniques. Despite the compact nature of integrated Brillouin platforms, a current bottleneck to the SWaP reduction of Brillouin-based MWP processors is that additional active and passive optical components required to construct an MWP link are currently implemented using discrete fibre-based components. To overcome this challenge and enable a fully-integrated MWP processor suitable for SWaP-constrained applications, the requisite active and passive components must be integrated in a chip-scale platform with the Brillouin waveguides. In this thesis, we work towards achieving complete photonic integration of Brillouin circuits and demonstrate performance and functionality benefits realised by combining integrated Brillouin processing with on-chip passive components, including silicon nitride (Si_3N_4) microring resonators (MRRs), as well as active silicon photonic electro-optic (E-O) devices.

We leverage the advantages of multi-material integration by combining Si_3N_4 MRRs with As_2S_3 Brillouin waveguides to form the basis of an MWP notch filter. Although As_2S_3 facilitates highly flexible Brillouin processing, As_2S_3 MRRs are limited in their central frequency and extinction ratio tunability. By merging the benefits of low-loss and reconfigurable Si_3N_4 MRRs with the reconfigurable nature of SBS in As_2S_3 waveguides, we demonstrated an MWP notch filter with three independently tunable notches.

Additionally, we present an MWP bandpass filter with low losses (5 dB) and a large RF rejection ratio (40 dB), enabled by on-chip Brillouin processing and a suppressed carrier phase modulation (SCPM) scheme. Brillouin processing was critical in enabling low losses through narrowband optical amplification, as opposed to typical MRR-based passive MWP processing techniques which introduce additional RF attenuation. The SCPM

scheme improved RF link performance by enabling optical carrier-sideband ratio control.

Finally, we heterogeneously integrate As_2S_3 waveguides in an active silicon photonic platform that contains E-O modulators and photodetectors; two key components of the MWP link. This platform overcomes the limitations of previous Brillouin-based MWP processing systems, where the on-chip functionality was exclusively limited to generating SBS gain or loss, resulting in dramatic SWaP reduction. We use this platform as the basis of a compact MWP notch filter that achieves fine spectral resolution of 37 MHz and large 51 dB RF rejection with notch central frequency tunability up to 15 GHz.

By combining the benefits of large Brillouin gain As_2S_3 waveguides with the advantages of additional materials, including silicon and Si_3N_4 , the work presented in this thesis provides a critical step towards the ultimate goal of achieving a fully-integrated Brillouin-based MWP processor with high performance and fine spectral resolution.

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Publications

This thesis with publications is based on the following articles and conference papers. As a requirement of submission, I must identify my contribution to each publication. It should be noted that the co-authors provided supervision, assistance and input across all aspects of the work that led to each publication. I thank them all for their time and effort.

Journal papers

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Supervisor Attestation

As supervisor for the candidature upon which this thesis is based, I can confirm that the authorship attribution statements above are correct.

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Chapter 1

Introduction

Microwave photonics (MWP) is a field that combines the disciplines of photonics and radio frequency (RF) electronics [1]. MWP systems utilise photonic techniques to transport, process or synthesise RF and microwave signals [2]–[5]. This is typically achieved by modulating an RF signal onto an optical carrier using an electro-optic (E-O) modulator. Subsequently, optical techniques are used to process the modulated signal which is then converted back into the RF domain using a photodetector (PD). Any desired RF functionality can be implemented using MWP processing techniques, including RF filters [6]–[10], phase shifters [11]–[13], time delays [14]–[16], frequency converters [17]–[20] and signal generation [21]–[24].

The key benefit of MWP technologies is that they can operate over broad frequency ranges, unmatched by traditional RF electronic processing approaches [25]. This wide tunability stems from the broadband and reconfigurable nature of photonic components that can be tuned over frequency ranges spanning hundreds of gigahertz (GHz). Additional advantages of MWP processing systems originate from the low-loss, lightweight and compact nature of optical fibres compared to coaxial cables that are used to transport RF signals. Coaxial cables offer limited flexibility and exhibit significant frequency-dependent losses [26]. Furthermore, optical fibres are immune to electromagnetic interference (EMI).

The broadband tunability and reconfigurability offered by MWP processing methods are particularly attractive when considering the increasing demand for bandwidth in the RF spectrum [27]. The growing utilisation of the spectrum causes it to become increasingly crowded and, at the same time, results in the operating frequencies of RF systems moving into millimeter wave bands beyond 30 GHz to find less crowded spectral bands. Therefore, RF systems that offer broadband frequency tunability and response reconfigurability are highly sought after as the spectrum becomes increasingly congested [5], [25].

In particular, filters are key components in broadband and reconfigurable RF systems as they attenuate the many undesired and interfering signals in the spectrum to ensure desired signals can be received [6], [7], [27]. Furthermore, significant opportunities exist for MWP filters to overcome the challenges of operating in the increasingly congested spectrum, as current RF filtering technologies cannot simultaneously meet future requirements for Size, Weight and Power (SWaP) and frequency tunability while maintaining satisfactory RF performance [28].

Although fibre-based MWP systems offer the attractive qualities outlined above, fur-

ther SWaP advantages have been achieved by implementing components used in MWP systems in a compact photonic integrated circuit (PIC) [25]. Many MWP processing systems have been demonstrated using chip-based methods, including micro-ring resonators (MRRs) [9], [15], [16], [29], [30], tapped delay lines [31]–[34], reconfigurable mesh networks [35]–[38] and stimulated Brillouin scattering (SBS) [10], [11], [20], [39]–[45]

On-chip SBS is particularly attractive for MWP processing, as it offers fine spectral resolution (10's MHz) that is unmatched by many other optical processing techniques mentioned above. For filtering applications, fine spectral resolution is critical to attenuate unwanted signals while preserving nearby signals of interest [27]. At the same time, the bandwidth of the SBS response can be increased by broadening the spectrum of the optical pump that is used to induce the gain resonance. This has been used to form the basis of MWP filters with reconfigurable bandwidths [44]–[48], as well as true time delays, which apply a dispersive phase slope over a broad frequency range [49], [50]. Furthermore, the SBS response can be tailored in magnitude by shaping the amplitude of the pump spectrum. In turn, this reconfigurability is translated to the RF domain, enabling versatile RF signal processing functionality that can be dynamically configured based on the desired use case.

Existing chip-based Brillouin MWP systems deliver dramatic SWaP reduction by inducing SBS gain in compact PICs, rather than kilometres of optical fibres [25], [51]. However, existing Brillouin-based MWP systems rely on a combination of chip-scale and discrete optical components to achieve MWP processing [20], [39], [42]. This limits the minimum achievable SWaP, stability, and production scalability of the system and is a barrier to the deployment of Brillouin-based MWP processing systems in real-world scenarios. Therefore, integrating Brillouin processors with passive and active optical components, including the key components of MWP processing systems, on a single integrated platform is critical to unleash the full potential of compact and high-performance chip-scale Brillouin MWP processors through photonic integration.

In this thesis, we work towards realising fully-integrated Brillouin-based MWP processors. We focus on Brillouin-based MWP filtering applications, as they are key building blocks of all RF systems. First, we explore the advantages of combining the flexibility of the generation of Brillouin gain in As_2S_3 chalcogenide waveguides with the low propagation losses of silicon nitride (Si_3N_4) micro-ring resonators (MRRs). We use on-chip SBS and multiple Si_3N_4 MRRs, leveraging their tunability to enable an MWP notch filter with three independently tunable notches, compared to previously reported As_2S_3 MRR platforms which only supported a single notch response [45]. We then report a chip-based Brillouin MWP bandpass filter that achieves low passband losses of 5 dB by utilising a suppressed-carrier phase modulation (SCPM) scheme. The SCPM scheme improves the RF link gain by 20 dB, compared to traditional MWP notch filters using phase modulation. Finally, we heterogeneously integrate As_2S_3 Brillouin waveguides in an active silicon photonic platform, combining critical fine spectral resolution of Brillouin processing with on-chip silicon E-O components in a single chip for the first time. We use this platform as the basis of an MWP notch filter with a narrow 37 MHz linewidth and a large 51 dB rejection ratio. Additionally, we integrated silicon photonic circuits to implement a SCPM scheme to increase the RF link gain of the filter. The simultaneous demonstration of link gain enhancement with silicon photonic components and fine spectral resolution processing in As_2S_3 Brillouin waveguides highlights the benefits of heterogeneous inte-

gration, enabling fine spectral resolution MWP processing in a compact footprint.

The thesis structure is as follows and is visually represented in Fig. 1.1.

Chapter 2 overviews integrated MWP processing systems. This includes a historical perspective of the developments of RF electronics and MWP. We also provide an overview of integrated photonic platforms, with a particular focus on MWP processing applications, as well as an introduction to on-chip optical processing components.

Chapter 3 outlines the theory of MWP processing systems. We outline modulation formats used for microwave photonic signal processing and the way each modulation format influences the RF output. Finally, we outline the key performance metrics of microwave photonic links.

Chapter 4 reviews chip-based Brillouin MWP processors. This includes a theoretical outline of the fundamental interactions that facilitate stimulated Brillouin scattering. We then provide a review of Brillouin-based chip-scale platforms that provide MWP processing functionality. Finally, we provide a comprehensive review of the state-of-the-art chip-based Brillouin MWP processors and highlight the advantages of SBS for MWP applications by achieving fine spectral resolution and a high degree of RF processing reconfigurability.

In Chapter 5, we present an MWP notch filter that demonstrates three RF notches, each of which is independently tunable in frequency. By cascading silicon nitride (Si_3N_4) MRRs with As_2S_3 Brillouin waveguides, we achieve increased spectral resolution with 3-dB bandwidth of 500 MHz. This is an improvement compared to previously reported MRR and Brillouin-based IMWP notch filters on a monolithic chalcogenide platform that was limited to a 3-dB bandwidth of ≈ 3 GHz and a single notch response limited by the losses in the chalcogenide ring resonator. The interferometric notch filtering scheme enabled deep RF rejection up to 50 dB.

Chapter 6 details a low-loss and large-rejection IMWP bandpass filter. We increase the RF link gain by 20 dB when compared to a traditional phase modulation scheme using a suppressed carrier optical modulation scheme. The filter exhibits a deep out-of-band rejection ratio of 40 dB, using only 14 dB of SBS gain, as the modulation scheme facilitates destructive RF interference in the filter stopband. We synthesise three independently tunable filter passbands which range from 1-20 GHz. The passband bandwidth can be controlled in the range 30-500 MHz.

Chapter 7 reports a novel MWP processing platform that combines large gain chalcogenide Brillouin waveguides and active silicon photonic electro-optic components on a single chip, bringing the essential functions of fine-spectral resolution MWP processing and E-O conversion into a single chip-scale platform for the first time. Using Brillouin processing in As_2S_3 chalcogenide Brillouin waveguides, we achieve fine frequency selectivity of 37 MHz and large RF rejection ratio of 51 dB. Furthermore, we extend the work from Chapter 6 and use a SCPM scheme to facilitate RF link gain enhancement by a factor of 16 dB.

Chapter 8 summarises the work we presented in this thesis and provides an outlook for future work for Brillouin-based MWP processors. In particular, we provide perspectives on novel platforms to explore integration with high-gain As_2S_3 waveguides to enable high-performance Brillouin-based MWP processors. Additionally, we outline pathways to achieve a fully-integrated Brillouin MWP processor.

Appendix A provides a perspective on the application of a fully-integrated Brillouin-

based MWP processor in phased array antenna (PAA) systems.

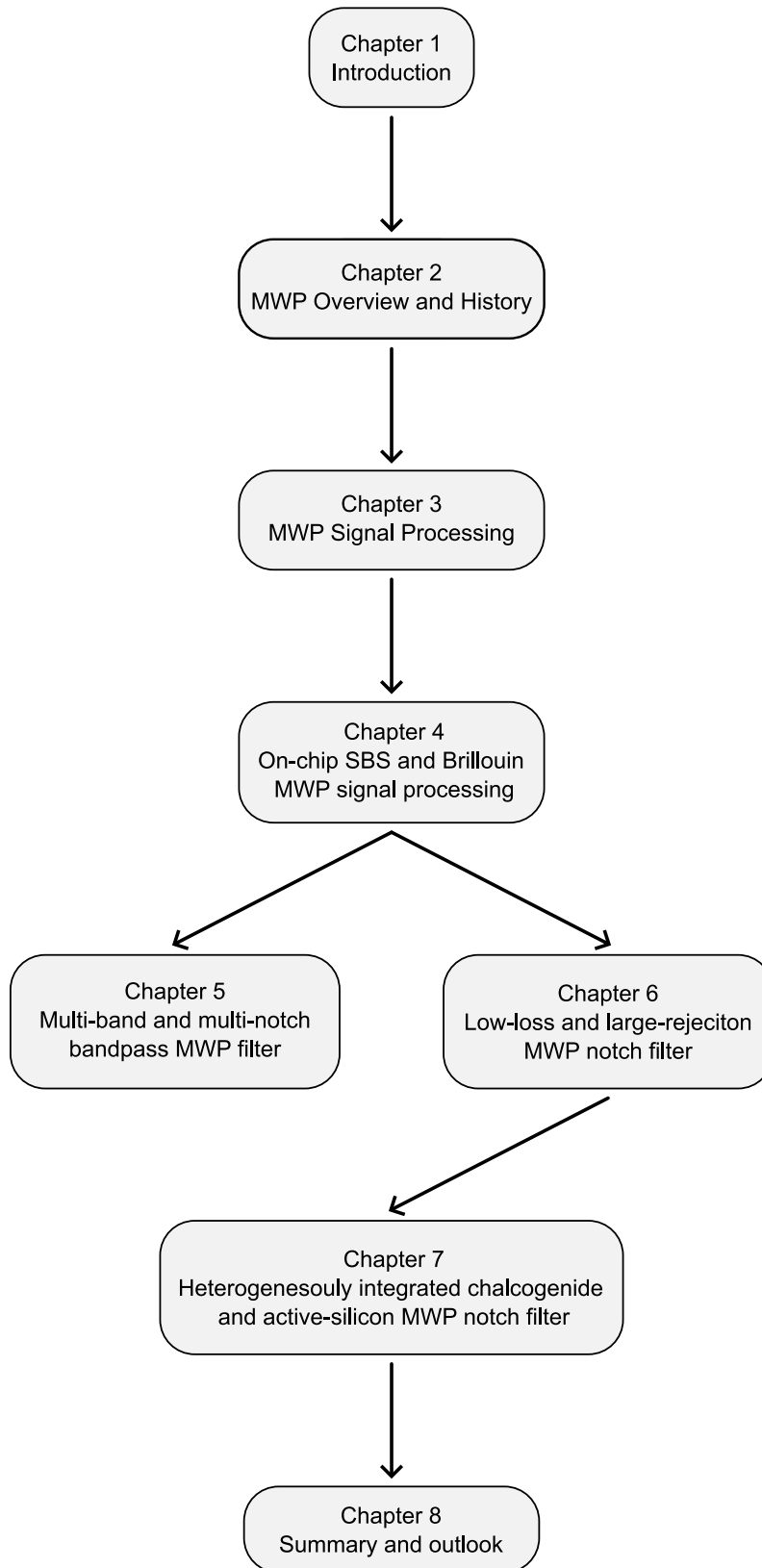


Figure 1.1: Structure and relationship between the chapters of this thesis.

Chapter 2

Integrated Microwave Photonics: Overview and Historical Perspective

This chapter gives an overview of microwave photonics (MWP) and its advantages. Additionally, we provide an overview of integrated photonic platforms and their application in MWP systems. It is aimed at readers who have a general background in optics and/or electronics but may not be familiar with MWP.

2.1 Radio Spectrum: A historical perspective

The electromagnetic spectrum has played a pivotal role in shaping modern society for over a century. It has paved the way for significant advances in applications including communications, sensing and navigation that are central to modern life. The spark gap generator, invented by Guglielmo Marconi, is considered to be the first device that enabled wireless communications [52]. In 1896, the historic demonstration of wireless radio transmission from the Isle of Wight to a boat 18 miles offshore revolutionised long-distance communications [53]. Subsequent developments in radio technology facilitated improvements to communications, navigation and sensing systems, all of which enabled significant advancements in military and civilian applications [54], highlighting the usefulness of harnessing the spectrum in a range of applications and propelling the future development of RF electromagnetic systems.

Today, the electromagnetic spectrum has seen an enormous increase in usage. This is linked to the development of electronic integrated circuits (ICs), a key driver in reducing the form factor and cost of wireless communications hardware while also increasing their production volume and availability, thus enabling the widespread deployment of connected and wireless sensors to be embedded in everyday products [55]. As a result, wireless radio devices underpin cellular networks, satellite communications and sensing applications. In particular, electromagnetic sensing spans many diverse fields, including remote agricultural monitoring [56], slope stability monitoring [57], driver assistance in automotive applications [58] and biosensing [59]. Additionally, global satellite navigation systems (GNSS) [60], including the ubiquitous global positioning systems (GPS), transmit accurate timing and location information over radio frequency electromagnetic waves that are critical for navigation, as well as precision timing for trading applications [61].

However, most commercial and domestic wireless devices operate in the frequency range below 6 GHz. Because of the increased usage of the finite available bandwidth in RF spectrum, it has become overcrowded. As a result, interference between the many RF signals in the spectrum can create significant issues with signal fidelity and even cause wireless systems to cease operation [62], [63]. More efficient spectrum access techniques have been employed to combat this issue, such as spread-spectrum techniques to reduce interference using different frequencies or spreading codes [63]. Additionally, beamforming methods have been harnessed by directing the radiation pattern from antenna arrays to reduce interference in the spatial domain [64]. At the same time, many electronic devices have been moving to higher frequency bands, for example 66 GHz and 81 GHz bands [65] and beyond, which are less crowded and offer more available bandwidth.

Due to the growing congestion in the RF spectrum, filters are key components in RF systems to maintain signal fidelity by attenuating unwanted signals [27]. RF filters require spectrally selective responses to attenuate undesired signals in the crowded RF spectrum, which can cause performance degradation, whilst preserving the magnitude of nearby signals of interest [7]. Additionally, applications place stringent SWaP requirements upon RF systems to the extent that devices are precluded from deployment if the SWaP requirements are not met in critical scenarios such as airborne and spaceborne applications. This emphasises the importance of realising RF systems with compact form factors. However, traditional electronic implementations of RF filters cannot simultaneously achieve the required broadband frequency tunability, fine spectral resolution and response reconfigurability in a suitably compact form factor [28]. For example, filters based on surface acoustic waves (SAW) and bulk acoustic waves (BAW) achieve fine frequency selectivity and are prevalent in many mobile devices owing to their compact form factor [66], they struggle to operate into millimetre wave bands (>30 GHz) and beyond, owing to intrinsic acoustic losses and challenges in the acoustic wave transducer design at high frequencies [67], [68]. Although yttrium iron garnet (YIG) filters exhibit widely tunable responses, their form factor is too large for deployment in SWaP-constrained applications [69], [70]. Similarly, switched filter banks offer highly reconfigurable responses but cannot be continuously tuned in frequency and require parallel hardware to achieve broadband frequency tunability [71]. Therefore, there is a great unmet need to implement spectrally selective and reconfigurable electronic systems that can be tuned over a broad frequency range for next-generation RF systems.

2.2 Microwave Photonics: Enabling Agile RF Systems

2.2.1 Overview and Applications

By leveraging the broadband nature of optoelectronic components, MWP systems overcome many challenges faced by traditional RF electronics and enable microwave systems with frequency agility and flexibility, unmatched by conventional electronic methods. The fundamental architecture of an MWP link is shown in Fig. 2.1 and operates in the following manner: (a) an RF input signal is (b) modulated onto a continuous-wave (CW) laser where the RF spectrum is translated to the optical sidebands of the laser, as shown in inset (i). (c) The modulated optical signal then undergoes processing in the optical domain, as shown in inset (ii), before (d) a photodetector (PD) converts it back to the RF domain

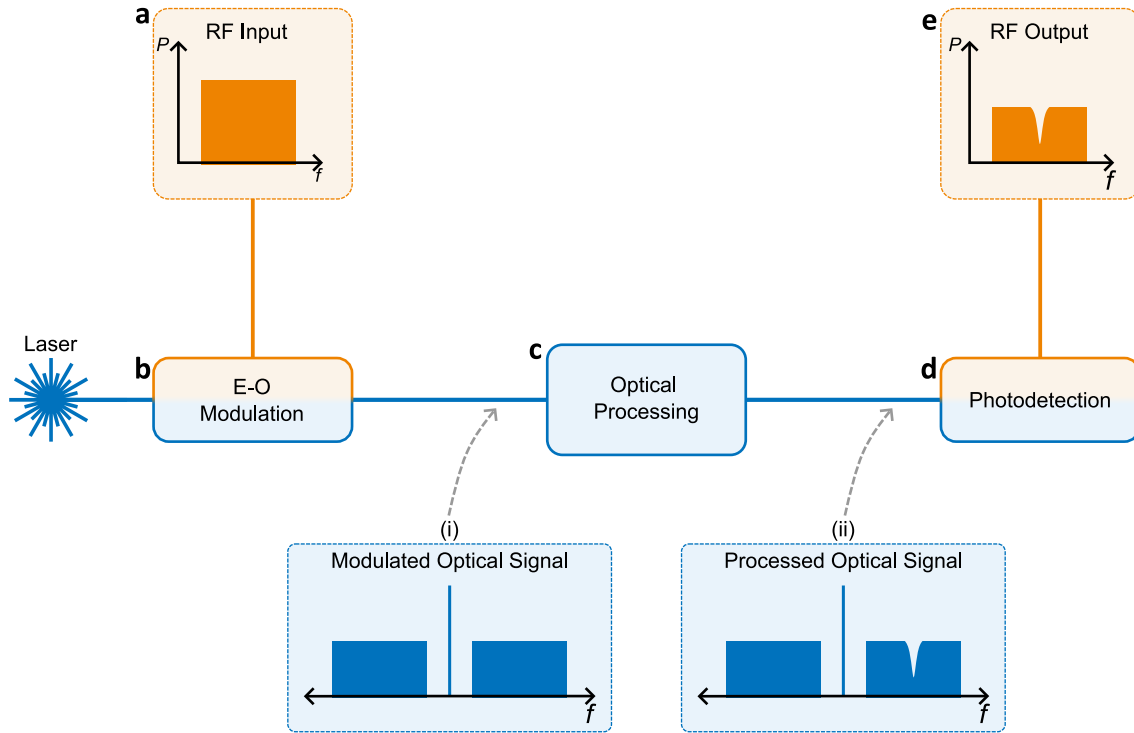


Figure 2.1: Simplified schematic of a microwave photonic link. (a) An RF signal is (b) modulated onto a continuous wave laser. The modulated signal is then (c) processed optically before (d) conversion back to the RF domain, where the (e) RF output signal is formed.

through square law photodetection [25], [72], forming the (e) RF output.

Although a notch filter is depicted in Fig. 2.1, MWP systems can use optical processing techniques to implement a wide variety of RF functionalities that includes filtering [27], as well as, phase shifting [11]–[13], true time delay [14]–[16], frequency conversion [17]–[20], and signal generation [21]–[24].

Furthermore, MWP-based processing offers specific advantages that cannot be achieved using traditional electronic methods. For example, the low losses (0.2 dB/km) of optical fibres allow RF signals to be sent over distances that would be impractical using coaxial cables. This is attractive for applications such as radio astronomy [73], exemplified in Fig. 2.2 (a), as the low losses of the fibres enable radio telescopes to be placed in remote areas with a low level of background interference. The RF signal received by the telescope can be transported over fibre to a central office for processing [73], [74]. Optical fibres are also immune to electromagnetic interference (EMI) and are much lighter than coaxial cable. As such, it is advantageous for MWP links to be deployed in scenarios where interference must be suppressed and weight must be minimised, such as in an aircraft as exemplified in Fig. 2.2 (b).

The lightweight and compact nature of optical fibres has made MWP processing techniques desirable for deployment in SWaP-constrained applications, such as phased array antenna (PAA) systems, and has been a focus for the MWP community since the 1980s [77]. Furthermore, the broadband nature of photonic systems, as well as the ability to implement true time delay (TTD) beamforming techniques are well suited to overcome the undesired effect of beam squint [78]–[84]. Beam squint is the phenomenon where

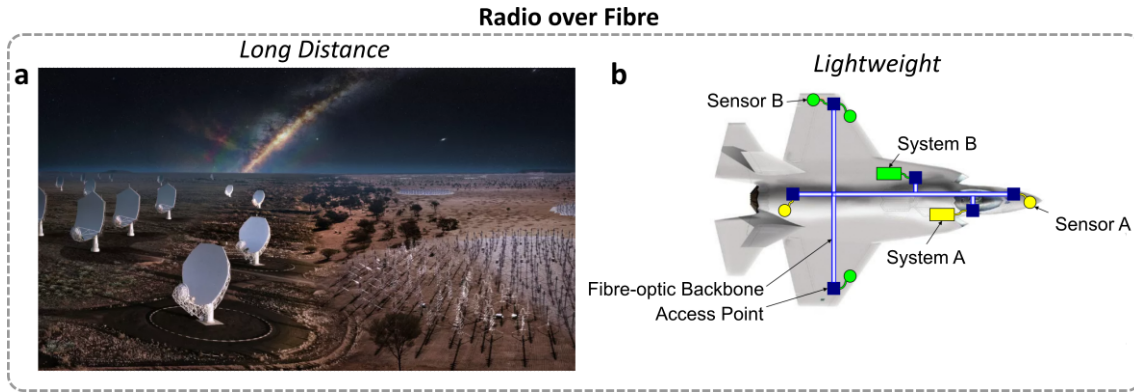


Figure 2.2: Radio over fibre applications. (a) Radio over fibre enables RF signals to be transported over long distances owing to the low propagation loss of optical fibre. This is attractive for applications such as radio astronomy. (b) The lightweight of optical fibres is advantageous for distributing RF signals in scenarios where weight is critical such as in aeroplanes. Figures from (a) [75], (b) [76].

the beam direction changes as a function of frequency [85], [86]. It occurs when phase shifts are used to steer the radiation pattern of broadband signals within PAAs and does not occur with TTD beamforming [81]. Recent advances in photonic integration have further reduced the SWaP of photonic TTD beamformers, further highlighting advantages of MWP systems, exemplified by the compact millimeter-scale TTD beamformer presented in Fig. 2.3 (a-b) [15]. The ability to use this device for beam steering is exemplified in Fig. 2.3 (c), which indicates a beam pattern for three different steered angles.

Photonic systems also facilitate high-performance RF signal synthesis functions. Photonic approaches of RF signal generation offer the ability to generate high-quality RF signals with low phase noise and high spectral purity over a wide frequency range [5], [25]. The optoelectronic oscillator (OEO) [87]–[89] is an effective method for producing such signals, utilising the low-loss of optical fibres to create a hybrid electrical-optical cavity with a high-quality factor (Q) that generates ultra-stable RF signals. An example of a compact chip-scale OEO is depicted in Fig. 2.4 (a-b). Owing to their broadband tunability, highlighted in Fig. 2.4 (c), and low phase noise, OEOs are attractive for use in future communication systems and high-performance radars [21].

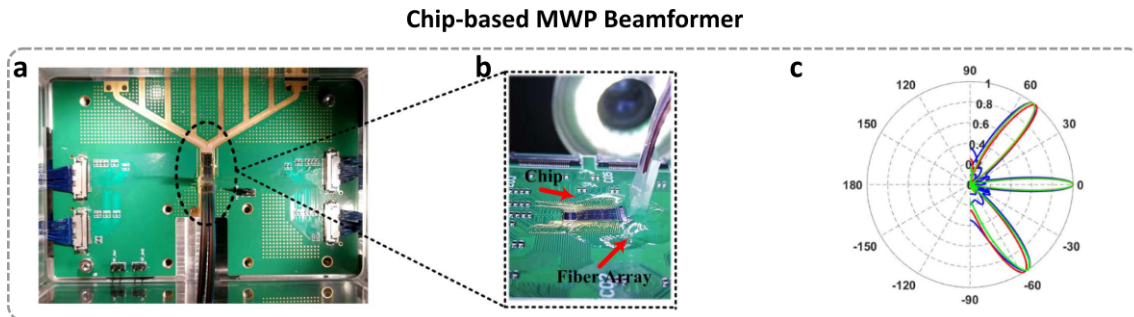


Figure 2.3: MWP beamformer. (a-b) Image of an integrated MWP true time delay beamformer. (c) Beam steering was demonstrated for three different angles. Figures from [15].

From the analysis above, MWP is well-positioned to play a critical role in future RF and wireless applications by enabling a new generation of RF and microwave systems with unprecedented tunability, flexibility and frequency agility.

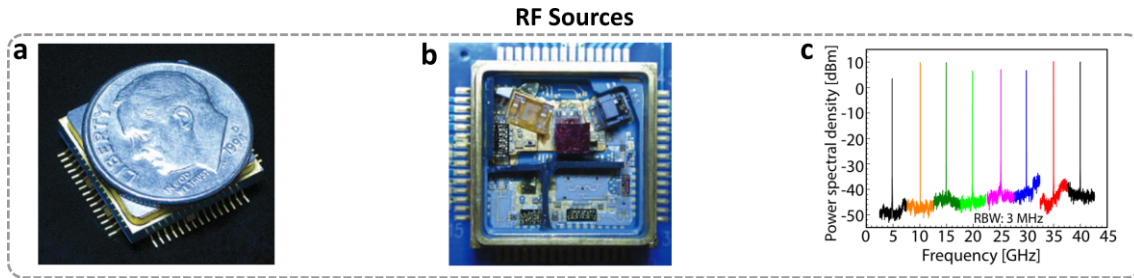


Figure 2.4: Optoelectronic oscillator (OEO). (a-b) Image of an integrated OEO. (c) Demonstration of the wideband RF frequency tunability offered by photonic systems. Figures from (a-b) [87] (c) [22].

2.2.2 Historical Perspective

The field of MWP traces its origin to the invention of the laser in 1960 [90], [91], which revolutionised the field of optics and photonics. However, the first free-space analog optical link was not created until the development of the continuous wave (CW) semiconductor laser in 1962 [92]–[95]. The key innovation was that the intensity of CW semiconductor laser was able to be directly modulated by an RF electrical field [2], [95]–[97]. Although, MWP links initially planned to use free-space optics, the development of low-loss fibres in the 1970s [98] soon led to the widespread adoption of optical fibres across all optical communication devices, including within MWP systems [2].

As technology continued to develop and offer components with improved performance and commercial availability, the performance of MWP links improved correspondingly. In terms of photodetectors, initial PIN diode photodetectors were replaced with uni-travelling-carrier (UTC) photodiodes [99], [100] after their first demonstration in 1997 [101]. UTC photodiodes offer benefits for microwave photonic applications compared to traditional PIN diode detectors [102] including larger saturation photocurrents [103], a higher degree of linearity [104] and large 3-dB bandwidths, with a 310 GHz device demonstrated in 2000 [105]. We outline how these characteristics improve the MWP link performance in Chapter 3.

The bandwidth of early MWP links was limited by the electro-optic (E-O) bandwidth of the direct modulation approach, where the input RF signal modulated the laser drive current [106], [107]. Despite many attempts, modulation bandwidths above 20 GHz could not be achieved using the directly modulated approach [2], as a relaxation oscillation between the population states in the gain medium and photons in the laser cavity typically limits the modulation bandwidth of semiconductor lasers [108]–[110]. The direct modulation approach also increases the relative intensity noise (RIN) of the optical signal [99], which is detrimental to the RF link performance.

In 1975 Martin reported the first integrated waveguide-based optical modulator [111] and was the first foundational step towards implementing modern E-O modulators that exhibit broadband operation [112], [113]. The device was based on a ZnSe crystal that used

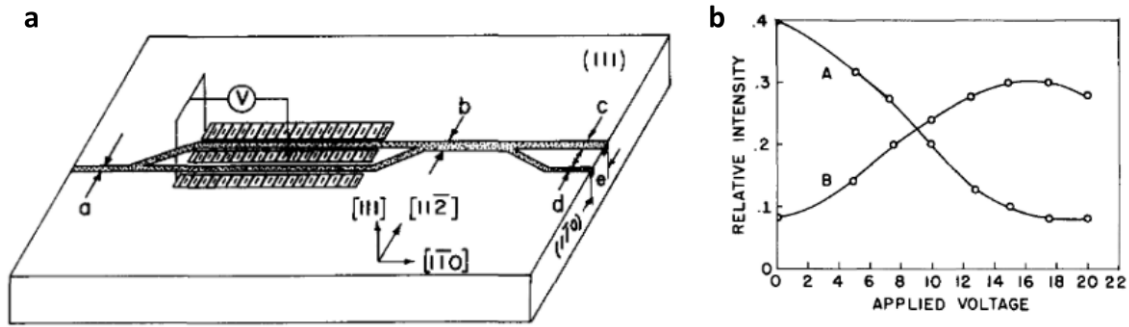


Figure 2.5: First integrated waveguide-based E-O modulator. (a) Device schematic (b) Control of optical intensity by applying a voltage. Figures from [111].

cadmium diffusion to achieve an index contrast to guide the optical mode and consisted of optical waveguides that formed a Mach-Zehnder interferometer, as well as electrodes that are used to apply an electric field to the waveguide. A high-level schematic of the device is shown in Fig. 2.5 (a). Control over the output signal intensity was achieved by applying a voltage to change the effective length of the arm of an interferometer through the Pockels effect. Control of the optical intensity through the applied voltage is shown for each waveguide in Fig. 2.5 (b).

Modulators that are implemented as a separate component to the laser source are referred to as external modulators. External modulators are almost exclusively used in MWP systems over the initial directly modulated approach. This is because external modulators offer advantages over the direct modulation approach, including large E-O bandwidths, with current state-of-the-art E-O modulators offering 3-dB E-O bandwidths exceeding 500 GHz [114], and they do not increase the laser relative intensity noise (RIN) during operation. At present, many forms of external E-O modulators exist. We introduce several common chip-based E-O modulators in Section 2.3.3 and outline how key parameters of E-O modulators affect the gain, linearity and noise performance [115], [116] of an MWP link in Section 3.3.

Despite continual improvements to the components used in MWP links, their performance did not meet the required performance in terms of gain, noise and linearity [95] that is demanded by applications and MWP link designers would compensate for any lack in performance using electronic amplification. To some extent, the RF loss and noise performance can be improved by electronically amplifying the input RF signal before it is modulated onto the optical carrier. However, it is difficult to achieve sufficient RF link performance using this approach, as RF amplification often comes at the expense of reducing linearity and input power handling [117].

It became clear that attention must be given to the performance of MWP links to ensure the widespread deployment of MWP systems [5]. Although the required link performance is dictated by the specific application, it is generally accepted that MWP links must have *intrinsic* link performance (i.e. without RF amplification) with a noise figure below 10 dB and linearity (SFDR₃) above 120 dB·Hz^{2/3} [5], [118]. A major focus of the MWP subsequently turned to improving the *intrinsic* link performance [119]. Early models of MWP links using intensity modulation and direct photodetection (IMDD) were developed [120]–[122] and experimentally verified [117]. With notable demonstrations from 2007

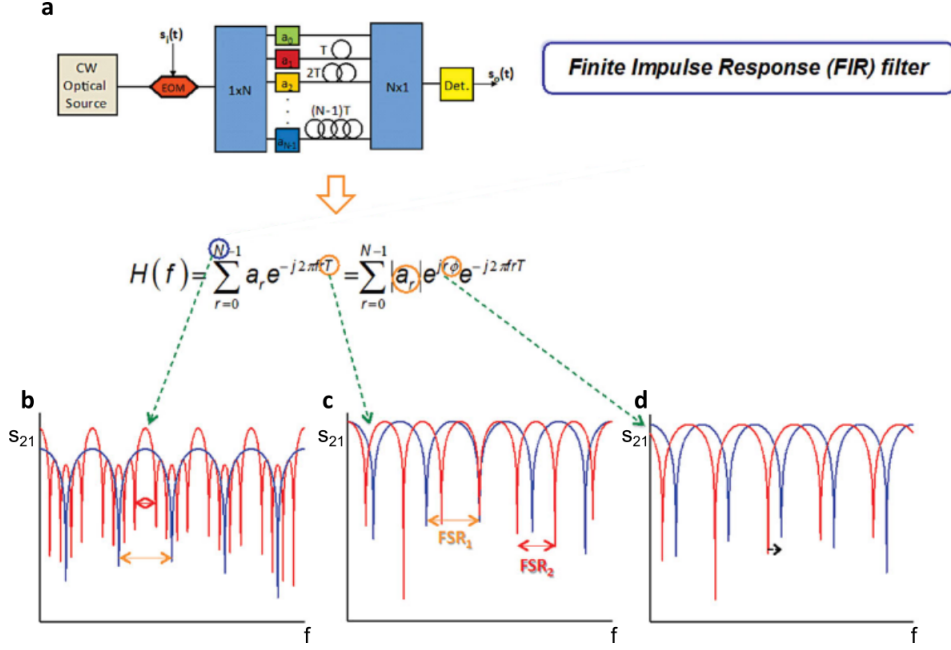


Figure 2.6: Discrete optical processing of microwave signals. (a) High-level implementation of a tapped delay line MWP system. (b) The effect on the number of filter taps on the filter response bandwidth. (c) The effect of the optical delay applied to each tone on filter FSR. (d) The effect of the phase shift applied to each optical tone on the filter central frequency. CW: Continuous-wave, EOM: Electro-optic modulator, Det: Photodetector. FSR: Free-spectral range. Figure modified from [5].

achieving optimised intrinsic link gain as high as 17 dB [123] and noise figures as low as 3.4 dB with a large dynamic range of 120 dB.Hz^{2/3} [124], using large optical powers and the low-biasing link optimisation technique, which we outline in Sec. 3.3.

Some of the first demonstrated MWP systems benefited greatly from the above-mentioned RF link performance enhancement and targeted RF signal distribution over long distances [2]. For example, the first proof of principle demonstration of transmitting television signals over fibres in 1976 noted that noise was observed in the transmitted video stream due to a poor signal-to-noise ratio [125]. However, by the 1990s, photonic technology was widely adopted and successfully distributed video signals on a large scale [126].

Investigations of RF signal processing functionalities using MWP systems were performed in parallel to the above-mentioned RF link optimisation research [127]–[130]. The seminal work of Willner and Van de Heuvel in 1976 [131] demonstrated that optical fibres are attractive candidates for implementing RF signal delays over a broad frequency range. Optical delay lines have been widely used since this time to implement discrete optical processing of microwave signals (DOPMS) [6], [26], often referred to as tapped delay lines.

The tapped delay line method was particularly prevalent among early MWP systems. It is analogous to discrete time signal processing used in digital signal processors (DSP) [132], where a weighted sum of delayed input signals are combined to form the system output. The transfer function of the resultant signal forms a passband for frequencies where the signals constructively interfere and a stopband is formed for frequencies that result in destructive interference. The response can be modified by changing the delay

and relative amplitude contributions of each discrete optical tone. Typically, the system transfer function takes the form [4], [5], [27]

$$H(\omega) = \sum_{r=0}^{N-1} a_n e^{-jr\omega T} = \sum_{r=0}^{N-1} |a_n| e^{j\phi} e^{-jr\omega T} \quad (2.1)$$

where $a_n = |a_n|e^{j\phi}$ is the complex weighting applied to each tone, T is the incremental delay applied to each filter tap and N is the number of filter taps.

Fig. 2.6 highlights the common high-level architectures that implement the tapped delay line technique. Fig. 2.6 (a) depicts a signal processing architecture that enables MWP filtering by splitting a single tone optical source into N components. The signal in each arm undergoes complex amplitude shaping and experiences a delay. Upon recombination and photodetection, the transfer function represented in eq. 2.1 is implemented when the resultant photocurrent from each filter tap is summed at the photodetector output. Fig. 2.6 (b) shows that the bandwidth of the filter response can be changed as a function of the number of filter taps. Increasing the number of taps reduces the bandwidth. Fig. 2.6 (c) shows that changing the successive delay between each filter tap changes the free-spectral range (FSR) of the filter response and Fig. 2.6 (d) shows that the phase shift applied in each arm can translate the frequency response of the filter.

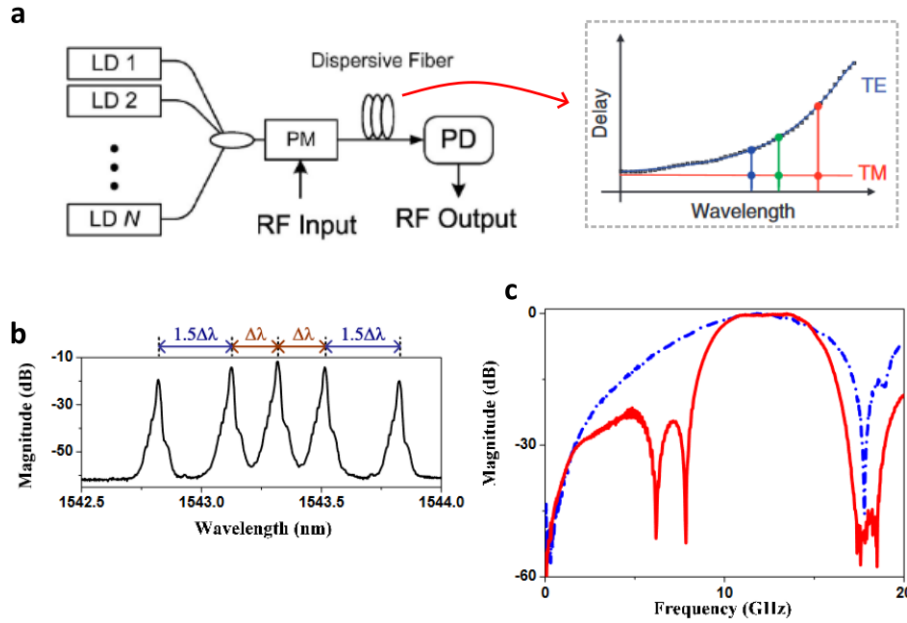


Figure 2.7: Discrete time MWP filter demonstration. (a) System architecture. A multi-tone optical source consisting of multiple laser diodes (LD) is combined and modulated by an RF input. The modulated signal propagates through a dispersive fibre, where each wavelength component experiences a different delay, as shown in the inset. The signal is converted back to the RF domain using a photodetector (PD). Inset: Schematic of wavelength-dependent delay of a dispersive fibre. (b) The optical spectrum used to implement the MWP filter which consists of multiple tones. (c) Measured RF filter response. The blue trace corresponds to the measured PM-IM conversion. The red trace corresponds to the measured MWP filter response. Figures from (a) [4]. Inset of (a) [31] and (b-c) [133].

Similarly, MWP processing systems can be implemented using discrete optical processing methods by using a highly-dispersive element to provide a wavelength-dependent delay to each optical tone. An advantage of this approach over the architecture presented in Fig. 2.6 (a) is that a single component can delay many optical tones, as opposed to requiring one delay line per tone [4].

Fig. 2.7 (a) shows a high-level architecture of a laser array and dispersive fibre used to implement an MWP filter. A key to understanding this principle is that the dispersive fibre delays each optical tone as a function of its wavelength, as shown in the inset of Fig. 2.7 (a). The optical spectrum consisting of multiple optical tones used to implement the MWP filter is shown in Fig. 2.7 (b) and the measured broadband flat-top RF filter response is shown by the red trace in Fig. 2.7 (c). Suitable components for processing each optical tone include dispersive fibre [133]–[135], fibre grating arrays [136]–[138], and stimulated Brillouin scattering (SBS) [139]–[141].

SBS is a nonlinear acousto-optic interaction that has gained significant attention as a method for implementing MWP signal processing in its own right [25], [27], [51], [142], owing to its narrowband linewidth (10's MHz), wideband frequency tunability and highly reconfigurable response [143]–[145]. We provide a brief introduction to the concept of SBS in Section 2.3.4 and provide an in-depth overview of SBS and its role in MWP applications in Chapter 4.

The ability to generate the narrowband Brillouin interaction in optical fibres led to many MWP system-level demonstrations with unprecedented spectral resolution, including MWP filters [46]–[48], [140], [141], [146]–[150], phase shifters [151]–[153], true time delays [49], [154], [155] and microwave sources [156]–[159].

The Brillouin resonance is widely tunable over a large frequency range, enabling widely tunable RF signal processing. This is depicted in Fig. 2.8 (a), where a passband, with a 20 MHz fine spectral resolution from the narrow linewidth of the Brillouin gain resonances, is tuned from 1 GHz to 20 GHz [149]. Additionally, the spectral shape of the Brillouin resonance can be tailored by modulating the spectrum of the pump laser, which generates the Brillouin gain [44], [147]. This is shown in Fig. 2.8 (b), where the passband bandwidth of an MWP filter is reconfigured from 250 MHz to 1 GHz by broadening the pump spectrum [147].

The highly reconfigurable nature of the Brillouin resonance lends itself to applications such as RF spectrum equalisation. In particular, alternative MWP processing techniques including the discrete-time methods that were outlined above, and micro-ring resonators would struggle to match the reconfigurability and spectral resolution offered by SBS in spectral equaliser applications. By tailoring the SBS gain spectrum, Grafer *et al.* demonstrated a Brillouin-based equaliser that was able to compensate for ripple and roll-off within the response of an RF system. The orange trace in Fig. 2.8 (c) represents the raw RF transfer function, which contains significant ripple. The blue trace in Fig. 2.8 (c) represents the transfer function after Brillouin equalisation was applied. The equalised frequency response exhibits significantly reduced roll-off and ripple than the originally measured response.

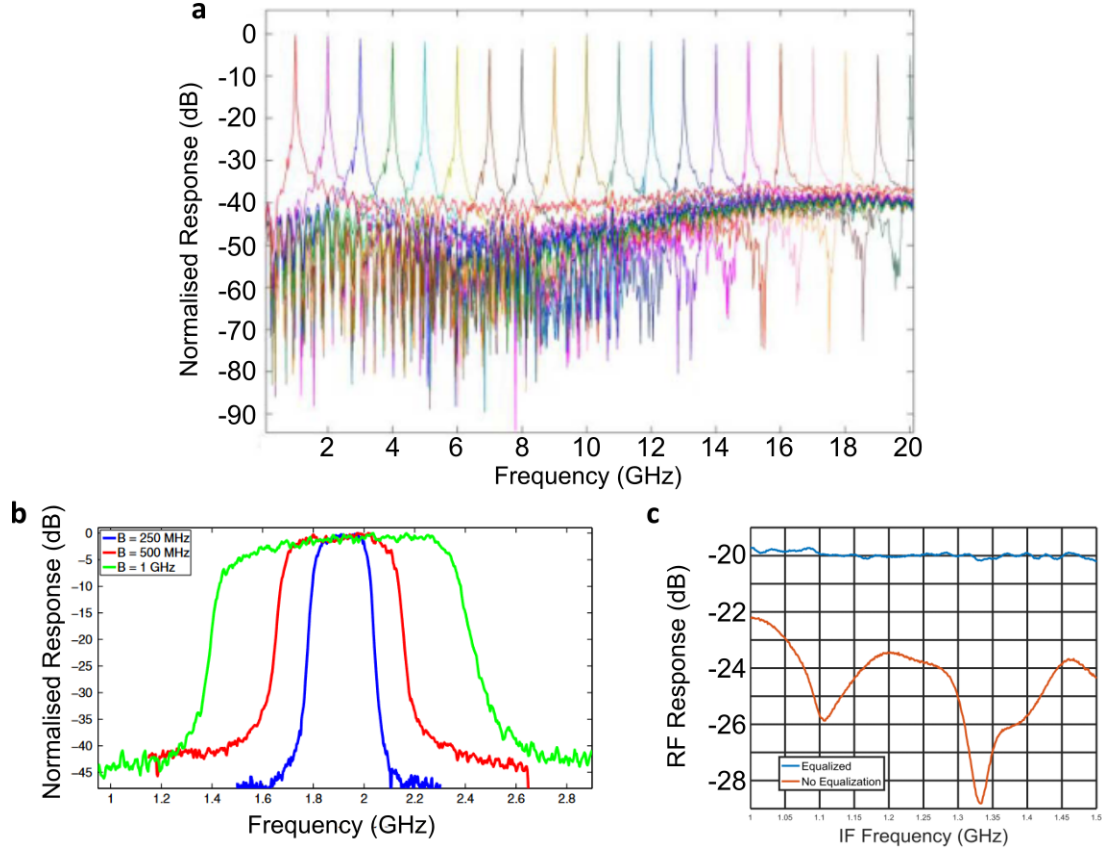


Figure 2.8: Brillouin-based MWP systems. (a) Passband tunability of a Brillouin MWP filter. (b) Broadening of the bandwidth of a Brillouin MWP filter. (c) Brillouin-based RF equalisation. Figures from (a) [149], (b) [147] and (c) [160].

2.3 Integrated Microwave Photonics

2.3.1 Development Timeline

Initial MWP processing systems relied exclusively on discrete fibre-based components, which limits their minimum SWaP. Owing to the significant SWaP reduction enabled by photonic integration, there are many ongoing global research efforts to implement an MWP systems exclusively using chip-scale components, rather than relying on discrete fibre-based components [5], [25].

Fig. 2.9 outlines a timeline that represents key developments in the field of microwave photonics. The development of optical sources, E-O components, MWP link performance improvement techniques and fibre-based processing that we outlined above in Section 2.2.2 laid the essential groundwork for developing integrated MWP signal processing devices and are indicated in Fig. 2.9. We now overview the development of integrated MWP systems.

Fully-integrated MWP systems must include on-chip E-O components. The initial demonstration of an integrated waveguide-based modulator, Fig 2.9 (a), highlighted the potential for SWaP reduction through integration. The subsequent development of the plasma-dispersion modulator by Soref and Bennett in 1987, and subsequent improvements enabling modulation bandwidths above 1 GHz in 2004 [161], Fig 2.9 (b), were key

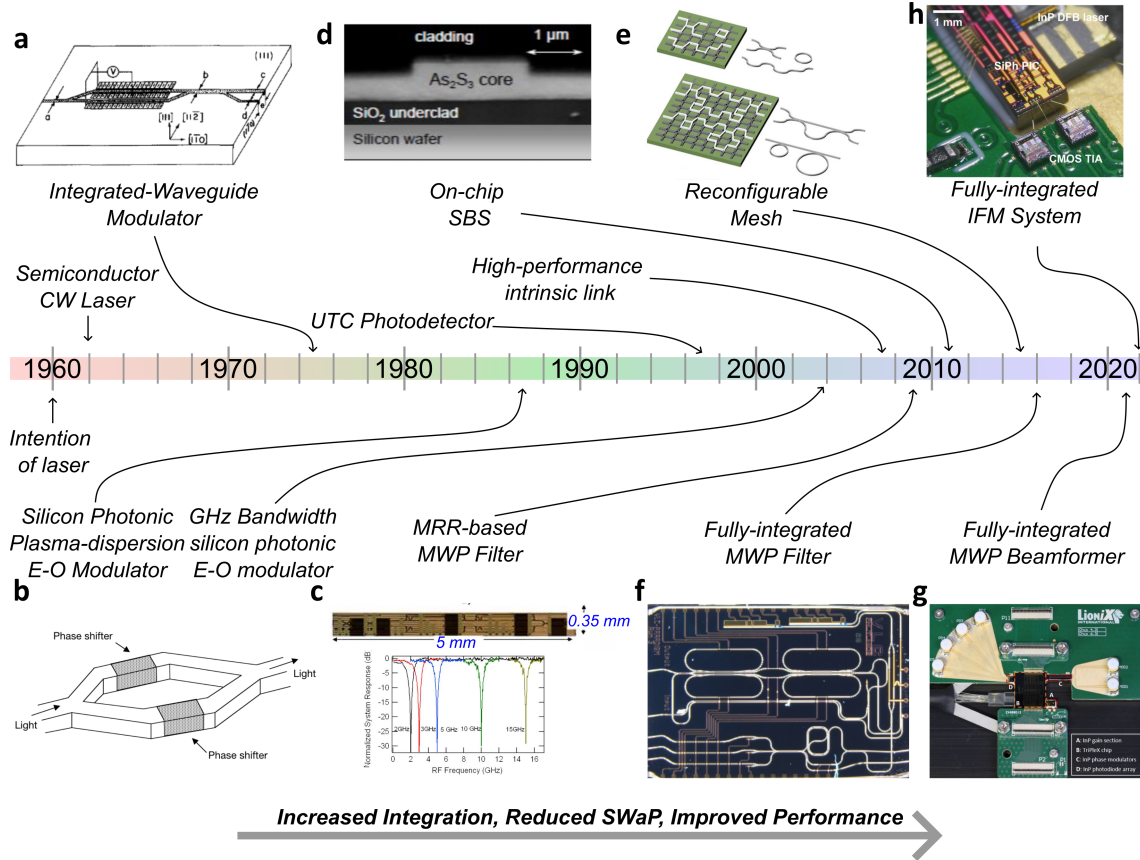


Figure 2.9: Timeline of key developments in microwave photonics. (a) First integrated waveguide modulator. (b) First GHz bandwidth silicon plasma-dispersion modulator. (c) First MRR-based MWP filter. (d) First demonstration of on-chip SBS. (e) First integrated reconfigurable mesh processor. (f) First fully-integrated MWP filter. (g) Fully-integrated InP-Si₃N₄ MWP beamformer (h) Fully-integrated InP-Si instantaneous frequency measurement system. IFM: Instantaneous frequency measurement. Figures from (a) [111], (b) [162], (c) [163], (d) [164], (e) [35], (f) [165], (g) [16], (h) [166].

to developing the widely-used high-speed silicon modulators today and drew attention to silicon as a photonic processing platform.

In addition to enabling compact E-O interfaces, the ability to guide light in compact chip-scale platforms has enabled novel classes of MWP signal processing that are not possible, impractical, or too bulky if implemented using optical fibres. For example, the first MWP filter based on a microring resonator (MRR) [167] was the silicon platform demonstrated by Rasras *et al.* in 2009 [163], Fig 2.9 (c), which exhibited a compact on-chip footprint of only 5 mm x 0.35 mm. The first demonstration of SBS in a chip-scale platform in 2011 by Pant *et al.* [164], Fig 2.9 (d), enabled kilometers of optical fibres to be replaced with centimeter-scale chalcogenide waveguides [51], [144]. Chip-scale technology also enables the implementation of compact reconfigurable mesh processors, which offer impressive reconfigurability with compact chip-scale form factors [38], [168], [169] that would otherwise be too large and impractical to implement using fibre-based mesh networks. Initially demonstrated by Zhuang *et al.* in 2015 for MWP signal processing applications on a silicon nitride PIC [35], Fig 2.9 (e), the mesh-processing method has

since been extended to enable a vast array of photonic processing functionalities [36], [170], [171], enabling highly-reconfigurable processing.

As the key components used to achieve MWP signal processing continued to mature, focus subsequently turned towards the implementation of fully-integrated MWP links [5], [25]. Fandiño *et al.* reported the first fully-integrated MWP filter [165], Fig 2.9 (f). The monolithic InP PIC included a laser, modulator, optical processing using a ring-assisted Mach-Zehnder interferometer (RAMZI) and a photodetector, highlighting the significant footprint reduction possible through photonic integration. Since this initial monolithically integrated demonstration, there has been a trend towards the use of hybrid and heterogeneously integrated approaches to enable several fully-integrated MWP processors [172]–[174]. The advantage of the hybrid and heterogeneous approaches is that it allows for the advantages of each material to be utilised [175]. For example, the fully integrated InP-Si₃N₄ beamformer presented by Tsokos *et al.* in 2021 [16], Fig 2.9 (g), utilised InP as a gain section and low-loss Si₃N₄ waveguides to implement MWP TTD processing. The InP-silicon platform presented by Tao *et al.* in 2022 [166], Fig 2.9 (h), utilised the compact nature of silicon components with InP optical sources to form the basis of an MWP instantaneous frequency measurement (IFM) system.

Considering the progression of MWP devices, the field now is in a position where individual processing components, such as MRRs, SBS and mesh networks, as well as integrated optical sources, modulators and detectors, are reaching a stage of maturity whereby system-level integration of MWP processing is being explored with success in achieving significant footprint reductions [16], [45], [166], [172]–[174], [176]. In parallel, the demonstrations of high-performance chip-scale MWP systems are highlighting the potential to achieve a fully integrated MWP processor with high performance [13], [177], [178]. Continuing to progress the photonic integration of MWP systems, whilst simultaneously improving performance is key to enabling the widespread deployment of MWP systems in real-world applications [25], [27].

2.3.2 Advantages of Integration

Let us consider the tapped delay line MWP filter exemplified in Fig. 2.6 to further highlight the advantages of chip-based MWP processing systems over their fibre-based counterparts. In 2012, Sancho *et al.* used millimeter length-scale photonic crystal waveguides in place of hundreds of meters of dispersive optical fibres, providing a wavelength-dependant delay in a significantly reduced and compact footprint [31]. Additionally, in 2020, Hu *et al.* used chip-scale optical microcomb sources to create a compact multi-wavelength source [33] rather than requiring multiple lasers. Recently in 2022, both the optical comb generation and optical processing were combined on a hybrid platform demonstrated by Shu *et al.* [34], as shown in Fig. 2.10 (a). They generated a frequency comb using an InP laser coupled to an AlGaAs-on-insulator microring resonator. Each comb line was processed on a silicon photonic chip using a series of on-chip delay lines [34].

The FSR of the filter response can be controlled, as shown in Fig. 2.10 (b), by modifying the comb line spacing which can be achieved by changing the start-up condition of the soliton through laser detuning and thermal tuning [34]. Furthermore, a demonstration towards real-world applications is depicted in Fig. 2.10 (c), whereby two 50 Mbs⁻¹ sig-

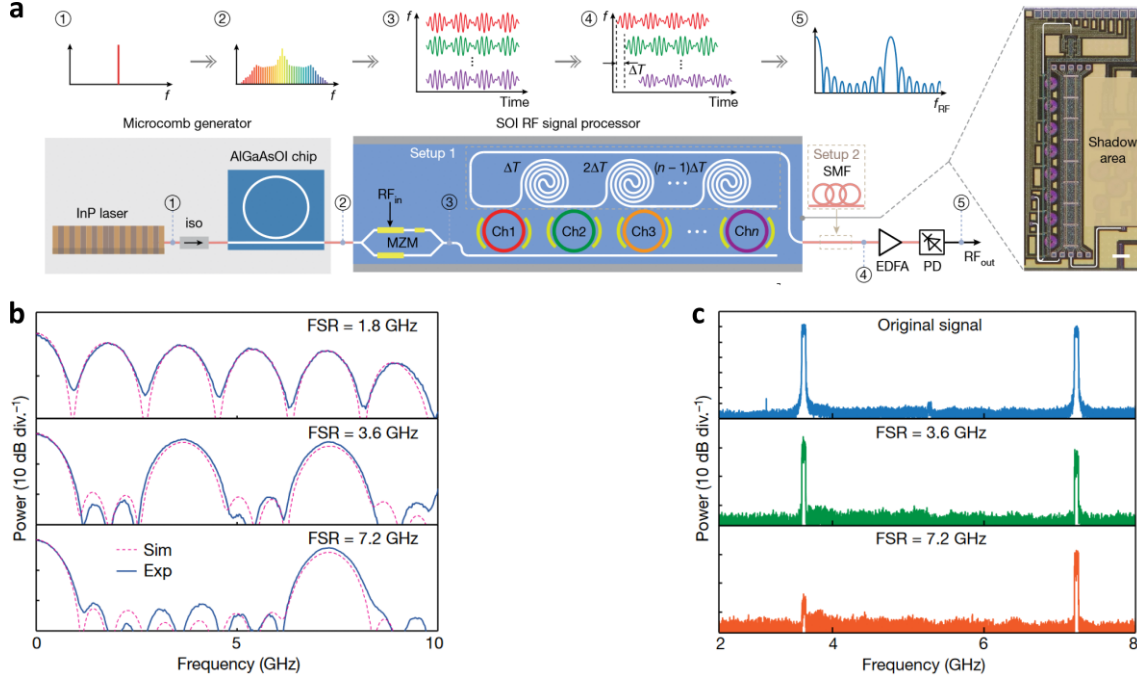


Figure 2.10: Discrete multi-tap MWP filter exhibiting a high degree of photonic integration. (a) Experimental setup. In particular, the optical comb source and required delay of each optical comb line are implemented on compact photonic chips. (b) Reconfigurability of RF response by changing the comb line spacing. (c) Demonstration of filtering towards real-world applications. Figures from [34].

nals are input to the filter at 3.6 GHz and 7.2 GHz. The signal at 3.6 GHz is attenuated by modifying the FSR of the MWP filter.

In addition to the advantages of wideband functionality and reconfigurability, MWP filters must also exhibit high-performance to be suitable for use in real-world applications [13], [25], [27], [30]. On-chip MWP link optimisation techniques have been demonstrated, which deliver promising results in compact footprints. We outline MWP link metrics, as well as link optimisation techniques in Chapter 3. A recent promising example of an integrated MWP processor that achieved exceptionally high performance was presented recently by Daulay *et al.* [30]. On-chip processing in a silicon nitride PIC, as shown in Fig. 2.11 (a), was used to flexibly synthesise an MWP filter with notch and bandpass responses, Fig. 2.11 (b-c). The combination of a high-power optical source and a low-biased E-O modulator enabled a high-performance link with 10 dB gain, 15 dB noise figure, Fig. 2.11 (d), and 116 dB.Hz^{2/3} SFDR₃ [30]. Although these results rely on a combination of on-chip processing and discrete fibre-based off-chip components, it highlights the possibility of achieving high-performance MWP processing systems. At present, many fully-integrated platforms have focussed on demonstrating novel functionality and increased level of integration, and do not necessarily achieve the required link performance [15], [16], [165], [172]. The remaining challenge now is to achieve the required RF processing functionality with satisfactory link performance in a *fully-integrated* platform.

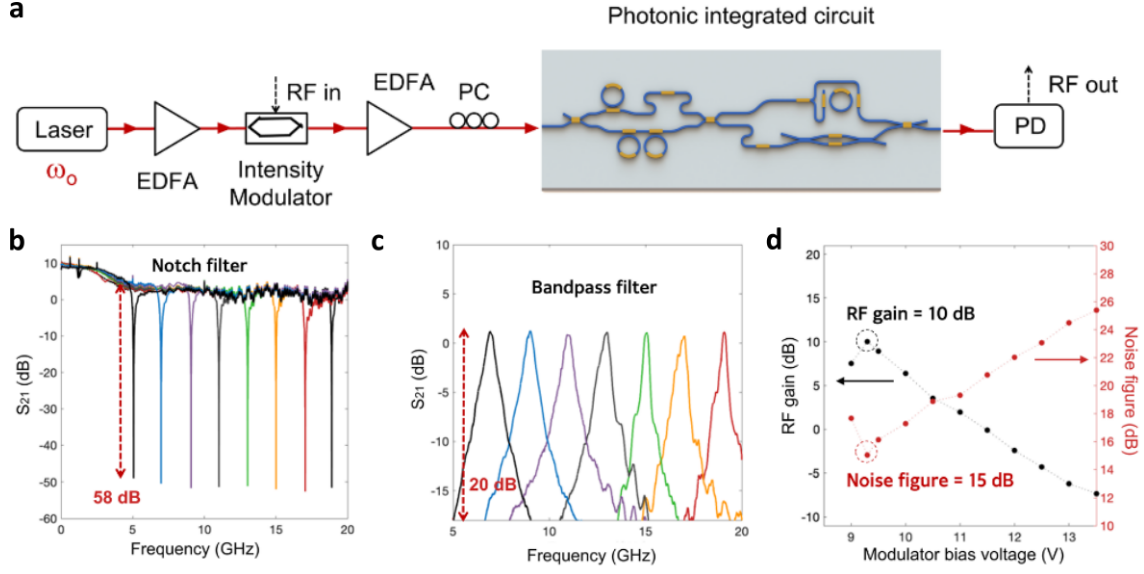


Figure 2.11: High-performance integrated MWP filter. (a) Experimental setup. (b) RF notch response. (c) RF bandpass filter response. (d) Measured gain and noise figure. Figures from [30].

2.3.3 Material Platforms

The remainder of this section provides an overview of key platforms used to implement integrated microwave photonic processors, including their current status and potential pathways to achieve fully-integrated MWP signal processors with high performance. A conceptual diagram of a fully-integrated MWP processor is depicted in Fig. 2.12, which includes an optical source, modulator, signal processor and photodetector implemented on the same chip-scale platform. It identifies that multi-material integration is required to achieve this goal, as it is challenging to implement all the above-mentioned functionalities using a single material. Key considerations to implement a platform targeting high-performance MWP applications include:

- **High power handling and low losses.**

We show in Chapter 3 that the performance of MWP systems improves with low optical losses and large optical powers.

- **Compatibility with key active components of MWP links.**

The components that are essential to an MWP link include optical sources and amplifiers, high-speed modulators and photodetectors. A material requires a direct band gap at the operating wavelength (typically 1550 nm) to facilitate on-chip lasers, amplifiers and detectors. To achieve high-speed E-O modulation, a material typically requires second-order optical nonlinearities ($\chi^{(2)}$).

- **Optical signal processing components.**

The components that implement optical signal processing in an MWP link ultimately determine the underlying characteristics of the RF response. Key considerations for on-chip processing components include the linewidth of the response,

which ultimately limits the spectral resolution of the underlying MWP system. Additionally, the reconfigurability of the applied resonance, which includes the rate and frequency range over which the resonance can be tuned, must be considered. Furthermore, the waveguide size and minimum bend radius, which generally decreases with increasing refractive index because the wavelength in the waveguide is inversely proportional to the refractive index, play an important role in determining the SWaP of MWP systems.

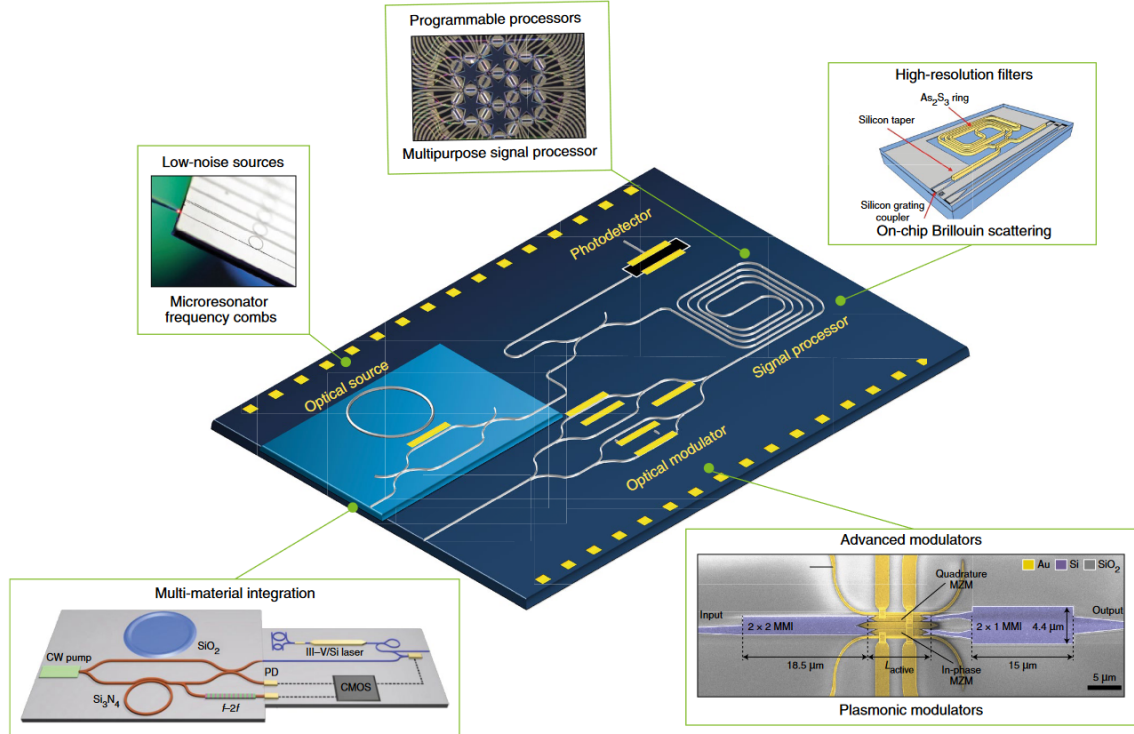


Figure 2.12: Conceptual figure of integrated microwave photonic processing. The chip contains the key components of an MWP link including an optical source, modulator, signal processor and photodetector. Figure taken from [25].

Silicon

Silicon is perhaps the most developed platform for integrated photonics, leveraging the multi-billion dollar electronics industry which developed CMOS fabrication processes. Silicon has a large refractive index of 3.45 at 1550 nm [179], which enables devices with a compact footprint as silicon photonic single-mode waveguides have small dimensions that are typically 450 nm wide and 220 nm high [179]. Owing to its widespread use, a comprehensive set of passive functions have been demonstrated in the silicon photonic platform, including directional couplers, multimode interferometers (MMIs), grating couplers, waveguide crossing and ring resonators. Comprehensive reviews on silicon photonic platforms can be found in refs. [112], [167], [180]–[183].

However, silicon can not intrinsically facilitate E-O modulation, a key functionality for MWP and telecommunications applications, as crystalline silicon lacks $\chi^{(2)}$ nonlinearity

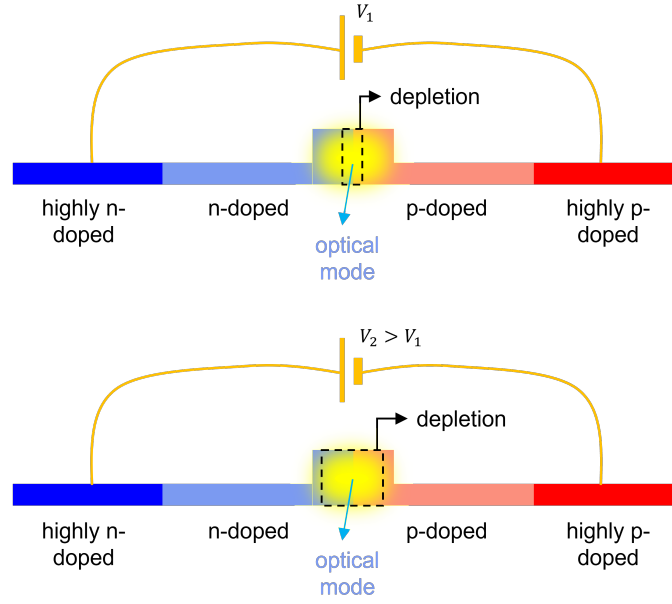


Figure 2.13: Simplified cross-section of typical foundry-compatible silicon photonic plasma-dispersion modulator. The device consists of a doped silicon rib waveguide that guides the optical mode (yellow shaded region). The depletion region across the diode (black dashed box) is altered as a function of the voltage applied across the junction. Figure from [184].

due to its centrosymmetric crystalline nature [185]. Therefore, silicon cannot intrinsically exploit the Pockels effect to achieve electro-optic modulation. Additionally, silicon lacks a direct band gap at 1550 nm and therefore cannot directly provide optical laser sources, gain, or photodetectors at 1550 nm. Despite this, the silicon photonic platform has been successfully augmented with the necessary active functionalities for an MWP link using the techniques outlined below.

High-speed modulation can be achieved in silicon through the plasma-dispersion effect [112], [186] which involves the injection or depletion of electronic charges in a waveguide that induces a refractive index change. We present the cross-section of a typical silicon depletion modulator in Fig. 2.13. The optical mode is represented by the shaded yellow region. The depletion region (black dashed box) across a PN junction overlaps with the optical mode and the width of the depletion region changes with the applied voltage [112], which in turn changes the refractive index of the waveguide and results in phase modulation. In contrast to Pockels modulators, the plasma-dispersion effect induces an intensity change along with a refractive index change, known as chirp [106], [187], which is not desirable if pure phase modulation is required [112].

In 1987, the seminal work of Soref and Bennett [188] was the first to demonstrate the plasma-dispersion effect [112], [186] in silicon, thus introducing high-speed E-O modulation to the silicon photonic platform. Demonstrations of phase modulators [189], [190] and the first high-speed Mach-Zehnder (intensity) modulators with 1 GHz bandwidth [161] were soon to follow. In parallel, it was discovered that SiGe grown through molecular beam epitaxy (MBE) is a suitable candidate for silicon photonic photodetectors [191] and were subsequently integrated with silicon photonic waveguides [192]. This

provided silicon photonic platforms with high-speed photodetection capabilities.

Alternatively, silicon photonic platforms can be augmented to achieve Pockels modulation by using organic electro-optic materials [112], [195], [196]. These devices are typically implemented by depositing an organic electro-optic material in a silicon slot waveguide, as shown in Fig. 2.14 (a), which offers a large overlap between the optical mode and the driving RF field, shown in Fig. 2.14 (b-c). The Pockels effect in these organic electro-optic materials exhibits high E-O conversion efficiency, enabling modulators with half-wave voltages below 1 V [193]. However, the bandwidth of such devices is typically limited by the resistor-capacitor (RC) time constant of the device structure, as opposed to the E-O response time of the organic material itself [112], [197], [198]. Plasmonic organic hybrid modulators overcome this limitation by using a metal plasmonic slot waveguide [114], [194], [197], as opposed to a silicon waveguide. As a result, they exhibit compact lengths (tens of μm), as depicted in Fig. 2.14 (d), enabled by the increased confinement of the plasmonic mode. Owing to the reduced length, the RF path of the modulator appears as a lumped element and therefore offers large E-O conversion bandwidths that exceed 500 GHz [114], as is not limited by the RC time constant of its travelling wave counterparts.

Significant developments have been made to the silicon photonic platform since the 1980s. Today, many commercial foundries offer dedicated and multi-project wafer services for research and commercial applications from foundries including IMEC [199], AIM Photonics [200], CEA-Leti [201], TSMC [202], Global foundries [203] and AMF [204]. These foundries typically offer all components, including standard waveguides, doped waveguides for thermal heaters and high-speed E-O modulators, as well as high-speed photodetectors and metal routing layers to interface on-chip heaters and other electrical signals. The typical layer stack for a foundry-produced silicon photonic chip is shown in Fig. 2.15. Foundry offerings for silicon photonic platforms continue to mature and take advantage of the progress made towards integrating optical sources through

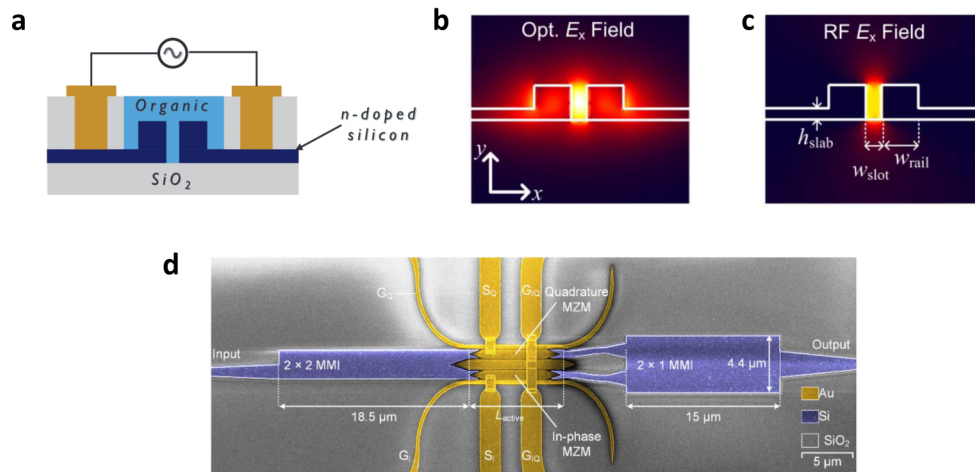


Figure 2.14: Silicon organic hybrid modulators. (a) Cross-section of a typical silicon organic hybrid modulator. The device consists of a silicon slot waveguide, filled with an electro-optic polymer. (b) Simulated optical field of the silicon slot waveguide. (c) Simulated RF field of the modulator. (d) False-coloured image of a plasmonic organic hybrid modulator. Figures from (a) [112], (b-c) [193], (d) [194].

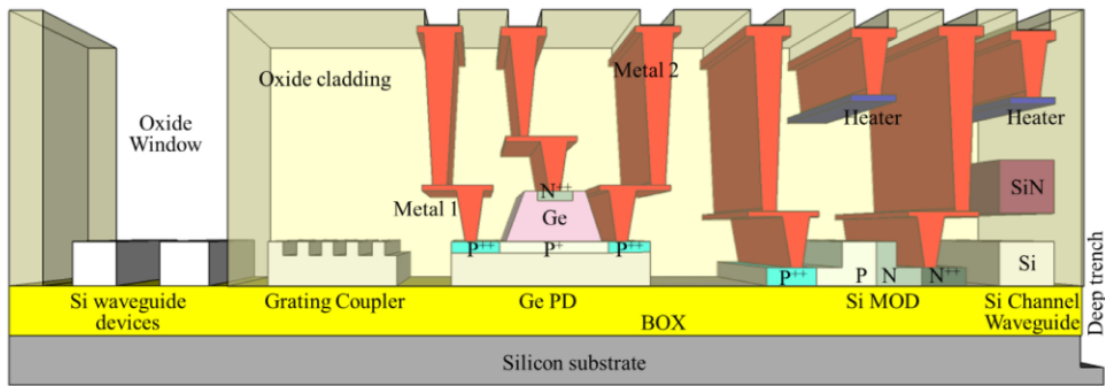


Figure 2.15: Cross-section of typical foundry-compatible silicon photonic platform. Figure from [181].

hybrid and heterogeneous approaches [205]–[212].

The silicon photonic platform is attractive for integrated MWP applications owing to its maturity and ability to integrate lasers, modulators and photodetectors in a single platform. However, further work must be done to improve the RF link performance of silicon photonic-based MWP processors which is currently limited by the relatively inefficient plasma-dispersion modulation process, as well as the limited on-chip power handling due to two-photon absorption (TPA) in silicon [213].

Indium Phosphide

Indium phosphide and other III-V materials were used extensively since the 1980s for developing integrated photonic devices [214]. The term III-V refers to devices that contain elements from groups III and V of the periodic table. The key benefit of the InP platform is that it offers a direct band gap at 1550 nm, which enables lasers, optical amplifiers and photodetectors to be implemented using a single material. Demonstrations of optical and MWP filters that use high-speed phase modulators to rapidly change their resonant frequency in 2011 [8], [215] were among some of the first InP-based MWP devices reported and demonstrated a high degree of photonic integration. Since then, the first MWP processor to integrate lasers, modulators, optical processing, and photodetectors on a single platform was presented by Fandiño *et al.* in 2016 [165], as shown in Fig. 2.16 (a). This InP-based MWP filter highlighted the advantages of photonic integration through its compact form factor. Since then, several fully-integrated MWP platforms have been presented, including the optoelectronic oscillator (OEO) packaged onto a circuit board, Fig. 2.16 (b) [216], and the frequency converter presented by Bottenfield *et al.* [217].

A challenge faced by monolithically integrated InP systems is the relatively high propagation loss of the InP waveguides, and can be up to an order of magnitude larger than their silicon photonic counterparts [218]. To overcome the losses of InP, semiconductor optical amplifiers (SOAs) may be used to amplify the optical signal. However, SOAs introduce excessive noise, which is translated into electrical noise in the MWP link output and could consequently limit the dynamic range of the resultant MWP processor [5]. As a result, there has been a trend to towards heterogeneous and hybrid integration of InP devices [219], where InP-based materials are used for optical gain, modulators or photodetectors

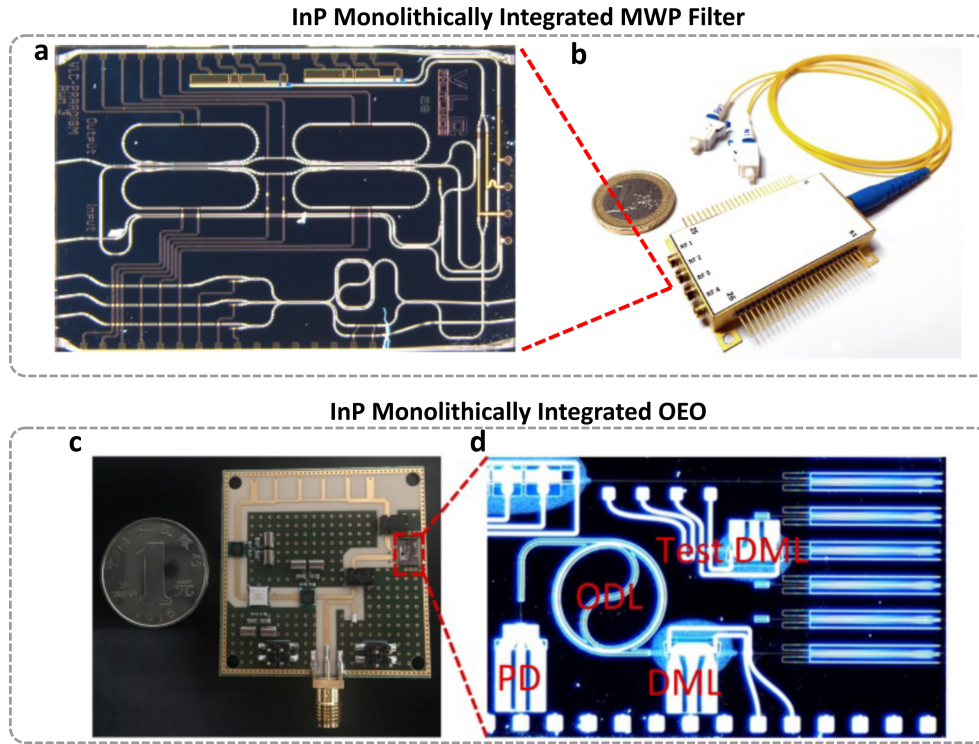


Figure 2.16: Fully-integrated InP-based MWP Processing platforms. (a-b) MWP filter (c-d) An optoelectronic oscillator (OEO). Figure (a-b) modified from [165], (c-d) modified from [216].

and integrated into other lower loss platforms, including silicon, silicon nitride or lithium niobate.

Silicon Nitride

Silicon nitride (Si_3N_4) is a widely used optical material owing to its wide transmission window that spans the wavelength range 400 nm to 2350 nm [220]. This wide transmission window makes Si_3N_4 attractive for a diverse set of applications [220]–[226] including MWP [5], nonlinear optics [227], [228], biosensing [229], [230], optical gyroscopes [231], atomic traps [232] and optical atomic clocks [233]. Silicon nitride is available from companies including Ligentec [234], LioniX [235], IMEC [236] or AIM Photonics [237].

Si_3N_4 has attracted significant attention from the MWP community as it can handle large optical powers (at least 7 W [238]) and it demonstrates low optical losses (0.12 dB/m [239], 0.034 dB/m [240]). Both properties are required in high-performance MWP links, as we discuss in Section 3.3. It is worth noting that although sub-dB/m losses of Si_3N_4 waveguides have been measured [240], such losses are not readily available in platforms where Si_3N_4 is integrated with other materials, such as silicon and metals [241].

Like silicon, Si_3N_4 does not offer a direct band gap at 1550 nm and lacks $\chi^{(2)}$ nonlinearities. Therefore, it cannot be used as a source of optical gain or to achieve electro-optic modulation. To augment the Si_3N_4 platform with active optical functionality, significant research efforts have been undertaken to integrate active optical materials (such as InP) with passive Si_3N_4 circuits and have delivered great success. Notably, high-performance

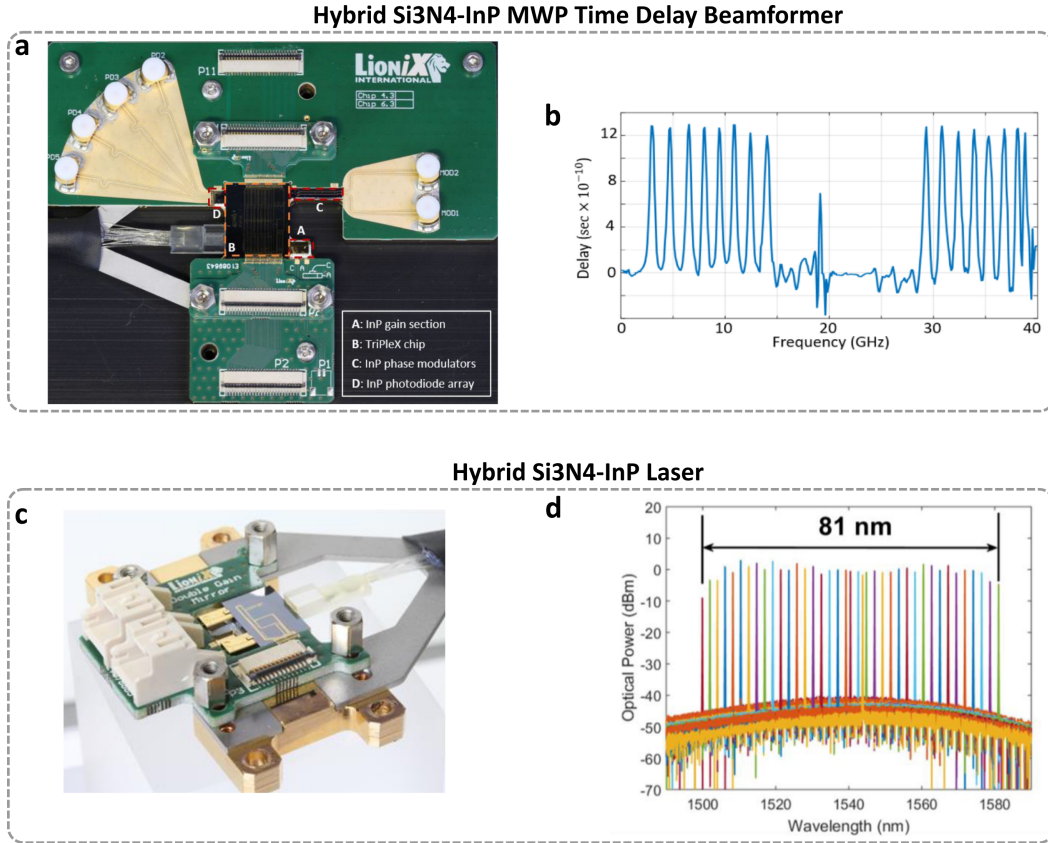


Figure 2.17: Integrated Si₃N₄-InP platforms. (a) Fully-integrated Si₃N₄-InP MWP true time delay beamformer. (b) Measured time delay vs frequency using the fully-integrated MWP beamformer. (c) Integrated Si₃N₄-InP laser (d) Measured output spectrum while tuning the wavelength of the laser. Figures taken from (a-b) [16] (c-d) [242].

lasers [243], [244] have been demonstrated using InP SOAs to provide optical gain and utilise the low-losses of Si₃N₄ waveguides to form a high-Q cavity. This can be achieved by interfacing the InP and Si₃N₄ waveguides through various techniques including butt-coupling [245]–[248], micro transfer printing [249], or wafer bonding [250], [251]. An exemplary demonstration of an InP-Si₃N₄ laser by Epping *et al.* [242] achieved large on-chip power (20.7 dBm) and ultra-low RIN noise (< -165 dBc/Hz). Both properties are highly attractive for MWP applications. A similar approach of butt-coupling Si₃N₄ and InP PICs has enabled fully-integrated MWP processing platforms, such as the four-element beamformer [16] that use low-loss Si₃N₄ waveguides for optical processing and use active InP chips for optical gain, E-O modulation and photodetection.

Lithium Niobate

Lithium niobate (LiNbO₃) is an attractive material for integrated photonics owing to its wide transmission window (400 nm to 5 μ m) and large nonlinear optic and acousto-optic coefficients [252]. For MWP applications in particular, the discovery of the ferroelectric

properties of LiNbO_3 in 1949 by Matthias and Remika [253]¹, led to its extensive use for E-O modulators, boasting highly efficient and linear electro-optic conversion with a broadband electro-optic frequency response [252]. Owing to its widespread use in photonics, many review articles cover the fundamentals of lithium niobate photonics and the diversity of applications; the following references represent a small selection that the interested reader may refer to broad reviews on the lithium niobate platform [113], [252], [256]–[259].

Perhaps the most common commercial use of LiNbO_3 for MWP applications is the titanium-diffused LiNbO_3 modulator. In these devices, an index contrast is achieved by through titanium diffusion [262], as exemplified by the device cross-section in Fig. 2.18 (a). In these modulators, the applied RF field propagates in the same direction as the optical field, modulating the refractive index through the Pockels effect in LiNbO_3 . A typical cross-section of a Ti-diffused LiNbO_3 modulator is shown in Fig. 2.18 (b). Although the weak confinement provided by Ti diffusion gives mode volumes similar to optical fibres at near-infrared wavelengths, which enables low-loss interfaces to fibres (< 0.5 dB) [257], the E-O conversion efficiency is limited due to the relatively large spacing between RF electrodes, exposing the optical mode to a weaker modulating RF field [263].

A newer class of thin-film lithium niobate (TFLN) devices offer higher confinement than their diffused counterparts. TFLN devices can be implemented using several techniques that are illustrated in Fig. 2.18 (c). The strip-loaded configuration is achieved by depositing a material with a higher refractive index above the LiNbO_3 film [256]. Alternatively, the LiNbO_3 film may be directly etched [264], [265]. We depict an etched TFLN

¹The element niobium (Nb) was referred to as columbium (Cb) at the time by American chemists, as it is in Ref. [253]. Despite element 41 being named columbium first, the name niobium was adopted in Europe. In 1950, the International Union of Pure and Applied Chemistry (IUPAC) ruled that element 41 was to be called niobium, ending 100 years of controversy [254], [255].

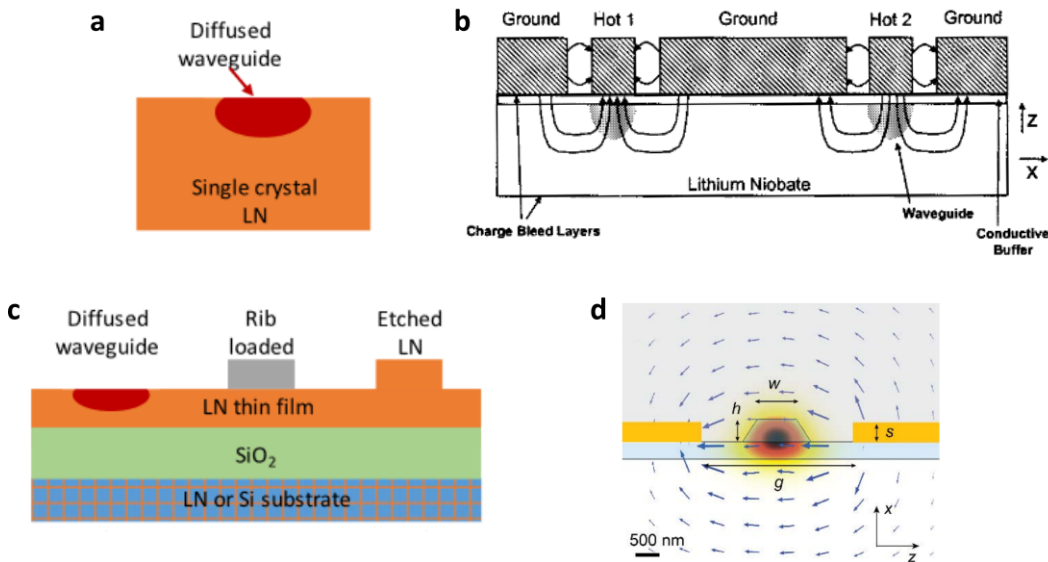


Figure 2.18: Lithium niobate modulators. (a) Diffused waveguide. (b) E-O modulator based on a diffused waveguide. (c) Thin-film lithium niobate configurations. (d) Optical field (coloured) and RF field (arrows) of a TFLN E-O modulator. Figures from (a) [258], (b) [260], (c) [258] and (d) [261].

modulator in Fig. 2.18 (d). The optical field is indicated by the coloured contours and the RF field is indicated by the arrows. As a result of the increased optical confinement, the RF field has a larger overlap with the optical field and increases the E-O conversion efficiency [263] compared to their Ti-diffused counterparts. Current state-of-the art TFLN modulators offer low half-wave-voltages of only 1.6 V at 50 GHz [263], which is essential to enable high-performance MWP links (we discuss this further in Section 3.3).

Because lithium niobate does not have a direct band-gap at 1550 nm, optical sources, amplifiers, and photodetectors must be added to the platform through heterogeneous, or hybrid integration techniques. A number of demonstrations have enabled active E-O components and optical sources on lithium niobate platforms using micro transfer printing [266], wafer bonding [267], or multi-chip hybrid integration approaches [268], [269].

Chalcogenide

Chalcogenide glasses exhibit a broad transparency window from visible to far infrared range [270]. Chalcogenide glasses contain one or more of the chalcogen elements in the periodic table (group 6a), covalently bonded with network formers such as arsenic (As), germanium (Ge), gallium (Ga), antimony (Sb), silicon (Si) or phosphorus (P) [271]. Their large refractive index, as well as high optical nonlinearities and low two-photon absorption, make them attractive for exploiting nonlinear optical effects [272]. The unique properties of chalcogenide glasses make them attractive for use applications including imaging, chemical sensing and microscopy, as outlined by Sanghera *et al.* [273]. Perhaps the most prevalent use of chalcogenide glasses has been in CD's and DVD's due to their phase change properties [270].

In particular, As_2S_3 chalcogenide glasses have found use in MWP signal processing applications, owing to their ability to facilitate stimulated Brillouin scattering (SBS), an acousto-optic interaction that provides uniquely narrow linewidth optical resonances. We provide an extensive background on the SBS process and how chalcogenide waveguides facilitate the Brillouin interaction for MWP signal processing systems in Chapter 4. In addition to facilitating SBS, passive MRRs have been demonstrated in As_2S_3 glasses [45], [274], [275]. Although As_2S_3 glasses cannot implement key functions of MWP links such as modulation and photodetection, they have been heterogeneously integrated with silicon photonic platforms [176] and Ti-diffused lithium niobate platforms [276]–[281]. Both represent attractive pathways towards augmenting MWP processors with fine spectral resolution from SBS and active optical functionalities through heterogeneous integration.

2.3.4 Optical Processing

On-chip optical processing is achieved using core components that are common between the material platforms chosen for MWP signal processing. In this section, we provide an overview of on-chip optical processing components that are used in this thesis.

Stimulated Brillouin scattering

Stimulated Brillouin scattering (SBS) is an optical nonlinear process that has gained significant attention from the MWP community owing to its uniquely narrow resonant linewidth of roughly 30 MHz, as well as its highly reconfigurable response that can be

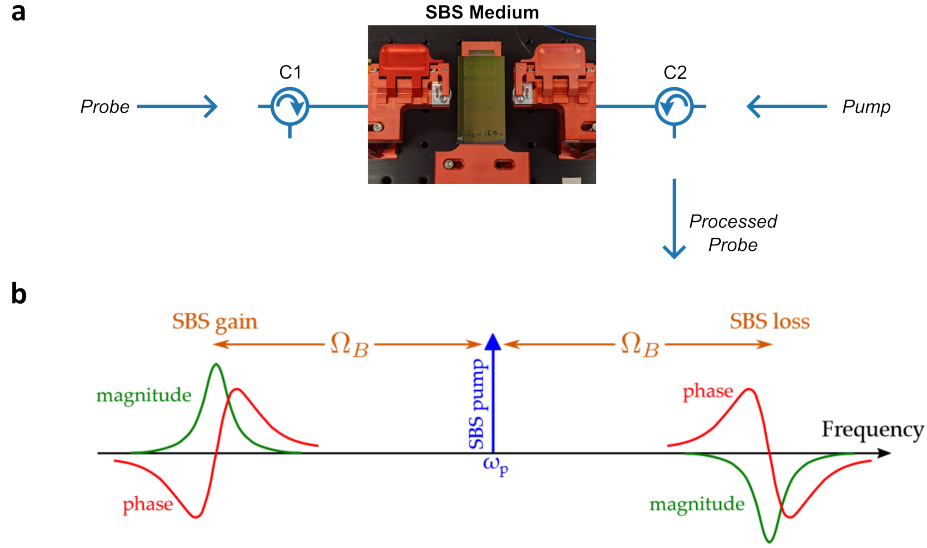


Figure 2.19: Overview of stimulated Brillouin scattering. (a) A simplified experimental setup used to measure backwards SBS. A pump and probe are counter-propagated against each other in the Brillouin medium. Circulators are used to separate the counter-propagating pump and probe. (b) The pump induces a gain and a loss resonance, each are separated from the pump by a frequency ω_B . The magnitude and phase response of the gain and loss resonances are labelled. C: Circulator. Figure from (b) [284].

tailored based on the optical pump spectrum and its ability to be induced in chip-scale platforms [25], [51], [81], [142].

SBS is induced through an interaction between an optical pump and probe wave, which creates an acoustic wave [51]. In this explanation, let us consider the canonical case of backward Brillouin scattering, where the optical pump and probe waves travel in different directions known as backward SBS. We outline the different configurations in which SBS can be induced in Chapter 4. Backward SBS is typically induced using an experimental setup similar to Fig. 2.19 (a). When the pump and probe have different frequencies, a travelling optical interference pattern is formed. If the pump signal has sufficient intensity, a travelling acoustic wave is created through electrostriction [282]. The change in density manifests as a change in refractive index through the photoelastic effect, forming a travelling grating [283]. If the pump and the probe are separated in frequency by the Brillouin frequency shift, Ω_B , such that a strict phase-matching condition is met [282], the pump is reflected and Doppler shifted by the moving grating such that it appears as a coherent amplification of the probe. The value of Ω_B is determined by the strict phase matching condition for this stimulated process [282]. A loss resonance also occurs, which is associated with the annihilation of acoustic phonons, and occurs at a frequency Ω_B above the Brillouin pump, as depicted in Fig. 2.19 (b). We provide a theoretical outline of SBS in Chapter 4, along with a comprehensive review covering the use of on-chip SBS in MWP systems.

Microring Resonators

The optical microring resonator (MRR) is one of the most commonly used optical signal processing components owing to its compact size. An MRR consists of an input wave-

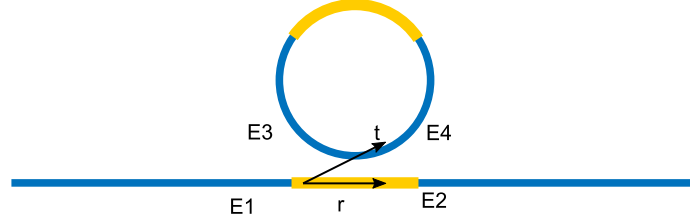


Figure 2.20: A microring resonator. The device consists of an input waveguide that is coupled to a looped resonator. The coupling coefficient into the resonator (t), and the coupling coefficient through the input waveguide (r) are labelled.

uide coupled to a looped resonator. Resonance occurs when the round-trip optical path length inside the resonator is an integer number of wavelengths. A full description of the theory and application of MRRs can be found in references [167], [285], [286].

The principle of the MRR relies on coupling from the input waveguide into the looped waveguide, as shown in Fig. 2.20. The input (E_1), output (E_2), and circulating (E_3 , E_4) electric fields are related by the matrix [167], [285]

$$\begin{bmatrix} E_4 \\ E_2 \end{bmatrix} = \begin{bmatrix} r & jt \\ jt & r \end{bmatrix} \begin{bmatrix} E_3 \\ E_1 \end{bmatrix} \quad (2.2)$$

where r is the self-coupling coefficient of the input waveguide and t is the cross-coupling coefficient from the input waveguide into the looped waveguide and satisfy the relation $r^2 + t^2 = 1$. When travelling through the ring, the optical signal experiences attenuation and accumulates phase, such that the circulating field components are related by the expression

$$E_3 = ae^{j\phi} E_4 \quad (2.3)$$

where a is the single-pass amplitude transmission and ϕ is the accumulated phase in the ring, which relates to the waveguide effective index (n_{eff}), the path length of MRR (L) and wavelength (λ) by [167], [285]

$$\phi = \frac{2\pi n_{eff} L}{\lambda} \quad (2.4)$$

Combining Eq.'s 2.2 and 2.3 we can express the transfer function of the MRR as

$$\frac{E_2}{E_1} = e^{j(\pi+\phi)} \frac{a - re^{-j\phi}}{1 - rae^{+j\phi}} \quad (2.5)$$

The separation between adjacent resonances in the MRR response is defined as the free spectral range (FSR) and can be calculated as [167], [285]

$$FSR = \frac{\lambda^2}{n_g L} \quad (2.6)$$

where n_g represents the group index. The bandwidth of each resonance, often referred to as the full width at half maximum (FWHM), is a key parameter for MWP-based processing applications and can be calculated as [167], [285]

$$FWHM = \frac{(1 - ra) \lambda^2}{\pi n_g L \sqrt{ra}} \quad (2.7)$$

By analysing Eq. 2.5, the response of the MRR is affected by the intrinsic losses and accumulated phase in the MRR cavity, as well as the coupling coefficients of the ring. In fact, the MRR response can be separated into three distinct regimes that are based on the ratio of power coupled into the MRR (t^2) to the intrinsic losses in the MRR ($1 - a^2$). The ring is under-coupled when $t^2 < 1 - a^2$, and results in a zero-phase shift at the resonant frequency of the ring. When the power coupled into the ring equals the intrinsic losses of the ring, $t^2 = 1 - a^2$, the transmission at the output of the bus waveguide drops to zero, and a sharp phase transition is observed around the resonant frequency. Finally, the ring is over-coupled when $t^2 > 1 - a^2$ and a π phase shift is observed at the resonant frequency of the MRR with respect to the phase outside the MRR resonance. We plot the amplitude and phase response of the MRR for the under-coupled, critically coupled, and over-coupled conditions in Fig. 2.21 (a-c), respectively. Additionally, the dependence of MRR rejection ratio on the coupling coefficient is plotted in Fig. 2.21 (d), where each distinct coupling regime is labelled. In MWP processing applications, the intrinsic losses in the cavity are typically fixed. However, the MRR cavity path length and coupling conditions can be tuned to reconfigure the MRR response.

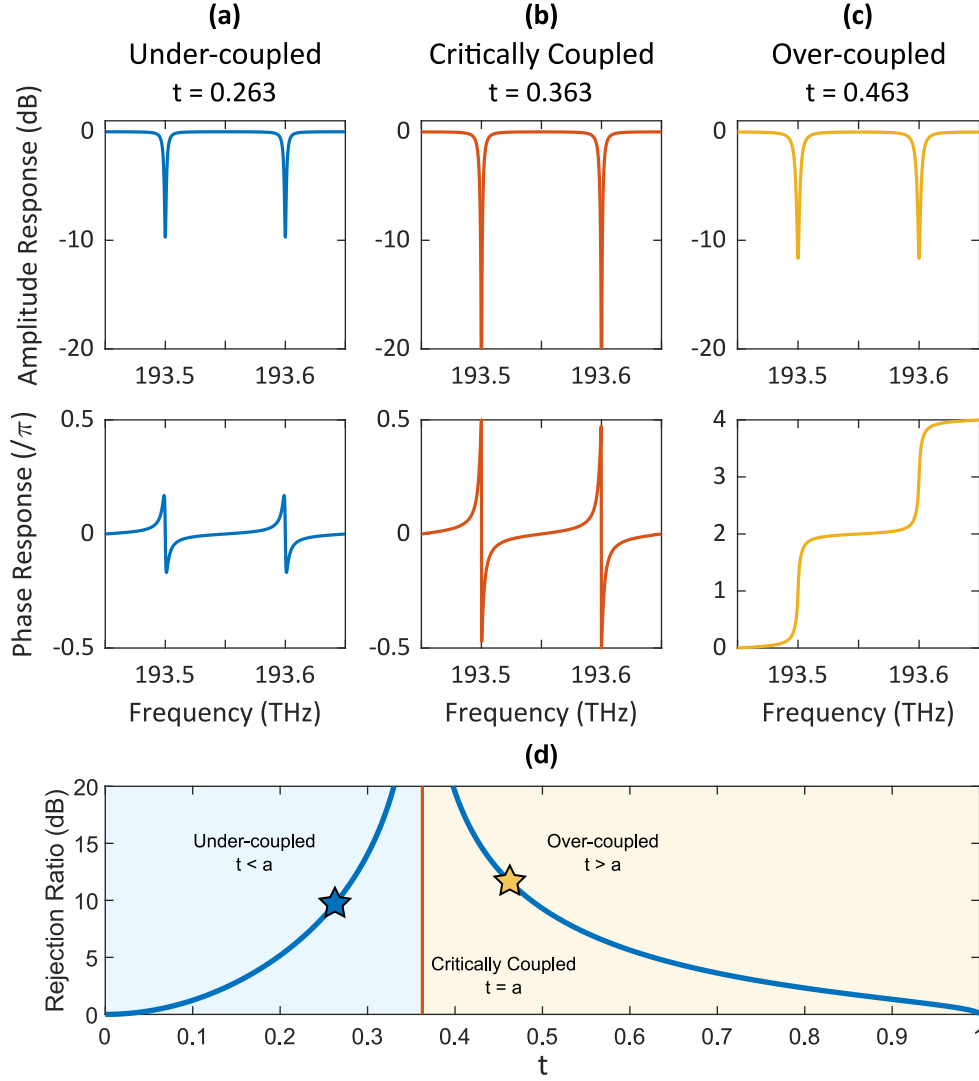


Figure 2.21: Simulated response of a MRR. In the simulations, we used the values $\alpha = 1$ dB/cm, $n_{\text{eff}} = 3.5$, FSR = 100 GHz. The coupling values for each trace are labelled in the title. The amplitude response is shown on the top row and the phase response is plotted on the bottom row. (a) Under-coupled response. (b) Critically coupled response (c) Over-coupled response. (d) Simulated rejection ratio as a function of coupling condition. The stars represent the corresponding under and over-coupled conditions presented in (a) and (c).

Chapter 3

Integrated Microwave Photonic Processing

This chapter introduces MWP signal processing. We overview the key components used to achieve MWP signal processing, including electro-optic modulators, and photodetectors. We outline how the frequency response of an MWP link can be altered through the use of different optical modulation formats and how RF filters can be synthesised as a result. We also outline the key metrics used to quantify the performance of MWP links and techniques that can be employed to improve their performance.

3.1 MWP Link Architecture and Components

3.1.1 Modulation

Electro-optic modulators (EOMs) transduce the RF input signal of an MWP link into the optical domain and are therefore critical components in MWP systems. In the most general sense, the function of an EOM is to modulate the amplitude, or phase of the input optical signal by the modulating electronic signal. In Chapter 2, we identified that modulation can be achieved by exploiting effects including the Pockels effect and plasma-dispersion effect, and that external modulators are most often used in MWP links. For the remainder of this subsection, we derive expressions for the electric field at the output of several modulator types that are commonly used in MWP signal processing systems.

Phase Modulation

The most basic modulator configuration is the phase modulator. Phase modulation (PM) is achieved when an applied electronic input changes the optical path length, and hence the accumulated phase, of an optical waveguide. A high-level setup of how a PM is driven is shown in Fig. 3.1. If we express the continuous wave (CW) optical input as shown in Fig. 3.1 (a) with peak electric field E_0 , angular frequency ω_c , as $E_{PM,in}(t) = E_0 e^{j\omega_c t}$, the output phase modulated signal is given by

$$E_{PM,out}(t) = E_0 e^{j\omega_c t} e^{j\phi(t)} \quad (3.1)$$

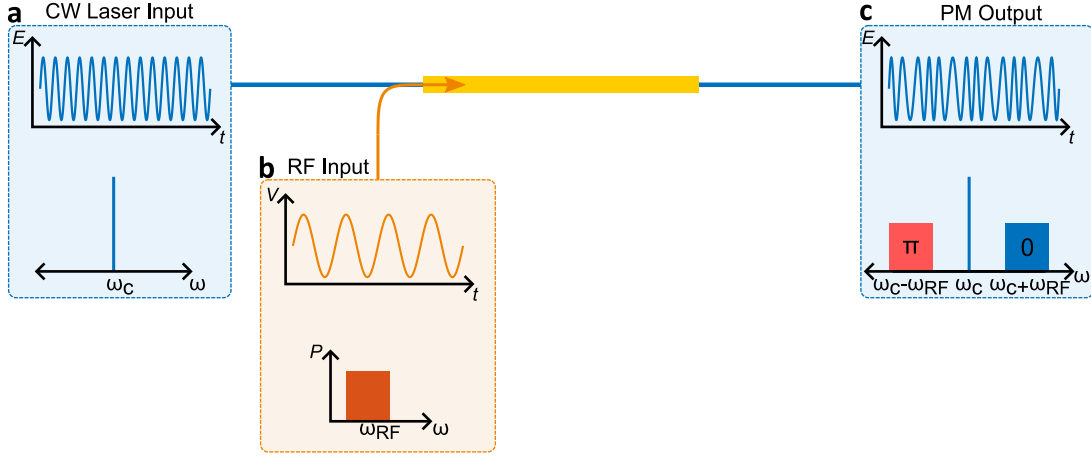


Figure 3.1: Optical phase modulation. (a) Optical input signal to the modulator. (b) The RF input used to modulate the phase of the optical input. (c) Phase modulated optical output. The yellow strips represent electrical contacts that drive a phase modulator.

where $\phi(t)$ is the additional optical phase shift accrued in the modulator and t represents time. If the PM is driven with a CW RF tone with peak voltage V_{RF} and angular frequency ω_{RF} , given by $V_{PM}(t) = V_{RF} \sin(\omega_{RF}t)$, as shown in Fig. 3.1 (b), the output of the phase modulator, Fig. 3.1 (c), can be expressed as

$$E_{PM,out}(t) = E_0 e^{j\omega_c t} e^{j\phi_{RF} \sin(\omega_{RF}t)} \quad (3.2)$$

The modulation depth is related to the applied peak input voltage by the following expression

$$\phi_{RF} = \frac{\pi V_{RF}}{V_{\pi RF}} \quad (3.3)$$

where $V_{\pi RF}$ is the frequency-dependant voltage required to induce a π phase shift, known as the half-wave voltage. Applying the Jacobi-Anger expansion to eq. 3.2, the output of the phase modulator can be expressed as

$$E_{PM,out}(t) = E_0 \sum_{n=-\infty}^{\infty} J_n(\phi_{RF}) e^{j(\omega_c + n\omega_{RF})t} \quad (3.4)$$

where J_n is the Bessel function of the first kind, and n refers to the sideband order. Assuming small-signal modulation conditions ($V_{RF} \ll V_{\pi RF}$), the spectral components that correspond to $|n| > 2$ can be ignored. Therefore, the resultant electric field at the PM output can be expressed as

$$E_{PM,out}(t) = E_{l,PM} e^{j\phi_l} e^{j(\omega_c - \omega_{RF})t} + E_{c,PM} e^{j\phi_c} e^{j\omega_c t} + E_{u,PM} e^{j\phi_u} e^{j(\omega_c + \omega_{RF})t} \quad (3.5)$$

$$E_{l,PM} e^{j\phi_l}(t) = -E_0 J_1(\phi_{RF}) \quad (3.6)$$

$$E_{c,PM} e^{j\phi_c} = E_0 J_0(\phi_{RF}) \quad (3.7)$$

$$E_{u,PM}e^{j\phi_u} = E_0J_1(\phi_{RF}) \quad (3.8)$$

The spectrum consists of upper ($E_{u,PM}$) and lower ($E_{l,PM}$) modulated sidebands that are equally spaced in frequency around the carrier signal ($E_{c,PM}$). By comparing eq.'s 3.6 - 3.8 and using eq. 3.3, we see that the carrier and sidebands are determined by the voltage and frequency of the RF drive signal and the modulator $V_{\pi RF}$.

Intensity Modulation

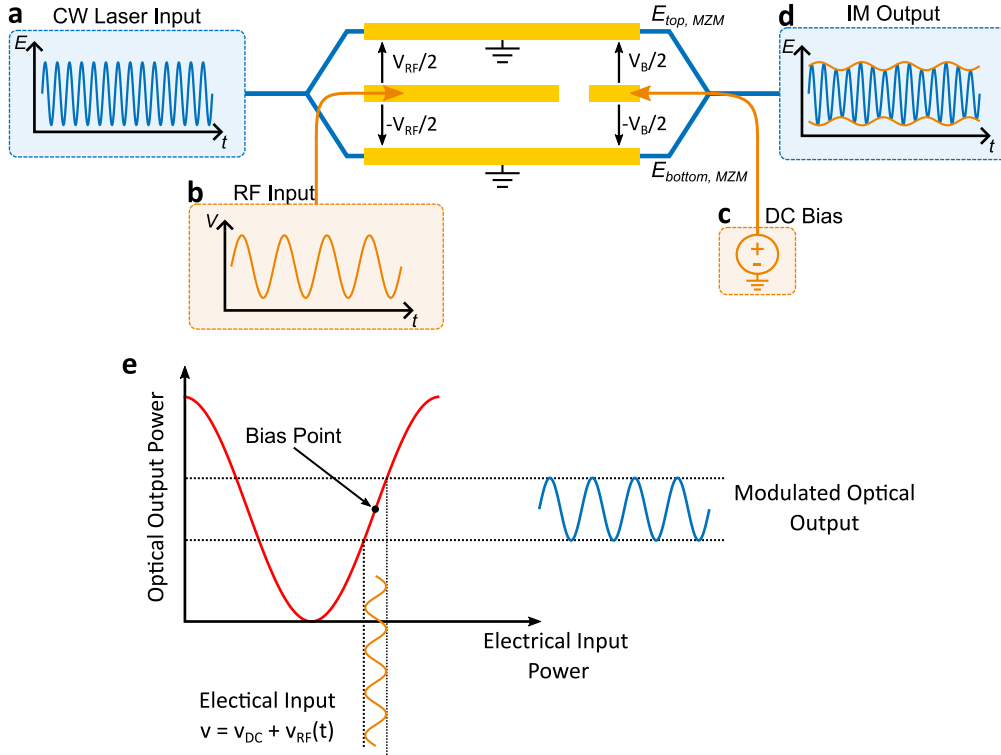


Figure 3.2: Schematic diagram of an intensity modulator. (a) An optical input signal is (b) modulated by the input RF signal. (c) The modulator is biased using a DC voltage source. (d) The intensity of the output optical signal is modulated by the RF input. (e) The output optical power is dependent on both the RF electrical input, and the DC bias of the modulator. The yellow strips represent electrical contacts that drive a phase modulator.

Intensity modulation (IM) is achieved by embedding phase modulators in an interferometer. When the outputs from each arm are in phase, they sum constructively when recombined at the modulator output. In contrast, destructive interference occurs when the superimposed signals differ in phase by π [1]. The Mach-Zehnder interferometer structure is almost exclusively used to achieve IM. Intensity modulators are therefore often referred to as Mach-Zehnder modulators (MZMs).

The key features of an MZM are highlighted in Fig. 3.2. A CW optical signal, Fig. 3.2 (a), is modulated by an input RF tone, Fig. 3.2 (b). In this case, each arm is modulated by a voltage $\pm V_{RF}/2$, and experiences a DC bias of $\pm V_B/2$ [1]. The configuration of an MZM where a phase modulator drives only a single arm of the MZM is also possible and leads

to the same RF output of the MWP link, as discussed in Section 6.1 of Ref [1]. We can express the output of the electric field in the top and bottom arms of the interferometer as [1]

$$E_{top,MZM}(t) = \frac{E_0}{\sqrt{2}} e^{j\omega_c t} e^{j\frac{\phi_{RF}}{2} \sin \omega_{RF} t} e^{j\frac{\phi_B}{2}} \quad (3.9)$$

$$E_{bottom,MZM}(t) = \frac{E_0}{\sqrt{2}} e^{j\omega_c t} e^{-j\frac{\phi_{RF}}{2} \sin \omega_{RF} t} e^{-j\frac{\phi_B}{2}} \quad (3.10)$$

We introduced the MZM DC bias angle (ϕ_B), shown in Fig. 3.2 (c), which is related to the applied DC voltage (V_B) and DC half-wave voltage ($V_{\pi_{DC}}$) by the following expression

$$\phi_B = \frac{\pi V_B}{V_{\pi_{DC}}} \quad (3.11)$$

The output of the MZM can be expressed as [187]

$$E_{out,MZM} = \frac{1}{\sqrt{2}} (E_{top,MZM}(t) + E_{bottom,MZM}(t)) \quad (3.12)$$

Substituting eq. 3.9 and 3.10 into eq. 3.12 and using the results from eq.'s 3.5 - 3.8, we can express the electric field at the output of the MZM, shown in Fig. 3.2 (d), as

$$E_{out,MZM}(t) = E_{l,MZM} e^{j\phi_l} e^{j(\omega_c - \omega_{RF})t} + E_{c,MZM} e^{j\phi_c} e^{j\omega_c t} + E_{u,MZM} e^{j\phi_u} e^{j(\omega_c + \omega_{RF})t} \quad (3.13)$$

Where,

$$E_{l,MZM} = -jE_0 J_1 \left(\frac{\phi_{RF}}{2} \right) \sin \left(\frac{\phi_B}{2} \right) \quad (3.14)$$

$$E_{c,MZM} = E_0 J_0 \left(\frac{\phi_{RF}}{2} \right) \cos \left(\frac{\phi_B}{2} \right) \quad (3.15)$$

$$E_{u,MZM} = jE_0 J_1 \left(\frac{\phi_{RF}}{2} \right) \sin \left(\frac{\phi_B}{2} \right) \quad (3.16)$$

Similar to PM, the IM spectrum consists upper ($E_{u,MZM}$) and lower ($E_{l,MZM}$) modulated sidebands that are equally spaced in frequency around the carrier signal ($E_{c,MZM}$). However, a key distinction between PM and IM can be drawn from eq. 3.14-3.16, as the power of each spectral component of the optical output is dependent on the DC bias voltage. Of particular importance for MWP links is that the optical carrier power can be controlled as a function of DC bias voltage. The sinusoidal dependency on optical output power as a function of DC bias voltage is illustrated in Fig. 3.2 (e). We discuss how the RF output of an MWP link differs for PM and IM in Section 3.2 and how the MZM bias voltage and average optical power impact key MWP link performance metrics in Section 3.3.

Asymmetric Dual-Sideband Modulation

So far, we have outlined PM and IM formats which are widely adopted for use in MWP links. However, intensity and phase modulators do not offer the ability to alter the amplitude *and* phase of the resultant carrier and sidebands, which is often required to implement MWP signal processing systems. Alternatively, the magnitude and phase of the sideband and carrier can be controlled using dual-drive Mach-Zehnder modulators (DDMZMs), or dual-parallel Mach-Zehnder modulators (DPMZMs). We will refer to the broader class of modulation formats that have two sidebands with unequal amplitudes as asymmetric dual-sideband modulation (aDSB) [27]. We outline how DDMZMs and DPMZMs can generate reconfigurable aDSB modulation formats, enabling flexible MWP signal processing applications.

Dual Drive Mach-Zehnder Modulator

A DDMZM, depicted in Fig. 3.3, shares a similar topology to the MZM. Both are typically driven by a CW optical input, Fig. 3.3 (a), however, one key difference is that the phase modulator in each arm of the DDMZM can be driven by independent RF signals.

The output of the PM in the top and bottom arms can be expressed as

$$E_{top,DDMZM} = \frac{E_0}{\sqrt{2}} e^{j(\omega_c t + \phi_{RF,top}(t) + \phi_{B,top})} \quad (3.17)$$

$$E_{bottom,DDMZM} = \frac{E_0}{\sqrt{2}} e^{j(\omega_c t + \phi_{RF,bottom}(t) + \phi_{B,bottom})} \quad (3.18)$$

where $\phi_{RF,top}(t)$, $\phi_{RF,bottom}(t)$, $\phi_{B,top}$ and $\phi_{B,bottom}$ are the RF and DC phase shifts induced in the top and bottom arms of the DDMZM, respectively. Typically, and in the following analysis, we assume that a CW RF signal with a phase difference of θ_{RF} modulates each arm of the DDMZM, as shown in Fig. 3.3 (b), and each arm can be driven by an independent DC bias, as shown in Fig. 3.3 (c). Using these assumptions, we can express the output of the phase modulator in each arm as

$$E_{top,DDMZM} = \frac{E_0}{\sqrt{2}} e^{j(\omega_c t + \phi_{RF} \sin(\omega_{RF} t + \theta_{RF}) + \phi_{B,top})} \quad (3.19)$$

$$E_{bottom,DDMZM} = \frac{E_0}{\sqrt{2}} e^{j(\omega_c t + \phi_{RF} \sin(\omega_{RF} t) + \phi_{B,bottom})} \quad (3.20)$$

We can express the output of the DDMZM, Fig. 3.3 (d), as [187]

$$E_{out,DDMZM} = \frac{1}{\sqrt{2}} (E_{top,DDMZM} + E_{bottom,DDMZM}) \quad (3.21)$$

Substituting eq.'s 3.19 and 3.20 into eq. 3.21, and assuming a bias voltage is not applied to the lower arm of the DDMZM ($\phi_{B,bottom} = 0$) we can express the DDMZM output as

$$E_{out,DDMZM} = \frac{E_0}{2} \left(e^{j\phi_{B,top}} e^{j\phi_{RF} \sin(\omega_{RF} t + \theta_{RF})} + e^{j\phi_{RF} \sin(\omega_{RF} t)} \right) e^{j\omega_c t} \quad (3.22)$$

Applying the Jacobi-Anger expansion, we can simplify eq. 3.22 to

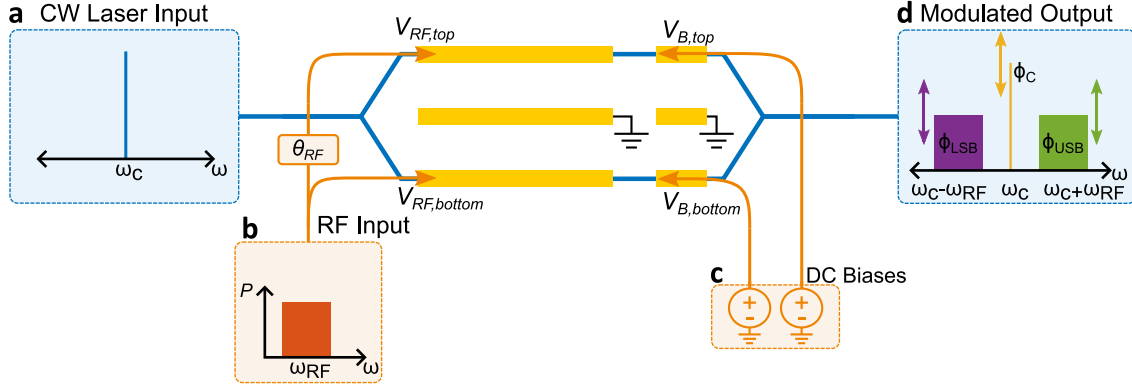


Figure 3.3: Typical use of a DDMZM. (a) Optical input (b) Each arm is driven by an RF signal with a relative phase shift of θ_{RF} between each arm. (c) DC bias for each arm. (d) The output spectrum of the modulator contains a carrier, upper and lower sidebands. The amplitude and phase of each spectral component change with DC bias voltage. The yellow strips represent electrical contacts that drive a phase modulator.

$$E_{out,DDMZM} = \frac{E_0}{2} \sum_{n=-\infty}^{\infty} J_n(\phi_{RF}) e^{jn\omega_{RF}t} (e^{j\phi_{B,top}} e^{jn\theta_{RF}} + 1) e^{j\omega_c t} \quad (3.23)$$

We can therefore express the optical field of the carrier and first-order sidebands as

$$E_{l,DDMZM} = \frac{E_0}{2} J_{-1}(\phi_{RF}) (e^{j(\phi_{B,top} - \theta_{RF})} + 1) e^{j(\omega_c - \omega_{RF})t} \quad (3.24)$$

$$E_{c,DDMZM} = \frac{E_0}{2} J_0(\phi_{RF}) (e^{j\phi_{B,top}} + 1) e^{j\omega_c t} \quad (3.25)$$

$$E_{u,DDMZM} = \frac{E_0}{2} J_1(\phi_{RF}) (e^{j(\phi_{B,top} + \theta_{RF})} + 1) e^{j(\omega_c + \omega_{RF})t} \quad (3.26)$$

The most common values for θ_{RF} for MWP applications are π and $\pi/2$. We analyse the DDMZM output for each case below.

We use eq.'s 3.24-3.26 to plot the normalised power and phase of the electric fields of the DDMZM output are shown for $\theta_{RF} = \pi$ in Fig. 3.4 (a-b), respectively. We see that the power of the upper and lower sideband are equal, but differ in phase by π . Additionally, the carrier power is maximum when the sideband power is minimum, and vice versa. The relationship between the carrier and sideband magnitude and phase for the DDMZM with $\theta_{RF} = \pi$ matches the MZM case. This can be understood by noting that a π phase shift results in a sign inversion of the RF signal and therefore the DDMZM in Fig. 3.3 results in the same configuration as the MZM in Fig. 3.2 (a) with the substitution $V_{RF,top} = V_{RF}/2$, $V_{RF,bottom} = -V_{RF}/2$, $V_{B,top} = V_B/2$ and $V_{B,bottom} = -V_B/2$.

We use eq.'s 3.24-3.26 to plot the normalised power and phase of the electric field of the carrier and sidebands of the DDMZM output when $\theta_{RF} = \pi/2$ in Fig. 3.5 (a-b). The output modulation formats vary between single-sideband modulation (SSB) for $\phi_{B,top} \in \{\pi/2, 3\pi/2\}$, carrier-suppressed dual-sideband modulation at $\phi_{B,top} = \pi$, and finally dual-sideband with carrier modulation at $\phi_{B,top} = 0$. These cases are shown in Fig. 3.5 (c). We annotated each sideband with its phase relative to the carrier.

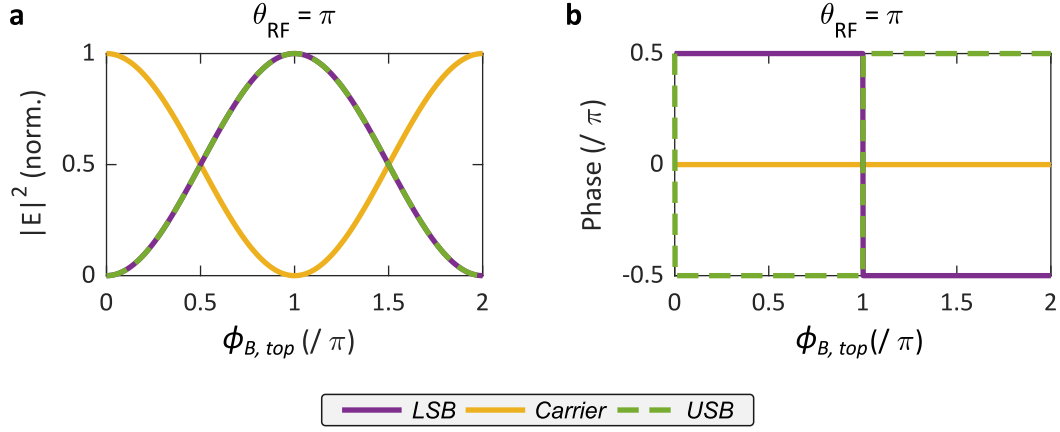


Figure 3.4: (a) Power and (b) phase of the lower sideband, carrier and upper sideband at the output of a DDMZM as a function of DC bias angle of the upper arm of the DDMZM ($\phi_{B,top}$). These plots assume a π phase shift between RF drive arms.

There are two critical distinction to be made when comparing the output spectrum for a DDMZM between the $\theta_{RF} = \pi/2$ and π cases. First, the upper and lower sideband amplitudes of the DDMZM output are not necessarily equal for all bias voltages when $\theta_{RF} = \pi/2$, whereas they are equal for all applied voltages when $\theta_{RF} = \pi$. A second distinction is the relative phase difference between the sidebands for both cases are different, which dramatically affects the RF response. We illustrate this with an example in Section 3.2.

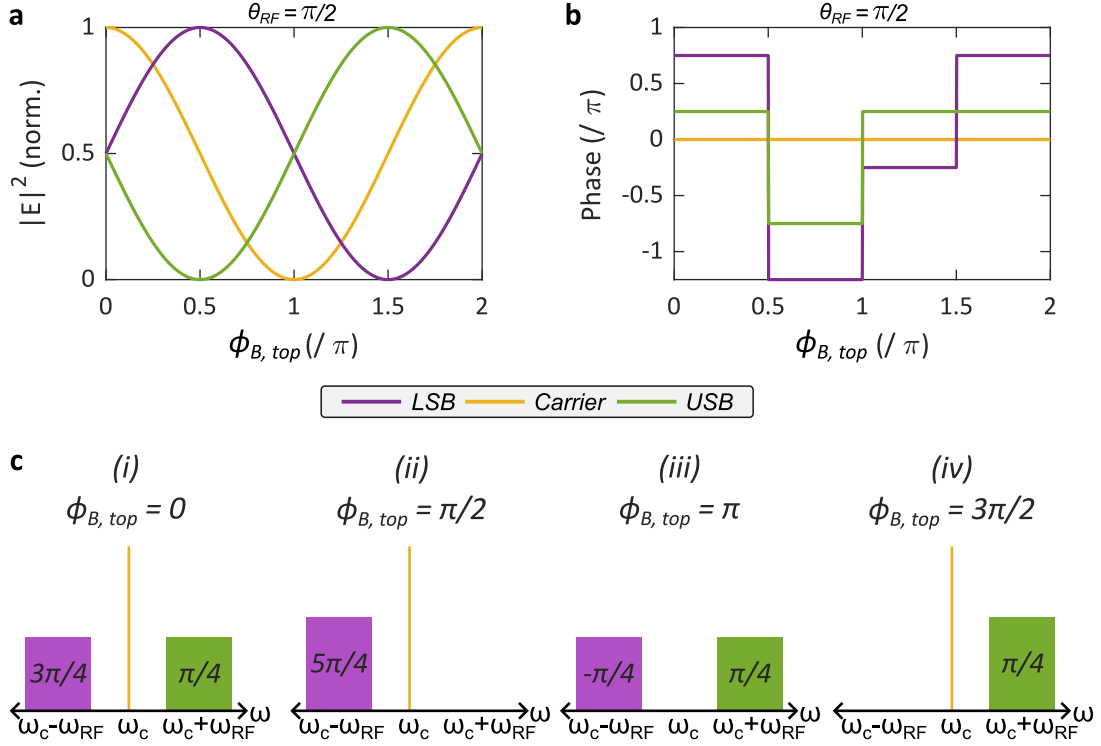


Figure 3.5: (a) Power and (b) phase of the lower sideband, carrier and upper sideband at the output of a DDMZM as a function of bias angle in the upper arm of the DDMZM. These plots assume a $\pi/2$ phase shift between RF drive arms. (c) The output spectrum that corresponds to bias voltages for the cases $\phi_{B, top} \in \{0, \pi/2, \pi, 3\pi/2\}$.

Dual Parallel Mach-Zehnder Modulator

A DPMZM provides the most flexible control over the amplitude and phase of its output spectral components. When driven with a $\pi/2$ RF hybrid coupler the output modulation format can be controlled between IM, SSB, PM and aDSB [30], as shown in Fig. 3.6. The output of a DPMZM can be expressed as [177], [287]

$$E_{out, DPMZM} = \frac{\sqrt{2}}{2} \left[E_{l, DPMZM} e^{j\phi_l} e^{-j\omega_{RF}t} + E_{c, DPMZM} e^{j\phi_c} + E_{u, DPMZM} e^{j\phi_u} e^{j\omega_{RF}t} \right] e^{j\omega_c t} \quad (3.27)$$

where the complex coefficient of the carrier and sidebands is given by the expressions

$$E_{l, DPMZM} e^{j\phi_l} = E_0 J_1(\phi_{RF}) \left(j \sin(\phi_{B,1}) - \sin(\phi_{B,2}) \cdot e^{j\phi_{B,3}} \right) \quad (3.28)$$

$$E_{c, DPMZM} e^{j\phi_c} = E_0 J_0(\phi_{RF}) \left(\cos(\phi_{B,1}) + \cos(\phi_{B,2}) \cdot e^{j\phi_{B,3}} \right) \quad (3.29)$$

$$E_{u, DPMZM} e^{j\phi_u} = E_0 J_1(\phi_{RF}) \left(-j \sin(\phi_{B,1}) - \sin(\phi_{B,2}) \cdot e^{j\phi_{B,3}} \right) \quad (3.30)$$

Owing to the ability to independently control the amplitude and phase of the carrier and sidebands, DPMZMs have found extensive use in many MWP applications [27]. Despite the flexibility the DPMZM provides, some limitations of DPMZMs include bias drift [288], and they typically exhibit higher losses than intensity or phase modulators.

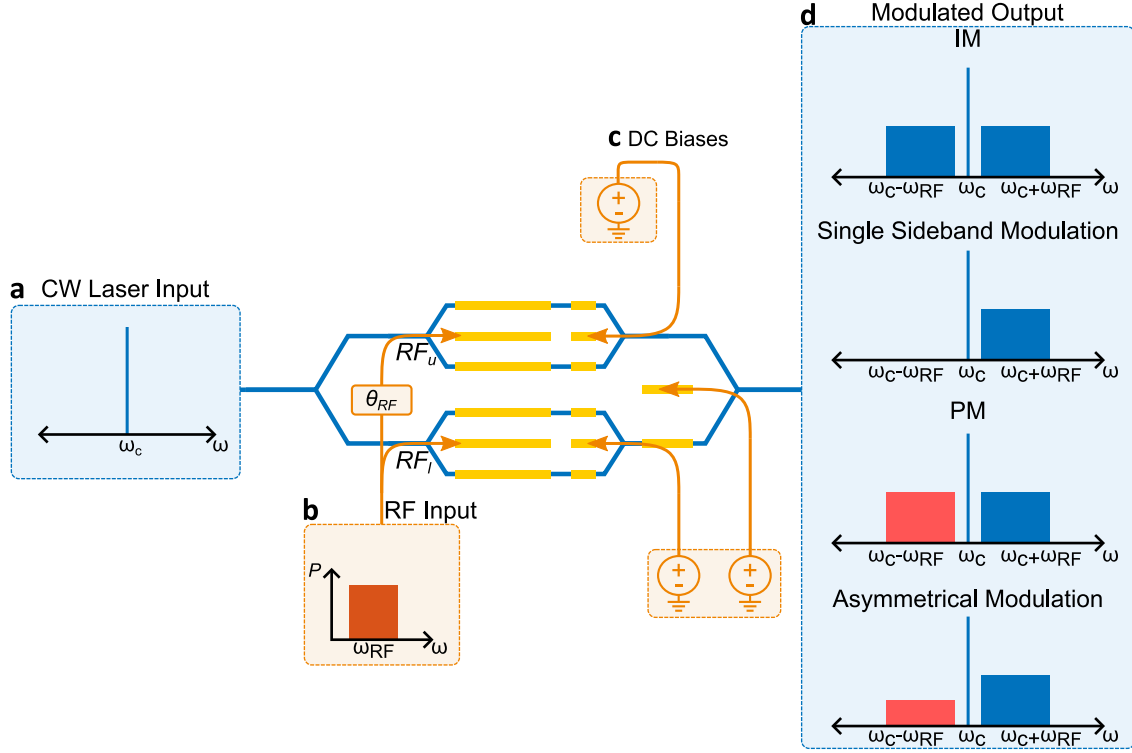


Figure 3.6: Dual parallel Mach-Zehnder modulator. (a) An input signal is modulated by a (b) RF input signal. Based on the (c) modulator bias, the (d) optical output spectrum can be configured to provide a flexible range of output modulation schemes. The yellow strips represent electrical contacts that drive a phase modulator.

3.1.2 Photodetection

Photodetection is a square-law process that converts the intensity of the optical signal to an electrical current, referred to as photocurrent. The photodetection process can be represented by the following equation [289], [290]

$$i_{PD}(t) = r_{PD}P_{opt,pd} = r_{PD}EE^* \quad (3.31)$$

where r_{PD} represents the responsivity of the photodetector (A/W), $P_{opt,pd}$ and E are the optical power and electric field at the photodetector and the operator $*$ denotes the complex conjugate. Additionally, we can use the relationship between current and electrical power ($P = i^2R$) to express the RF output power of the photodetector as

$$P_{PDRF} = i^2R_L = r_{PD}^2P_{opt,pd}^2R_L \quad (3.32)$$

where R_L denotes the electrical load impedance and is typically 50Ω for RF and MWP systems. For MWP processing applications, important specifications of a photodetector also include the response bandwidth and maximum optical power handling.

Fig. 3.7 shows a high-level schematic of an MWP link. An RF signal undergoes E-O modulation, Fig. 3.7 (a), whereby the RF spectrum is translated to the optical domain. We showed in Section 3.1.1 that the output optical signal from a modulator under small-signal conditions consists of a carrier, lower and upper sideband, which can be expressed as

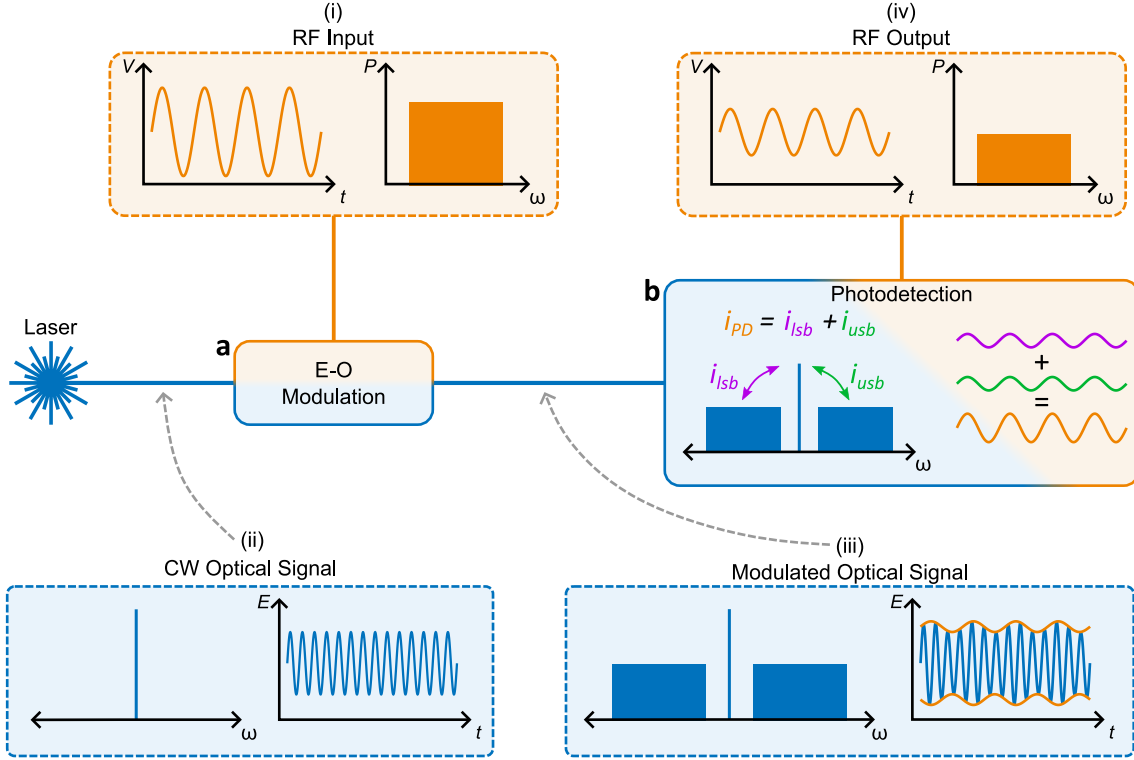


Figure 3.7: High-level MWP link schematic highlighting the function of the photodetector. (a) An RF input signal modulates a CW laser. (b) Upon photodetection, the beat between the lower sideband and the carrier, as well as the upper sideband and the carrier form photocurrents which are added to create the RF output of the MWP link.

$$E_{pd}(t) = E_l e^{j\phi_l} e^{j(\omega_c - \omega_{RF})t} + E_c e^{j\phi_c} e^{j\omega_c t} + E_u e^{j\phi_u} e^{j(\omega_c + \omega_{RF})t} \quad (3.33)$$

where $E_{l,c,u}$, $\phi_{l,c,u}$ correspond to the magnitude and phase of the lower sideband, carrier, and upper sideband at the photodetector, respectively. Let us now consider the PD output when such an optical signal undergoes photodetection, as shown in Fig. 3.7 (b). Using the result from eq. 3.31, we can express the RF photocurrent as

$$i_{pd}(t) = r_{PD} \left(E_l e^{j\phi_l} e^{j(\omega_c - \omega_{RF})t} + E_c e^{j\phi_c} e^{j\omega_c t} + E_u e^{j\phi_u} e^{j(\omega_c + \omega_{RF})t} \right) \times \text{c.c.} \quad (3.34)$$

where c.c. denotes the complex conjugate. Expanding this expression we get

$$\begin{aligned} i_{PD}(t) = r_{PD} \Big[& (|E_l|^2 + |E_c|^2 + |E_u|^2) + \\ & (E_c E_l e^{j(\omega_c - (\omega_c - \omega_{RF}))t} e^{j(\phi_c - \phi_l)} + E_c E_u e^{j(\omega_c + \omega_{RF} - \omega_c)t} e^{j(\phi_u - \phi_c)} + \text{c.c.}) + \\ & (E_u E_l e^{j(\omega_c + \omega_{RF} - (\omega_c - \omega_{RF}))t} e^{j\phi_u - \phi_l} + \text{c.c.}) \Big] \end{aligned} \quad (3.35)$$

Simplifying this expression, we get

$$\begin{aligned} i_{PD}(t) = r_{PD} \Big[& (|E_l|^2 + |E_c|^2 + |E_u|^2) + \\ & (E_c E_l e^{j\omega_{RF}t} e^{j(\phi_c - \phi_l)} + E_c E_u e^{j\omega_{RF}t} e^{j(\phi_u - \phi_c)} + \text{c.c.}) + \\ & (E_u E_l e^{j2\omega_{RF}t} e^{j\phi_u - \phi_l} + \text{c.c.}) \Big] \end{aligned} \quad (3.36)$$

applying the identity $z + z^* = 2\text{Re}(z)$, we get

$$\begin{aligned} i_{PD}(t) = r_{PD} \big[& (|E_l|^2 + |E_c|^2 + |E_u|^2) + \\ & 2E_c E_l \cos(\omega_{RF}t + \phi_c - \phi_l) + 2E_c E_u \cos(\omega_{RF}t + \phi_u - \phi_c) + \\ & 2E_u E_l \cos(2\omega_{RF}t + \phi_u - \phi_l) \big] \end{aligned} \quad (3.37)$$

The photocurrent output consists of three frequency components, an average DC component, the fundamental output, which has the same frequency as the RF input, ω_{RF} , and a harmonic term with frequency $2\omega_{RF}$. Noting the relation between photocurrent and optical power from eq. 3.31, we can use eq. 3.37 to express the average and modulated optical power as a sum of average power, $P_{opt,avg}$, as well as time varying components, $P_{opt,\omega_{RF}}$ and $P_{opt,2\omega_{RF}}$, given by

$$i(t) = r_{PD} (P_{opt,avg} + P_{opt,\omega_{RF}} + P_{opt,2\omega_{RF}}) \quad (3.38)$$

Each spectral component of the photocurrent and optical field can therefore be represented in phasor form as

$$I_{pd}(0) = r_{PD} P_{opt,avg} = r_{PD} (|E_l|^2 + |E_c|^2 + |E_u|^2) \quad (3.39)$$

$$I_{pd}(\omega_{RF}) = r_{PD} P_{opt,\omega_{RF}} = 2r_{PD} (E_c E_u e^{j(\phi_u - \phi_c)} + E_c E_l e^{j(\phi_c - \phi_l)}) \quad (3.40)$$

$$I_{pd}(2\omega_{RF}) = r_{PD} P_{opt,2\omega_{RF}} = 2r_{PD} E_u E_l e^{j(\phi_u - \phi_l)} \quad (3.41)$$

we use $I_{pd}(\omega)$ to denote the frequency domain representation of the RF output photocurrent with angular frequency ω . $I_{pd}(\omega_{RF})$ is the RF output, which is particularly important for MWP signal processing applications as it corresponds to the RF input transported through or processed by the MWP link.

By analysing eq. 3.40, we can see that the fundamental RF output component with angular frequency ω_{RF} consists of two main photocurrent contributions. One contribution is from the beat between the carrier and upper sideband, and the second contribution is from the beat between the carrier and lower sideband, as shown in Fig. 3.7 (b).

3.2 Implementation of MWP Filtering

We now demonstrate how the modulation schemes presented in Section 3.1.1 can be used to implement an MWP filter. The key feature of an RF filter is that its response has distinct passbands, where a signal propagates through the system, and has stopbands, where a signal is attenuated relative to the passband. In this section, we consider coherent optical filters, whereby optical resonances used to impart a frequency-selective amplitude change and phase shift to the modulated optical signal that in-turn shapes the RF response upon photodetection. Examples of optical processing techniques include MRRs and SBS, which were outlined in Chapter 2.

The key to understanding how the RF response changes with an applied optical resonance is to analyse the amplitude and relative phase of the RF photocurrent contributions

inside and outside the applied optical resonance. We show that the MWP modulation scheme significantly impacts the RF response of the MWP link, to the extent that bandpass and notch MWP filter configurations can be achieved by changing the optical modulation format, even if the applied optical transfer function does not change. In the following analysis, we express the optical transfer function with respect to optical power between the modulator and PD as $H_{\text{opt}}(\omega) = |H_{\text{opt}}(\omega)|e^{j\phi_{\text{opt}}(\omega)}$. We also refer to the RF transfer function of the MWP link in terms of the scattering parameter, s_{21} [291].

3.2.1 Single Sideband Modulation

Consider the case of SSB, which can be achieved using either DDMZM or DPMZM. Alternatively, one can use a phase or intensity modulator and remove one sideband with an optical filter such that $E_1 = 0$, as shown in Fig. 3.8 (a). We can express the RF output photocurrent by substituting the electric field that we derived for the PM output in eq. 3.6 - 3.8, noting that $E_1 = 0$, which gives the expression

$$I_{SSB}(\omega_{RF}) = r_{PD} 2E_0^2 J_0(\phi_{RF}) J_1(\phi_{RF}) |H_{\text{opt}}(\omega_c + \omega_{RF})|^{\frac{1}{2}} e^{j\phi_{\text{opt}}(\omega_c + \omega_{RF})} \quad (3.42)$$

This reveals that there is only one photocurrent contribution in the RF output of the PD, as illustrated in Fig. 3.8 (b). Furthermore, we see that the transfer function of the MWP link ($s_{21} \propto I^2$) is directly proportional to the optical transfer function through the MWP link

$$s_{21SSB}(\omega_{RF}) \propto |H_{\text{opt}}(\omega_c + \omega_{RF})| e^{j\phi_{\text{opt}}(\omega_c + \omega_{RF})} \quad (3.43)$$

This is illustrated in Fig. 3.8 (c), as the RF transfer function (orange) is proportional to the optical transfer function (red line) of the MWP link. Specifically, in this example, the depth of the optical resonance is equal to the depth of the notch in the RF response. SSB can be used to measure the optical transfer function of an MWP link because the RF transfer function of the MWP link is proportional to the optical transfer function. However, the direct mapping of the optical to RF response requires large optical notch depths to synthesise an MWP filter with large out-of-band rejection, which limits the flexibility of the underlying MWP filter.

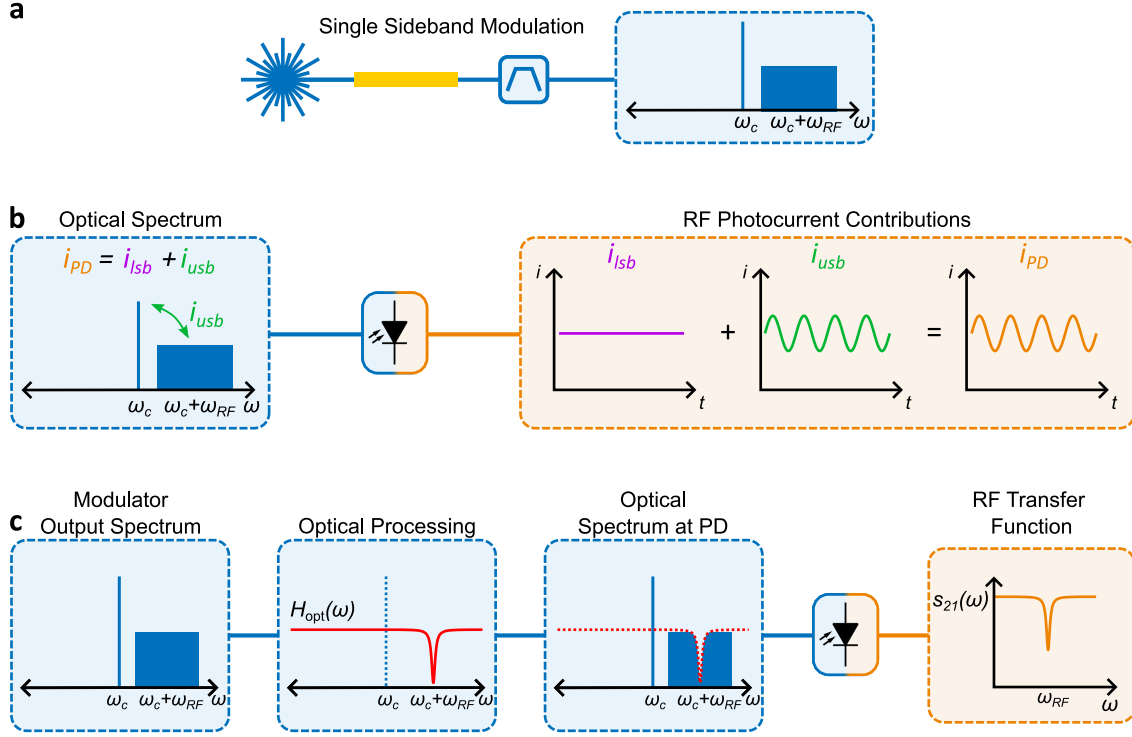


Figure 3.8: RF output of MWP link using SSB modulation. (a) High-level schematic of single sideband modulation. (b) RF photocurrent contribution. (c) Optical and RF transfer functions.

3.2.2 Phase Modulation

Fig. 3.9 (a) depicts the output spectrum of a phase modulator. By substituting the electric field at the output of a phase modulator that we derived in eq's. 3.6 - 3.8 into the photodetector output in eq. 3.40, and assuming the lower sideband experiences a unity transfer function, we find the RF photocurrent is given by

$$I_{PM}(\omega_{RF}) = r_{PD} E_0^2 J_0(\phi_{RF}) J_1(\phi_{RF}) \left(|H_{opt}(\omega_c + \omega_{RF})|^{\frac{1}{2}} e^{j\phi_{opt}(\omega_c + \omega_{RF})} - 1 \right) \quad (3.44)$$

From this, we see that the transfer function of the MWP link ($s_{21} \propto I^2$) is related to the optical transfer function as follows

$$s_{21_{PM}}(\omega_{RF}) \propto \left(|H_{opt}(\omega_c + \omega_{RF})|^{\frac{1}{2}} e^{j\phi_{opt}(\omega_c + \omega_{RF})} - 1 \right)^2 \quad (3.45)$$

We examine the case where no optical processing is applied, which corresponds to $H_{opt}(\omega) = 1$, in Fig. 3.9 (b), where the net RF photocurrent consists of two contributions that are equal in magnitude but π out-of-phase and therefore sum *destructively*. However, if one sideband changes in amplitude or phase, a residual RF photocurrent exists as the resultant photocurrents no longer destructively interfere. This is illustrated for the case $|H_{opt}(\omega_c + \omega_{RF})| < 1$ in Fig. 3.9 (c).

We illustrate a typical RF transfer function of an MWP link that contains a single resonance in Fig. 3.9 (d). Outside the resonance, the sideband amplitudes are equal and therefore destructively interfere. However, when the upper sideband is attenuated, the sideband

amplitudes are no longer equal and destructive interference does not occur, which results in a residual signal at the output of the PD. As a result, an RF bandpass filter response is formed.

Furthermore, we plot the MWP link gain as a function of $|H_{\text{opt}}(\omega)|$ in Fig 3.10. In this plot we assume $\phi_{\text{opt}} = 0$. We observe that relatively small changes in $|H_{\text{opt}}(\omega)|$ can create large changes in the RF transfer function due to the interferometric nature of the PD photocurrents. This is attractive for MWP filtering applications, as large RF rejection ratios can be achieved with only a modest change in optical amplitude [174], [292].

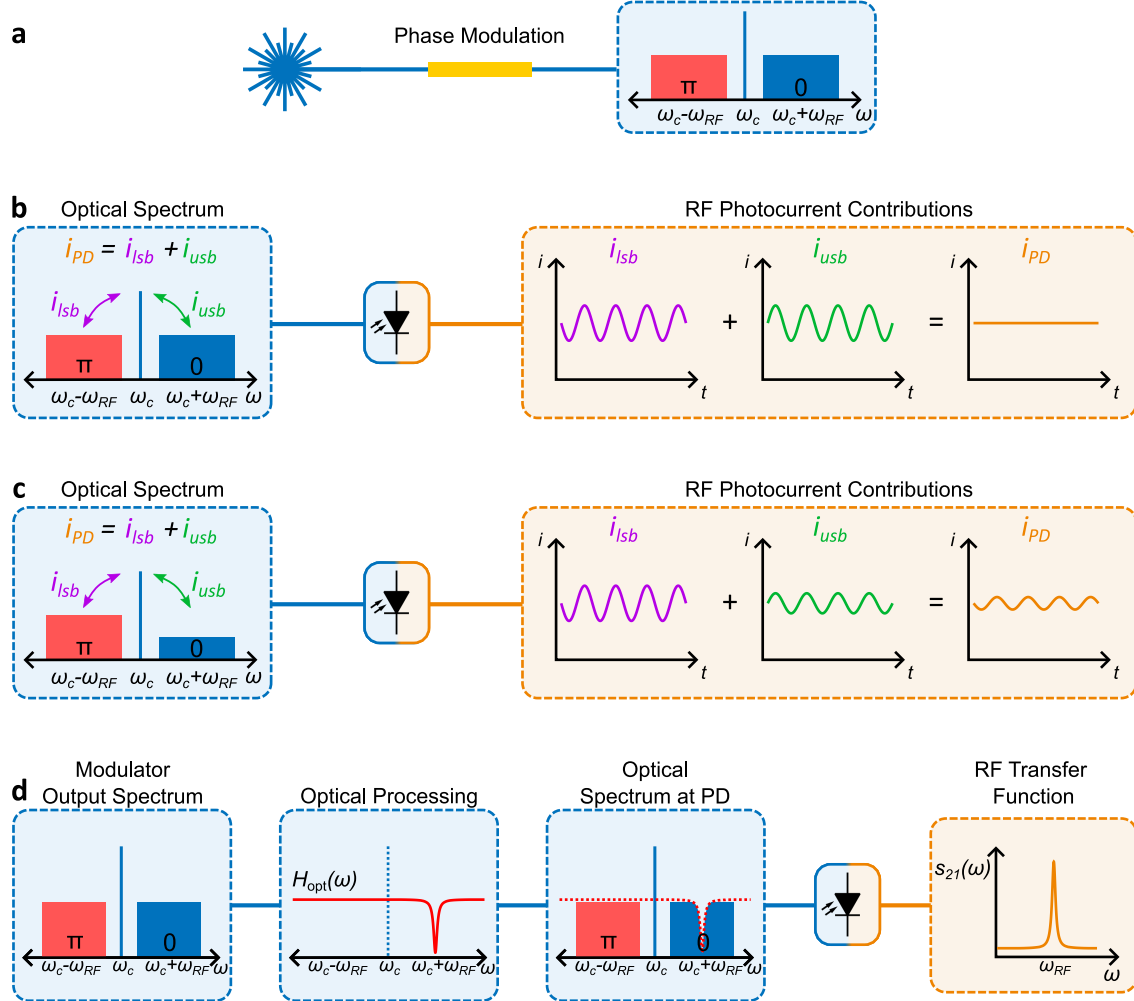


Figure 3.9: RF output of MWP link using phase modulation. (a) High-level schematic of phase modulation. (b) RF photocurrent contributions when the sideband amplitudes are equal and destructively interfere. (c) RF photocurrent contributions when the sideband amplitudes are unequal and do not destructively interfere. (d) Optical and RF transfer functions.

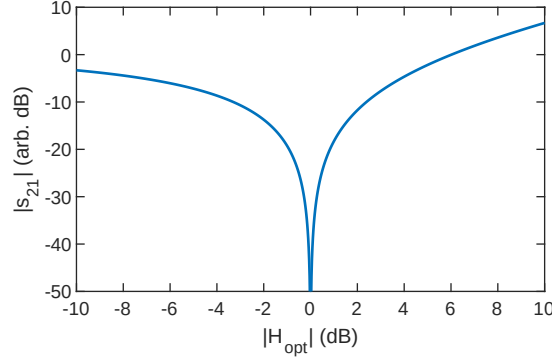


Figure 3.10: RF transfer function vs optical transfer function for phase modulation.

3.2.3 Intensity Modulation

The output photocurrent of an MWP link using IM can be found by substituting the expressions we derived for the electric field at the output of an intensity modulator in eq's. 3.14 - 3.16 into the RF photocurrent expression in eq. 3.40. We find the output RF photocurrent is given by

$$\begin{aligned}
 I_{IM}(\omega_{RF}) &= r_{PD} E_0^2 J_0\left(\frac{\phi_{RF}}{2}\right) \cos\left(\frac{\phi_B}{2}\right) J_1\left(\frac{\phi_{RF}}{2}\right) \sin\left(\frac{\phi_B}{2}\right) \left(j |H_{opt}(\omega_c + \omega_{RF})|^{\frac{1}{2}} e^{j\phi_{opt}(\omega_c + \omega_{RF})} - j^* \right) \\
 &= j r_{PD} E_0^2 J_0\left(\frac{\phi_{RF}}{2}\right) J_1\left(\frac{\phi_{RF}}{2}\right) \cos\left(\frac{\phi_B}{2}\right) \sin\left(\frac{\phi_B}{2}\right) \left(|H_{opt}(\omega_c + \omega_{RF})|^{\frac{1}{2}} e^{j\phi_{opt}(\omega_c + \omega_{RF})} + 1 \right) \\
 &= \frac{j r_{PD} E_0^2}{2} J_0\left(\frac{\phi_{RF}}{2}\right) J_1\left(\frac{\phi_{RF}}{2}\right) \sin(\phi_B) \left(|H_{opt}(\omega_c + \omega_{RF})|^{\frac{1}{2}} e^{j\phi_{opt}(\omega_c + \omega_{RF})} + 1 \right)
 \end{aligned} \tag{3.46}$$

Using this result, we find the transfer function of the MWP link ($s_{21} \propto I^2$) relates to the optical transfer function as follows

$$s_{21_{IM}}(\omega_{RF}) \propto \left(|H_{opt}(\omega_c + \omega_{RF})|^{\frac{1}{2}} e^{j\phi_{opt}(\omega_c + \omega_{RF})} + 1 \right)^2 \tag{3.47}$$

In the absence of any optical processing, the PD output consists of two in-phase photocurrents that sum *constructively*, illustrated in Fig. 3.11 (b). However, by only processing one of the IM sidebands, the RF transfer function is not proportional to the optical transfer function of the MWP link, as depicted by the reduced depth of the RF notch compared to the optical transfer function in Fig. 3.11 (c). This limits the achievable RF rejection ratio of the resulting IM-based MWP filter.

However, RF notch filters with deep RF rejection can be synthesised by processing both IM sidebands such that the photocurrents destructively interfere at one frequency in the RF domain. To understand this, let us consider the case where the lower sideband experiences a transfer function of $H_{opt}(\omega_c - \omega_{RF}) e^{j\phi_{opt}(\omega_c - \omega_{RF})}$, as opposed to the unity transfer function described above. We can therefore modify eq. 3.47 to become

$$s_{21_{IM}}(\omega_{RF}) \propto \left(|H_{opt}(\omega_c + \omega_{RF})|^{\frac{1}{2}} e^{j\phi_{opt}(\omega_c + \omega_{RF})} + |H_{opt}(\omega_c - \omega_{RF})|^{\frac{1}{2}} e^{j\phi_{opt}(\omega_c - \omega_{RF})} \right)^2 \tag{3.48}$$

We illustrate the dual sideband processing technique in Fig. 3.11 (d). A notch is formed in the RF domain if the upper sideband ($\omega_c + \omega_{RF}$) and lower sideband ($\omega_c - \omega_{RF}$) are equal in magnitude and π out-of-phase. This creates a spectrally-localised destructive interference of RF photocurrents and enables large RF notch depths to be formed with modest changes of optical magnitude, similar to the PM case we outlined above in Section 3.2.2.

The dual sideband processing technique has been applied to enable RF notch filters with large RF rejection ratios by using a pair of MRRs, with one over-coupled MRR providing a π phase shift on resonance to one sideband and another under-coupled MRR providing 0 phase shift on resonance to the other sideband [9]. A similar principle has also been exploited by Brillouin-based MWP filters, which overlap the over-coupled MRR, which induces π phase shift, and an SBS gain resonance to compensate for the attenuation of the MRR. The net result is a pair of sidebands with π phase difference and equal amplitudes, giving large notch depths [39], [45], [178].

In Section 3.3 we show that the DC photocurrent of an MWP link output largely determines its output noise power. For systems using IM, we can determine the DC photocurrent by substituting eq.'s 3.14-3.16 into eq. 3.39

$$\begin{aligned}
 I_{IM}(0) &= r_{PD} P_{opt,avg} \\
 &= r_{PD} E_0^2 \left(J_0^2 \left(\frac{\phi_{RF}}{2} \right) \cos^2 \left(\frac{\phi_B}{2} \right) + J_1^2 \left(\frac{\phi_{RF}}{2} \right) \sin^2 \left(\frac{\phi_B}{2} \right) \right) \\
 &= r_{PD} E_0^2 \cos^2 \left(\frac{\phi_B}{2} \right) \\
 &= r_{PD} \frac{P_{opt,pd}}{2} (1 + \cos(\phi_B))
 \end{aligned} \tag{3.49}$$

where we used the approximations $J_0(\phi_{RF}) \approx 1$ and $J_1(\phi_{RF}) \approx 0$ by assuming small signal modulation. We note that some derivations of MZM average power have the term $(1 - \cos(\phi_B))$ where we use the term $(1 + \cos(\phi_B))$. The difference simply arises from deriving the optical propagation through the MZM assuming the use of hybrid couplers, which impart a $\pi/2$ phase shift to the cross-coupled port. In this derivation, we derived the expression using Y-branch couplers that do not impart a $\pi/2$ phase shift between arms [187]. This distinction is discussed in Section 6.2 in Ref. [1]. Due to the periodic nature of the cosine function, the difference in sign here causes a translation of the DC photocurrent with respect to bias voltage and has no meaningful impact on the result.

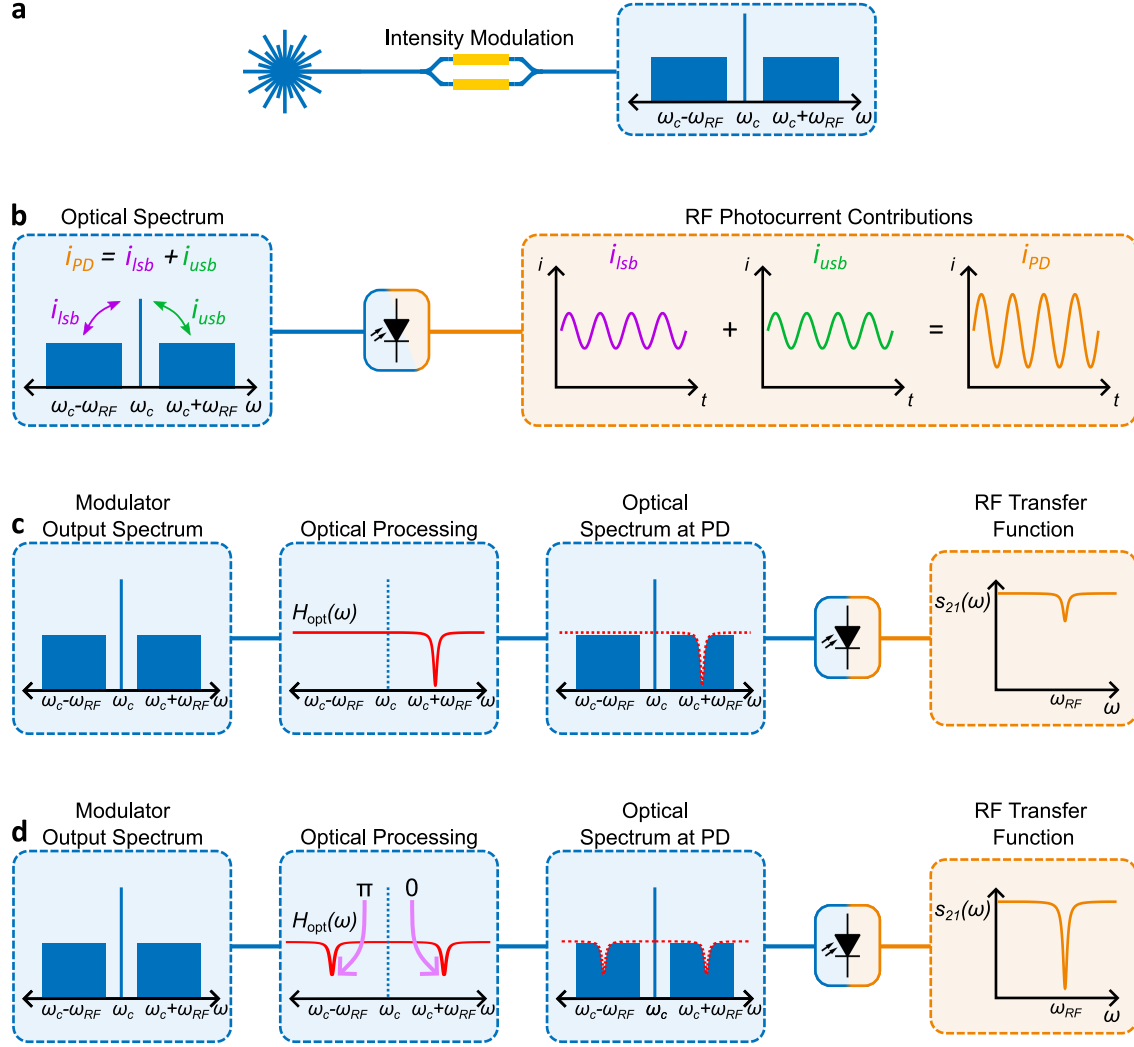


Figure 3.11: RF output of MWP link using intensity modulation. (a) High-level schematic of intensity modulation. (b) RF photocurrent contributions. (c) Optical and RF transfer functions. (d) Optical and RF transfer functions when processing both sidebands imparting equal magnitude response and a π phase difference between the upper and lower sidebands.

3.2.4 Asymmetric Dual-Sideband Modulation

An advantage of using a DPMZM, or DDMZM for MWP signal processing is that the amplitude and phase relationship between the carrier and sidebands can be controlled by the applied bias voltage, highlighted in Fig. 3.12 (a), allowing for a flexible range of RF responses to be synthesised.

An RF notch response can be synthesised by configuring the sideband amplitudes such that they are unequal and the sideband phases such that the resulting photocurrents have a π phase difference. If an optical resonance is then applied such that the sideband amplitudes are equal at a given frequency offset from the carrier, a notch response is formed through destructive interference of RF photocurrents, as shown in Fig. 3.12 (b).

Alternatively, an RF bandpass response can be synthesised by configuring the sideband amplitudes such that they are equal and the sideband phases such that the resulting pho-

to currents have a π phase difference. If an optical resonance is then applied such that the sideband amplitudes are unequal at a given frequency offset from the carrier, a band-pass response is formed as the destructive interference condition is no longer satisfied, as shown in Fig. 3.12 (c).

We can summarise the RF transfer function of an MWP link for the cases outlined above as a function of the amplitude difference between each π out-of-phase sideband. In the case that the electric fields of the upper and lower sidebands and therefore the corresponding photocurrent contributions differ by a factor $\frac{1}{\sqrt{\Delta|H_{\text{opt}}|}}$, we can express the net RF photocurrent output of the photodetector as

$$I_{\text{aDSB}}(\omega_{\text{RF}}) \propto i_{\text{USB}} - \frac{i_{\text{USB}}}{\sqrt{\Delta|H_{\text{opt}}|}} \quad (3.50)$$

$$\propto i_{\text{USB}} \times \left(\frac{\sqrt{\Delta|H_{\text{opt}}|} - 1}{\Delta|H_{\text{opt}}|} \right)$$

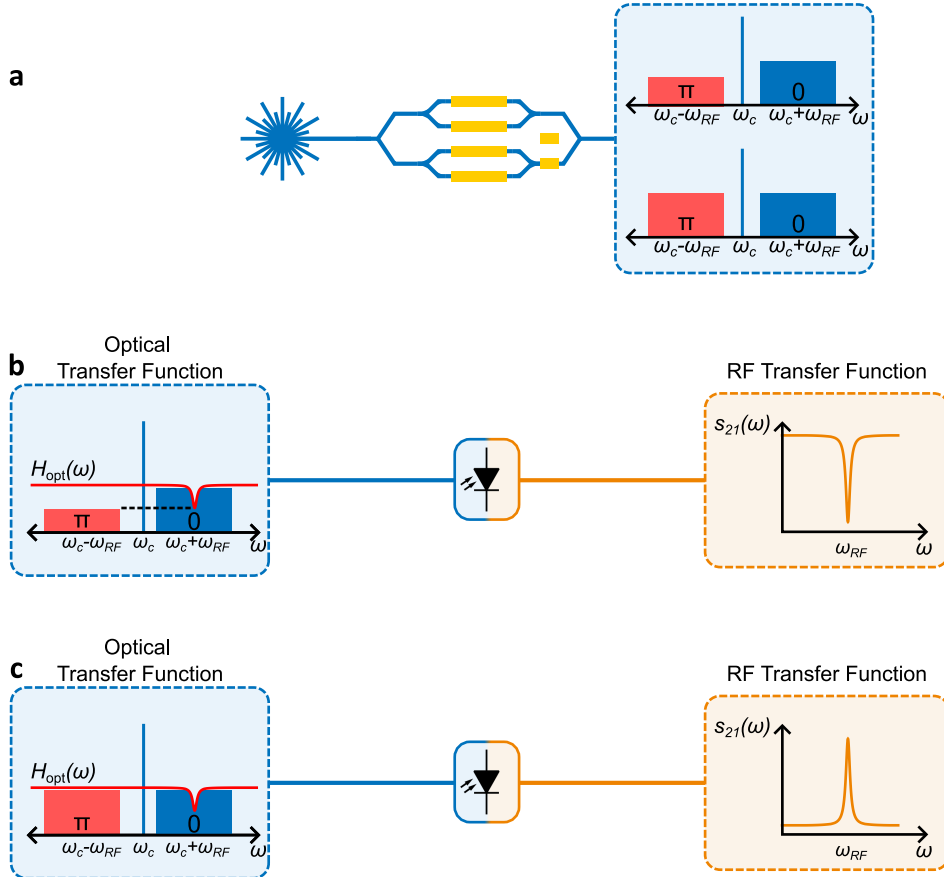


Figure 3.12: RF output of MWP link using asymmetric dual-sideband modulation formats. (a) High-level schematic of modulation format generation using a DPMZM. (b) Implementation of notch filtering functionality. (c) Implementation of bandpass filtering functionality.

where i_{USB} refers to the RF photocurrent contribution from the beat between the carrier and the upper sideband. We can therefore express the resultant RF transfer function ($s_{21} \propto I^2$) as

$$s_{21_{\text{aDSB}}}(\omega_{\text{RF}}) \propto \left(\frac{\sqrt{\Delta|H_{\text{opt}}|} - 1}{\Delta|H_{\text{opt}}|} \right)^2 \quad (3.51)$$

To provide a concrete example of how the MWP response can change between band-pass to notch responses, let us consider an MWP link using a DDMZM with a $\pi/2$ phase shift between the RF signal in each arm of the modulator, as depicted in Fig. 3.13 (a).

We calculated the net RF photocurrent of the PD, $I(\omega_{\text{RF}})$, by substituting the electric field of the DDMZM output expression we derived in eq's 3.24-3.26 (using $\theta_{\text{RF}} = \pi/2$), into the RF component of the PD output given by eq. 3.40. We plot the magnitude of the PD output photocurrent contributions in Fig. 3.13 (b). The green trace represents the photocurrent contribution from the beat between the carrier and upper sideband, the purple trace represents the photocurrent contribution from the beat between the carrier and lower sideband. The dashed line represents the resultant PD output which is the vector sum of both photocurrent contributions. The phase of each photocurrent contribution is plotted in Fig. 3.13 (c) and the difference in phase between the two photocurrent contributions ($\Delta\angle I(\omega_{\text{RF}})$) is plotted in Fig. 3.13 (d).

To achieve destructive interference, we require the photocurrent contributions to have a relative phase difference that satisfies $\Delta\angle I(\omega_{\text{RF}}) = (2n + 1)\pi$, which occurs for $0 < \phi_{B,\text{top}} < \pi/2$ and $1.5\pi < \phi_{B,\text{top}} < 2\pi$, as depicted in Fig. 3.13 (d). Additionally, we require the RF photocurrent contributions to be equal in magnitude.

Fig. 3.13 (e) illustrates how a notch response can be achieved by considering the case where $0 < \phi_{B,\text{top}} < \pi/2$, such that the sideband amplitudes are unequal and resultant photocurrents are π out-of-phase. By tuning the depth of an optical loss resonance such that the sideband amplitudes are equal at a specific frequency, a notch response is formed.

Conversely, a bandpass response can be synthesised if the $\phi_{B,\text{top}} = 0$, such that the sideband amplitudes are equal and hence destructively interfere. When an optical loss resonance is applied to one sideband, a bandpass response is formed because the condition for destructive interference is no longer satisfied, as depicted in Fig. 3.13 (f).

Therefore, the RF link response of an MWP filter can alternate between bandpass and notch configurations by changing the DC bias voltage of the DDMZM. A notch response occurs when the RF photocurrents destructively interfere either when an optical resonance is applied, and a bandpass response occurs when the RF photocurrents destructively interfere outside the optical resonance. This technique has been used to create MWP filters with reconfigurable bandpass and notch responses [150], [172], [290].

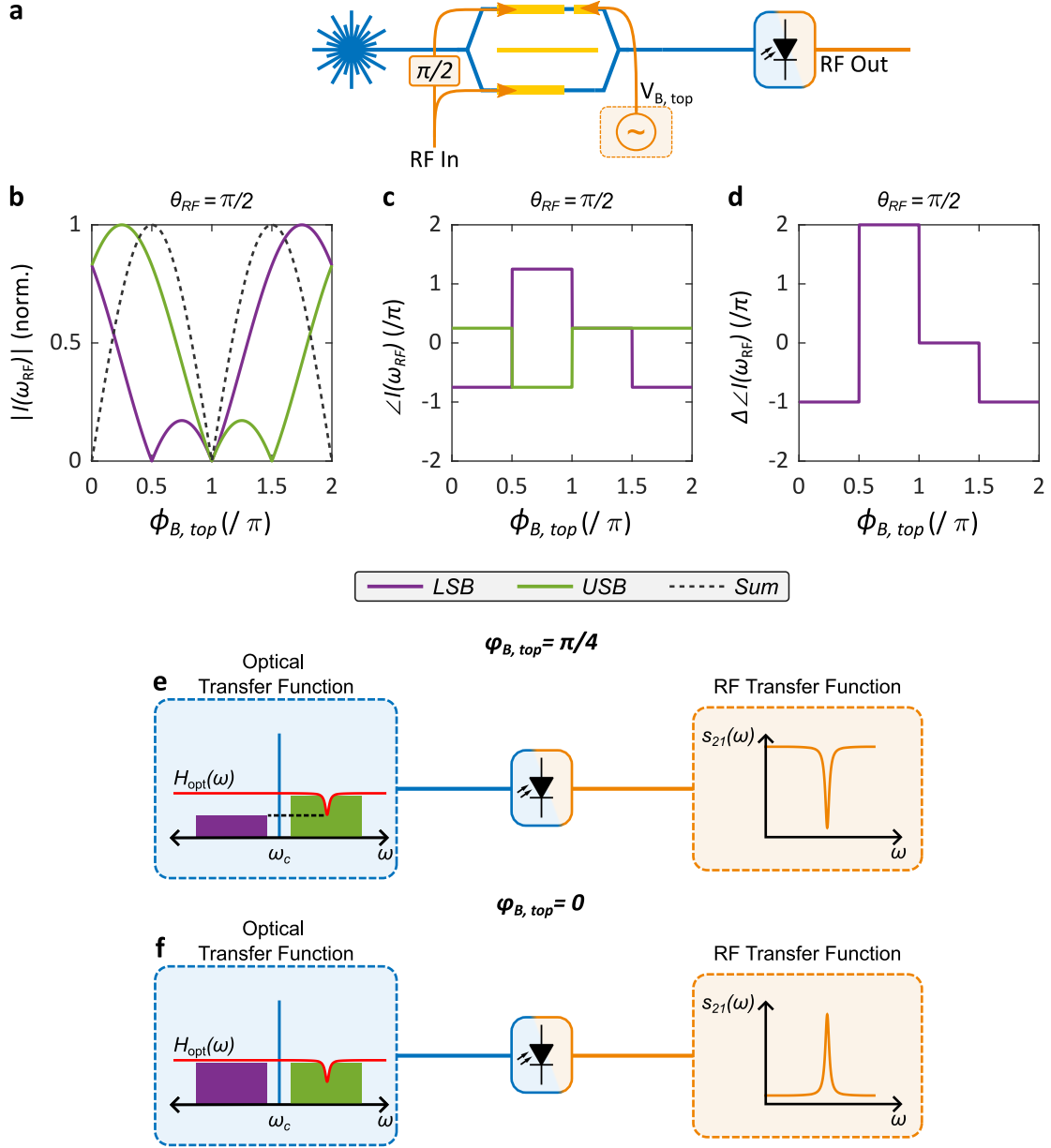


Figure 3.13: Example RF output of MWP link using DDMZM. (a) High-level schematic of modulation format generation using a DDMZM. (b) Magnitude of RF photocurrent contributions from the lower sideband (purple), upper sideband (green) and the overall response (black dashed). (c) Calculated phase of each photocurrent contribution. (d) Difference in phase between the upper and lower sideband photocurrent contributions. (e) An example implementation of MWP notch filter. (f) An example implementation of MWP bandpass filter.

3.2.5 Summary

We have presented how MWP bandpass and notch filters can be achieved by using common modulation formats including SSB, PM, IM, as well as through reconfigurable aDSB modulation formats. In doing so we developed expressions that relate the RF gain to the optical transfer function of an MWP link in eq.'s 3.43, 3.45, 3.47, 3.48 and 3.51. To com-

pare the difference in RF link gain between these different scenarios, we plot the MWP link transfer functions for different modulation formats with respect to the applied optical transfer function in Fig. 3.14. In the following analysis, we note that RF rejection ratio of a filter can be determined by comparing the $|s_{21}|$ for the case where no optical processing is applied ($|H_{\text{opt}}| = 0$ dB) which corresponds to the vertical dashed line, to the case where a specific optical magnitude change is applied. In these plots, we assume that the optical transfer function does not impart any phase change ($\phi_{\text{opt}} = 0$).

For SSB modulation, the relative $|s_{21}|$ is 0 dB in the absence of any optical processing, and the optical transfer function is directly mapped to the RF response, as shown in Fig. 3.14 (a).

Using PM allows for the synthesis of MWP bandpass filters with large rejection ratios using relatively small magnitudes of applied optical transfer functions, as shown in Fig. 3.14 (b). This is enabled through the destructive interference of photocurrents at the PD output. This technique overcomes the challenge of achieving large RF rejection that is faced by SSB modulation. However, the RF link gain for the PM-based MWP filter scheme is less than the link gain for the SSB case because the components of the net RF photocurrent are π out-of-phase and therefore reduce the magnitude of the PD output.

Conversely, an IM-based MWP link consists of two in-phase photocurrent contributions that sum constructively, enabling a larger passband RF link gain than PM-based MWP links. As a result, the RF gain of an IM-based MWP link is 6 dB greater than an MWP link using SSB modulation and considerably larger than the corresponding PM link gain, as shown Fig. 3.14 (c). However, we showed in Section 3.2.3 that the RF rejection of an IM-based MWP filter is limited by the presence of the second in-phase sideband contribution. Furthermore, we outlined that the advantages of large RF gain, offered by IM-based schemes, as well as the large RF rejection, offered by destructive interference-based detection schemes such as PM, can be combined using an IM-based MWP link and

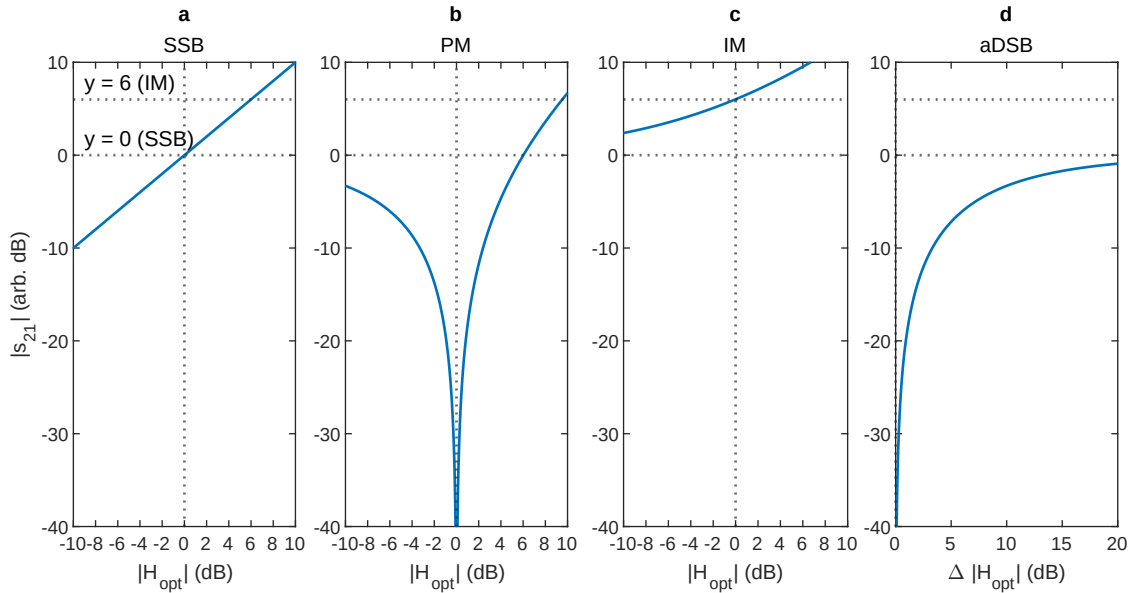


Figure 3.14: Summary of RF transfer function as a function of optical transfer function for (a) Single sideband modulation (SSB). (b) Phase modulation (PM). (c) Intensity modulation (IM). (d) Asymmetrical dual-sideband modulation (aDSB).

synthesising an optical response that enables spectrally localised destructive interference of RF photocurrents.

Finally, we highlighted how a DDMZM, or a DPMZM can be used to synthesise aDSB modulation to flexibly alternate between MWP bandpass and notch responses. Although the flexibility in synthesising reconfigurable RF responses is attractive, it comes at the expense of broadband partial cancellation of the RF photocurrents, owing to the broadband π out-of-phase photocurrents. We present the normalised MWP link gain as a function of the maximum depth of the applied optical resonance, $\Delta|H_{\text{opt}}|$, in Fig. 3.14 (d). For small values of $\Delta|H_{\text{opt}}|$, the modulation scheme essentially acts as PM and hence the RF link gain decreases rapidly. For large values of $\Delta|H_{\text{opt}}|$, the modulation scheme essentially acts as SSB modulation, and hence tends towards a relative gain of 0 dB.

3.3 MWP Link Metrics

MWP processing devices must also meet demanding performance specifications to be considered for deployment in real-world applications [25]. In this section, we outline key performance metrics of MWP links and outline techniques that may be applied to improve their performance.

3.3.1 Gain

The first metric discussed is the RF link gain, which is defined as the ratio of RF output power (P_{out}) to RF input power (P_{in}), given by

$$g = \frac{P_{\text{out}}}{P_{\text{in}}} \quad (3.52)$$

We use the circuit model presented in Fig. 3.15, consisting of an input voltage source with voltage V_S in Fig 3.15 (a), and an output current source with current I_{det} in Fig. 3.15 (b). In this analysis, we assume the use of impedance-matching resistors at the input and output of the MWP link, $R_S = R_L = R_{\text{match},PD}$ [72], [177], [287]. The source and load impedance is typically 50 Ω . We can therefore express the RF input and output power of the MWP link as

$$P_{\text{in}} = \frac{\langle V_S^2(t) \rangle}{4R_S} \quad (3.53)$$

$$P_{\text{out}} = \frac{\langle I_{\text{det}}^2(t) \rangle R_L}{4} \quad (3.54)$$

For MWP links using intensity modulation (IM) and direct photodetection, the link gain is given by [72]

$$g_{\text{MWP}} = \left(\frac{\pi r_{pd} R_L P_{\text{opt}} \sin(\phi_B)}{4L V_{\pi_{\text{RF}}}} \right)^2, \quad (3.55)$$

where r_{pd} is the responsivity of the PD, R_L is the RF load resistance (which is equal to the source resistance R_S), P_{opt} is the laser output power, L is the loss of the MWP link, ϕ_B ,

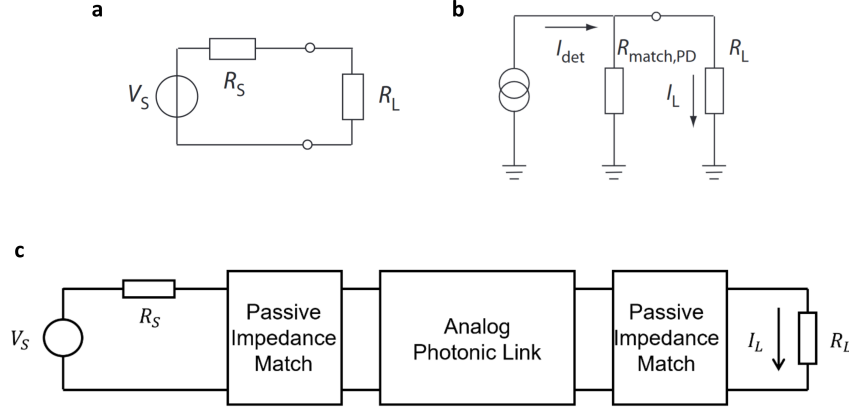


Figure 3.15: Circuit model used to analyse MWP link performance. (a) RF input. (b) RF output. (c) Overall link model. Figures (a-b) from [72], (c) from [177].

and $V_{\pi_{RF}}$ are the bias angle and half-wave voltage of the E-O modulator, respectively. We note that it is often convenient to express the MWP link gain in decibel form

$$G_{MWP} = 10 \log_{10} (g_{MWP}) \quad (3.56)$$

We can analyse eq. 3.55 to determine the factors that must be considered to increase the gain of an MWP link which include

- **High-efficiency O-E interfaces** ($V_{\pi_{RF}}, r_{PD}$)

Conversion between the RF and optical domains is the dominant source of loss in MWP links [72]. Therefore, it is highly desirable to use modulators with low $V_{\pi_{RF}}$ and photodetectors with high responsivity to enable low-loss MWP links. Typical values for $V_{\pi_{RF}}$ in most MWP demonstrations are approximately 3 V [293], and PDs typically have responsivities of approximately 0.5 A/W [294], [295]. However, novel modulators based on thin-film lithium niobate [257], [259], [264] have demonstrated low half-wave voltages as low as 1.6 V at 50 GHz [263]. Furthermore, PDs with high responsivity approaching 1 A/W [296], [297] have recently been demonstrated in the literature. These devices highlight future possibilities for improving the RF link performance through E-O interfaces with improved efficiency.

- **Large Optical Power at the photodetector** (P_{opt}, L)

We see from eq. 3.55 that the RF link gain depends on the square of the optical power at the photodetector (P_{opt}/L). Therefore, it is important to use high-power lasers, and optical amplifiers in MWP link. However, the use of optical amplifiers must be treated with care to avoid excessive added optical noise [298], [299]. Furthermore, to maximise optical power at the PD, it is essential to design the MWP link carefully to minimise optical losses [72].

3.3.2 Noise

Thermal Noise

Thermal noise arises from thermal fluctuations of charge within a resistor [1]. The thermal noise power in a bandwidth B at temperature T is given by

$$p_{N,th} = k_B T B \quad (3.57)$$

where k_B is the Boltzmann constant (J/K), T is assumed to be (290 K) for room-temperature operation and gives a noise power spectral density of $p_{N,th} \approx -174$ dBm/Hz.

Shot Noise

Shot noise arises from the random arrival time of photons, which correspondingly generates a random current photocurrent [72]. The shot noise photocurrent can be described as a Poisson distribution, with a flat PSD that is given by [300]

$$p_{N,shot} = 2qr_{PD}P_{opt,avg}BR_L \quad (3.58)$$

where q is the electronic charge and $P_{opt,avg}$ is the average optical power at the photodetector.

Relative Intensity Noise

Relative intensity noise (RIN) arises from random power fluctuations to the laser output power, typically from spontaneous emission within the laser cavity, and has an associated power in the electrical domain after photodetection given by [72], [300]

$$p_{N,RIN} = 10^{\frac{RIN}{10}} r_{pd}^2 P_{opt,avg}^2 BR_L \quad (3.59)$$

Total Noise

The circuit model of noise sources in an MWP link is shown in Fig. 3.16. In this analysis, we assume a load matching resistor ($R_{match,PD}$) is used at the PD, which leads to a noise power at the PD output port of

$$p_n = (1 + g_{MWP})p_{N,th} + \frac{1}{4}p_{N,shot} + \frac{1}{4}p_{N,RIN} \quad (3.60)$$

The first term in eq. 3.60 corresponds to thermal noise which consists of the thermal noise from the input of the MWP link (hence scaled by a factor g_{MW}) and the thermal noise from the matching resistor. The second term refers to the shot noise contribution, and the final term corresponds to the RIN noise contribution.

When considering the noise characteristics of any RF system, it is common to refer to the noise power spectral density (PSD), which represents the noise power in a 1 Hz bandwidth ($B = 1$). We plot the noise PSD of an MWP link as a function of the received optical power in Fig. 3.17 using the values $r_{pd} = 0.5$ A/W and $RIN = -160$ dBc/Hz. It is clear that for large optical powers (large MWP link gain), the RIN noise is the dominant contribution to p_n .

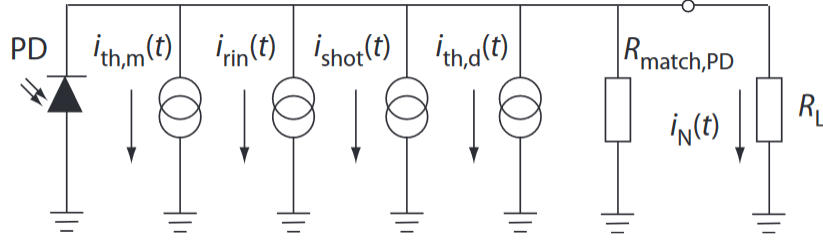


Figure 3.16: Noise source circuit model of a generic MWP link [25].

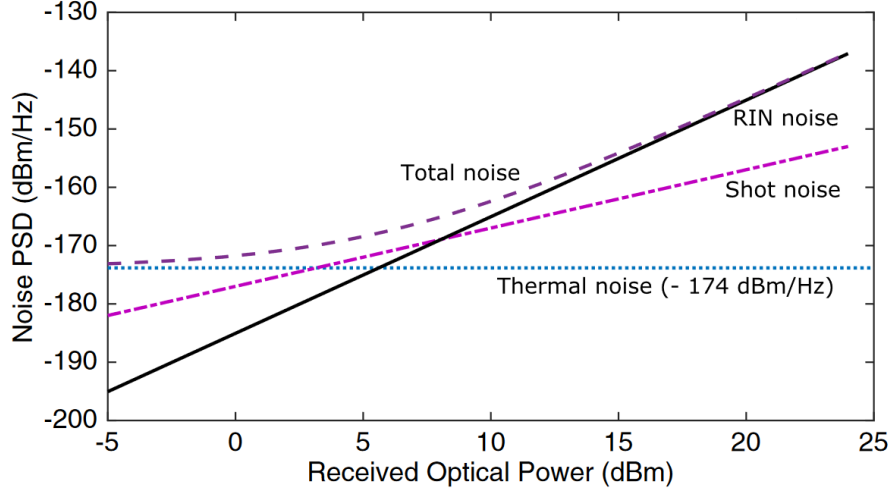


Figure 3.17: Noise contributions in an MWP link. The thermal, shot, and RIN noise contributions are plotted separately as a function of optical power at the PD. The total RF output noise is the sum of the thermal, shot and RIN noise contributions. Figure from [27].

Noise Figure

The key figure of merit to assess the noise degradation of an RF component is the noise figure (NF), which quantifies the degradation of the signal-to-noise ratio (SNR) in an RF system and is defined as [1], [72].

$$NF = 10 \log_{10} \left(\frac{SNR_{in}}{SNR_{out}} \right) \quad (3.61)$$

where SNR is defined as the ratio of signal power (P_{sig}) to noise power (p_N) and is given by

$$SNR = \frac{P_{sig}}{p_N} \quad (3.62)$$

When referring to NF, it is assumed that thermal noise is the only input noise contribution. Substituting eq. 3.52 and 3.57 and 3.62 in eq. 3.61, we can express the noise figure in terms of known quantities as

$$NF = 10 \log_{10} \left(\frac{p_N}{g_{MWP} k_B T B} \right) \quad (3.63)$$

It is often convenient to express these quantities in decibel scale, and refer to measurements in terms of noise PSD, which leads to the expression

$$NF = PSD - G_{MWP} + 174 \quad (3.64)$$

where 174 represents the thermal noise power spectral density at 290 Kelvin.

3.3.3 Nonlinearity

The third important metric of an MWP link is its linearity. This must be considered to reduce the effect of unwanted harmonic and intermodulation distortions which are introduced by nonlinearities in the MWP link. While many components in an MWP link introduce nonlinearities, the main source of nonlinearities is the E-O modulator, owing to its nonlinear transfer function [72] which we derived in eq. 3.49.

In this subsection, we will use a Taylor series to express the output power of the MWP link as a function of input voltage [1]

$$P_{out}(v) = \sum_{m=0}^{\infty} \frac{(v - v_b)^m}{m!} \cdot \left. \frac{d^m P_{out}}{dv^m} \right|_{v=v_b} \quad (3.65)$$

The linearity of an MWP link is most often quantified by the spurious-free dynamic range (SFDR), and is measured using the two-tone test [72]. Using the Taylor approximation of eq. 3.49, can express the output power of the MWP link with two input tones, given by $V_{in} = V_b + V \sin(\omega_{RF,1}t) + V \sin(\omega_{RF,2}t)$, as [1]

$$\begin{aligned} P_{out}(v) = & (a_0 + a_2 V^2) \\ & + \left(a_1 V + \frac{9a_3 V^3}{4} \right) \sin(\omega_{RF,1}t) + \left(a_1 V + \frac{9a_3 V^3}{4} \right) \sin(\omega_{RF,2}t) \\ & - \left(\frac{a_2 V^2}{2} \right) \cos(2\omega_{RF,1}t) - \left(\frac{a_2 V^2}{2} \right) \cos(2\omega_{RF,2}t) \\ & + a_2 V^2 \cos((\omega_{RF,1} - \omega_{RF,2})t) - a_2 V^2 \cos((\omega_{RF,1} + \omega_{RF,2})t) \\ & - \frac{a_3 V^3}{4} \sin(3\omega_{RF,1}t) - \frac{a_3 V^3}{4} \sin(3\omega_{RF,2}t) \\ & + \frac{3a_3 V^3}{4} \sin((2\omega_{RF,1} - \omega_{RF,2})t) + \frac{3a_3 V^3}{4} \sin((2\omega_{RF,2} - \omega_{RF,1})t) \\ & - \frac{3a_3 V^3}{4} \sin((2\omega_{RF,1} + \omega_{RF,2})t) - \frac{3a_3 V^3}{4} \sin((2\omega_{RF,2} + \omega_{RF,1})t) \\ & + \dots \end{aligned} \quad (3.66)$$

We see that the output consists of harmonically related terms, in the form $n\omega_{RF,1}$ and $m\omega_{RF,2}$, as well as intermodulation distortion (IMD) terms, in the form $n\omega_{RF,1} \pm m\omega_{RF,2}$, as shown in Fig. 3.18. The IMD terms with frequencies $2f_{RF,1} - f_{RF,2}$ and $2f_{RF,2} - f_{RF,1}$ are of particular concern, as they are the closest frequency components the fundamental input tones, and are therefore the most difficult to remove by filtering [72].

The spurious free dynamic range (SFDR_n) is used to quantify the linearity of a microwave photonic link. SFDR_n measures the difference between the fundamental signal

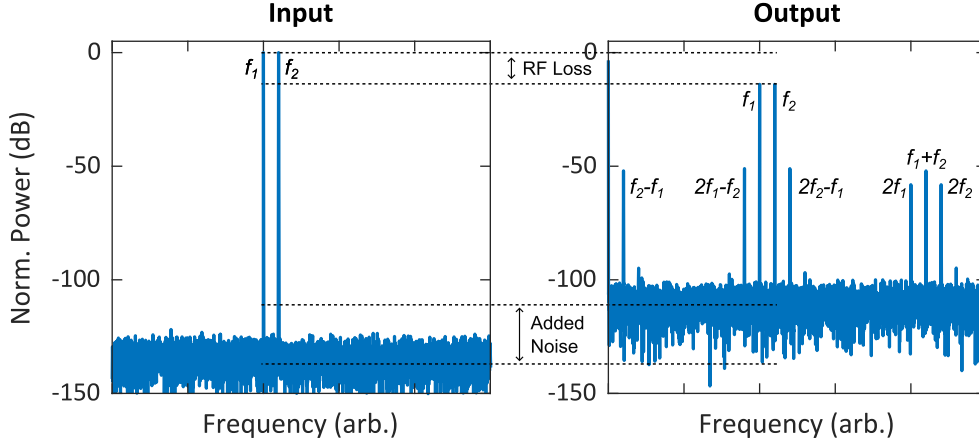


Figure 3.18: Typical output spectrum of MWP link for the two-tone test. The input spectrum consists of two tones. The output spectrum consists of the same two tones, with the intermodulation products labelled. The RF link loss and added noise in the MWP link are labelled.

power and the noise PSD when the n^{th} order IMD terms are equal to the noise PSD, as labelled in Fig. 3.19. The SFDR_n corresponds to the maximum signal-to-noise ratio (SNR) that is achievable without the need to filter the n^{th} order intermodulation terms [27].

Because the fundamental and IMD components scale at different rates with respect to RF input power, as shown in eq. 3.66, their lines of fit intersect at a point that is referred to as the n^{th} order intercept point. The intercept point can be expressed with respect to input power, and output power. The 3rd order input intercept point (IIP_3) and output intercept point (OIP_3) are illustrated in Fig. 3.19, and are related by

$$\text{OIP}_n = \text{IIP}_n + G \quad (3.67)$$

the SFDR_n is related to the OIP_n , noise figure (NF) and gain (G) are related by the expression

$$\text{SFDR}_n = \frac{n-1}{n} (\text{OIP}_n - \text{NF} - G + 174) \quad (3.68)$$

3.3.4 Link Optimisation Techniques

Now that we have examined the key metrics of MWP links, we explore methods that can be employed to improve RF link performance. For example, consider the case of enhancing link performance by increasing optical power at the PD. By examining eq. 3.55, one can observe that the optical link gain scales with the square of the optical power incident on the PD. However, from eq. 3.60 the overall MWP link noise scales with the square of the average photocurrent, which is proportional to the optical power at the PD, and no net performance benefit is achieved. Therefore, different approaches must be made to improve the noise figure of the MWP link. For the remainder of this section, we discuss common strategies that are used to improve the noise figure of the MWP link.

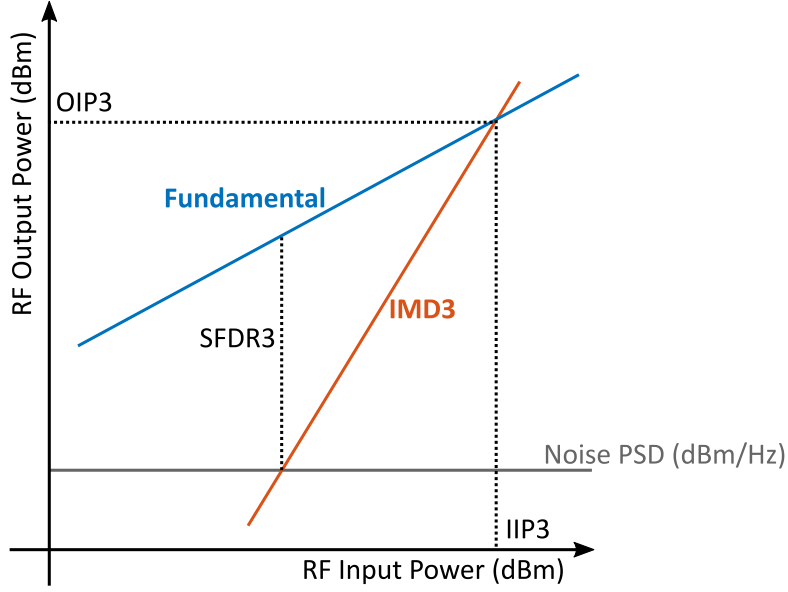


Figure 3.19: Illustration of spurious free dynamic range. The fundamental tone output power, IMD_3 power, and noise power spectral density are labelled.

Low Biasing

The NF of MWP links can be reduced using the technique known as low biasing [1], [72]. In this analysis, we define a new bias angle $\phi'_B = \pi - \phi_B$. Substituting these values in eq. 3.49 gives

$$P_{opt,avg} = \frac{P_{opt}}{2} (1 - \cos(\phi'_B)) \quad (3.69)$$

This substitution gives no change to the magnitude of the MWP link gain.

As ϕ'_B is moved from quadrature towards null bias ($\phi'_B = 0$), we see from eq. 3.55 that the RF link gain decreases proportionally to $\sin^2(\phi'_B)$. The RIN noise is the dominant noise source in the MWP link for large optical powers. Examining eq. 3.59 and eq. 3.69, we see the RIN noise is proportional to the square of the average optical power, $(1 - \cos(\phi'_B))^2$. We plot the RF link gain dependency (blue solid line) and the RIN noise dependency (orange dashed line) on the DC bias voltage in Fig. 3.20. We can see that the RF noise power reduces faster than the RF link gain as ϕ'_B reduces, thus reducing the bias voltage from quadrature leads to an improvement in NF.

Optical Carrier Suppression

Adjusting the optical carrier-sideband power ratio (CSR) similarly has the effect of improving RF link performance. Let us consider the expression for the RF photocurrent output of an MWP link given by eq. 3.40, and defining the optical carrier-sideband power ratio as

$$CSR = \left(\frac{E_c}{E_l} \right)^2 \quad (3.70)$$

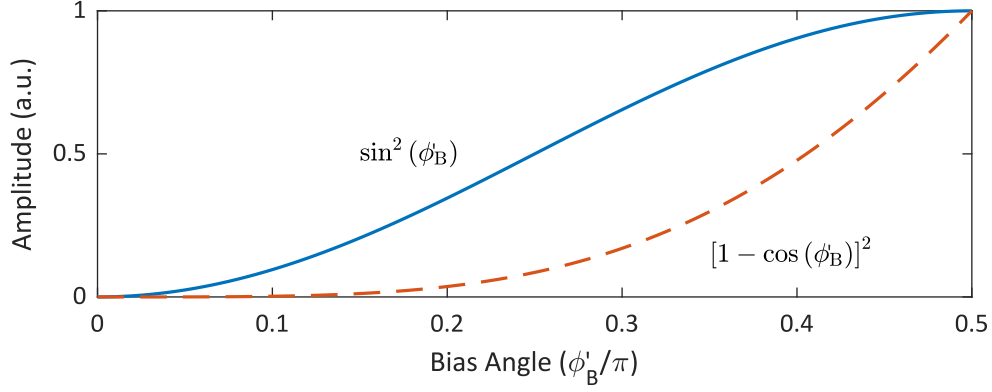


Figure 3.20: Effect of changing the modulator bias (ϕ'_B) voltage on RF link gain, which is proportional to $\sin^2(\phi'_B)$, and RF noise PSD, which is proportional to $(1 - \cos(\phi'_B))^2$. The noise contribution is reduced faster than the RF link gain as the bias is moved towards the null point ($\phi'_B = 0$).

We can express the output RF photocurrent in terms of the CSR for modulation schemes where the sidebands have comparable amplitudes, ($E_l \approx E_u$), as

$$i(\omega_{RF}) \propto \frac{E_c}{\sqrt{CSR}} (e^{j\phi_u - \phi_c} + e^{j\phi_c - \phi_l}) \quad (3.71)$$

Expressing eq. 3.71 in terms of RF link gain ($g_{MWP} \propto I^2(\omega_{RF})$), the RF link gain, optical power (P_0), and CSR can be expressed as

$$g_{MWP} \propto \frac{P_{opt}^2}{CSR} \quad (3.72)$$

Following from eq. (3.72), the RF link gain can be increased by reducing the CSR and maintaining constant optical power on the PD. Furthermore, because the RF link gain improves with CSR reduction and the optical power at the PD, and therefore RF noise PSD, remains constant, the NF is also reduced. The improvement to RF link gain and reduction in noise figure with carrier suppression is shown in Fig. 3.21 (a-b), respectively.

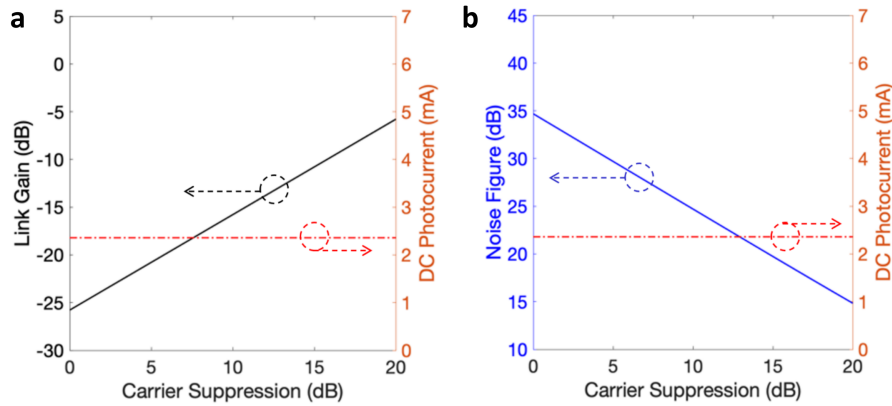


Figure 3.21: RF link gain improvements as a function of optical carrier suppression. (a) Increase in RF link gain. (b) Reduction in RF noise figure. Figure modified from [301]

Chapter 4

On-chip Brillouin-based Microwave Photonic Signal Processing

This chapter provides an overview of on-chip stimulated Brillouin scattering (SBS), a particularly powerful tool used for MWP signal processing owing to the reconfigurability and fine-spectral resolution of the Brillouin response. We start the chapter with a historical perspective on SBS before we provide a theoretical overview of the underlying principles which facilitate Brillouin interactions. We then review recent advances in chip-scale Brillouin microwave photonic signal processing systems, including reconfigurable filters, frequency converters, tunable phase shifters, true time delays and microwave sources, which are key functionalities across many RF systems. Finally, we discuss a roadmap for developments of Brillouin-based microwave photonic processors to reach the required performance and compact footprint for implementation in future SWaP-constrained RF systems that require dynamically reconfigurable responses. Parts of this chapter are based on the work in [81]

- **Matthew Garrett**, Moritz Merklein and Benjamin J. Eggleton, "Chip-Based Brillouin Processing for Microwave Photonic Phased Array Antennas," *IEEE Journal of Selected Topics in Quantum Electronics*, 2023, **29**, 1, 1-20.

4.1 Introduction to stimulated Brillouin scattering

4.1.1 Historical Perspectives of On-chip SBS

Stimulated Brillouin scattering is a third-order optical nonlinearity mediated by interactions between optical photons and acoustic phonons within a transparent material [282]. The first reported prediction of light scattering was made by Leon Brillouin in 1922 [302], however many agree that Mandelstam suggested that light could be scattered off thermally generated phonons in 1918, and published his results later in 1926 [303]. As depicted in Fig. 4.1, it was not until after the invention of the laser [90], [91] that stimulated Brillouin scattering was first observed in quartz and sapphire by Chiao, Townes and Stoicheff [304], as sufficient optical intensities to initiate Brillouin scattering could not be generated without a laser source.

The advent of low-loss optical fibres led to the generation of SBS within fibres, and SBS was initially seen as a nuisance for telecommunication applications [305]. However, the interaction between optical and acoustic fields facilitated by Brillouin scattering has found advantages in a range of applications. For example, the long-lived phonon lifetime gives rise to the narrow response of the Brillouin gain resonance (typically 10's MHz), which has led to the demonstration of many MWP processing functionalities [25], [27], [81], [144], [306]–[309], as fine spectral resolution is required in MWP systems but difficult to achieve using many alternative optical processing methods. We review integrated Brillouin-based MWP processors in Section 4.3. Additionally, long storage lifetimes of electronic information can be achieved by encoding optical information into the acoustic wave generated from SBS, which has formed the basis of Brillouin-based memory devices [310]. Furthermore, by examining the linewidth, Stokes magnitude and Brillouin frequency shift (BFS) of the Brillouin response of a material, its structural properties can be inferred. As a result, SBS has found widespread use in sensing [311], spectroscopy [312] and imaging [313] applications.

The first demonstration of SBS in microstructured photonic crystal fibres [314] unlocked the potential to increase the optical intensity through device miniaturisation. Following this demonstration, there has been a renewed interest in generating SBS in compact and chip-scale platforms [51], as shown in Fig. 4.1. Demonstrations of SBS in compact CaF_2 resonators [315], soft chalcogenide glass waveguides [316], and silicon-on-insulator platforms [317]–[320] were soon to follow. More recently, Brillouin interactions have been recorded in other mature photonic platforms based on silicon nitride [39], [227], [321], AlGaAs [322] and silicon oxynitride [323]. Although these platforms generate significantly less Brillouin gain than chalcogenide glass waveguides or silicon-on-insulator platforms, they each offer great potential in enabling a range of versatile nonlinear optical processing platforms.

4.1.2 Theoretical Outline of Brillouin scattering

In this section, we provide a phenomenological overview of the underlying principles which lead to the generation of stimulated Brillouin scattering. For a comprehensive review and derivation of SBS fundamentals and theory, we refer the reader to Refs. [282], [283], [305], [324]–[327]. For recent reviews of Brillouin-based platforms and broad applications of SBS we refer the reader to Refs. [51], [142], [328].

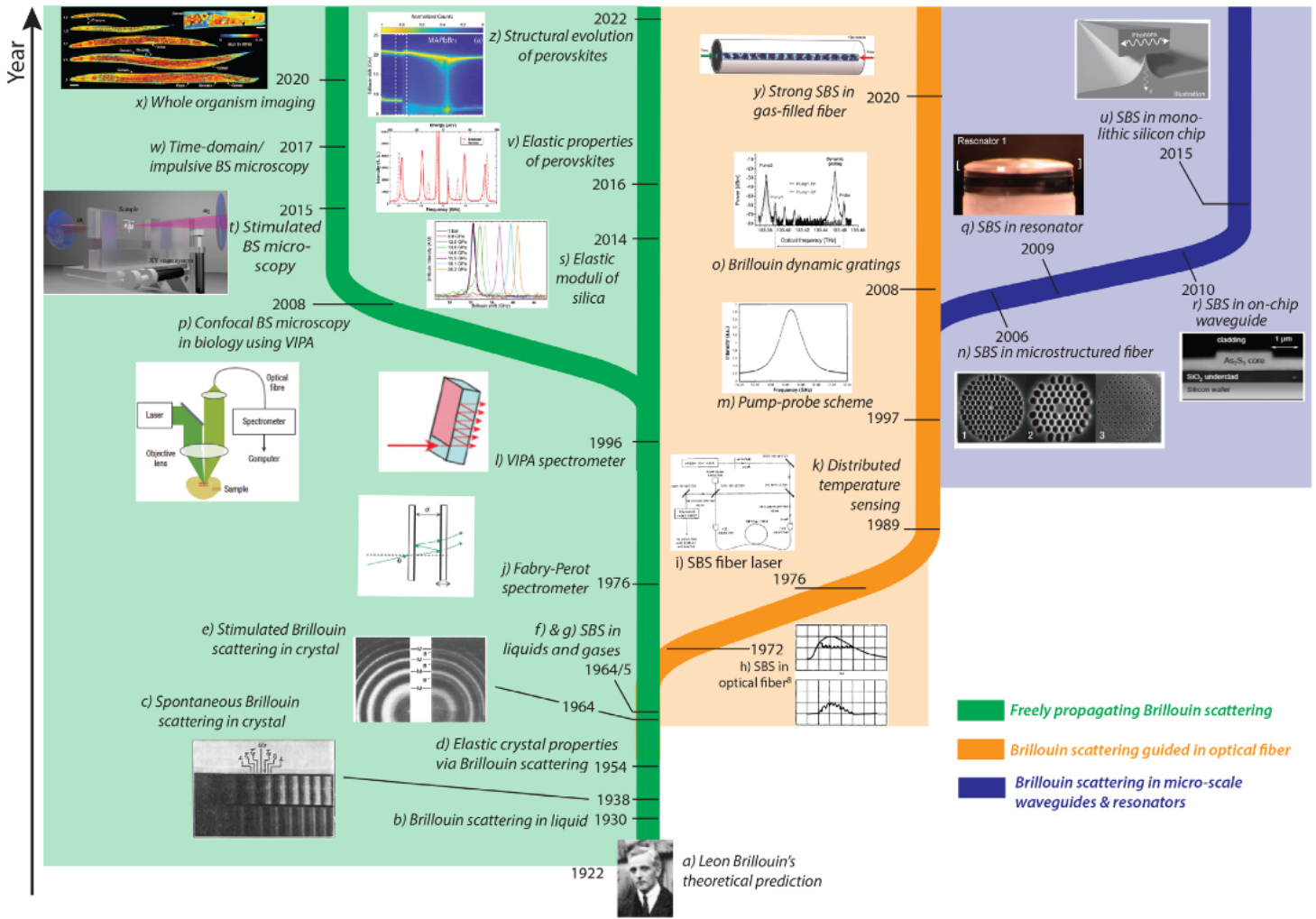


Figure 4.1: Historical evolution of platforms and applications of stimulated Brillouin scattering from its prediction to the current day. Figure taken from [142].

Let us first consider an optical waveguide, shown in Fig. 4.2, which supports confined optical modes with a given angular frequency ω and wave number k . The waveguide can also support a set of acoustic modes with frequency Ω and acoustic wave number q . The acoustic modes can be in the form of a longitudinal pressure wave, where the motion is along the direction of propagation and have a relatively large acoustic wave number, as shown in Fig. 4.2 (a). Alternatively, the acoustic mode can be a transverse shear wave, where the displacement occurs perpendicular to the direction of the waveguide and therefore has small acoustic wave numbers, as the motion is mostly transverse to the direction of optical propagation, as shown in Fig. 4.2 (b). To reduce losses of transverse acoustic waves, it is common to suspend the waveguide in air which increase the acoustic confinement of the transverse acoustic wave and minimises leakage to the substrate [329], [330].

Brillouin scattering occurs through the coherent interaction between an acoustic mode supported in a waveguide and optical modes which propagate through the waveguide. More specifically, what differentiates the Brillouin interaction from other types of acousto-optic effects, is that in Brillouin scattering an acoustic wave is optically excited in a stim-

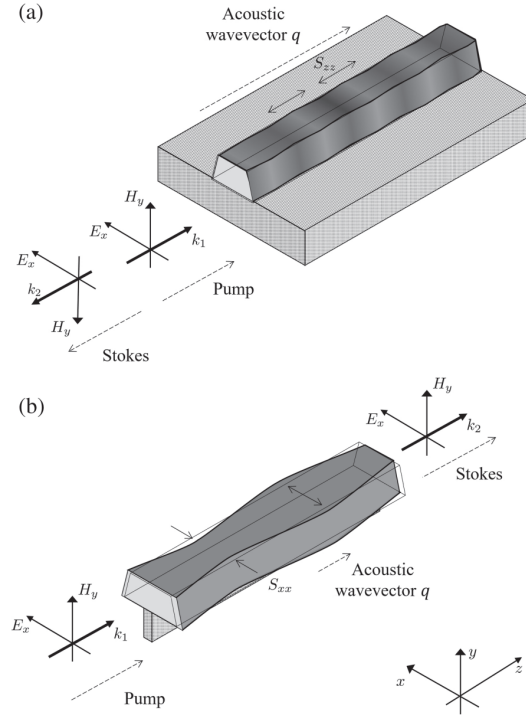


Figure 4.2: Overview of common waveguide geometries used to generate Brillouin interaction using integrated photonic platforms. (a) Shows the case for backward SBS, where the optical pump and probe are counter propagating and create a longitudinal acoustic wave. (b) Shows the case for forward SBS, where the optical pump and probe co-propagate create a transverse acoustic wave. Figure taken from [282].

ulated process through electrostriction and the photoelastic effect [282]. Brillouin scattering can be seeded in two distinct regimes [283]. The first regime is generated only by a single strong optical signal, referred to as a pump. The acoustic wave is seeded from thermally generated phonons within the interaction medium. Boyd referred to this as the generator configuration [283]. The second regime is known as an SBS amplifier. In this configuration, the pump and a weaker laser signal, referred to as the probe, as both injected into the interaction medium. Power is coupled from the pump to the probe when the probe is offset from the pump by the BFS in the medium. In this thesis, we are primarily concerned with the amplifier configuration, which is used in Chapters 5, 6 and 7, and will be assumed for the majority of this thesis.

Consider the case shown in Fig. 4.3 of two counter-propagating optical waves within a waveguide, a pump wave with angular frequency ω_p and a stokes wave with frequency ω_s , Fig. 4.3 (a). A beat pattern with wavelength $2\pi/\Delta k$ is formed that moves with velocity $v = \Delta\omega/\Delta k$ when the pump and probe differ in wave number Δk and differ in angular frequency $\Delta\omega = \omega_p - \omega_s$ [325], Fig. 4.3 (b). Through electrostriction, a strain is induced in the material which manifests as an acoustic wave [282], Fig. 4.3 (c). The moving acoustic wave locally modifies the density of the waveguide, creating a travelling optical grating due to the photoelastic effect. The pump is subsequently reflected and Doppler is shifted off this travelling acoustic wave such that the reflected pump and probe waves have the same frequency. The amplified Stokes wave further enhances the acoustic wave,

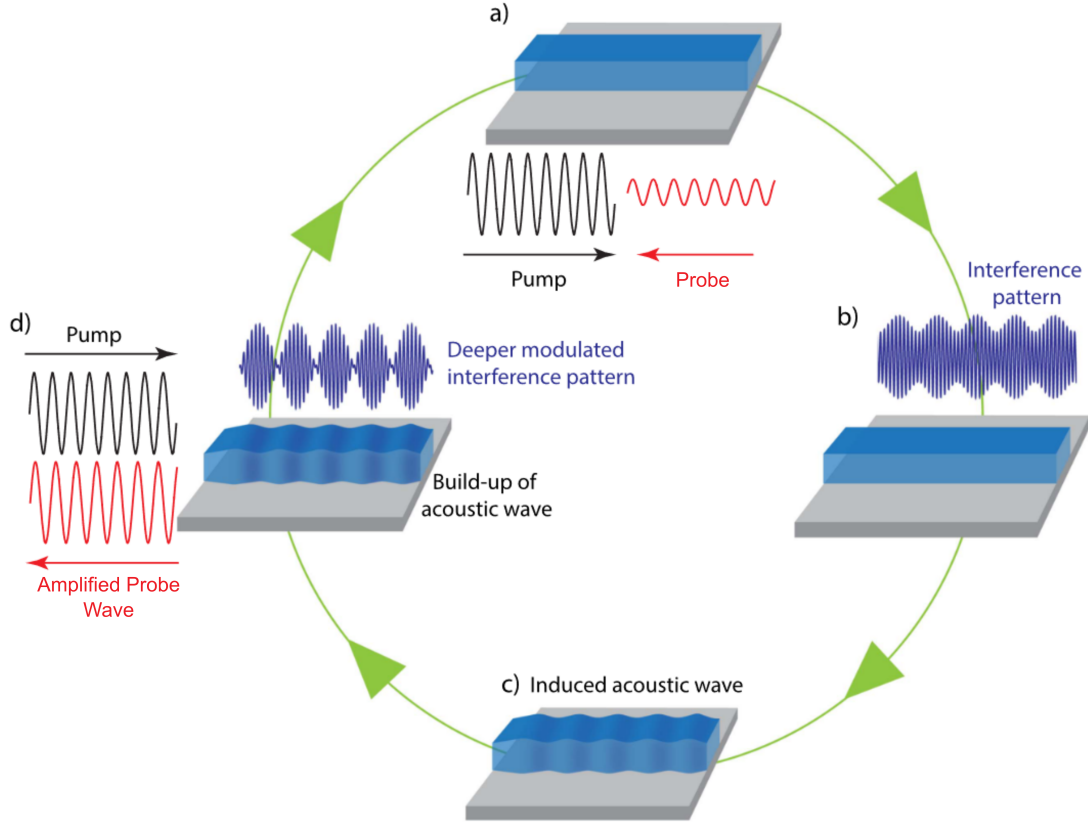


Figure 4.3: (a) Counter-propagating pump and probe (seed) signals. (b) As the frequencies of the pump and probe are detuned, an interference pattern with a beat-note is formed. (c) An acoustic wave is formed due to the optical interference pattern. (d) The pump is reflected by the acoustic wave, which further reinforces the optical beat pattern, stimulating the process. Figure modified from [310].

which in turn further scatters the stokes wave leading to a resonantly enhanced stimulated process [283], Fig. 4.3 (d).

The scattering process described above requires the conservation of energy and momentum, which for a pump, stokes and acoustic waves with wave numbers and angular frequencies (k_p, ω_p) , (k_s, ω_s) , (q, Ω) , are related by the expressions [282]

$$q = k_p - k_s \quad (4.1)$$

$$\Omega = \omega_p - \omega_s \quad (4.2)$$

Further, the interactions between optical and acoustic modes can be expressed as a sum of vectors using the dispersion diagrams presented in Fig. 4.4. In the case of backward Brillouin scattering, where the pump and probe are counter propagating, we see that the acoustic wave is longitudinal in nature, shown in Fig. 4.4 (a), and travels in the direction of the pump, indicated by the vectors in Fig. 4.4 (b). Given the difference in wave numbers between optical and acoustic waves is several orders of magnitude, we can approximate the Brillouin frequency shift (BFS), and wave number by [282]

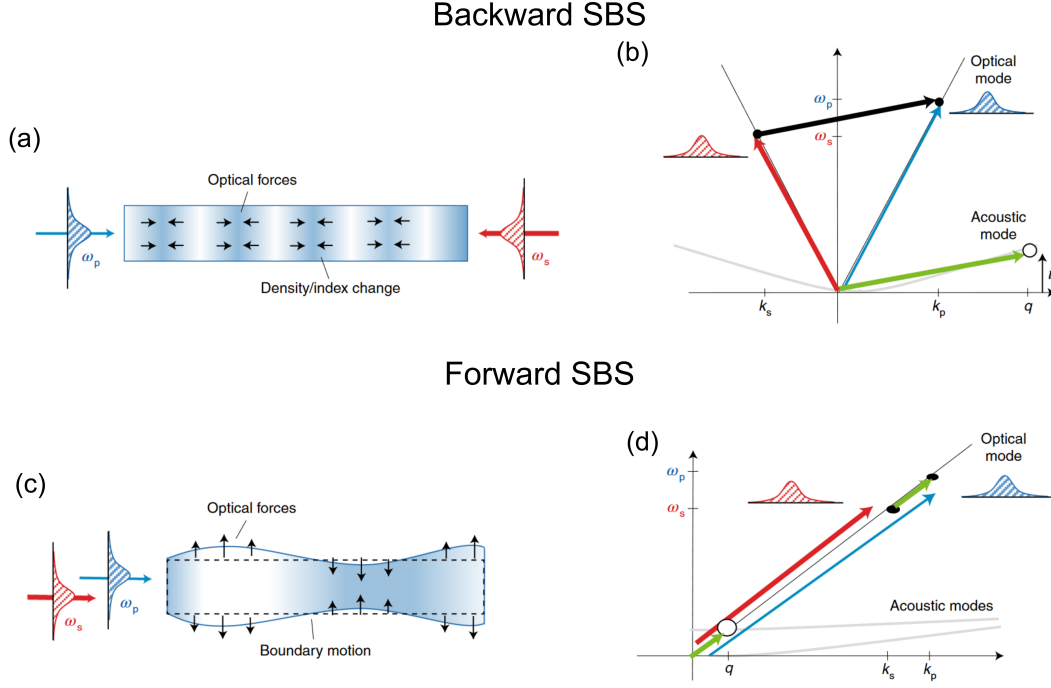


Figure 4.4: Dispersion diagrams for backward and forward SBS. Backward SBS showing schematically the (a) properties of the longitudinal acoustic wave and (b) the dispersion diagram for optical and acoustic modes. Forward SBS showing schematically the (c) properties of the transverse acoustic wave and (d) the dispersion diagram for the optical and acoustic modes. Figure taken from [51].

$$\Omega \approx 2k_p c_p \quad (4.3)$$

$$q \approx 2k_p \quad (4.4)$$

where c_p is the phase velocity of the acoustic wave. Typical values of Ω are approximately 11 GHz for standard single-mode optical fibres [305], 8 GHz for soft chalcogenide glass waveguides [164] and 13 GHz for multi-layer silicon nitride waveguides [39].

SBS can also occur when the pump and probe waves are co-propagating [283], [325]. This results in a transverse acoustic wave, as shown in Fig. 4.4 (c), with small wave vectors, as shown in Fig. 4.4 (d).

At the Brillouin resonant frequency, the probe signal experiences the peak magnitude of SBS gain, g , which can be expressed as a function of the Brillouin gain parameter, G , as [325]

$$g = e^G \quad (4.5)$$

$$G = \frac{g_B}{A_{eff}} P_p L_{eff} \quad (4.6)$$

P_p is the pump power in the Brillouin medium (W), and L_{eff} the effective Brillouin interaction length is related to the optical propagation loss (/m) and length of the Brillouin

medium (m) by

$$L_{eff} = \frac{1}{\alpha} (1 - e^{-\alpha L}) \quad (4.7)$$

g_B is the intrinsic Brillouin gain factor which can be expressed as [164], [305]

$$g_B = \eta \frac{4\pi n^8 p_{12}^2}{\lambda_p^3 c \rho \Omega_B \Delta \Omega_B} \quad (4.8)$$

where η is the acousto-optic overlap, n is the material refractive index, p_{12} is the elasto-optic coupling coefficient, λ_p is the pump wavelength, ρ is the material density, $\Delta \Omega_B$ is the Brillouin gain linewidth.

Analysing eq. 4.8, we can see that a material with a large refractive index is desirable, as it leads to a larger intrinsic Brillouin gain factor. Additionally, the elasto-optic coefficient impacts the intrinsic Brillouin gain factor of a material and can vary significantly between materials, for example, chalcogenide glasses exhibit values of approximately 0.24, whereas silicon approximately 0.017 [325]. Furthermore, from eq. 4.7, waveguides with low-losses are desirable to increase the effective length of the Brillouin interaction.

While many of the terms in eq. 4.8 are fixed material properties, the acousto-optic overlap depends strongly on the waveguide design and must be carefully considered when designing Brillouin-based processing platforms. The acousto-optic overlap is bounded between 0 and 1 and depends on the optical intensity, $\|E\|^2$, and the acoustic mode field, $\hat{\rho}$, and be approximated using the relation [326]

$$\eta = \frac{\int \|E\|^2 \hat{\rho} dA}{\int \|E\|^2 dA \int |\hat{\rho}|^2 dA} \quad (4.9)$$

A typical cross-section of a As_2S_3 waveguide used to facilitate SBS is shown in Fig. 4.5 (a). The acoustic mode and optical mode used in the Brillouin interaction are shown in Fig. 4.5 (b-c), respectively. Pant *et al.* simulated a large acousto-optic overlap of $\eta = 0.95$ using a similar geometry. We refer the reader to ref. [326] for a detailed analysis regarding how the acousto-optic overlap impacts the generation of Brillouin gain in both guided and leaky acoustic mode regimes.

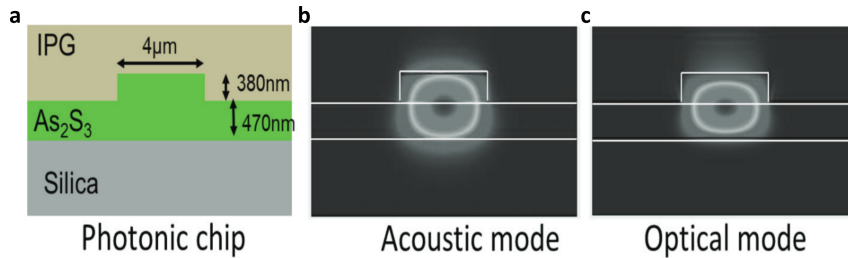


Figure 4.5: (a) Refractive index of As_2S_3 waveguide. (b) Optical field of As_2S_3 waveguide. (c) Acoustic field of As_2S_3 waveguide. Figure modified from [331].

One can derive from the coupled mode equations presented in refs. [283], [332] the following expressions for the Brillouin magnitude, $M(\omega)$, and phase, $P(\omega)$, response for a pump with angular frequency ω_p which are sufficient for many MWP signal processing applications [333]

$$M(\omega) = \exp \left(G \frac{\Gamma_B^2}{4(\omega_p - \Omega_B - \omega)^2 + \Gamma_B^2} \right), \quad (4.10)$$

$$\Phi(\omega) = -G\Gamma_B \frac{(\omega_p - \Omega_B - \omega)}{4(\omega_p - \Omega_B - \omega)^2 + \Gamma_B^2}. \quad (4.11)$$

where Γ_B is the linewidth of the Brillouin resonance and is inversely proportional to the phonon lifetime, τ_{ph} [283], [325]

$$\Gamma_B = \frac{1}{\pi\tau_{ph}} \quad (4.12)$$

4.2 Integrated platforms for Brillouin scattering

Integrated Brillouin Platforms

Brillouin scattering has been induced in numerous chip-scale platforms [308] using the material platforms outlined in Figs. 4.6 - 4.10 and summarised in Table 4.1, which details the key metrics used to characterise Brillouin platforms. The first approach is based on soft chalcogenide glass waveguides, which provides good confinement for both optical and acoustic modes because the speed of sound in the cladding is faster than in the core [51]. The second class of devices use silicon photonic waveguides, in which the acoustic mode is not strongly confined for traditional SOI rib and ridge waveguides [51]. The acoustic losses are reduced by suspending silicon waveguides above the substrate, so that it is essentially surrounded by air, thereby increasing mechanical isolation [317]–[319]. The third approach is the use of high Q factor resonant cavities, where long optical lifetimes enhance the generation of Brillouin gain, allowing for the use of materials with low Brillouin gain coefficients if they have low loss [227], [337]. Most recently, Si_3N_4 platforms have been shown to generate modest amounts of SBS gain, limited by the acoustic properties of the Si_3N_4 waveguide core [39], [321].

Table 4.1: Summary of Integrated Brillouin Processing Platforms.

Platform	Type	Brillouin Gain (dB)	G_{SBS} ($m^{-1}W^{-1}$)	L (cm)	α (dB/cm)	NL Loss (Y/N)	Ref
Chalcogenide As_2S_3 rib waveguides	B	52	500	23	0.5	N	[334]
Chalcogenide As_2S_3 strip waveguides	B	22.5	750	5.6	0.7	N	[176]
Chalcogenide $Ge_{25}Sb_{10}S_{65}$ strip waveguides	B	17.6	338	7	0.2	N	[335]
SOI underetched waveguides	F	0.6	3218	0.27	2.6	Y	[319]
SOI membrane suspended waveguides	F	6.9	1152	2.9	0.18	Y	[320]
SOI resonantly enhanced membrane suspended waveguide	F_{res}	30	380	1.5	0.13	Y	[336]
Si_3N_4 MRR	B_{res}	-	0.1	7.4	0.004	N	[227]

B - backward Brillouin scattering, F - forward Brillouin scattering, F_{res} - forward Brillouin scattering in a resonant device, B_{res} - backward Brillouin scattering in a resonant device. NL Loss - Nonlinear loss.

4.2.1 Chalcogenide Glass

The generation of backward SBS in As_2S_3 rib waveguides [164], which was the first demonstration of on-chip SBS, shown in Fig. 4.6, significantly reduced the interaction length required to generate SBS gain compared to previous fibre-based systems, and inspired the development of additional on-chip Brillouin microwave processing platforms based on chalcogenide glasses[335], [338]. Brillouin devices based on chalcogenide glasses combine strong acoustic confinement with negligible nonlinear loss, and have achieved more than 50 dB of Brillouin gain [334] using waveguides with lengths of a few centimeters that can be routed into compact and low-loss spirals. For example Morrison *et al.*, achieved 22.5 dB of gain in an on-chip area less than 1 mm^2 [176], as shown in Table 4.1.

Hybrid Brillouin platforms use multiple materials to enhance the functionality and performance of traditional As_2S_3 Brillouin devices. For example, integration of As_2S_3 waveguides with SOI platforms [176], shown in Fig. 4.7 (a), combines the benefits of the large magnitudes of SBS gain offered by chalcogenide waveguides with the availability of passive and active components available in CMOS silicon photonics. Fig. 4.7 (b) illustrates a typical SOI circuit that was fabricated by the IMEC foundry; this device aims to reduce the form factor of Brillouin processors by integrating As_2S_3 waveguides with active E-O components. Additionally, Si_3N_4 and As_2S_3 PICs [50], [340], as shown in Fig. 4.7 (c), have been cascaded to enable MWP notch filters with reduced pump power requirements, and TTDs using separate carrier tuning; we expand on these devices in Section 4.3.

Although integration of multiple materials enhances functionality, the transitions between different materials must be carefully designed to minimise reflections of the Brillouin pump [341], which increases the spurious content in the RF output [340]. For example, Lai *et al.* measured a 4% reflection from the interface between lensed tip fibre and monolithic As_2S_3 chip [341], which is lower than the reflection at the interface between air and bulk As_2S_3 ($\approx 17\%$) owing to the use of lensed-fibres and the anti-reflection coating at the facet of the As_2S_3 chip [10]. Further reducing reflections of the Brillouin pump is desirable to minimise spurious content in the output of the MWP link [340]. To reduce the reflection experienced by the Brillouin pump, a mode-matched and vertically tapered transition between As_2S_3 and germanosilicate waveguides [341] facilitated an adiabatic transition between materials, as shown in Fig. 4.7 (d). This device achieved minimal

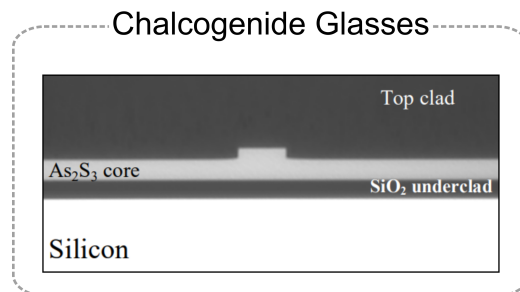


Figure 4.6: Hardware platforms that have been used for on-chip Brillouin processing of microwave signals. (a) As_2S_3 chalcogenide waveguides [339] © 2007, Optica Publishing Group.

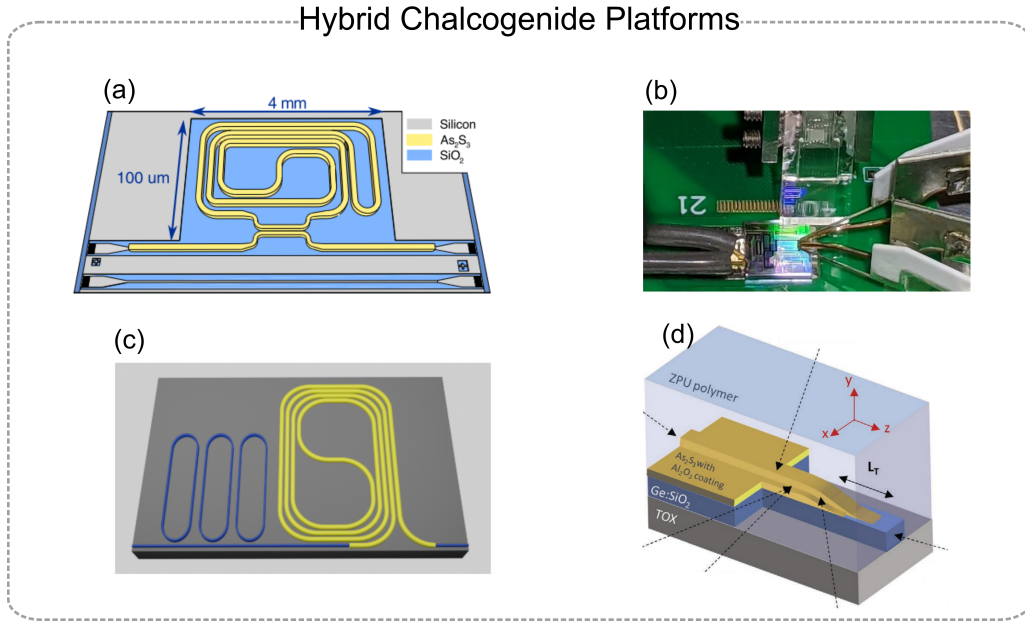


Figure 4.7: Hardware platforms that have been used for on-chip Brillouin processing of microwave signals. (a) Compact hybrid silicon and As_2S_3 devices [176]. (b) Hybrid integrated SOI- As_2S_3 Brillouin processor currently under development. (c) Hybrid Si_3N_4 ring resonator and As_2S_3 Brillouin platform [340]. (d) Vertically tapered $\text{Ge}:\text{SiO}_2$ - As_2S_3 waveguides for reduced reflection interfaces [341]. (a) from [176], © 2017, Optica Publishing Group. (c) from [340]; licensed under a Creative Commons Attribution (CC BY) license. (d) from [341], © 2021 Wiley-VCH GmbH.

optical back-reflections of only $\approx 0.04\%$ and promises to improve the performance of integrated Brillouin MWP processors [340].

4.2.2 Silicon

Whereas SBS is typically observed in the backward direction in optical fibres and chalcogenide planar rib waveguides, it manifests in the forward direction for SOI devices. This is because the p_{11} photoelastic constant of silicon, which couples transversal electrostrictive forces in forward SBS, is much stronger than the p_{12} photoelastic constant, which is the relevant photoelastic constant in backwards SBS processes [51], [282]. As mentioned above, standard silicon waveguides lack intrinsic acoustic guidance, which causes acoustic modes to leak rapidly from the waveguide core, inhibiting the generation of Brillouin gain [329]. However, the phonon mode can be confined if the waveguide is suspended, thereby reducing acoustic losses and supporting the propagation of transverse acoustic waves, leading to the generation of forward Brillouin scattering [329]. As seen in Table 4.1, silicon-based Brillouin platforms have large Brillouin gain coefficients and shorter interaction lengths than chalcogenide devices. However, as mentioned, they suffer from nonlinear loss at high pump powers, limiting the magnitude of Brillouin gain that can be generated [51]. Despite the aforementioned limited intrinsic acoustic guidance and nonlinear losses, SBS has been demonstrated in several SOI platforms [317], [319], [320].

Acoustic confinement has been achieved in an SOI platform using an underetched waveguide [319], as shown in Fig. 4.8 (a). However, the magnitude of the generated

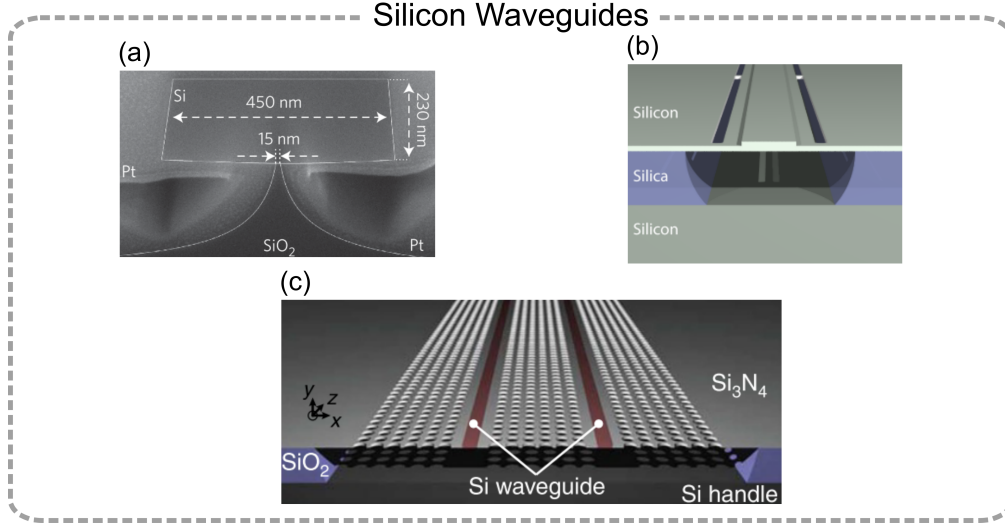


Figure 4.8: Hardware platforms that have been used for on-chip Brillouin processing of microwave signals. (a) Underetched silicon waveguide for forward Brillouin scattering [319]. (b) Suspended silicon waveguide [342]. (c) Photonic-Phononic Emit Receive (PPER) waveguides [318]. (a) from [319], Reprinted by permission from Nature Publishing Group. (b) from [342], Reprinted by permission from Nature Publishing Group. (c) from [318]; licensed under a Creative Commons Attribution (CC BY) license.

Brillouin gain saturated at approximately 1 dB, limited by optical losses and free-carrier absorption at higher pump powers [319]. Subsequently, 30 dB of on-chip Brillouin gain was generated in an SOI platform using a Brillouin active membrane waveguide structure [336], shown in Fig. 4.8 (b). However, the Brillouin-active waveguide was placed in an optical resonator to increase the magnitude of the generated Brillouin gain, limiting the functional bandwidth of the device.

One challenge faced by forward Brillouin devices is that the strong co-propagating pump must be filtered to avoid the generation of spurious terms caused by beating between the pump and probe at the PD [11]. The photonic-phononic emitter-receiver (PPER) structure demonstrated by Shin *et al.* [318] shown in Fig. 4.8 (c), overcomes this challenge by placing the Brillouin pump and probe in spatially separated waveguides. In this approach, the pump is injected into the emitter waveguide and is modulated by an RF signal with a frequency that coincides with the Brillouin frequency shift (BFS). This generates an acoustic wave through forward Brillouin scattering [318], which modulates the phase of the optical signal in the adjacent, receiver waveguide through photoelastic coupling [318]. The signatures of acoustic waves with narrow linewidths as low as 2.4 MHz [42] have been observed, facilitating MWP filters with fine frequency resolution, as we discuss later in this section.

4.2.3 Silica Resonators

High Q optical resonators enable increased interaction times between the pump and probe, allowing for materials with low Brillouin gain coefficients to be used, and facilitates narrow linewidth optical tones that are suitable for generating stable RF signals. For example, silica wedge resonators are attractive for SBS-based processing due to their high Q and

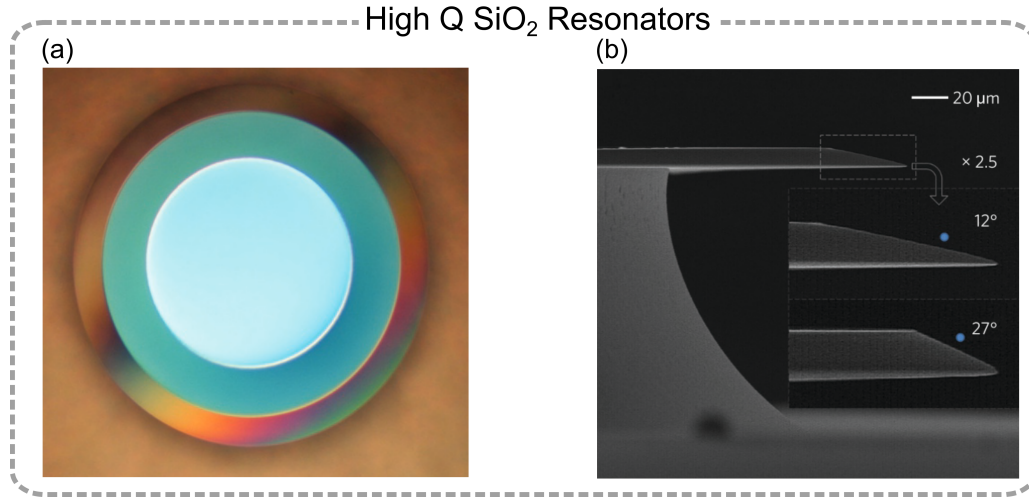


Figure 4.9: Hardware platforms that have been used for on-chip Brillouin generation of microwave signals. (a) Optical image and (b) scanning electron microscope image of silica micro-disk resonator from [337], Reprinted by permission from Nature Publishing Group.

compact form factors [337], as shown in Fig. 4.9 (a-b).

4.2.4 Silicon Nitride

Si_3N_4 platforms achieve Brillouin interactions through two distinct mechanisms. The first mechanism relies on leaky acoustic modes to facilitate Brillouin gain [227], [321]. Gundavarapu *et al.* demonstrated a Brillouin laser in a Si_3N_4 MRR [227], shown in Fig. 4.10 (a). The leaky nature of the acoustic mode in the Si_3N_4 resonator platform is evident by comparing the confined optical mode in Fig. 4.10 (b) with the acoustic mode in Fig. 4.10 (c). Due to the leaky nature of the acoustic wave, the Brillouin interaction occurs in the SiO_2 cladding. Because the acoustic wave decays faster than it leaks from the Si_3N_4 core, a sustained Brillouin interaction is achieved and long optical lifetimes, owing to low propagation losses enable narrow linewidth lasing [227].

Alternatively, Gyger *et al.* demonstrated SBS in Si_3N_4 waveguide [321] using the structure shown in Fig. 4.10 (c). By increasing the waveguide dimensions, the acoustic wave leaked away from the core at a slower rate [321], [326] compared to the work presented by Gundavarapu *et al.*. The increased acousto-optic interaction can be seen by comparing the optical and acoustic modes in Fig. 4.10 (e-f) and noting that a larger portion of the acoustic mode exists in the Si_3N_4 waveguide, where the optical mode is guided, as opposed to the SiO_2 cladding. Despite this, only modest Brillouin gain was observed.

Conversely, in the guided acoustic mode regime, the Brillouin interaction is mediated through a strong overlap between guided acoustic and optical modes [39], which offers a much greater Brillouin gain coefficient than its leaky mode counterparts. Fig. 4.10 (g) shows an optical image of the multi-layer Si_3N_4 platform [343] optimised for SBS. The guided nature of the acoustic wave between the Si_3N_4 layers, with a strong acousto-optic overlap can be seen by comparing the optical and acoustic modes in Fig. 4.10 (h-i). Despite the increased Brillouin gain coefficient offered by the guided acoustic wave on this platform, the Brillouin gain coefficient exhibited by all Si_3N_4 devices is orders

of magnitude lower than alternative platforms based on silicon or chalcogenide glasses [51]. Despite the weaker Brillouin interaction, the low-loss and high power handling of Si_3N_4 allows for limited demonstrations of MWP functions when combined with off-chip components [39].

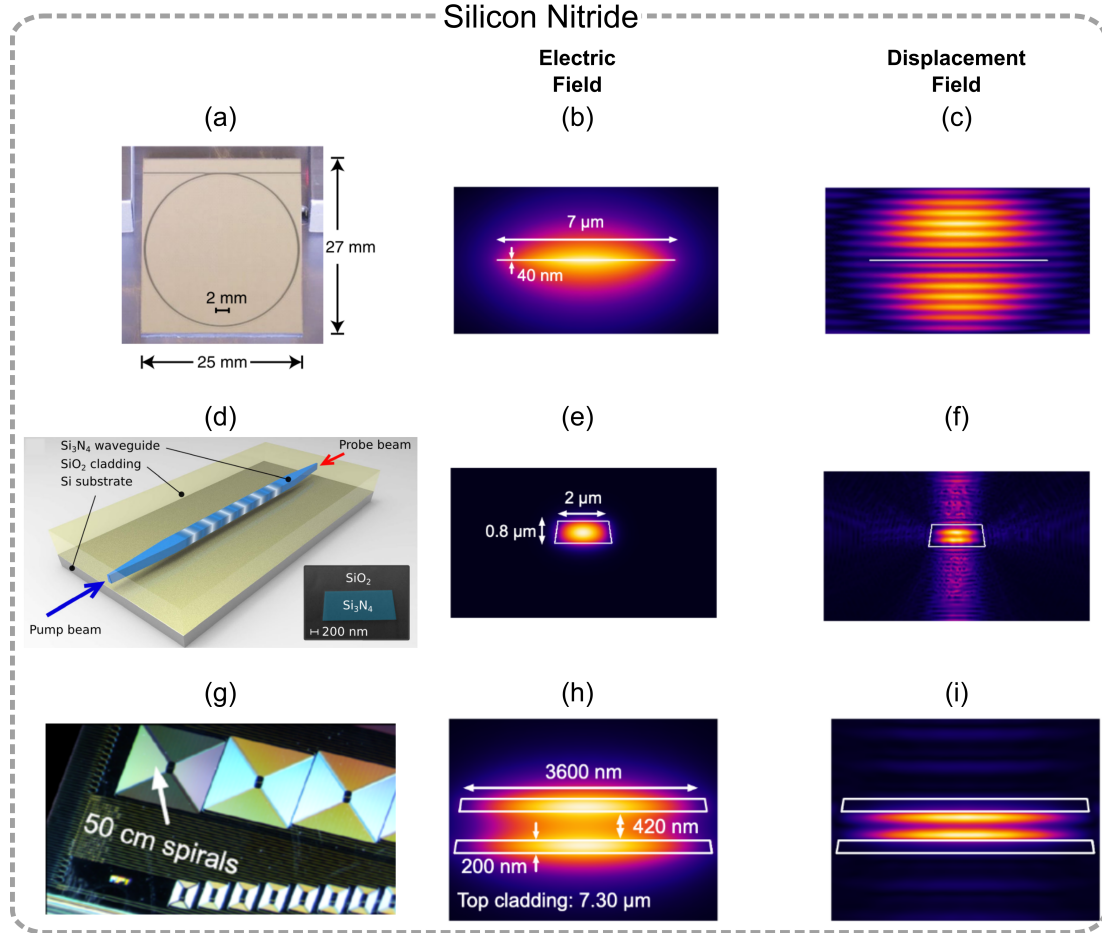


Figure 4.10: Silicon Nitride platforms that have been used to generate Brillouin gain (a) Si_3N_4 micro-resonators [227] Reprinted by permission from Nature Publishing Group. The corresponding electrical and displacement modal fields are plotted in (b-c), respectively. (d) LPCVD Si_3N_4 waveguide, taken from [321]. The corresponding electrical and displacement modal fields are plotted in (e-f), respectively. (g) multi-layer Si_3N_4 waveguide optimised for Brillouin scattering, taken from [39]. The corresponding electrical and displacement modal fields are plotted in (h-i), respectively. (b-c), (e-f), (h-i) taken from [39].

4.3 Component-level demonstration

4.3.1 Brillouin Microwave Photonic Filters

Microwave filters are essential components used to attenuate unwanted spurious or interfering signals in the RF spectrum [27]. MWP filters are required to exhibit wideband frequency tunability and fine frequency selectivity to attenuate unwanted signals whilst preserving nearby signals of interest. Therefore, Brillouin processing is particularly well suited for use in MWP filters as its fine spectral resolution enables narrow passbands or notches [25], [27].

As we outlined in chapter 3, the modulation scheme used in MWP filters has a significant impact on filter functionality and performance [27]. By altering the optical modulation scheme, both bandpass and notch filter functionalities can be realised [306]. Initial demonstrations of chip-based Brillouin MWP filters used single sideband modulation (SSB), and consequently were limited by a one-to-one spectral mapping of Brillouin gain to RF filter rejection ratio [306], [344], [345]. This is undesirable as it demands a considerable amount of pump power to generate large RF suppression [346]. For the remainder of this section, we outline approaches that combine advanced modulation formats [27], [306] with amplitude and phase control offered by SBS to enable high-performance MWP filters that do not require excessive Brillouin pump powers.

Bandpass filters enable transmission of a particular frequency range and are used to suppress out-of-band signals over a broad frequency range. In Brillouin-based processing

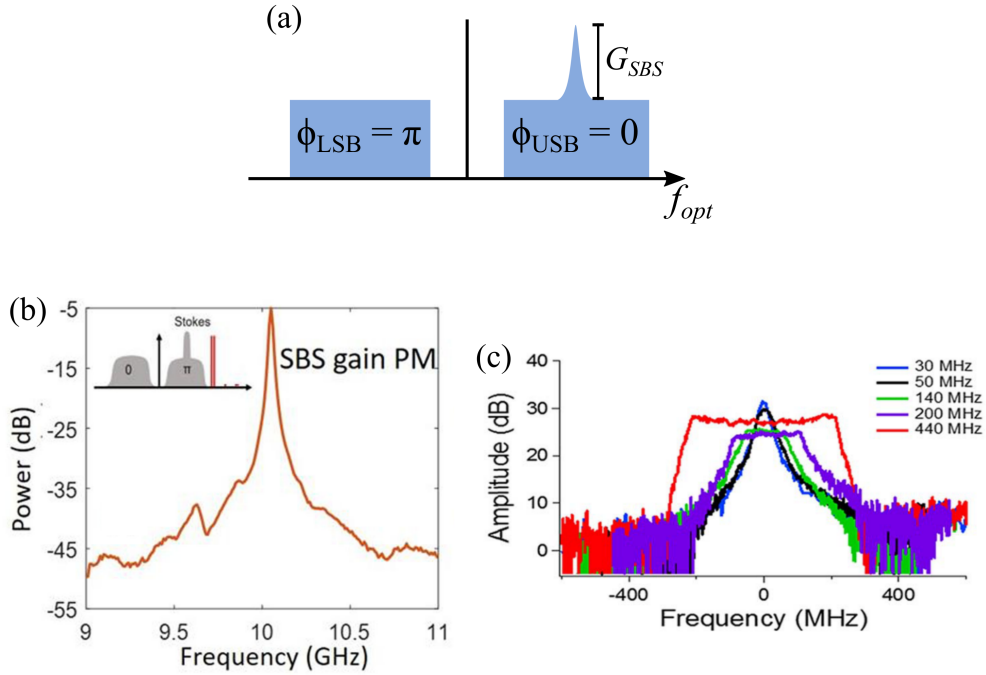


Figure 4.11: Brillouin bandpass filters. (a) An example of an optical spectrum used to achieve bandpass filtering. (b) Bandpass filter response achieved using phase modulation and Brillouin gain. (c) Passband reconfigurability by tailoring the on-chip Brillouin gain spectrum. Figure (b) from [292] and Figure (c) from [44]

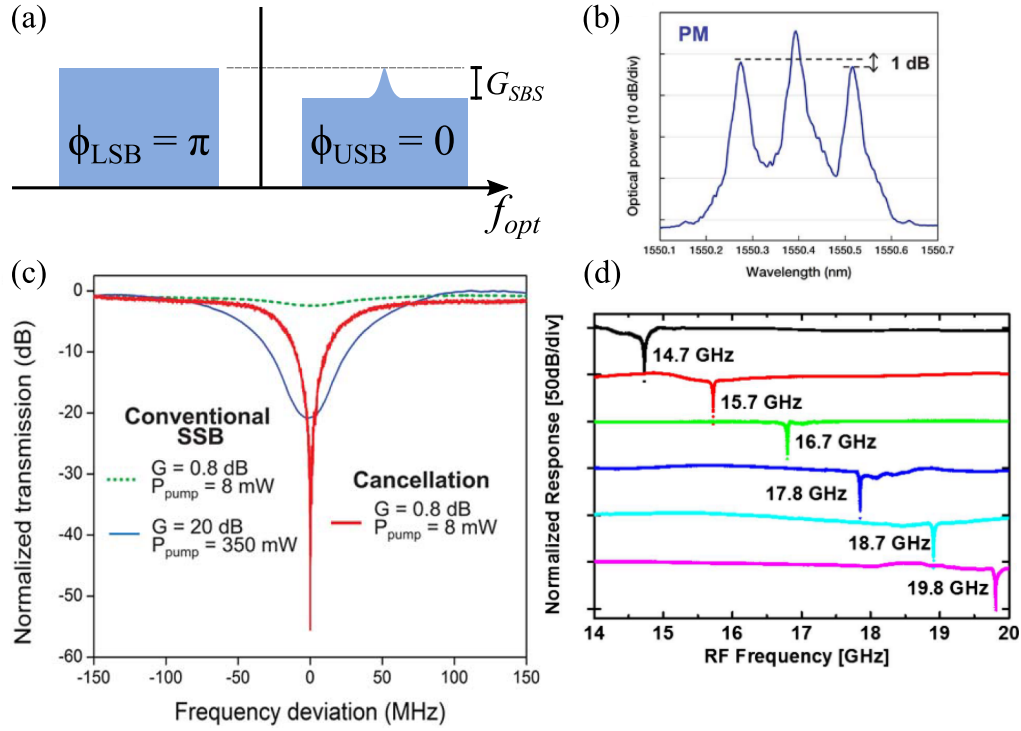


Figure 4.12: Brillouin notch filters. (a) An example of an optical spectrum used for RF notch filtering. (b) Optical spectrum showing the amplitude difference between the upper and lower sideband before Brillouin gain is applied. (c) Demonstration of using only 1 dB of Brillouin gain to generate an RF notch with rejection exceeding 50 dB. (d) RF notch filter response that uses only 1 dB of SBS gain to achieve an RF rejection exceeding 50 dB. The notch frequency is tuned from 14.7–19.8 GHz. (Figure (b-c) from [347], © 2015, Optica Publishing Group. Figure (d) from [348], © 2015, Optica Publishing Group.)

systems, bandpass filters are realised using a modulation scheme that results in broadband destructive interference of output RF photocurrents in the absence of any Brillouin resonance. However, when SBS is applied to one sideband, it breaks the condition for destructive interference and forms a narrow passband [292], [306]. Fig. 4.11 (a) depicts optical phase modulation scheme, that provides this functionality. Fig. 4.11 (b) shows the bandpass response when Brillouin gain is applied to one optical sideband, as shown in the inset. Fig. 4.11 (c) shows how the generation of Brillouin gain enables reconfigurable MWP filters, demonstrating passband bandwidth broadening from 30 MHz to 440 MHz [44] by broadening the Brillouin pump spectrum using an E-O modulator.

Notch filters are used to attenuate unwanted signals at a specific frequency. RF notches are formed by Brillouin-based devices, which use the SBS resonance to ensure RF photocurrents destructively interfere [347], [348]. One way to achieve this is shown in Fig. 4.12 (a), where the upper and lower sidebands of the probe are π out of phase and mismatched in amplitude by the magnitude of applied Brillouin gain [347], [348]. At the Brillouin resonant frequency, the upper and lower sideband are equal in amplitude, but π out of phase, and therefore their RF photocurrent contributions destructively interfere upon photodetection. Outside the Brillouin resonance, the upper and lower sidebands are not equal and therefore do not cancel. Fig. 4.12 (b) shows the modulation scheme used

by Marpaung *et al.* that formed the basis of a notch filter using 1 dB of Brillouin gain. The filter demonstrated notch depths exceeding 50 dB [347], shown in Fig. 4.12 (c). Similarly, Casas-Bedoya *et al.* required only 1 dB of forward Brillouin gain to demonstrate a widely tunable filter response using a SOI platform, as shown in Fig. 4.12 (d) [348]. One advantage of the modulation scheme in Fig. 4.12 (a) is that the pump power required to form a notch is greatly reduced when compared with non-interferometric approaches. However, the broadband π phase difference between upper and lower sidebands creates an RF gain penalty when the upper and lower photocurrent contributions are added.

Combining on-chip linear optical devices, for example MRRs, with Brillouin processing [45], [340] is an attractive method to implement an MWP notch filter without a gain penalty in the passband. In this notch filter scheme, an over-coupled MRR imparts a spectrally localised π phase shift to the upper sideband at its resonant frequency. Brillouin gain is then applied at the MRR central frequency to overcome the reduction in signal amplitude caused by the MRR resonance. The upper sideband therefore experiences a

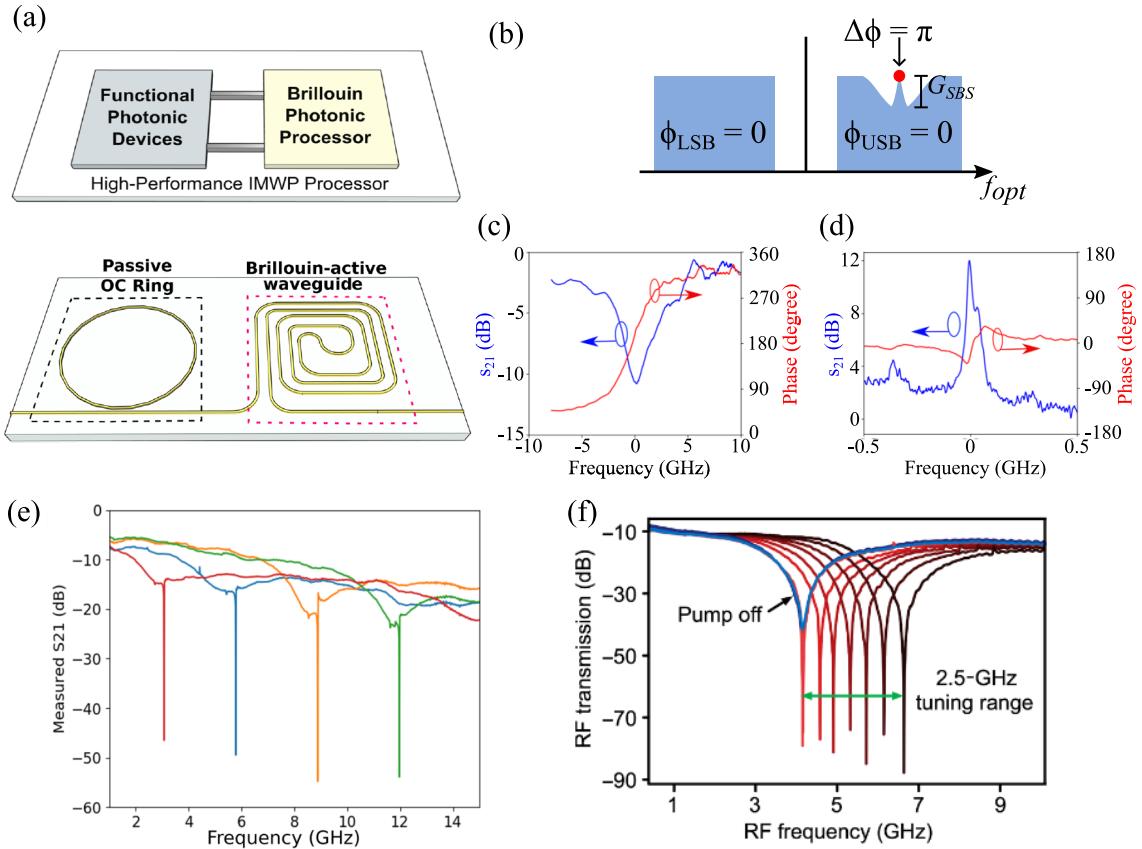


Figure 4.13: RF notch filter created by cascading Brillouin resonances with ring resonators. (a) Artistic impression of the photonic chip that facilitates linear MRR-based processing and nonlinear Brillouin processing. (b) Optical spectrum used to achieve a cascaded MRR-Brillouin MWP notch filter (c) Measured phase and amplitude response of the over-coupled micro-ring resonator. (d) Measured phase and amplitude response of the Brillouin response. (e) Measured MWP notch filter response. (f) MRR and Brillouin MWP notch filtering using Si_3N_4 MRRs and as the Brillouin medium. Figures (a-e) from [45] and Figure (f) from [39].

localised π phase shift with an unchanged amplitude, while the lower sideband is equal in amplitude and has unchanged phase, as shown in Fig. 4.13 (a). An RF notch is formed upon photodetection as the photocurrent contributions from upper and lower sidebands are equal in magnitude at the overlapped MRR and SBS resonant frequency, but π out of phase. The key to the passband link gain improvement offered by this scheme is attributed to the fact that the SBS and MRR responses do not impart a phase or amplitude change outside of their resonance. This results in constructive interference of RF photocurrents in the passband from the in-phase IM sidebands, and is not the case for the Brillouin-based notch filters discussed above, where the sidebands are out of phase with each other and do not sum constructively in the passband which increases RF losses [347], [348].

The first demonstration of this concept used optical fibre as the SBS medium [349] and was subsequently demonstrated in a monolithically integrated As_2S_3 platform [45], shown in Fig. 4.13 (a). The measured magnitude and phase response of the As_2S_3 MRR is shown in Fig. 4.13 (c), where a π phase shift is observed at the MRR resonance. The SBS gain response in As_2S_3 waveguides is shown in Fig. 4.13 (d), which does not impart any phase shift at its resonant frequency. The RF filter response using the monolithically integrated As_2S_3 platform exhibits tunable notch frequencies by changing the probe frequency relative to the MRR resonance, as shown in Fig. 4.13 (e). More recently, Botter *et al.* used this scheme in Si_3N_4 waveguides to form the basis of the MRRs and Brillouin medium [350]. The RF notch responses for this platform exhibited large 66 dB RF rejection using only 0.4 dB of SBS gain, as shown in Fig. 4.13 (f).

PPER waveguides, as outlined in Section 4.2, offer an alternate approach for implementing MWP bandpass [41], [351] and notch filters [42], which circumvent the need to filter the high-power Brillouin pump, as demonstrated by Gertler *et al.*. An MWP filter with a passband centred at the BFS, Ω_B , is realised by encoding the RF input signal onto the pump laser through intensity modulation [41], as shown in Fig. 4.14 (a). If the RF input contains a signal with frequency Ω_B , an acoustic wave is generated, which modulates the phase of the CW laser in the adjacent receiver waveguide. An optical bandpass filter removes one sideband of the phase modulated signal so that when this filtered optical signal is converted to the electrical domain, the RF signal is present in the device output. However, input RF signals that do not coincide with the BFS do not induce an acoustic wave, and are therefore suppressed by the MWP filter.

Fig. 4.14 (b) shows the bandpass RF filter response of the fabricated device with a BFS of 3.86 GHz and 3-dB bandwidth of 3.5 MHz [352]. The passband frequency is changed by using phase modulation to encode the RF information onto the pump, and introducing a second laser to the pump spectrum, which has a frequency offset from the original pump laser [41]. Tuning of the passband frequency is shown in Fig. 4.14 (c). MWP notch filters have been demonstrated using PPER waveguides [42], as shown in 4.14 (d). In this scheme, the PPER waveguides select a narrow region of the RF spectrum, which is subsequently subtracted from the original input signal, forming a notch filter through destructive interference. Using this technique, an RF notch filter, as shown in Fig. 4.14 (e), with rejection of up to 57 dB has been observed with bandwidth as narrow as 2.7 MHz [42]. The PPER structure has recently been extended to facilitate acousto-optic interactions between multiple waveguides, allowing the exploration of rich possibilities for on-chip signal processing [353] and waveform manipulation [354].

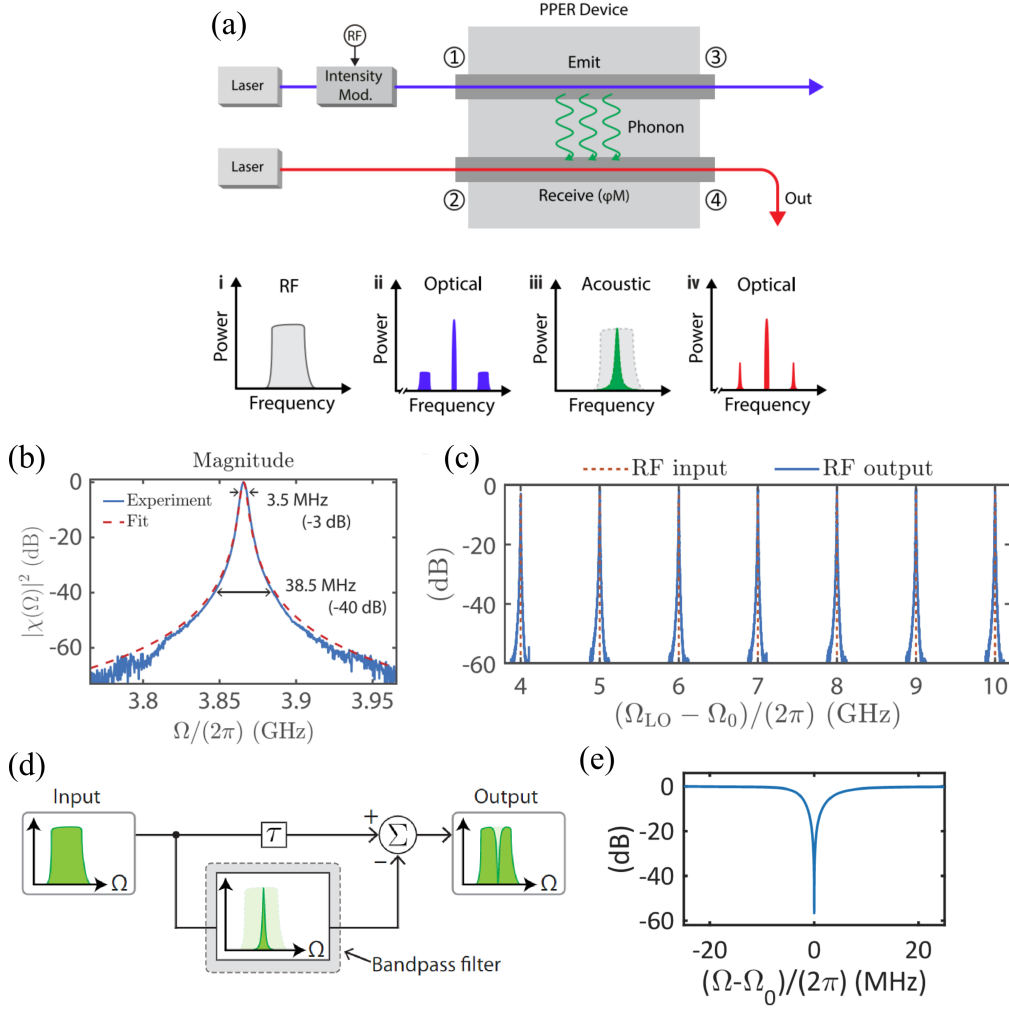


Figure 4.14: (a) Schematic outlining the generation of forward SBS in PPER waveguides. (b) Bandpass filter response using PPER waveguides. (c) Tunable filter responses using PPER waveguides to create a bandpass filter. (d) Architecture for interference-based notch filter using PPER waveguides. (e) RF notch filter response using PPER device and RF interference scheme. (Figure (a) from [43], © 2018, IEEE. Figure (b-c) from [41]; licensed under a Creative Commons Attribution (CC BY) license. Figure (d-e) from [42]; licensed under a Creative Commons Attribution (CC BY) license.)

4.3.2 Brillouin Phase Shifters

Phase shifters and TTDs are essential components responsible for beamforming in PAAs. The ideal RF phase shifter provides a 360° tunable phase shift with flat amplitude response over a wide RF frequency range. Brillouin-based MWP phase shifters typically use SSB modulation and apply a phase shift to the optical carrier from the phase shift associated with the Brillouin resonance, given by (4.11). This phase shift translates to a broadband phase shift in the RF output [307].

Pagani *et al.* demonstrated a MWP phase shifter using on-chip SBS [355] by applying equal amounts of Brillouin gain and loss to impart a phase shift to the optical carrier with minimal amplitude fluctuations, as shown in Fig. 4.15 (a). The generated RF phase shift was tuned by changing the Brillouin pump powers, as shown in Fig. 4.15 (b). A

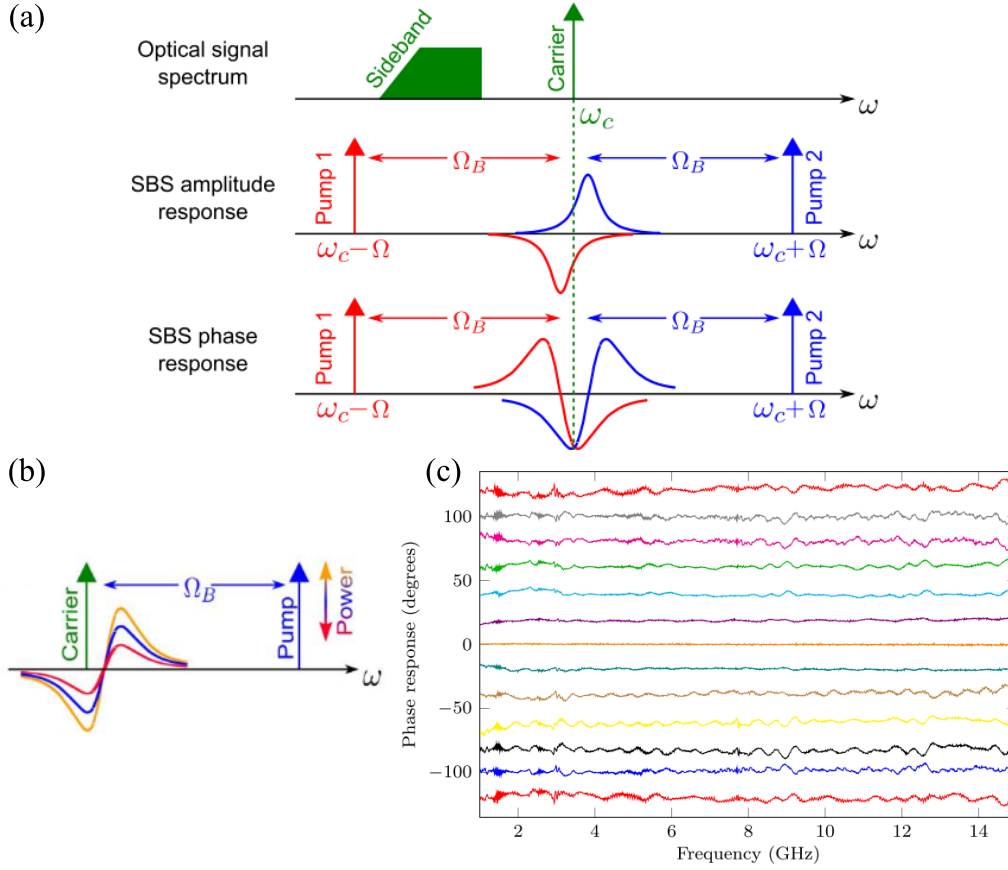


Figure 4.15: RF phase shifter using on-chip SBS (a) Operating principle. Brillouin gain and loss resonances are aligned such that the maximum phase excursion is applied to the optical carrier (b) The RF phase shift is tuned by changing the Brillouin pump power. (c) Measured broadband 240° phase shift. (Figures from [355], © 2014, Optica Publishing Group.)

continuous RF phase shift of 240° was observed, shown in Fig. 4.15 (c), while keeping RF power fluctuations within only ± 1.5 dB. Although good performance was demonstrated, achieving 360° phase tunability, which is required for beam steering in PAAs, demands more than 50 dB of Brillouin gain, leading to excessively large optical pump power [11].

RF interferometric techniques can be harnessed to provide a full 360° phase shift, without requiring high Brillouin pump powers; this approach uses RF vector addition to amplify the RF phase shift generated from an optical phase shift [356]. This scheme, summarised in Fig. 4.16 (a), consists of two RF photocurrent contributions. The first contribution, RF_1 , is assigned a phase of 0. The second, RF_2 , has a relative phase shift of π and is used for interferometric phase amplification. The output of this system, RF_{net} , is given by the vector summation of these two signals. When RF_1 and RF_2 are added, the overall phase shift, Θ_2 , is larger than the optical phase shift, Θ_1 , applied to RF_1 , at the expense of a reduced RF link gain [50]. The phase enhancement factor is defined as the ratio of the enhanced phase shift to the raw phase shift [11].

Fig. 4.16 (b) shows the un-enhanced $\pm 7.3^\circ$ phase shift over the RF frequency range 5-20 GHz generated from 2 dB of Brillouin gain using suspended silicon waveguides [11]. When the phase is amplified by the interferometric phase enhancement technique, a

360° phase shift is achieved using an enhancement factor of 25, as shown in Fig. 4.16 (c). Subsequently, this scheme was demonstrated using As_2S_3 waveguides and a 360° phase shift was generated up to 20 GHz using only 19 dBm of coupled on-chip pump power [357], compared to a total of 27 dBm on-chip pump power required to generate the 240° phase shift [355] in a similar As_2S_3 platform. The demonstrations of interferometric phase enhancement highlight the primary benefit of this interferometric technique, enabling 360° RF phase shifts while using minimal amounts of Brillouin gain.

4.3.3 Brillouin True Time Delays

While RF phase shifters may be used to implement analog beamforming for narrowband signals, beam squint occurs as the signal bandwidth increases [85]. TTDs are therefore preferred for wideband signals as they eliminate beam squint. MWP systems generate TTDs by applying a linear dispersive phase slope over a frequency range that includes the optical carrier and modulated sideband. However, this requires a delay to be applied from the optical carrier to the highest frequency component on the RF sideband, which is greater than the bandwidth of most photonic components [358]. Alternatively, the separate carrier tuning (SCT) [359] method is well suited to generate TTDs for MWP applications and is implemented by applying a linear phase slope, such as SBS slow-light effects [310], [316], [360], [361], to an optical sideband and separately applying the appropriate phase shift to the optical carrier, as shown in Fig. 4.17 (a).

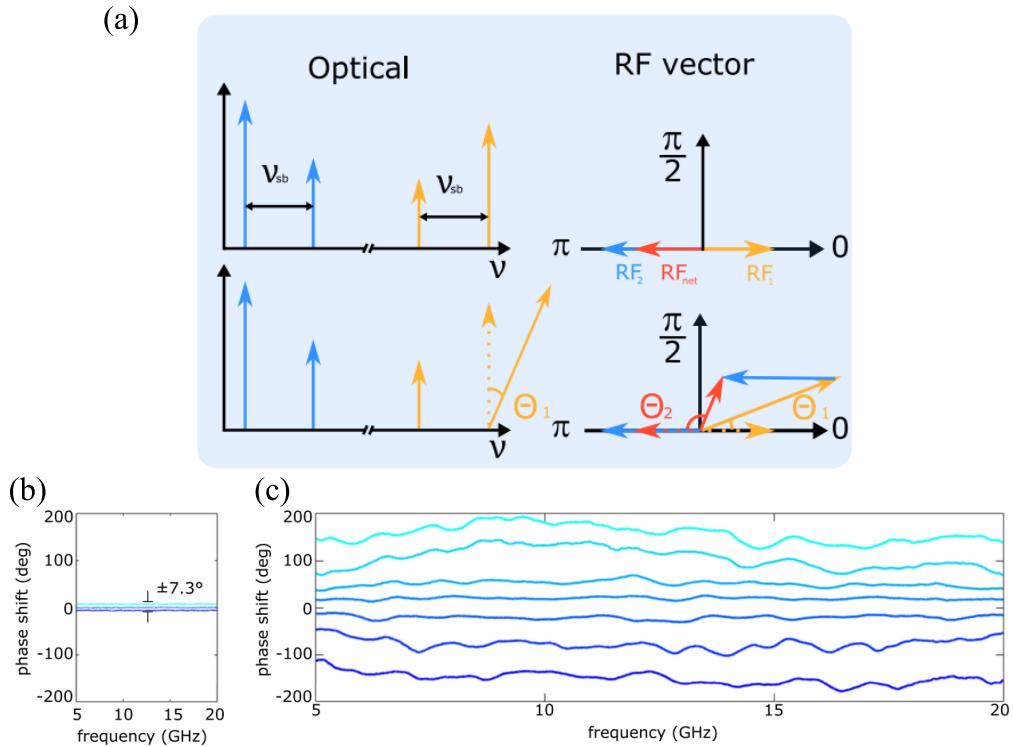


Figure 4.16: SBS-based phase shifter using RF interference scheme for phase amplification. (a) Operating principle. (b) Raw Brillouin phase response showing only $\pm 7.3^\circ$ phase shift. (c) Amplified phase response obtaining a full 360° phase shift from the raw response in (b). (Figures from [11], © 2019, Optica Publishing Group.)

The SCT method has been demonstrated using chip-based Brillouin devices by Aryanfar *et al.* [40], shown in Fig. 4.17 (b). The TTD was verified using a two-tap filter, which is an unbalanced interferometer with one arm containing the TTD and the other maintaining constant delay. The change of free spectral range (FSR) of the interferometer in Fig. 4.17 (c) corresponds to a delay of 4 ns over a 100 MHz bandwidth [40]. The phase shift applied to the optical carrier is indicated in Fig. 4.17 (d) by the change of the null frequency in the interferometer pattern.

The delay generated by an MWP system can be enhanced through interferometric techniques [50] using a similar principle as the phase enhancement technique for MWP phase shifters. Liu *et al.* [307] demonstrated dispersive phase slope enhancement which increased Brillouin slow light delay, but lacked carrier phase tunability to meet the SCT condition for generating a TTD. McKay *et al.* demonstrated TTD enhancement using the SCT method [50], shown in Fig. 4.18 (a). In this approach, two widely separated optical carriers, C_1 and C_2 , are used and both undergo single sideband modulation. However, the sidebands of each carrier are π out of phase relative to each other. Broadband Brillouin gain is applied to the sideband of C_1 , which causes a dispersive phase slope. A Si_3N_4 MRR applies the appropriate phase shift to C_1 to implement SCT, and a second MRR applies the same phase shift to C_2 to ensure that C_1 and C_2 maintain a π phase offset. The overall phase shift is increased through the summation of the output photocurrents RF_{SBS} and RF_{int} , which originate from the two optical carriers.

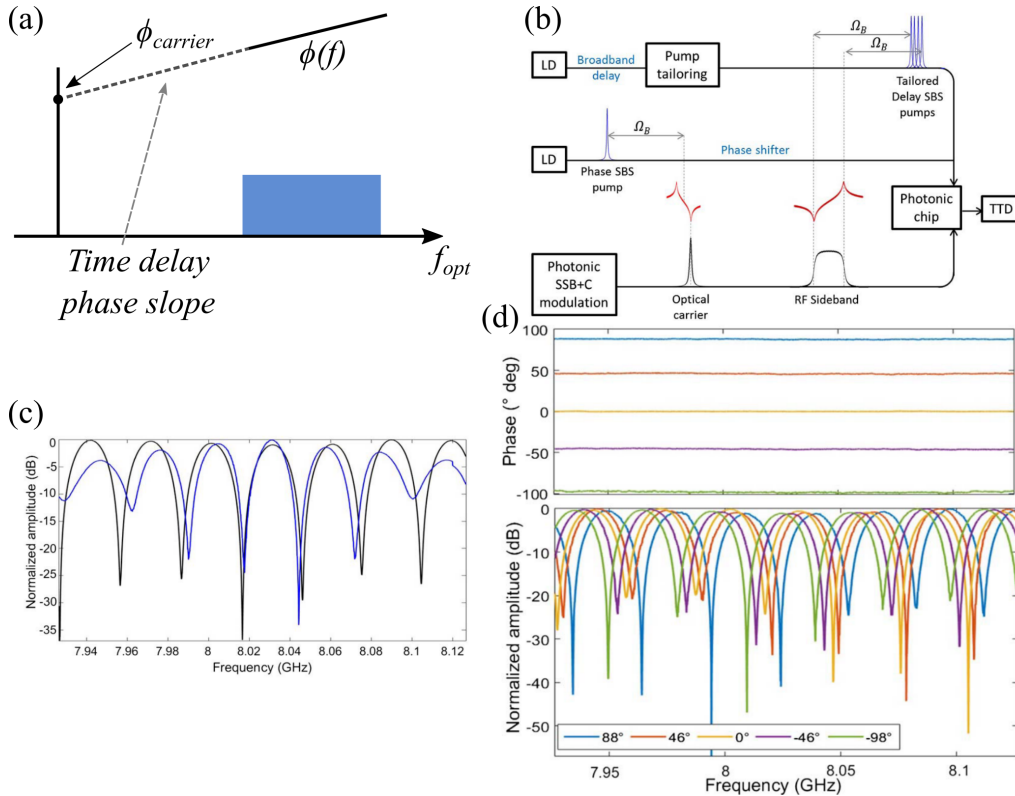


Figure 4.17: SBS-based true time delay using separate carrier tuning (a) Operating principle. (b) Experimental setup. (c) The difference in FSR of two tap filter corresponds to a delay of 4 ns. (d) A phase shift applied to the optical carrier changes the null location of the interference pattern. (Figure (b-d) from [40], © 2017, Optica Publishing Group.)

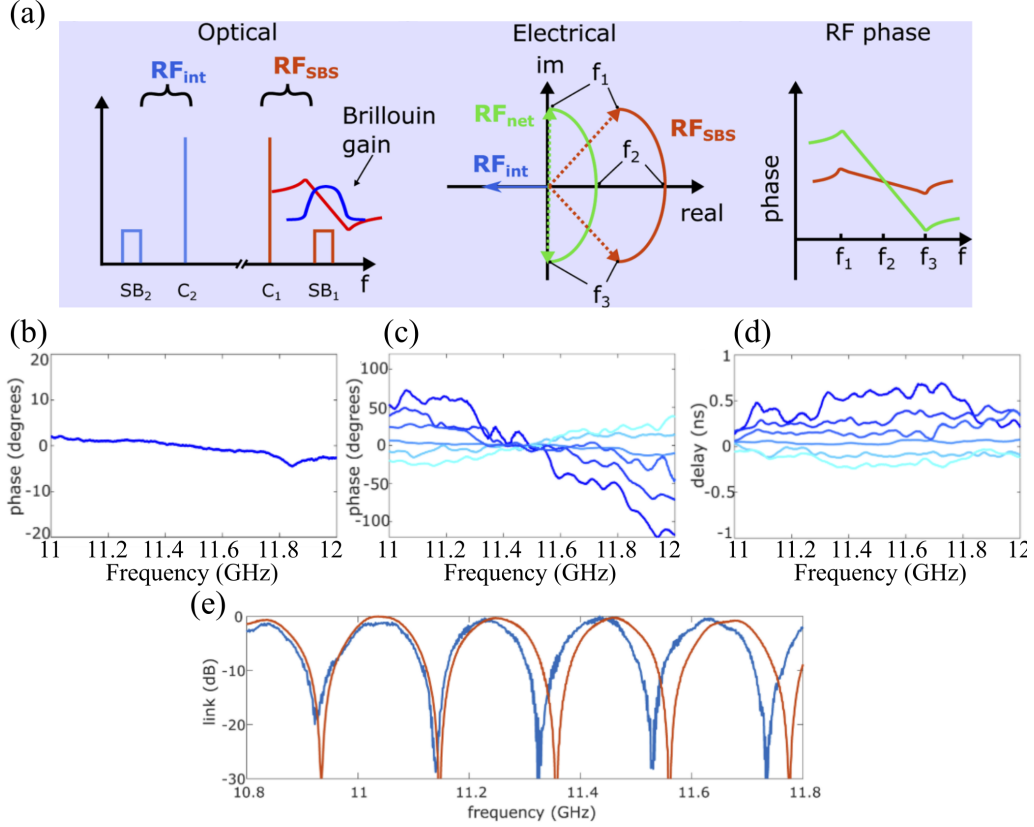


Figure 4.18: SBS-based true time delay enhancement using an RF interferometric technique. (a) Operating principle of interferometric true time delay (b) Raw Brillouin phase response over 1 GHz. (c) Enhanced phase response over 1 GHz. (d) Enhanced time delay over 1 GHz. (e) Two tap filter, showing change of free spectral range due to the imposed time delay. (Figures from [50], © 2020, Optica Publishing Group.)

Fig. 4.18 (b) shows the raw phase slope from the Brillouin resonance over a 1 GHz bandwidth, between 11 and 12 GHz. Fig. 4.18 (c) shows the phase slope created when using enhancement factors of up to 30. The corresponding delay for each phase slope in Fig. 4.18 (c) is shown in Fig. 4.18 (d). We note that fluctuations in the phase and delay responses are amplified due to the phase enhancement scheme, and can in principle be reduced by finely tailoring the Brillouin gain spectrum to achieve the desired response [50]. Fig. 4.18 (e) shows the two tap-filter response with and without the TTD applied. The change to the FSR corresponds to a delay of 330 ps [50].

If even larger delays are required for PAAs with more elements Brillouin-based light storage [362], quasi-light storage [363], and Brillouin dynamic grating based approaches [364]–[366] have been shown to achieve larger delay-bandwidth products compared to the Brillouin slow-light based approaches outlined above. However, so far those techniques have only been shown for optical signals and they still have to be demonstrated in an MWP link and only the former has been shown in an integrated platform [362]. Besides the large fractional delay achieved using Brillouin light storage it has been shown that the strict phase-matching condition between optical and acoustic waves in the Brillouin process can be used to induce non-reciprocal delays [367] which might open possibilities for novel processing methods in PAAs.

Alternatively, recent approaches utilise the polarisation dependence of Brillouin gain in both spun birefringent [368], and standard single mode [369] fibres to introduce narrow sub-MHz spectral dips within the Brillouin gain resonance. The spectral dip gives rise to a steep phase slope, which could form the basis of an MWP phase shifter or TTD with the potential to reduce the link gain penalty compared to the phase and delay enhancement techniques outlined above. Initial demonstrations of polarisation pulling using SBS in As_2S_3 waveguides by Athanasios *et al.* [370] highlights possibilities of implementing such optical processing techniques on PICs in a compact form factor.

4.3.4 Brillouin Frequency Converters

RF frequency converters, often referred to as mixers, are used to convert between RF and a lower intermediate frequency (IF) [371], which is a critical function in PAAs. A mixer is a three port device, as shown in Fig. 4.19 (a), where the frequency of the input RF signal is translated according to its difference in frequency from a second CW input tone, referred to as a local oscillator (LO). When downconverting a spectrum that contains a desired signal at frequency, f_{RF} , and a signal equally spaced but on the other side of the LO at frequency, f_{IM} , they are both converted to the same IF frequency, as shown in Fig. 4.19 (b-c). The undesired signal in the IF spectrum is referred to as an image and causes distortion. The process of suppressing the image signal in the IF spectrum is referred to as image rejection [372], and is depicted in Fig. 4.19 (d).

Brillouin-based MWP mixers inherently perform image rejection by ensuring that the output photocurrents destructively interfere in the absence of a Brillouin resonance. Zhu *et al.* demonstrated a chip-based SBS image-rejection mixer [371], which utilised a dual-parallel Mach-Zehnder modulator (DPMZM) by placing the RF input on the top sub-MZM and the LO input on the lower sub-MZM, shown in Fig. 4.20 (a). The DPMZM was biased to ensure destructive interference outside the Brillouin resonance, as shown in Fig. 4.20 (b). Frequency conversion occurs when Brillouin gain is applied to the signal of interest on the upper sideband, and SBS loss is applied to the signal of interest on the lower sideband, as shown in Fig. 4.20 (c). When SBS gain and loss are applied to the signals of interest, the photocurrent amplitudes are unequal and cause an output to form. An image rejection of 45.3 dB [371] was achieved using 15.1 dB of SBS gain and 13.3 dB of SBS loss, shown in Fig. 4.20 (d). This system exhibited a conversion gain of -9 dB over the 3.2-13.2 GHz frequency range [371]. Although this demonstration provides low-loss conversion and deep image rejection, it only supported frequency conversion of narrow-band signals. However, broadband image-rejection frequency conversion is required for

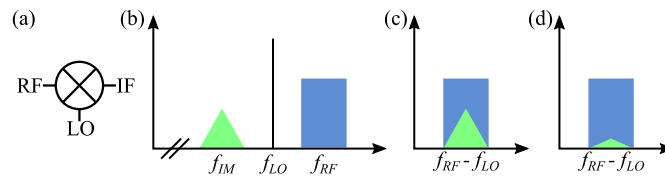


Figure 4.19: (a) Schematic symbol for an RF mixer. (b) Example spectrum to be converted in frequency. (c) Downconverted spectrum without image rejection. (d) Downconverted spectrum with image rejection. Note the signal at the image frequency is attenuated.

compatibility with modern PAAs. McKay *et al.* demonstrated broadband image rejection downconversion using on-chip SBS by modulating the Brillouin pump over a broad bandwidth. In doing so, a broadband image rejection of 28 dB and 16 dB over of 100 MHz and 400 MHz bandwidths respectively was achieved [20], as shown in 4.20 (e).

4.3.5 Brillouin Radio frequency sources

RF sources are essential components in the PAA, and when used in conjunction with an MWP frequency converter, enable PAAs with widely tunable central frequencies. RF sources with low phase noise are critical in wireless communication [374] and radar [22], [375] systems. However, RF frequency synthesizers typically generate an output that is

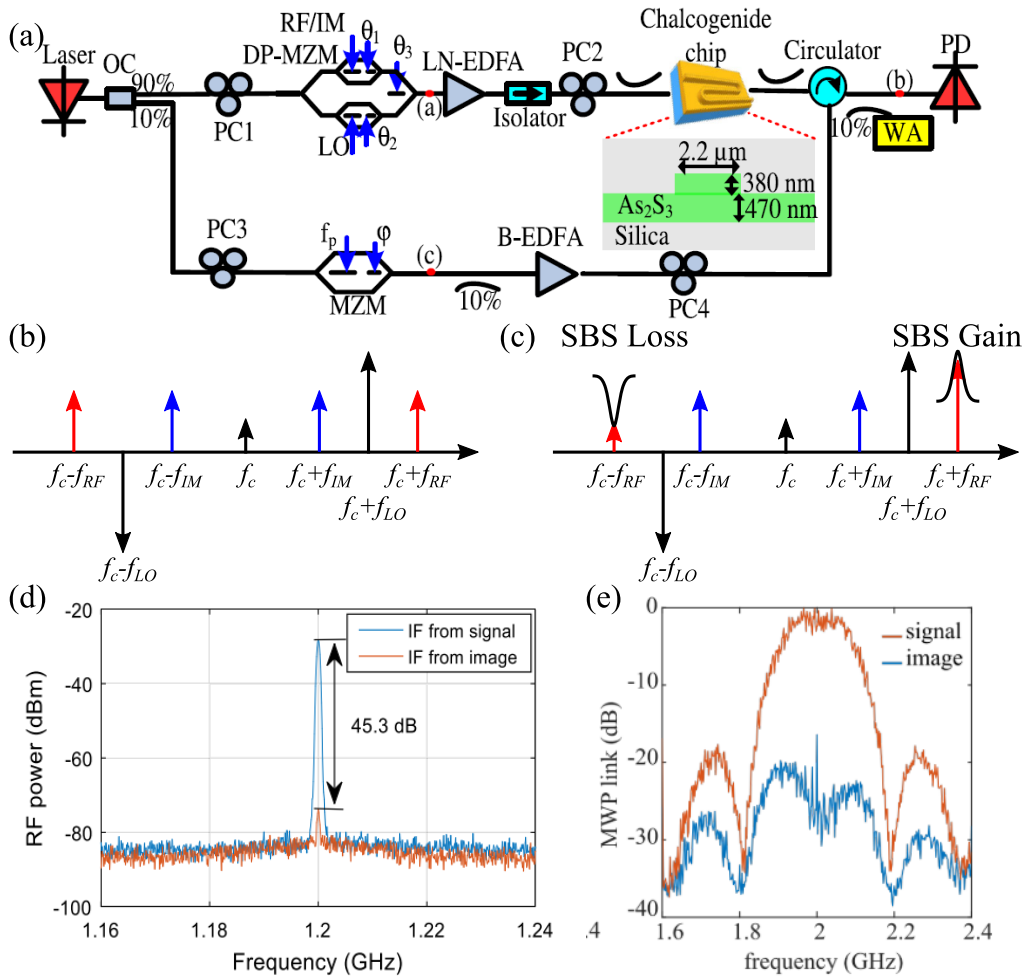


Figure 4.20: Brillouin-based RF frequency converters. (a) Experimental setup for Brillouin-based image rejection mixer. (b) Optical spectrum without Brillouin processing. (c) Optical spectrum after Brillouin processing. Gain and loss are applied to the RF signal of interest on the upper and lower sideband, respectively. (d) The IF spectrum of a narrowband Brillouin mixer. An image rejection of 45.3 dB was achieved. (g) The IF spectrum of a broadband Brillouin downconverter showing 16 dB of image rejection from over a 400 MHz band. (Figure (a, d) from [371], © 2020, Optica Publishing Group. Figure (e) from [20], © 2021, IEEE.)

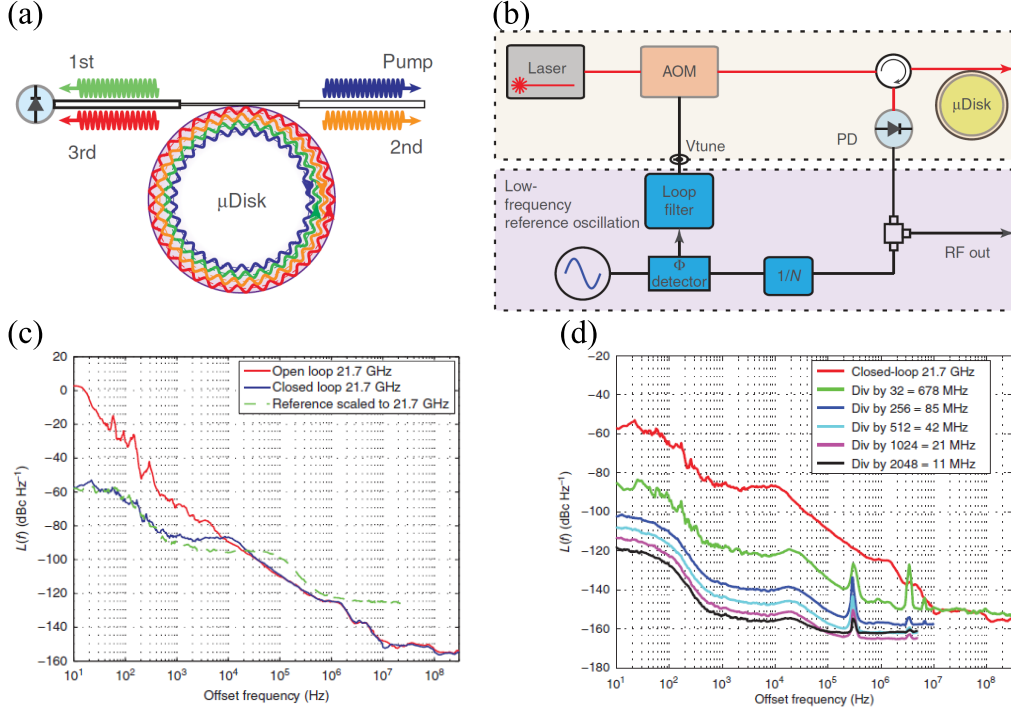


Figure 4.21: Brillouin-based RF source using silica resonator. (a) Principle of Brillouin RF source. An optical pump generates higher-order Brillouin responses, forming the tones used to create the RF signal. (b) Simplified schematic of Brillouin RF source. (c) Single-sided phase noise of RF source with and without closed loop operation. (d) Phase noise improvements to the 21.7 GHz output when its frequency is divided. (Figures from [373], licensed under a Creative Commons Attribution (CC BY) license.)

up-converted from a stable, low frequency electronic source. The up-conversion process increases the phase noise by a factor $20 \log_{10}(N)$ [373] when the frequency is multiplied by a factor of N . In practice, the phase noise is increased by additional noise from RF amplification stages [376]. Conversely, MWP-based RF sources directly generate outputs at radio frequencies by heterodyning two, or more optical tones. Brillouin laser cavities have been used to enable significant reductions to the linewidth, in addition to close-in frequency noise and RIN noise [227] of optical tones. Brillouin lasers have formed the basis of RF sources that achieve phase noise comparable to [373], or lower than the phase noise of high-performance RF microwave sources [377]. Additionally, photonic-based approaches have the distinct advantage of being able to generate RF tones over wide frequency ranges [22], [377].

Li *et al.* used Brillouin lasing in a high Q silica disk resonator [373] as the basis of a low phase noise RF source [373]. The main advantage of this scheme is that common mode jitter is reduced because the Brillouin laser lines share the same optical path [373], reducing the phase noise of the RF output. The operating principle of this device is shown in Fig. 4.21 (a), where cascaded Stokes waves are generated in the silica disk. The first and third order tones co-propagate and are separated in frequency by twice the BFS, 21.7 GHz [373]. An RF signal with a frequency of 21.7 GHz is formed when both tones mix at a PD. Although the close-in phase noise ($< 10\text{kHz}$) of this device is relatively large when free-running, it is reduced using closed-loop control to form a phase-locked loop,

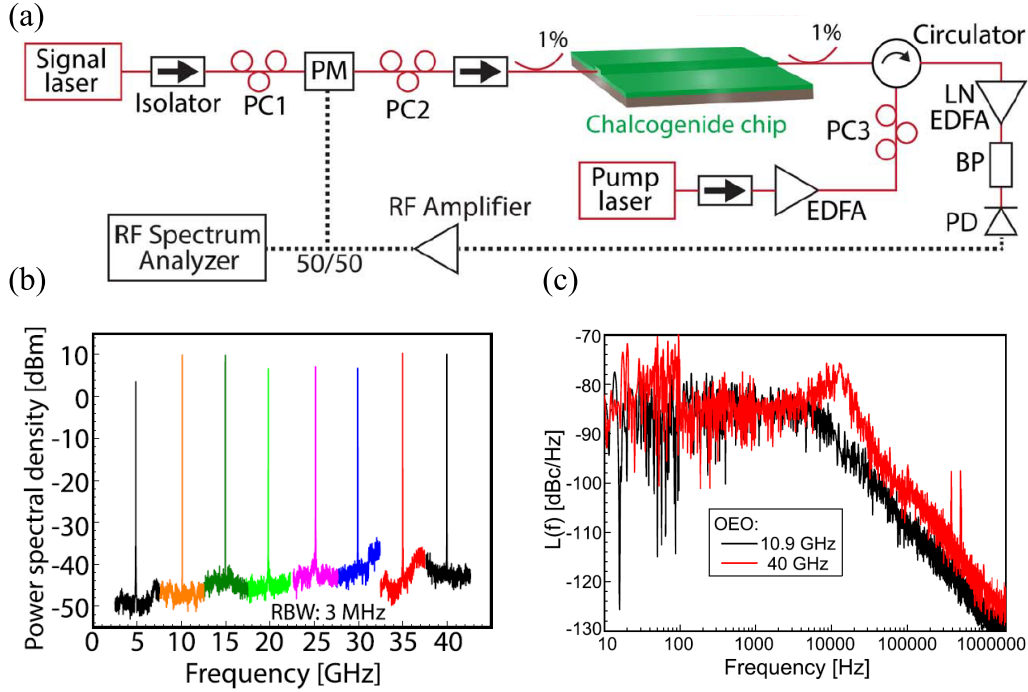


Figure 4.22: Chip-based Brillouin optoelectronic oscillator (OEO). (a) Experimental setup for Brillouin based OEO. (b) Wideband tunability of SBS OEO. Images taken from [22]. (c) Phase noise for generated signals at 10.9 GHz and 40 GHz. (Figures from [22], © 2016, Optica Publishing Group.)

as shown in Fig. 4.21 (b). The feedback loop allows for the use of a stable low-frequency quartz oscillator as a frequency reference [373], which improves the output phase noise, shown in Fig. 4.21 (c). A low white-phase-noise floor of -160 dBc/Hz was obtained [373]. When the output frequency is divided, the phase noise improves, as shown in Fig. 4.21 (d). We note that high Q Si_3N_4 MRRs, demonstrated by Gundavarapu *et al.*, have been used to form an RF source through cascaded Brillouin generation. This approach offers narrow Stokes linewidths of only 0.7 Hz [227], and is an attractive method to generate low phase noise RF signals. It is desirable to generate microwaves with continuous frequency tunability for use in PAAs, however, the frequency of the RF output is limited to discrete frequencies determined by the BFS [227], [373].

Li *et al.* extended on the work presented in Ref. [373] by demonstrating a continuously tunable RF source using two lasers locked to distinct modes of a common Brillouin laser cavity [377]. A reduction in phase noise was achieved through optical frequency division.

Optoelectronic oscillators (OEOs) [87] are an alternate method to generate low phase noise RF signals. However, most OEOs rely on a narrowband filter to select one cavity mode, which are not widely tunable if implemented using traditional electronic methods [22]. On-chip SBS enables widely tunable filtering functionality, and has been used as a widely tunable MWP filter to select the desired OEO loop mode and enable wideband RF signal generation, as shown in Fig. 4.22 (a) [22]. In this approach, Brillouin gain amplifies one cavity mode to create phase-intensity modulation conversion and sustain an oscillation. Using this technique, an RF tone was synthesised with frequency up to 40 GHz, as shown in Fig. 4.22 (b). The advantage of optically generating microwave tones

over a wide frequency range is highlighted in Fig. 4.22 (c), where signals at 10.9 GHz and 40 GHz have similar phase noise.

4.3.6 Non-reciprocal devices

Non-reciprocal components are key to on-chip photonic processors, as they protect optical components from back-reflected light [378], which is particularly important for backward Brillouin photonic devices in which counterpropagating pump and probe signals are required [51]. Optical devices traditionally achieve non-reciprocity using magneto-optical materials which induce a non-reciprocal Faraday rotation [379]. While chip-scale isolators and circulators using magneto-optic garnets [380] have demonstrated large 30 dB isolation over a wide 16 nm 10-dB bandwidth [381], these devices are challenging to integrate into PICs [382].

Non-reciprocity requires the breaking of time reversal symmetry, which can be implemented using SBS [383]. Acousto-optic non-reciprocal devices rely on phase matched optical mode conversions that occur for optical waves travelling in one direction but not the other [383], [384]. Chip-based demonstrations of optical non-reciprocity have utilised resonant structures [382], [385], which limited the usable bandwidth, or required suspended waveguides to increase acoustic guidance [342], [386], which increased fabrication complexity. On-chip non-reciprocity has been achieved without the use of resonant structures [21], [387]. For example, Kittlaus *et al.* recently demonstrated an electrically induced acoustic wave using standard silicon waveguides and on-chip mode multiplexers to facilitate non-reciprocal intermodal acousto-optic scattering [388]. This device achieved over 16 dB of optical isolation over a 100 GHz bandwidth without the use of an optical cavity [388].

An alternate approach introduced by Liu *et al.* [389] avoids the requirement for a non-reciprocal element by exploiting backward inter-modal Brillouin scattering (BIBS) in combination with passive mode-selective filters. However, less than 1 dB of BIBS gain was observed [389]. Therefore, further developments to this platform are required to increase the magnitude of BIBS gain generated to make this platform suitable for MWP processing.

4.4 Future Challenges

In this chapter, we reviewed chip-scale Brillouin platforms and their application in MWP signal processing platforms. Based on this analysis, we now identify future research directions that work towards the ultimate goal of achieving a high-performance and fully-integrated Brillouin-based platform for MWP signal processing.

Fig. 4.23 provides an overview of the chronological evolution of on-chip Brillouin processors. While many platforms have been developed that induce SBS in chip-scale platforms, we limit the scope of this discussion to devices that target MWP applications as it is the focus of this thesis. The platforms that we identify here highlight the chronological evolution of on-chip Brillouin platforms for MWP processors and each platform was discussed in detail in Section 4.2.

By observing the number of citations per year for papers with the terms ‘stimulated Brillouin scattering’ and ‘microwave photonics’ in Fig. 4.23 (a), it is evident Pant *et al.*

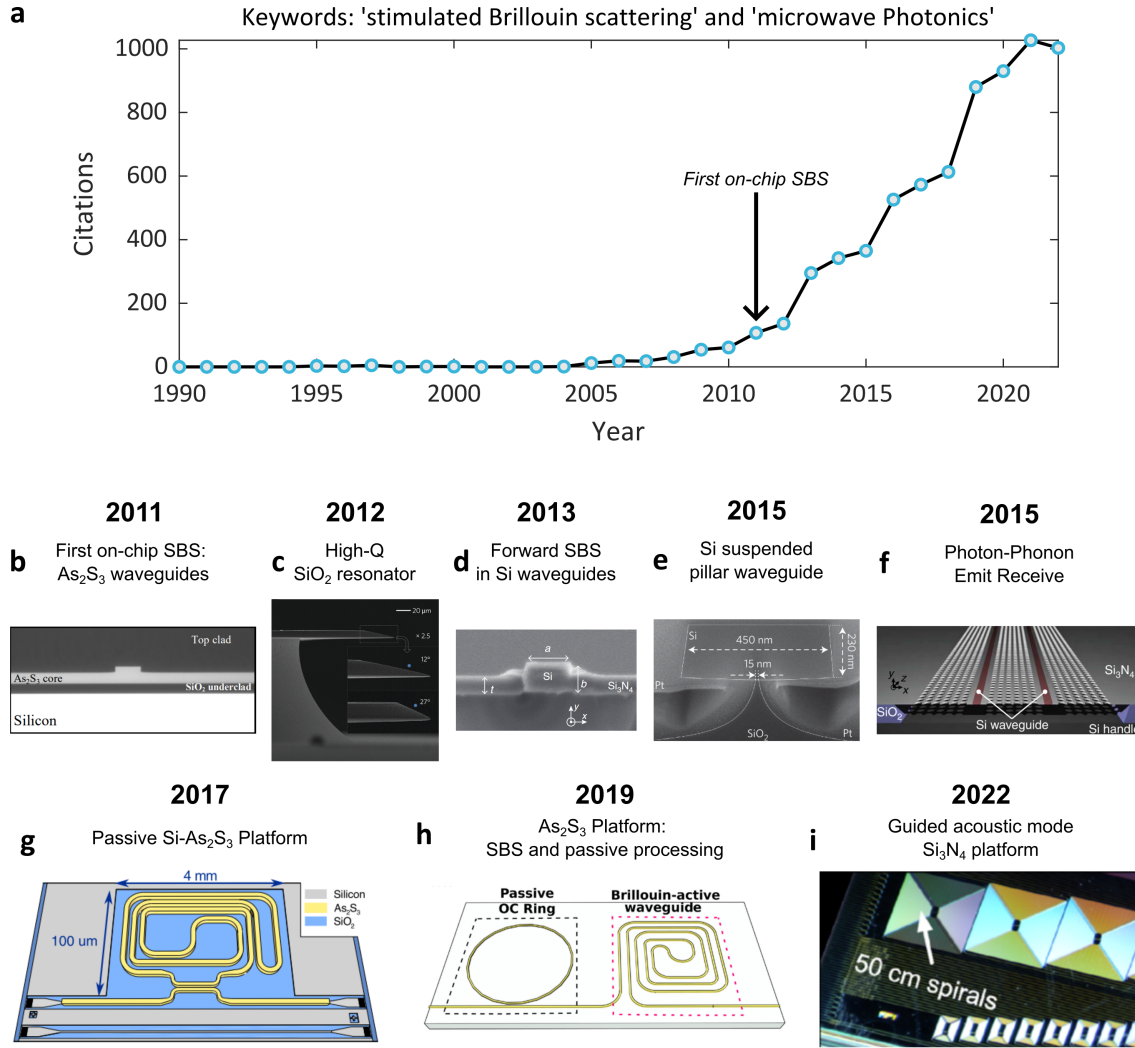


Figure 4.23: Chronological development of Brillouin-based MWP processing platforms. (a) Citations per year for papers the terms ‘stimulated Brillouin scattering’ and ‘microwave photonics’. The citation graph was collated using Clarivate Web of Science. (b) As_2S_3 waveguides (c) SiO_2 wedge resonator (d) suspended silicon waveguide (e) pillar silicon waveguide (f) Photon-phonon emit-receive waveguide (g) Passive Si- As_2S_3 platform (h) Monolithically integrated As_2S_3 platform with MRRs and Brillouin waveguides (i) Guided acoustic mode silicon nitride platform. Figures from (b) [339], (c) [337], (d) [317], (e) [319], (f) [318], (g) [176], (h) [45], (i) [39]

drew significant interest from the MWP community by generating SBS in compact As_2S_3 waveguides [164], depicted in Fig. 4.23 (b).

Despite the many advances since the initial demonstration of on-chip SBS in 2011, Brillouin platforms have so far been implemented as standalone devices which generate SBS. As a result, they do not offer any other functions required in an MWP link such as modulation, photodetection, and cannot serve as an electrically pumped optical source. Consequently, these MWP systems rely on discrete fibre-based components to achieve the required functions of the MWP link, which increases their SWaP and is not desirable when considering deployment in SWaP-constrained applications. Therefore, developing Brillouin platforms with active on-chip components is of great interest to the MWP com-

munity and is key to enabling many SWaP-constrained applications [81].

Silicon is an attractive platform for implementing Brillouin devices with active photonic components, owing to the maturity of the silicon platform. Additionally, silicon platforms offer a pathway to integrate key components of the MWP link including lasers, modulators and photodetectors, as outlined in Chapter 2. Silicon photonic platforms that induce forward Brillouin interactions have been extensively developed and offer large Brillouin gain coefficients if the silicon waveguides are suspended, as depicted in Fig. 4.23 (d-f). A unique approach was taken by Morrison *et al.*, Fig. 4.23 (g), by heterogeneously integrating Brillouin As_2S_3 waveguides in a silicon photonic platform, which overcame the challenges of generating large magnitudes of SBS gain in the silicon photonic platform. Although many on-chip Brillouin platforms have been demonstrated, none take advantage of the active on-chip functionalities offered by silicon photonics.

Furthermore, combining nonlinear Brillouin processing with linear MRR-based processing allows systems to benefit from the fine spectral resolution and flexibility offered by SBS, with additional processing functions implemented by MRRs. This unique combination of linear and nonlinear processes has enabled RF link gain improvements for MWP notch filters [39], [45], [340] and time delay enhancement for TTDs [50]. However, there is an apparent trade-off between generating large magnitudes of SBS gain and fabricating narrow linewidth MRRs in a monolithic platform. For example, Liu *et al.* demonstrated a monolithically integrated As_2S_3 platform that includes MRRs and large Brillouin gain As_2S_3 waveguides [45], depicted in Fig. 4.23 (h). However, the As_2S_3 MRRs have wide 3-dB bandwidths of 3 GHz which are limited by the propagation losses of the MRR. Conversely, the TriPleX Si_3N_4 platform presented by Botter *et al.* [39], Fig. 4.23 (i), can enable MRRs with 3-dB linewidths in the order of 100's MHz [178], owing to the low propagation losses of Si_3N_4 waveguides. However, the magnitude of the generated Brillouin gain in Si_3N_4 platforms is limited for reasons that we outlined in Section 4.2.4, which limits the potential Brillouin-based MWP functionalities that can be exploited by the platform. Considering the trade-off between narrow linewidth MRRs and the ability to generate large magnitudes of SBS gain, it is attractive to explore integration approaches of large SBS gain As_2S_3 waveguides with ultra low-loss materials, including silicon nitride and lithium niobate, to unlock the full potential of combined Brillouin and MRR-based processing.

Chapter 5

Multi-band chip-based MWP notch filter for ultra-deep interference rejection

Unwanted high-power and frequency-agile radio-frequency (RF) signals cause saturation and nonlinear distortions to sensitive RF receivers. RF notch filters with reconfigurable responses over multiple frequency bands are highly sought after to prevent these detrimental effects. Microwave photonic (MWP) notch filters have shown increased frequency agility compared to their electronic counterparts. However, demonstrated filter schemes focus only on single notch responses, leaving them unable to simultaneously attenuate multiple interferers over a wide frequency range. In this chapter, we demonstrate a high-performance chip-based MWP notch filter with three independent notches widely tunable over 20 GHz. The filter exhibits low RF passband losses of 8 dB and peak notch depth greater than 40 dB with 500 MHz spectral resolution. Specifically, each notch is formed through cascaded optical processing from on-chip, low-loss Si_3N_4 microresonators and Brillouin gain on an As_2S_3 photonic chip. We use this filter to demonstrate high-performance analog RF filtering by substantially attenuating multiple interferers, and show effective image rejection during RF downconversion in a communications receiver to enable a large reduction of error vector magnitude (EVM) from 60% to 15%. Finally, we provide performance analysis and design perspectives for future photonic integration of the proposed filter subsystem.

This chapter is based on the works:

- **Matthew Garrett**, Yang Liu, Moritz Merklein, Duk-Yong Choi, Kunlun Yan, Stephen J. Madden, and Benjamin J. Eggleton, "Multi-Band and Frequency-Agile Chip-Based RF Photonic Filter for Ultra-Deep Interference Rejection," *Journal of Lightwave Technology*, 2022, **40**, 1672-1680.
- M. Garrett, Y. Liu, D. Choi, K. Yan, S. J. Madden, and B. J. Eggleton, "Ultra-Deep Multi-Notch Microwave Photonic Filter utilising On-Chip Brillouin processing and Microring Resonators," in 2021 Conference on Lasers and Electro-Optics Europe and European Quantum Electronics Conference, OSA Technical Digest (Optica Publishing Group, 2021), paper cd.4.5.
- M. Garrett, Y. Liu, M. Merklein, D. Choi, K. Yan, S. J. Madden, and B. J. Eggleton, "Chip-based RF Photonic Notch Filter for Deep Rejection of Multi-Band Interfering Signals," 2021 International Topical Meeting on Microwave Photonics (MWP), 2021, pp. 1-4.
- M. Garrett, Y. Liu, M. Merklein, D. Choi, K. Yan, S. J. Madden, and B. J. Eggleton, 'Microwave Photonic Notch Filter for Suppression of Multi-Band Interferers', in ANZCOP21, 2021, p. 77.

5.1 Introduction

Radio frequency (RF) notch filters that operate over multiple frequency bands with reconfigurable responses are highly sought after for use in RF front ends [27]. Large stopband rejection is required to mitigate agile frequency hopping interference used for wireless communications [390] and to prevent damage to sensitive electronics from high power jamming signals. Conventional RF electronic filters that rely on fixed cavities [391] or surface acoustic waves [67], [392] are typically limited to a single RF resonance and are fundamentally unable to demonstrate wideband response reconfigurability and flexibility. On the other hand, filters that are based on digital signal processing may be easily reconfigured, however, their operating bandwidth is limited by the sampling rate of commercially available analog-digital converters [393]. Therefore, RF filters with multiple reconfigurable stopbands are demanded for use in next-generation wireless communication [19], [394] and spectrum sensing systems [376].

Microwave photonics (MWP) [25] provides an attractive solution to implement widely tunable RF notch filters [27] whilst maintaining large notch depth and dynamic reconfigurability [150], [395]. These desirable characteristics are enabled by the large fractional bandwidth of optical frequencies compared to radio frequencies and advanced optical response engineering. Recent implementations of MWP systems on photonic integrated circuits (PICs) offer advantages of small size, weight and power (SWaP) over their optical-fibre-based predecessors [16], [35], [36], [45], [352], [396]–[398]. As a result, chip-based MWP notch filters have attracted great interest due to their frequency agility and reconfigurability, whilst maintaining high-performance [390].

Previously reported chip-based MWP filters based on tapped delay line architectures [33], [399]–[401] have achieved multiple passbands over a wide RF frequency range. However, the complex and precise control of the tap parameters intrinsically limits the filter frequency tuning to sub-GHz ranges, due to the small free spectral range of the periodic filter response. MWP notch filters based on optical micro-ring resonators (MRRs) with compact footprints [9], [172], [178], [290] have demonstrated broad central frequency tunability up to 40 GHz with a filter rejection greater than 60 dB [402]. However, the implementation of these filter schemes often requires complex control of the amplitude and phase of optical sidebands, which leads to a significant increase in RF insertion loss.

Brillouin processing [51] has emerged as an attractive candidate for MWP filters [147], [396], [403], [404], owing to optical parametric generation of gain, narrow resonant bandwidth (≈ 30 MHz) and ultra-wide frequency tunability for high-resolution signal processing, and compatibility with chip-scale integration [176], [227], [322], [342], [389]. On-chip Brillouin processing has recently been cascaded with an over-coupled MRR [45] on an As_2S_3 photonic chip to form a single RF notch filter through destructive interference of RF photocurrents. However, the lack of frequency tunability of the As_2S_3 MRR and its wide resonance linewidth (>2 GHz due to high linear losses) prohibits the feasibility to achieve multiple RF notches with wideband independent tunability and fine frequency resolution, which are required for applications within the congested modern RF spectrum.

In this chapter, we overcome these limitations with a hybrid approach [50], combining tunable and low-loss Si_3N_4 MRRs with on-chip active As_2S_3 Brillouin waveguides on two fibre-coupled PICs. This approach provides a basis for multiple filter notches with tunability over 20 GHz, fine 3-dB spectral resolution of 500 MHz and a stopband rejection

tion exceeding 40 dB. Three independent RF notch responses are formed by cascading multiple low-loss on-chip Si_3N_4 MRRs for phase inversion and Brillouin resonances for RF notch depth enhancement. We demonstrate this MWP filter with three ultra-high-rejection and independently tunable notches while sustaining low RF loss of 8 dB, noise figure (NF) of 20.8 dB and spurious free dynamic range (SFDR) of $92.2 \text{ dB}\cdot\text{Hz}^{2/3}$. With this level of performance, we experimentally demonstrate analog RF signal filtering of multiple frequency hopping interferers and image rejection for RF down-conversion in communications receivers. Finally, we provide analysis and techniques to further minimise performance degradation caused by undesired Brillouin pump reflections and discuss the perspectives to implement such filter subsystems via heterogeneous photonic integration.

5.2 Operating Principle

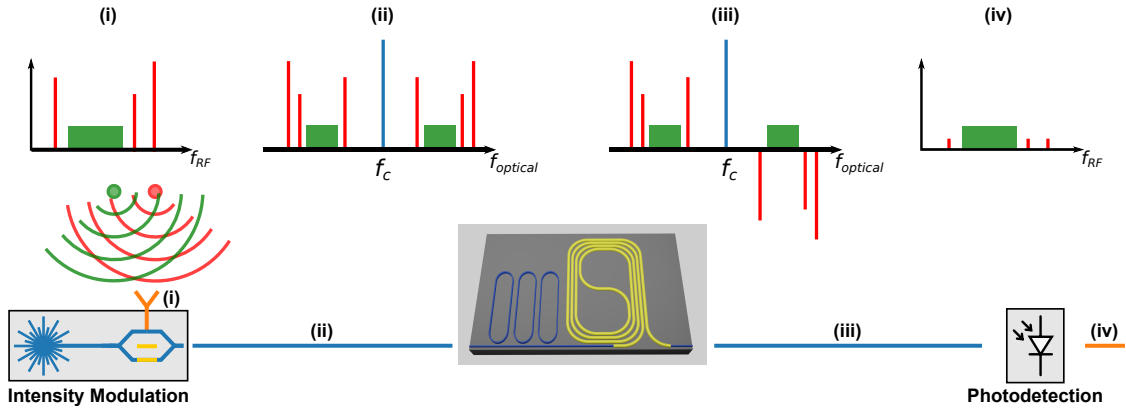


Figure 5.1: Simplified filter architecture. The RF spectrum containing a signal of interest among strong unwanted signals is modulated onto an optical carrier through intensity modulation. A π phase shift is selectively applied to one optical sideband after on-chip optical processing (artist's impression). The amplitude of the interfering signals in the filter stopband at the output of the photodetector.

Figure 5.1 provides a high-level overview of a filter architecture based on destructively interfering RF photocurrents. The input RF spectrum is modulated onto an optical carrier through intensity modulation (IM). On-chip optical processing is utilised for the generation of deep RF notches through cascaded MRR and Brillouin processing. The optical spectrum at the photodetector (PD) contains spectrally localised regions with a π phase shift that corresponds to the filter stopband. Destructive interference of the RF beat contribution from the lower and upper sidebands constitutes the filter stopband, which attenuates the undesired signals at the PD output.

Destructive interference of RF photocurrents is the key phenomenon that enables the formation of deep RF notches. This requires the superposition of two currents with equal amplitude, but a relative phase offset of π . In small signal conditions, MWP systems using IM have two main RF photocurrent contributions, one from the beat between the carrier and upper sideband, the other from the beat between the carrier and the lower sideband. Accordingly, the net output photocurrent, i , is given by:

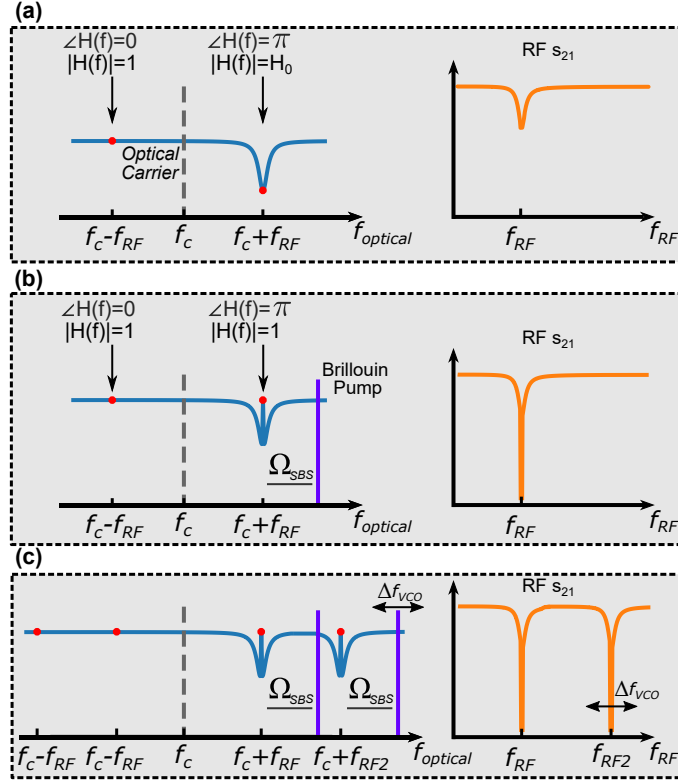


Figure 5.2: Destructive interference concept for RF photonic multi-notch filter. (a) MRR only based approach with limited RF rejection. (b) Brillouin enhancement of notch depth. (c) Independently tunable notches using multiple MRR and Brillouin resonances.

$$i(f_{\text{RF}}) \propto |H(f_c + f_{\text{RF}})| \cos(2\pi f_{\text{RF}}t + \angle H(f_c + f_{\text{RF}})) + |H(f_c - f_{\text{RF}})| \cos(2\pi f_{\text{RF}}t + \angle H(f_c - f_{\text{RF}})), \quad (5.1)$$

where f_{RF} and f_c are the frequencies of the modulating RF tone and optical carrier and $H(f)$ is the electric field transfer function through the MWP link at frequency f .

By analysing Eq. (5.1), destructive interference occurs when the optical transfer function through the MWP link is engineered such that the upper and lower sidebands are equal in amplitude, but differ in phase by π :

$$|H(f_c + f_{\text{RF}})| \cos(2\pi f_{\text{RF}}t + \angle H(f_c + f_{\text{RF}})) = -|H(f_c - f_{\text{RF}})| \cos(2\pi f_{\text{RF}}t + \angle H(f_c - f_{\text{RF}})). \quad (5.2)$$

The unique combination of active on-chip Brillouin processing and passive optical MRR processing enables destructively interfering RF photocurrents using a simple intensity modulator is further detailed in Fig. 5.2. An over-coupled MRR processes the upper sideband in Fig. 5.2 (a) and imparts the required π phase shift and amplitude attenuation of H_0 at its resonant frequency, $H_{\text{MRR}}(f_c + f_{\text{RF}}) = H_0 e^{j\pi}$. This forms an RF notch; however, the notch depth is limited by the unequal magnitude of the upper and lower sideband photocurrent contributions. Subsequently, we apply Brillouin gain to ensure the amplitude of the upper and lower sidebands are equal, $H_{\text{SBS}}(f_c + f_{\text{RF}}) = \frac{1}{H_0}$. This results in a net transfer function with unity gain and π phase shift, $H(f_c + f_{\text{RF}}) = e^{j\pi}$ at the MRR and Brillouin resonant frequency, forming a deep notch, shown in Fig. 5.2 (b).

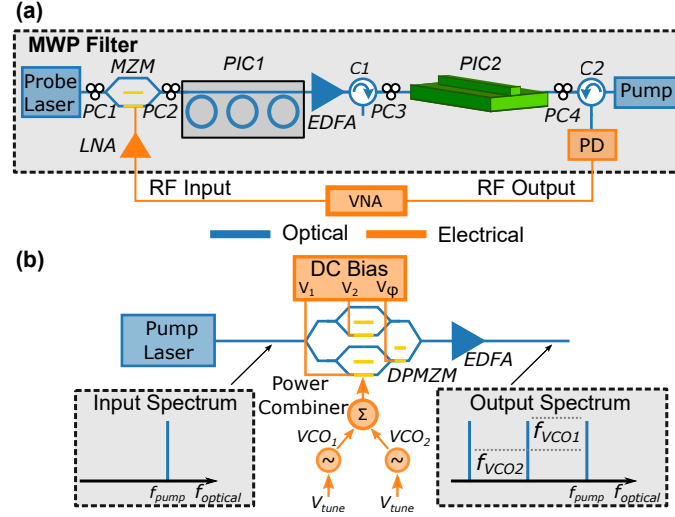


Figure 5.3: (a) Experimental setup for MWP filter with multiple-notches. (b) Pump generation scheme that enables simultaneous operation over multiple RF bands. MZM: Mach-Zehnder Modulator, DPMZM: Dual Parallel MZM, EDFA: Erbium-doped fibre amplifier, LNA: Low Noise Amplifier, VCO: Voltage Controlled Oscillator, PC: Polarisation Controller

We achieve the above operation for multiple RF notches by overlapping the MRR resonance and the Brillouin gain resonance for each RF notch, as shown in Fig. 5.2 (c). We note it is possible to achieve deep notches through destructively interfering photocurrents using other MWP filter architectures that rely on passive processing. However, these schemes commonly rely on specific biasing of a dual-drive Mach-Zehnder modulator (DDMZM) [290] or a dual-parallel Mach-Zehnder modulator (DPMZM) [396] that lead to a reduction of the overall RF gain due to wideband RF partial destructive interference in the filter passband. In contrast, the scheme presented in this chapter utilises simple intensity modulation and does not place any constraints on modulator bias voltages to achieve deep RF notch depths. Accordingly, link gain, linearity and noise figure of the MWP link in the passband can be improved by tuning the modulator bias without reducing the RF notch depth.

5.3 Demonstration of frequency agile and multi-band filtering

The experimental setup for demonstrating frequency agile and multi-band RF performance is outlined in Fig. 5.3 (a). The probe laser undergoes intensity modulation from the amplified RF input. The modulated probe is subsequently processed by a series of over-coupled MRRs before optical amplification by an erbium-doped fibre amplifier (EDFA). Finally, Brillouin amplification occurs in a 12 cm As_2S_3 waveguide with propagation loss of ≈ 0.5 dB/cm and ≈ 3.5 dB loss per facet, before a deep notch is formed in the RF domain. We used two PICs as this study focuses on proving the principle of this multi-notch filtering scheme. We use a bank of voltage controlled oscillators (VCOs) and a DPMZM to generate multiple pump tones through single sideband modulation (SSB), as shown in

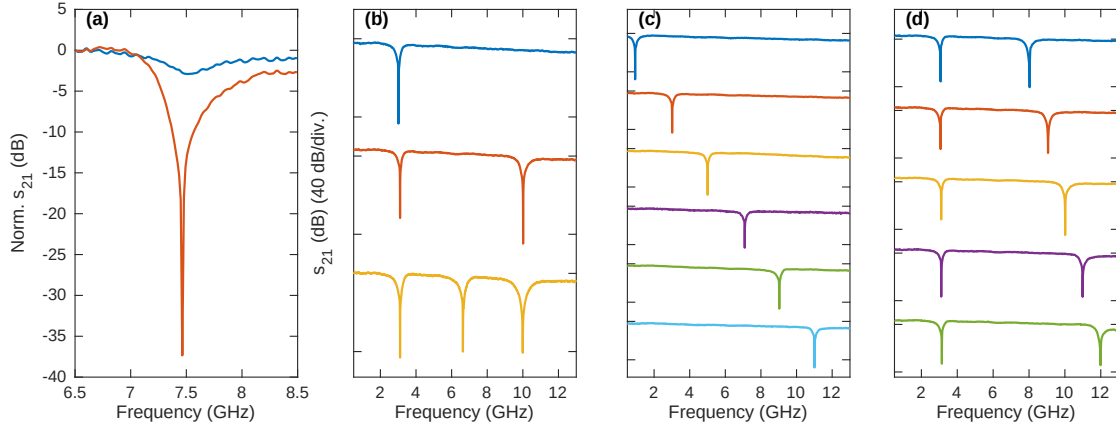


Figure 5.4: (a) Measured MRR transfer function (blue) and measured RF notch (orange) using IM and Brillouin enhancement. (b) Dynamic control of number of notches in filter response. (c) Filter wideband frequency tunability. (d) Independently tunable notch frequencies in multi-notch configuration.

Fig. 5.3 (b). An advantage of this scheme is that a single laser and simple electronics are used to enable multi-band operation, without the need for multiple independent pump lasers [405] or large form factor RF sources [44]

We experimentally demonstrate frequency agile filtering over multiple RF bands in Fig. 5.4. First, we compare the optical transfer function of the MRR measured using SSB modulation to the RF notch response formed using the proposed scheme in Fig. 5.4 (a). A 3 dB optical resonance from the MRR is enhanced to a 38 dB notch depth with 500 MHz 3-dB bandwidth when combined with Brillouin amplification. This highlights the enhancement in RF rejection offered through this filtering scheme. The RF notch bandwidth is largely determined by the 3-dB bandwidth of the MRR. This may be reduced by using low-loss Si_3N_4 fabrication techniques which can provide 3-dB bandwidths less than 100 MHz [406]. In Fig. 5.4 (b), notches at 3 GHz, 6.1 GHz and 10 GHz are dynamically and independently activated by adding additional overlapping MRR and Brillouin resonances. We estimate an on-chip pump power of 20 dBm when operating with three RF notches. Under these conditions, up to 15 Brillouin pumps of similar power can be used while staying below the power handling threshold of PIC2.

Wideband frequency tunability is demonstrated with a single notch from 1-12 GHz in Fig. 5.4 (c) by changing the frequency of the probe laser while maintaining the frequency of the Brillouin and MRR resonances. Fig. 5.4 (d) demonstrates the ability to independently control the frequency of each notch, where we fix one notch frequency at 3 GHz while the second notch is swept over the entire X-band (8-12 GHz). This is achieved by varying the second MRR resonant frequency and the pump frequency controlled by the VCO. We note that the maximum frequency of this filter is not limited to 13 GHz, as shown in Fig. 5.4 (b-d). However, aliasing effects are present from the lower sideband sweeping through a different MRR resonance, which creates an additional notch that does not overlap with a Brillouin resonance.

MWP filters must also meet the demanding noise and linearity performance requirements to process signals in the crowded RF spectrum. The key metric characterising the noise performance of RF devices is the noise figure (NF), which measures the signal-noise

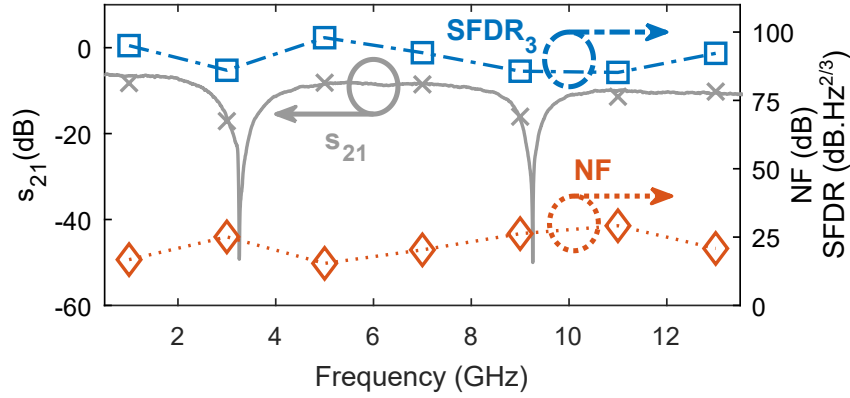


Figure 5.5: Measured spurious free dynamic range (blue) and noise figure (orange) for the shown RF response (grey).

ratio degradation at the output of the MWP filter and is defined by [1]:

$$NF = PSD - G + 174, \quad (5.3)$$

where PSD is the measured device power spectral density (dBm/Hz), G is the gain of the MWP link (dB) and 174 represents the thermal noise PSD (≈ -174 dBm/Hz). The linearity of the filter is primarily characterised by the spurious free dynamic range ($SFDR_n$), which measures the output signal power where the n^{th} order distortion terms are equal to the noise PSD. In Fig. 5.5 we plot the measured $SFDR_3$ (blue) and NF (orange) when the filter response is configured as per the grey trace. At 13 GHz we observe a $SFDR_3$ of 92.2 dB.Hz^{2/3} and NF of 20.8 dB. The variance in the NF and SFDR is attributed to the partial suppression of the output when the signal is within the stopband of the filter. The main contribution to the NF is the loss associated with electro-optic (E-O) modulation and optical-electrical (O-E) conversion via photodetection. We note that the results presented here include a wideband RF amplifier at the input of the E-O modulator with 24 dB gain and NF of 3 dB, which was included to overcome the excessive loss from two PICs used in this proof of principle demonstration.

5.4 Filtering of interfering signals

The true benefit of MWP-based filtering over traditional electronic methods becomes apparent in scenarios that highlight the tunability and reconfigurability offered by MWP systems. In this section, we use an MWP notch filter to enhance the performance of an RF receiver in two key demonstrations that require frequency agile hardware. First, we filter an unwanted interfering signal up to 20 GHz using the simplified schematic and bypassing the mixer, shown in Fig. 5.6 (a).

Wideband operation up to 20 GHz is shown in Fig. 5.7. Initially we demonstrate a single notch response, tuning the RF notch frequency from 5-20 GHz with attenuation from 34-49 dB in Fig. 5.7 (a-d). Subsequently, we add a second notch in Fig. 5.7 (e-f) by using one additional MRR and one VCO to generate a second Brillouin pump tone. Filtering of multiple tones spanning across different RF frequency bands, from 7.6 - 18.6 GHz is

demonstrated with notch depth up to 34 dB. The variation in notch depth is primarily attributed to the relative frequency fluctuations between the Brillouin and MRR resonances, rather than fluctuations in pump power.

To demonstrate the versatile functionality of this filter, we subsequently emulate a traditional RF receiver architecture. We use a mixer to downconvert the RF output of the PD before generating the received constellation at the IF frequency. We then configure the experimental setup as per the simplified schematic in Fig. 5.6 (a) and pass the notch filter output through the RF mixer. In this demonstration, we use the MWP notch filter to perform image rejection of unwanted signals which can reduce the quality, or completely disrupt a wireless communications channel. We use a QPSK signal at 10 GHz with a baud rate of 50 MHz as the desired data signal and perform downconversion using an RF mixer and local oscillator (LO) with frequency of 9 GHz. We also inject an unwanted RF signal at 8.03 GHz that is mapped onto the QPSK signal in the IF output of the mixer.

Fig. 5.8 (a) shows that the MWP filter reduces the amplitude of the unwanted signal by 38.7 dB. The IF output in Fig. 5.8 (b) shows that the interfering signal overlaps with the signal of interest upon downconversion. The MWP filter reduces the amplitude of the unwanted signal, however, the measured rejection is limited by the power spectral density of the QPSK data. We use the EVM of the constellation as an indicator of the received RF data quality to evaluate the effectiveness of the MWP notch filter in removing unwanted interference. We stress that this measurement is independent of the RF noise figure analysis outlined in Fig. 5.5. An increase to the EVM of the received constellation is observed as we increase the power of the interfering signal in Fig. 5.8 (c). However, the EVM is reduced to 14.9% from 59.8% when the interfering signal is attenuated by the MWP notch filter. The distortion of the constellations in Fig. 5.8 (d) is attributed to

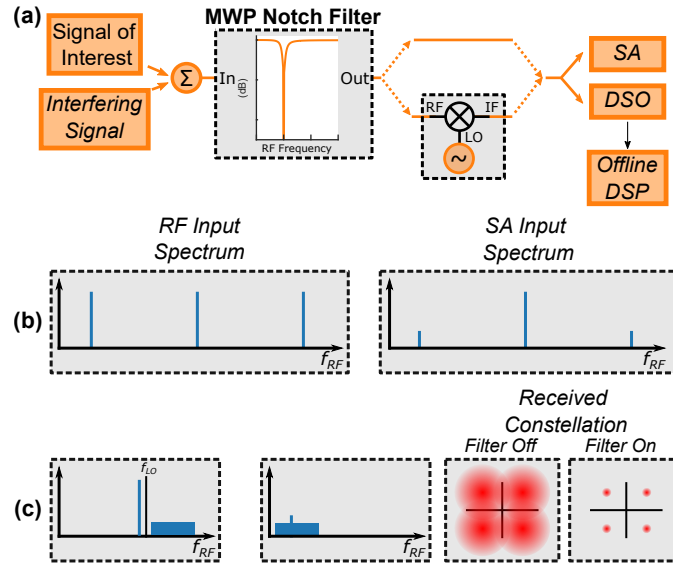


Figure 5.6: (a) Block diagram of system setup for MWP notch filtering demonstrations. When demonstrating the functionality of the filter, the upper RF mixer is bypassed. When demonstrating the filter for image rejection, the RF mixer is utilised. (b) Example RF spectrum before and after filter is utilised. (c) Example RF input for image rejection experiment, example IF output for image rejection experiment. The constellation diagrams show the expected improvement to the constellation when the MWP filter is applied.

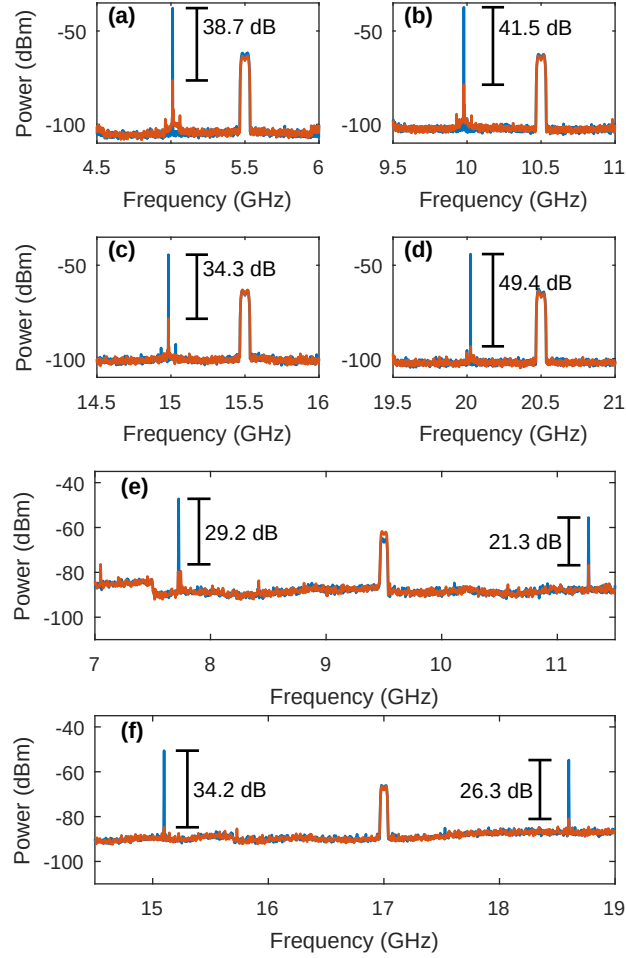


Figure 5.7: Demonstration of RF filtering over multiple bands. The blue trace corresponds to the spectrum when the filter is not applied. The orange trace corresponds to the spectrum when the filter is applied. (a-d) Single notch applied at 5, 10, 15 and 20 GHz. (C, X, Ku, K bands). (e-f) Two notches covering multiple frequency bands.

the increasing power of the CW interfering tone [407]. When the MWP filter attenuates the unwanted interferer, the received constellation maintains four distinct points and low EVM for all recorded interfering signal powers, as shown in Fig. 5.8 (e).

5.5 Performance analysis and discussion

Here, we provide analysis [298], [299], [408] of MWP link performance optimisation to consider the road map for further improvements to Brillouin-based MWP filters. Reflections of the strong Brillouin pump co-propagate with the probe, which increases the RF noise power spectral density (PSD) due to the increased photocurrent. The extent to which the noise PSD increases is determined by the power of the reflected pump. Additionally, the reflected pump may create spurious RF tones or fall within an MRR resonance, which

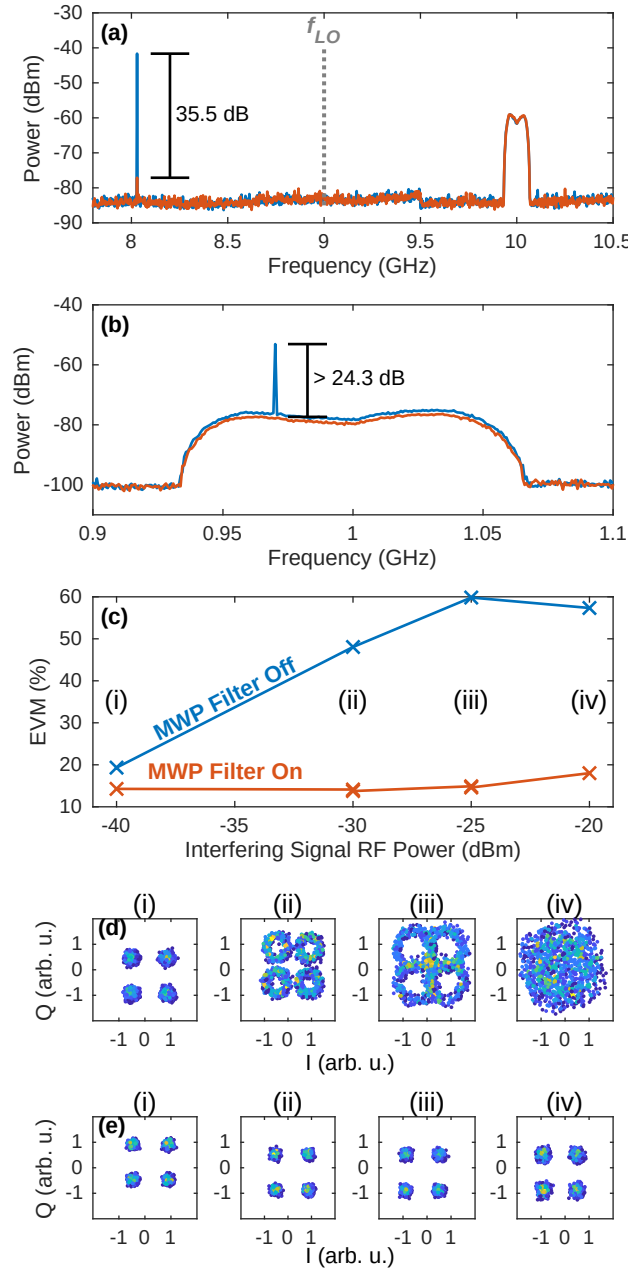


Figure 5.8: Data with notch in RF + IQ for image rejection (a) Measured RF spectrum with and without the MWP filter enabled. (b) IF output of the mixer displaying the interfering signal overlaps with the desired modulated carrier. (c) Comparison of the EVM vs interferer signal power for the case with the MWP filter on and off. (d) Constellation diagram without MWP filter enabled. (e) Constellation diagram with MWP filter enabled.

can degrade filter performance. Therefore, reflections of the Brillouin pump must be minimised to avoid an increase in the RF noise PSD. Accordingly, careful design of low-loss and index-matched transitions between the Brillouin medium (for example, chalcogenide glass) and other optical materials used for processing [181], [220], [409] or modulation [410] are required. In this section, we find the Brillouin pump power and optical reflectiv-

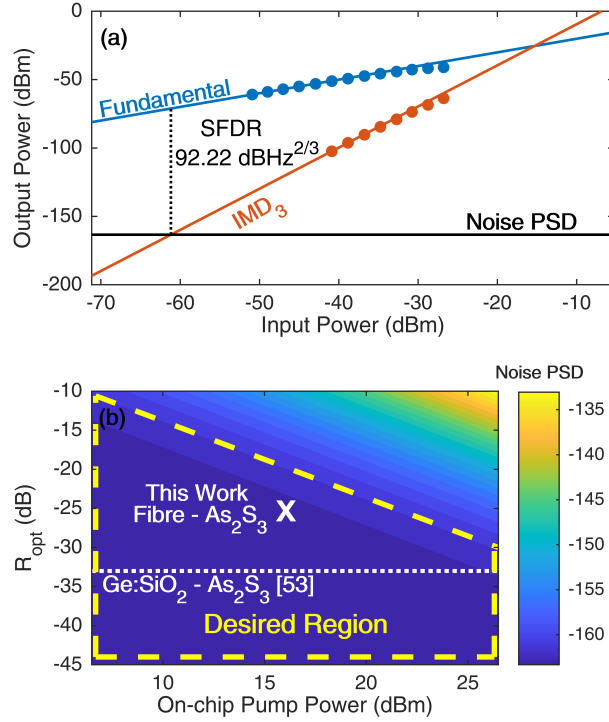


Figure 5.9: (a) Measured two-tone test at 13 GHz. (b) RF noise power spectral density for different on-chip Brillouin pump powers and optical reflection coefficients. The desired region corresponds to the case when the noise PSD contribution from the reflected pump is less than the noise PSD with the pump off.

ity necessary to ensure the RF noise PSD added from the reflected pump is less than the noise PSD contribution from the probe. This design rule is independent of the Brillouin medium material, hence can be applied to any transition within any Brillouin-based MWP system [51], [176], [227], [322], [342], [389].

We demonstrate the importance of maintaining low RF noise PSD in Fig. 5.9 (a), where we plot the two-tone test results at 13 GHz from the filter response presented in Fig. 5.5. The SFDR₃ corresponds to the output signal power when the intermodulation product power, IMD₃, is equal to the RF noise PSD. Any increase to the noise PSD will reduce the minimum detectable signal and SFDR₃ of the filter.

For a given pump power at the interface to the Brillouin medium, P_{pump} , the reflected power incident on the PD is given by:

$$P'_{pump} = \frac{P_{pump} \times R_{opt}}{L}, \quad (5.4)$$

where L is the optical loss between the PD and the facet of the Brillouin device and R_{opt} is the optical power reflection coefficient of the pump. The reflected pump increases the overall optical power at the PD, which in turn increases the generated photocurrent:

$$I_{PD} = r_{pd} \times \left(P_{probe} + \frac{P_{pump} \times R_{opt}}{L} \right), \quad (5.5)$$

where r_{PD} is the PD responsivity (A/W). This leads to a significant increase to the RF noise PSD, p_n (W/Hz), given by [72]:

$$p_n = (1 + g)kT + \frac{1}{4} \left(2qr_{PD}I_{PD}R_L + 10^{\frac{RIN}{10}} I_{PD}^2 R_L \right), \quad (5.6)$$

where g is the linear passband gain of the MWP filter, k is the Boltzmann constant, T is the temperature (K), q is the electronic charge (C), R_L is the RF load resistance (50 Ω) and RIN is the equivalent laser relative intensity noise (dB/Hz) that included the noise contribution from optical amplifiers for the probe and reflected pump.

By analysing Eq. (5.5) and Eq. (5.6), it is evident that the noise introduced from the reflected pump in a Brillouin-based MWP filter can be reduced in two ways, either by reducing the overall required pump power P_{pump} , or by reducing the reflection experienced by the pump, R_{opt} . Here L is considered as a constant insertion loss as indicated in Table 5.1. The reduction of the RF noise PSD with low R_{opt} and P_{pump} is depicted in Fig. 5.9 (b), where we plot the RF noise PSD for different R_{opt} and on-chip Brillouin pump powers. The desired region indicates where the noise contribution from the reflected pump is less than the noise from the signal laser, giving a noise PSD increase of less than 3 dB. In this analysis we used parameters similar to their experimental values, as shown in Table 5.1.

An advantage of this filtering scheme is that much less Brillouin pump power is needed to achieve similar performance to previously reported filter architectures. The fully tunable ring resonator used in this chapter is set such that only 3 dB of Brillouin gain per RF notch resonance is used. This requires significantly less pump power than other Brillouin-based filter architectures, which often use over 10 dB of Brillouin gain per RF resonance [45], [405]. This reduced pump power requirement results in a decreased RF noise PSD contribution from reflections of the weaker Brillouin pump. As seen in Fig. 5.9 (b), the pump power required for two RF notches, marked with a cross, ensures that the reflected pump does not significantly degrade the RF noise PSD. However, if additional notches are required, more Brillouin pump power must be used which can increase the RF noise PSD by more than 10 dB. Therefore, a reduction to the R_{opt} is required to enable operation with higher Brillouin pump power.

The R_{opt} can be further reduced by adiabatically transitioning the optical mode through the interface between multiple materials. A promising material platform has emerged from a recent demonstration through hybrid integration of As₂S₃ and germanosilicate waveguides [341], where the measured reflection coefficient was reduced to -33 dB and

Table 5.1: Parameters used to simulate pump reflection coefficient.

Parameter	Symbol	Value
Modulator half-wave voltage at 10 GHz	$V_{\pi_{RF}}$	2.9 V
Combined equivalent laser RIN	RIN	-135 dB/Hz
Loss from chip-facet to PD	L	6 dB
Probe power at PD	P_{probe}	640 μ W
PD responsivity	r_{PD}	0.5 A/W
RF Load Resistance	R_L	50 Ω

is shown by the dashed white line in Fig. 5.9 (b). Additionally, low facet coupling losses of 1.5 dB were observed, which makes this germanosilicate-As₂S₃ hardware platform a promising candidate to enable high-performance Brillouin-based MWP processing with large pump powers and greatly reduced RF noise PSD penalty. Heterogeneous integration of germanosilicate MRRs and As₂S₃ is an attractive candidate for realising the hybrid chip proposed in Fig. 5.1. The low propagation loss of germanosilicate waveguides [411] has the potential to enable high quality factor (Q) MRRs for narrow 3-dB RF bandwidth notches. Additionally, the vertically tapered transitions to As₂S₃ waveguides shown in [341] offer efficient, low reflection interfaces which enables large Brillouin gain and low-loss MRRs on the same chip.

Future implementations of this device require performance improvements to achieve operation without the need for an RF amplifier. Reduction of the losses between the MRRs and As₂S₃ Brillouin waveguides can be achieved through adiabatic optical mode transitions using hybrid photonic integration techniques [412]. However, this is out of the scope of this proof of principle demonstration. Therefore, we included an RF amplifier with gain of 24 dB to offset excess coupling losses from this experimental setup. We note that the performance of this MWP filter with two PICs and an RF amplifier is comparable to typical performance of MWP filters with a single PIC without an RF amplifier [301]. Reflections and losses can be reduced through the use of engineered transitions which optimise taper profiles and mode matching between materials, which will remove the need for RF amplification.

5.6 Conclusion

We demonstrated an MWP filter with multiple notches formed by cascading micro-ring resonators and on-chip Brillouin scattering. We achieved frequency agile filtering with operation up to 20 GHz, low passband losses of 8 dB whilst maintaining RF extinction ratios of greater than 40 dB. We demonstrated analog filtering functionality by showing simultaneous attenuation of unwanted interfering tones over 20 GHz. Subsequently, we used the filter for image rejection during down-conversion in a coherent communication receiver, improving the EVM of the received constellation from 59.8% to 14.9%. Finally, we provided analysis to guide the future development of Brillouin-based on-chip hybrid photonic platforms to understand and mitigate effects from undesirable reflections that arise from discontinuous waveguide interfaces. These results highlight the suitability of this filter scheme to enhance wideband and high-performance RF receiver systems for the next-generation RF systems.

Chapter 6

Low-RF-loss and large-rejection reconfigurable Brillouin-based MWP bandpass filter

We present a high-performance radio frequency (RF) photonic bandpass filter enabled by combining on-chip Brillouin scattering with a suppressed carrier phase modulation scheme. We achieve a low RF loss of 5 dB and a large stopband rejection of more than 40 dB, which represents a significant improvement of 20 dB to the RF passband gain and 31 dB to the RF rejection ratio over traditional modulation schemes under the same optical power consumption. We further demonstrate filter reconfigurability including multiple passbands, wide frequency (1-20 GHz) and bandwidth tunability (30-350 MHz) without compromising the RF performance.

This chapter is based on the works:

- **Matthew Garrett**, Yang Liu, Pan Ma, Duk-Yong Choi, Stephen J. Madden, and Benjamin J. Eggleton, "Low-RF-loss and large-rejection reconfigurable Brillouin-based RF photonic bandpass filter," *Optics Letters*, 2020, **45**, 3705-3708.
- M. Garrett, Y. Liu, D. Choi, P. Ma, S. J. Madden, and B. J. Eggleton, "High-Link-Gain and Deep-Rejection Chip-based Microwave Photonic Bandpass Filter using Moderate Brillouin Gain," in Asia Communications and Photonics (ACP) 2019, paper M3E.3.
- M. Garrett, Y. Liu, P. Ma, D. Choi, S. J. Madden and B. J. Eggleton, "Fully Reconfigurable Chip-Based Brillouin Microwave Photonic Multi-Passband Filter with High RF Link Gain," 2020 Conference on Lasers and Electro-Optics (CLEO), 2020, pp. 1-2.

6.1 Introduction

The increasing demand for available bandwidth in the radio frequency (RF) spectrum places great importance on frequency agile systems, such as software defined radio receivers [413], [414]. In these modern RF systems, reconfigurable and widely tunable RF bandpass filters that select signals of interest from adjacent unwanted signals, which exhibit low RF loss, large out of band attenuation and reduced size, weight, power and cost (SWaP-C) characteristics, are essential in emerging applications [415].

Integrated microwave photonics (IMWP) delivers impressive frequency agility, large instantaneous processing bandwidth and reduced form factor [5], [25] with the potential to overcome the limitations faced by traditional electronic filters [6], [7], [27]. Stimulated Brillouin scattering (SBS) induced in integrated photonic circuits [51] is an attractive candidate for a microwave photonic filter (MWPF), owing to the nature of signal amplification and the intrinsic narrow 30 MHz bandwidth of the SBS resonance that enables unmatched fine frequency selectivity. Additionally, the parametric generation of Brillouin gain resonances enables MWPFs with multiple passbands and reconfigurable bandwidths by tailoring the optical pump source [10], [46], [416].

MWPFs using SBS have been implemented with a number of modulation schemes and configurations [147], [148], [417]. Recently, there has been significant interest in applying these advances to integrated platforms that demonstrate high-performance filtering functionality with reduced form factor [43], [396]. MWPFs using conventional intensity modulation (IM) and single sideband (SSB) modulation schemes benefit from a tunable modulator bias that can produce low RF losses [292]. However, prohibitively large pump power is required to generate sufficiently large RF rejection. Conversely, the use of phase modulation (PM) can deliver excellent RF rejection due to the destructive interference of RF signals in filter stopbands [418], but it is challenging to optimize the RF performance due to the lack of bias control, especially when compared to tunable modulation schemes [45], [150].

In this chapter, we demonstrate a chip-based Brillouin microwave photonic bandpass filter that is capable of providing reconfigurable, high-resolution RF filtering with large rejection ratio and low RF loss. This is made possible by merging the advantageous features of PM, IM and SSB modulation formats using a novel suppressed carrier phase modulation (SCPM) scheme [419], [420] that, when combined with Brillouin sideband amplification, achieves a low RF loss (5 dB) and significant RF rejection (> 40 dB). Reconfigurability of the filter is demonstrated by dynamically programming the number of passbands. Each passband has independently tunable frequency (1-20 GHz) and 3-dB bandwidth (30-350 MHz). These results represent an important step towards meeting the demanding performance requirements of reconfigurable integrated MWPFs [165].

6.2 Principle and experimental setup

The operating principle of the high-performance MWPF filter enabled by on-chip Brillouin processing and the SCPM scheme is outlined in Fig. 6.1. In the case of MWPFs using SSB modulation, the RF filter response is determined by the beat between the carrier and processed sideband, resulting in a direct mapping of Brillouin gain to RF rejection ratio, as shown in Fig. 6.1 (a). However, if dual-sideband intensity modulation is used,

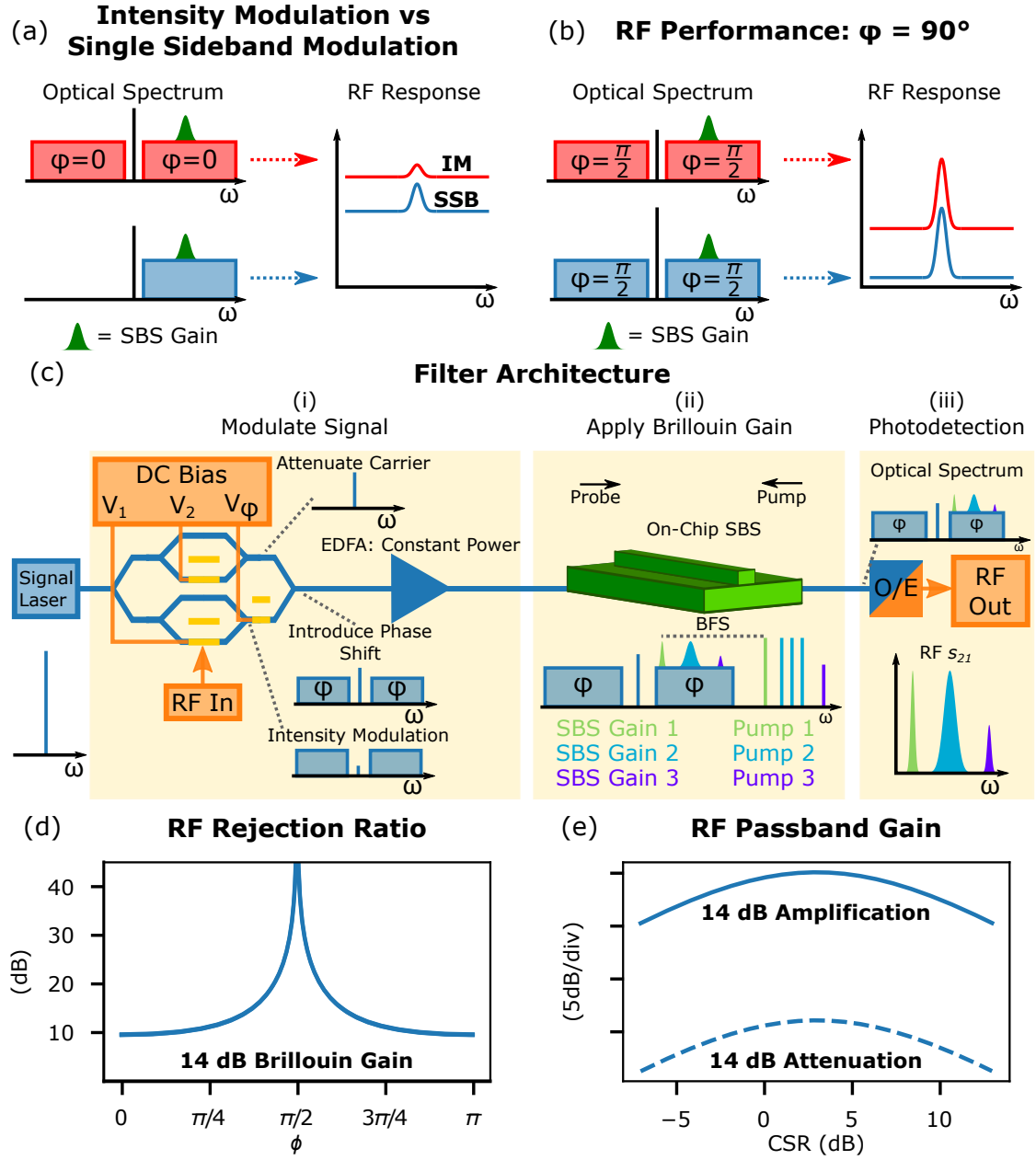


Figure 6.1: (a) MWPf performance comparison using IM and SSB modulation. (b) MWPf performance with $\frac{\pi}{2}$ phase offset between optical carrier and sidebands as the CSR changes. (c) Operating principle of the high-performance Brillouin-based microwave photonic filter. (d) Calculated RF rejection ratio (e) Calculated RF passband gain for MWPfs using sideband amplification and sideband attenuation.

the additional sideband increases the RF link gain, although the filter rejection ratio is reduced. Fig. 6.1 (b) shows that by introducing a phase shift between the carrier and sidebands, destructive RF interference can occur outside the Brillouin resonance and the RF rejection ratio is much greater than the applied Brillouin gain [10]. Additionally, Fig. 6.1 (b) shows that the RF gain of the filter can be increased by adjusting the carrier strength, while maintaining constant signal laser power. The SCPM scheme corresponds to the condition where the maximum RF rejection and minimum passband losses are simultane-

ously obtained.

Fig. 6.1 (c) outlines the architecture of the high-performance Brillouin RF photonic bandpass filter using the SCPM scheme and moderate optical power. (i) The SCPM scheme is generated by using a dual parallel Mach-Zehnder modulator (DPMZM) to modulate the signal laser. The lower arm is driven by an RF input and biased near a null point by V_1 . The upper arm is only driven by a DC bias voltage, V_2 , which controls the optical carrier strength. A phase shift, ϕ , is applied to the lower arm and is controlled by V_ϕ . The signal at the DPMZM output has an optical carrier with amplitude controlled by V_2 and sidebands with a phase offset from the carrier controlled by V_ϕ . Constant signal power is achieved by an erbium-doped fibre amplifier (EDFA) operating in constant power mode. (ii) An optical pump source, consisting of multiple lasers, applies Brillouin gain to the upper sideband of the signal laser. This allows for multiple passbands to be dynamically added without altering the filter architecture, or operating principle. (iii) After photodetection, a bandpass filter response is formed in the electrical domain.

The operating principle illustrated in Fig. 6.1 can be further understood by analysing the RF output of the filter. The electric field of the processed signal at the photodetector is given by

$$E(t) = E_1 \cos((\omega_c - \omega_{RF})t + \phi) + E_2 \cos(\omega_c t) + E_1 \sqrt{G_{SBS}} \cos((\omega_c + \omega_{RF})t + \phi), \quad (6.1)$$

where E_1 and E_2 are the amplitudes of the electric field of the sideband and carrier, controlled by V_1 and V_2 , the phase shift between carrier and sidebands, ϕ , controlled by V_ϕ and G_{SBS} is the linear Brillouin power gain. The RF photocurrent power at ω_{RF} is

$$P = r_{pd}^2 P_c P_{sb} (G_{SBS} + 2 \sqrt{G_{SBS}} \cos 2\phi + 1) \times R_L, \quad (6.2)$$

where r_{PD} is the photodetector responsivity, P_c and P_{SB} are the powers of the carrier and sideband and R_L is the load resistance (50 Ohms).

Eq. (6.2) is used to calculate the theoretical RF performance of the filter as a function of the phase and amplitude relationships between the optical carrier and sidebands. The RF output is calculated using a constant signal laser power and 14 dB Brillouin gain to match the following experimental conditions. Fig. 6.1 (d) shows that maximum RF rejection ratio is achieved when $\phi = \frac{\pi}{2}$. It should be noted that, in reality, the maximum RF rejection ratio is limited by several factors including slight deviations from the optimal phase offset, system noise sources and higher order harmonics from the modulator. Fig. 6.1 (e) compares the RF passband gain for filters that utilise selective sideband amplification and selective sideband attenuation when $\phi = \frac{\pi}{2}$ as a function of carrier-sideband power ratio (CSR), defined as $CSR = 10 \log_{10}(\frac{P_c}{P_{sb}})$. MWPFs that utilise selective sideband amplification (SBS gain) experience a larger RF gain than their counterparts that rely on attenuation (SBS loss and passive resonators). However, in both cases, the passband gain is maximised when $CSR = 3$ dB.

Fig. 6.2 outlines the experimental setup. The signal laser (RIO Grande) is modulated (JDSU 11 GHz DPMZM, EOSPACE 20 GHz PM) by the RF output of a vector network analyser (VNA) (Keysight N5224A), which is amplified by EDFA1 (Amonics) in constant power mode to give an on-chip signal power of 15 dBm. A laser array amplified by EDFA2 (Amonics), is used to generate the optical pump which creates the on-chip Brillouin gain used to amplify the signal. The pump counterpropagates with the modulated signal on an integrated As_2S_3 shallow-etched rib waveguide. The waveguide is 12

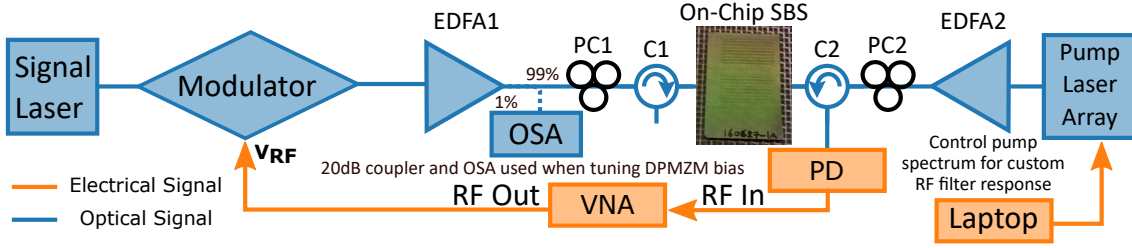


Figure 6.2: Experimental Setup. EDFA: erbium-doped fibre amplifier, OSA: optical spectrum analyser, PC: polarisation controller, C - circulator, PD: photodetector, VNA: vector network analyser.

cm long and $2.4 \mu\text{m}$ wide with fibre to fibre coupling loss of approximately 10 dB. The pump is generated by an array of independently tunable lasers, which enables customised tuning of the central frequency and bandwidth of the RF passbands using low frequency electronics. It should be noted that a single tone sweeping pump is not preferred for chip-based Brillouin filters, as the nanosecond flying time of the sweeping pump is too short to produce large Brillouin gain. A circulator (C2) is used to separate the counterpropagating pump and signal lasers. The filter response is measured by the VNA at the output of the photodetector (PD) (APIC 20 GHz). It is worth noting that these results are obtained without RF amplifiers.

6.3 Experimental results

6.3.1 Modulation format characterisation

To access the required modulation format, we examined RF filter responses with various CSRs and phase offsets, as shown in Fig. 6.3. The variation between the RF responses in Fig. 6.3 (a-c) shows that low RF passband losses and maximum RF rejection can be achieved by tuning the amplitude and phase of the modulated signal. Each RF response corresponds to a specific bias point labelled in Fig. 6.3 (d and e) which show the RF rejection ratio and passband gain, respectively, as a function of the V_2 and V_ϕ biases. In this analysis, we fixed V_1 , the bias of the lower arm of the DPMZM and reduced the signal and pump power to qualitatively observe filter performance.

The effect of phase tuning is evident by comparing Fig. 6.3 (a-b), which have small RF rejection ratio, to Fig. 6.3 (c), which has a much larger RF rejection. The RF rejection ratio increase is attributed to ϕ approaching $\frac{\pi}{2}$, as shown in Fig. 6.1 (d). However, a slight reduction in the passband gain is observed in Fig. 6.3 (c), because this phase offset has a component of destructively interfering photocurrents, as opposed to the $\phi = 0$ case. The effect of tuning the CSR is evident by comparing Fig. 6.3 (a) with Fig. 6.3 (b). The former exhibits lower gain than the latter, which can be attributed to the increase in the CSR; this is observed by comparing the optical spectrum in each case.

From Eq. 6.2, the maximum RF rejection can be achieved with $\phi = \frac{\pi}{2}$ and is annotated by the dashed curve in Fig. 6.3 (d). The maximum RF rejection ratio is maintained at all points along this path (due to the optimal phase tuning). If ϕ depended only on V_ϕ , this path would be a vertical line. However, when a DPMZM is used to implement the SCPM scheme, each bias voltage affects the value of ϕ . Therefore V_2 and V_ϕ must both change

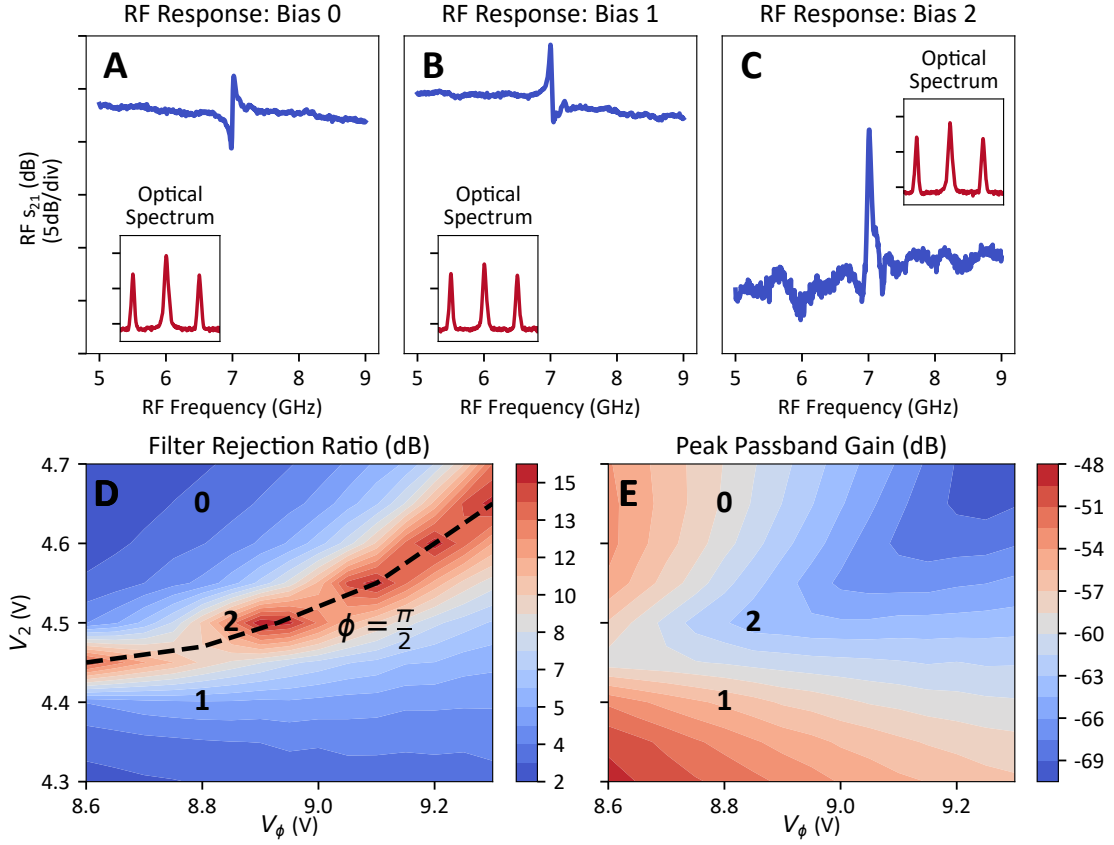


Figure 6.3: (a-c) RF response measured at the specific labelled bias point (**0**, **1**, **2**) below. (d-e) RF rejection ratio and the RF peak passband gain as the applied bias voltage to the dual parallel Mach-Zehnder modulator is adjusted.

to maintain a $\frac{\pi}{2}$ phase offset. The main benefit of this scheme is made clear by observing that the passband gain, shown in Fig. 6.3 (e), increases (by reducing the CSR) while the maximum RF rejection ratio is maintained. This enables the RF gain of the MWPF to be increased by tuning a bias voltage, which is impossible to do using a phase modulator.

It should be noted that Fig. 6.3 (a-b) exhibit a notch in the RF response of the filter adjacent to the passband peak. In these cases, ϕ deviates from its optimal value so perfect destructive interference in the RF stopband is not achieved. However, applying Brillouin gain will induce a phase and amplitude change that can result in perfect RF cancellation at a frequency detuned from central Brillouin resonance, causing this notch.

6.3.2 RF response tunability

Fig. 6.4 demonstrates wideband RF operation while sustaining high RF gain and significant RF rejection ratios; these results are achieved while constrained to moderate optical signal power of approximately 3 dBm at the photodetector and 14 dB of Brillouin gain. We demonstrate filtering with passband frequencies ranging from 1-20 GHz, low passband losses of 5 dB and significant filter rejection of 40 dB at 1.1 GHz offset from the passband (maximum rejection of 45 dB with 11 GHz passband frequency). If this filter was implemented using SSB modulation, the applied 14 dB of Brillouin gain would trans-

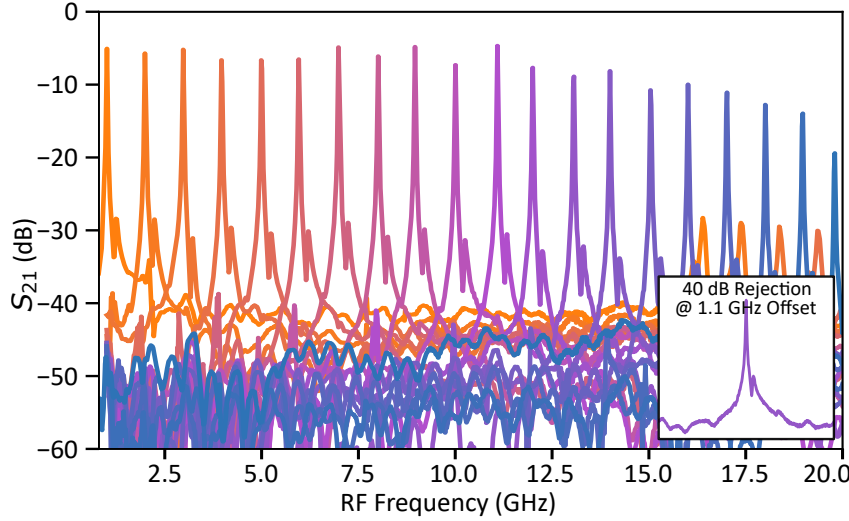


Figure 6.4: A single passband with low loss and large RF rejection ratio tuned from 1-20 GHz.

late to 14 dB RF rejection ratio [292]. Hence, we improve the RF rejection by up to 31 dB.

We note that the roll-off of the filter gain with increasing frequency in Fig. 6.4 is attributed to exceeding the maximum operating frequency of the modulator. This problem can be overcome by selecting electronic components with higher bandwidths, which highlights that the primary benefit of Brillouin MWPFs is that the RF operating range is only limited by the bandwidth of the electronic components used.

The secondary peak observed in each spectrum that is approximately 200 MHz offset from the main peak is due to an additional acoustic mode in the integrated As_2S_3 waveguide that induces a separate SBS resonance. The feature 16 GHz above the passband frequency in responses with central frequency up to 4 GHz is caused by Brillouin loss [421].

We measured a noise figure of 35 dB with an overall gain of -7 dB and an spurious-free dynamic range (SFDR_3) of $96 \text{ dB}\cdot\text{Hz}^{2/3}$. The noise figure and SFDR_3 are currently limited by Brillouin noise and optical amplifier noise. The effect of the Brillouin noise on RF filter performance can be further reduced by operating in the depletion regime at the expense of reduced filter linearity [147].

In general, optimising MWP filter performance to reduce the added noise penalty from the Brillouin gain resonance is critical to further improve the RF link metrics of Brillouin-based MWP filters. The underlying modulation scheme plays a critical role in determining the RF link performance of Brillouin-based filters [292]. Additionally, several approaches that mitigate the effect of added Brillouin noise in the RF passband have been achieved using filter schemes where the Brillouin gain response is within the stop band of the RF filter. One approach uses overlapping MRR and Brillouin resonances, as was the case in Refs. [39], [45] and in Chapter 5. Another approach uses two Brillouin loss resonances has been demonstrated to form a passband response within a given spectral window [148], [422], where the passband does therefore not contain added Brillouin noise.

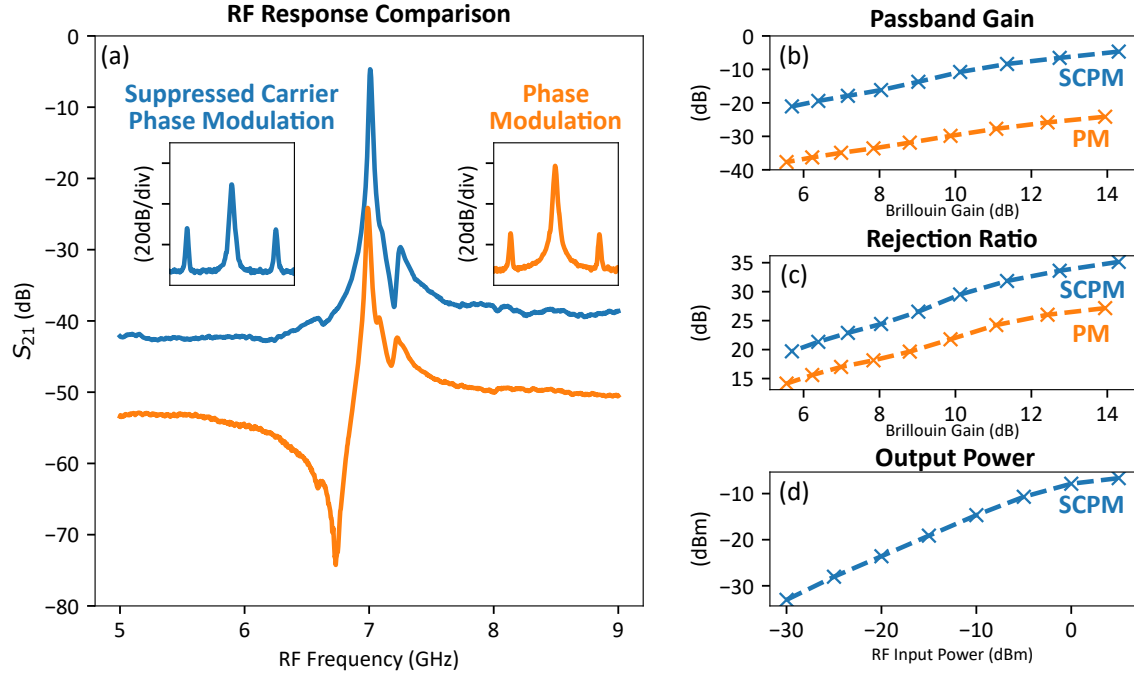


Figure 6.5: (a) RF passband response for SCPM and PM schemes. The inset shows the optical spectrum for each modulation scheme. (b-c) RF passband gain and rejection ratio as a function of Brillouin gain. (d) RF output power as a function of RF input power.

6.3.3 RF link performance comparison

Fig. 6.5 outlines the performance improvement of the SCPM format over the PM scheme with the same Brillouin gain and signal laser power. Fig. 6.5 (a) shows the improvement in the overall RF response enabled by the SCPM scheme. By comparing the optical spectrum of the PM and SCPM cases at the output of the modulator, shown in the inset of Fig. 6.5 (a), the reduction in CSR increases the overall gain of the MWPF by 20 dB. Additionally, the RF response of the filter using PM is asymmetric, as observed in Fig. 6.3 (a-b). This is corrected for in the SCPM scheme by finely tuning the modulator bias voltages.

Fig. 6.5 (b-c) compares the RF passband gain and rejection ratio of the SCPM and PM schemes as a function of applied SBS gain. Compared to the PM scheme, the SCPM scheme increases the RF gain by 20 dB and RF rejection by 7 dB. This increase is associated with the ability to precisely tune the DPMZM bias to provide the optimal phase offset and carrier suppression. Applying Brillouin gain here creates an RF rejection of 35 dB, which can be further improved to more than 40 dB, as shown in Fig. 6.4. Fig. 6.5 (d) shows that the filter operates in a linear regime of RF input power versus RF output power, before compressing at an input power of 0 dBm due to Brillouin pump depletion.

6.3.4 Multiple passbands with broadened bandwidth

Reconfigurability with multiple passbands is demonstrated in Fig. 6.6 (a). The frequency of each passband is independently tuned by modifying the frequency of the corresponding laser used to generate the Brillouin resonance. The trade-off between RF performance

and the optical pump power used to generate the passband necessitated an increase to the overall pump power. However, a slight reduction in the RF passband gain and rejection ratio is observed. The passband amplitude decays as the RF frequency exceeds the modulator bandwidth and the passband amplitudes vary because the power of each pump source can be independently tuned. Fig. 6.6 (b) adds further functionality by extending the 3-dB bandwidth of an RF passband up to 350 MHz. The passband response is widened by modulating the pump with multiple tones such that each sideband induces a distinct Brillouin resonance and the inset provides further details of a broadened passband response. This becomes increasingly challenging as the required pump power increases considerably as the passband broadens. These results were obtained with an on-chip pump power limit of 500 mW.

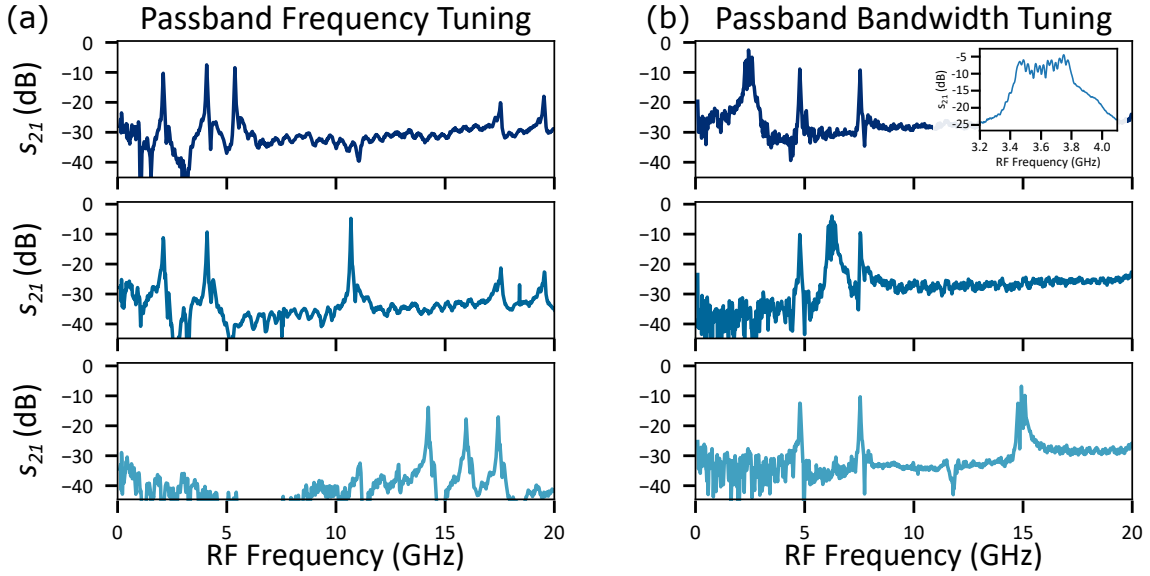


Figure 6.6: (a) Three passbands independently tuned over 1-20 GHz. (b) Three passbands independently tuned over 1-20 GHz and 3-dB bandwidth tuned from 30-350MHz. The inset shows details of a broadened RF passband.

6.3.5 Conclusion

In conclusion, we demonstrated a high-performance, fully reconfigurable chip-based MWPF using on-chip Brillouin scattering and a suppressed carrier phase modulation scheme. The filter simultaneously combines the low RF losses of conventional SSB modulation with the significant RF rejection ratio associated with phase modulation schemes. We experimentally demonstrated that the SCPM scheme provides a 20 dB improvement to the RF gain of a MWPF, compared to the gain achieved using a standard PM scheme[7], [423]. Additionally, the SCPM scheme increases the RF rejection ratio by 31 dB when compared to MWPFs using SSB modulation [46], [292]. This is an important advancement as it allows for low-loss Brillouin-based MWPFs with multiple, independently tunable passbands with large rejection ratios and reduced power consumption.

Chapter 7

Heterogeneously integrated Brillouin and active-silicon photonic MWP notch filter

Microwave photonics (MWP) has unlocked a new paradigm for Radio Frequency (RF) signal processing by harnessing the inherent broadband and tunable nature of photonic components. Despite numerous efforts made to implement integrated MWP filters, a key RF processing functionality, it remains a long-standing challenge to achieve a fully integrated photonic circuit that can merge the megahertz-level spectral resolution required for RF applications with key electro-optic components. Here, we overcome this challenge by introducing a compact 5 mm x 5 mm chip-scale MWP filter with active E-O components, demonstrating 37 MHz spectral resolution. We achieved this device by heterogeneously integrating chalcogenide waveguides, which provide Brillouin gain, in a complementary metal-oxide-semiconductor (CMOS) foundry-manufactured silicon photonic chip containing integrated modulators and photodetectors. This work paves the way towards a new generation of compact, high-resolution RF photonic filters with wideband frequency tunability demanded by future applications, such as air and spaceborne RF communication payloads.

This chapter is based on the works:

- **Matthew Garrett**, Yang Liu, Moritz Merklein, Cong Tinh Bui, Choon Kong Lai, Duk-Yong Choi, Stephen J. Madden, Alvaro Casas-Bedoya, Benjamin J. Eggleton, "Integrated microwave photonic notch filter using a heterogeneously integrated Brillouin and active-silicon photonic circuit", *Nature Communications*, 2023, **14**, 7544.
- M. Garrett, M. Merklein, Y. Liu, C. Bui, C. Lai, D. Choi, S. J. Madden, A. Casas-Bedoya and B. J. Eggleton, "Heterogeneous integration of Brillouin devices with active silicon photonic circuits," 2022 IEEE Photonics Conference (IPC), Vancouver, BC, Canada, 2022, pp. 1-2.
- M. Garrett, Y. Liu, M. Merklein, C. Bui, C. Lai, D. Choi, S. J. Madden, B. J. Eggleton and A. Casas-Bedoya, "Compact RF photonic notch filter based on heterogeneous integration of As_2S_3 with active silicon photonic circuits," 2023 Conference on Lasers and Electro-Optics (CLEO), 2023, pp. 1-2.

7.1 Introduction

Radio-frequency (RF) filters play a critical role in maintaining signal fidelity of RF and microwave systems by eliminating unwanted signals and preserving desired signals of interest [27]. RF filters require fine spectral resolution to operate within crowded RF spectral environments and must also exhibit broadband frequency tunability as operating frequencies of RF systems move toward millimeter wave bands above 30 GHz [25]. At the same time, they must also achieve small size, weight, and power (SWaP) to meet the demands for operation in future RF applications. However, state-of-the-art RF filter technology fails to simultaneously meet these requirements [28]. Common RF filter implementations based on surface acoustic waves (SAW) are compact, but face challenges operating at high frequencies. For example, SAW filters have large losses above 3.5 GHz [67], [424], whereas filters based on bulk acoustic waves (BAW) can operate into the X-band (8-12 GHz), yet they face fabrication challenges to extend into millimeter wave bands [67], [68]. Yttrium iron garnet (YIG) filters are widely tunable, however their form factor is too large for deployment in SWaP-constrained applications [69], [70]. Another approach is to use switched filter banks which offer compact form factors, although their frequency response cannot be continuously tuned [71].

Microwave photonic (MWP) filters overcome the frequency tunability limitations of traditional RF filtering technologies by utilising the broadband tunable and reconfigurable nature of optical components. Moreover, with advances in photonic integration, MWP processors that are fully integrated have the potential to offer the compact footprints required for next-generation microwave systems [25], [27], [30], [165], [172]. Although significant progress on integrated MWP filters has been made, they face difficulties in achieving the fine spectral resolution that is indispensable in RF communication and sensing applications [27]. One of the most common methods to implement on-chip MWP filters is to use integrated optical micro-ring resonators (MRRs), which exhibit compact form factor, as well as reconfigurable amplitude and phase responses [27], [167]. However, the achievable 3-dB bandwidth of MRR-based MWP filters is typically limited to hundreds of MHz due to waveguide losses [30], [172], [425]. Although ultra-high-quality factor (Q) MRRs can be fabricated, they are extremely sensitive to input signal power and temperature fluctuations due to thermo-optic effects [426], [427], which can add challenges in aligning the laser to the MRR resonance. Additionally, the free-spectral range (FSR) of the MRR can limit the frequency operating range of the resultant MWP processor. MWP filters based on optical tapped delay lines offer highly reconfigurable RF responses, yet are limited in spectral resolution to GHz levels as they require complex amplitude and phase control of many optical filter taps [31], [33], [428]. For example, Shu *et al.* demonstrated an MWP filter that generated an optical comb and processed each comb line on a photonic chip, but due to the limited number of optical comb lines, the filter 3-dB bandwidth was limited to 900 MHz [34]. Similar to MRR-based processors, the periodic response of tapped delay line MWP processors can limit their frequency operating range.

Stimulated Brillouin scattering (SBS) is an optical nonlinearity first introduced in 1922 [142] that has since re-gained interest for use in integrated MWP filters with fine spectral resolution following the first demonstration of SBS in a photonic chip in 2011 [164]. SBS is facilitated through a third-order nonlinear acousto-optic interaction [25], [51], [142], [389] and exhibits a narrow intrinsic optical resonant linewidth (10's MHz) due

to the long phonon lifetime. In addition to fine spectral resolution, SBS offers several advantages for MWP processing applications as the Brillouin response does not exhibit an FSR, which does not impose any limitations on the operating range of the resultant MWP system [25], [27]. Furthermore, the thermal stability of the Brillouin frequency shift (BFS), $\approx 1 \text{ MHz}/^\circ\text{C}$ in optical fibres [429], can be much greater than the thermal stability of MRRs, for example, a silicon MRR with Q of 10^7 exhibited a thermal sensitivity of $\approx 12 \text{ GHz}/^\circ\text{C}$ [427]. This highlights the advantages of SBS for MWP processing in terms of tunability and stability.

SBS has been generated in a number of chip-scale platforms [39], [51], [164], [317]–[319], [337] which enable critical chip-scale MWP processing systems, such as high-performance MWP filters [10], [39], [42], [44], [45], [347], [405], phase shifters [11], [333], true time delays [40], [50], frequency converters [371], [430], and high spectral purity microwave sources [22], [373]. Silicon photonic platforms, which have the advantage of being compatible with complementary metal-oxide-semiconductor (CMOS) processes and incorporate an extensive passive and active component library, exhibit large Brillouin gain coefficients [51], which is highly desirable for Brillouin-based MWP platforms. Despite these advantages, silicon nanowires on oxide lack intrinsic acoustic guidance, consequently requiring suspended or under-etched structures and exhibit nonlinear optical losses that limit the achievable on-chip Brillouin gain at optical telecommunication wavelengths. On the other hand, silicon nitride (Si_3N_4) has negligible nonlinear loss, but only offers minuscule SBS gain coefficients [39], [321] due to the modest elasto-optic coefficient that is intrinsic to Si_3N_4 [321]. Conversely, soft As_2S_3 chalcogenide glasses offer large Brillouin gain coefficients [164] and facilitate more than 30 dB of SBS gain in a compact footprint [334], owing to the large elasto-optic coefficient [325] and simultaneous confinement of acoustic and optical modes [164], [326], [334]. However, chalcogenide waveguides can only be used as a *standalone* element in MWP filter systems with key MWP components, such as electro-optic (E-O) modulators and photodetectors, remaining off-chip. Previously, we introduced a novel platform that heterogeneously integrated chalcogenide Brillouin circuits with passive silicon waveguides on a CMOS-compatible SOI platform [176]. However, the passive platform did not take full advantage of integration with silicon photonics because it lacked an E-O interface. To reduce the SWaP of Brillouin-based MWP processors, integrating chalcogenide Brillouin circuits with the required components of an MWP link, including E-O modulators and photodetectors (PDs), is essential [51].

In this chapter, we heterogeneously integrate As_2S_3 Brillouin waveguides in a silicon photonic platform, that includes active E-O modulators and photodetectors, to form the basis of a novel and compact MWP processing platform with fine spectral resolution. We demonstrate an integrated MWP notch filter using on-chip silicon devices and Brillouin gain in As_2S_3 waveguides for fine spectral resolution processing, achieving 51 dB out-of-band rejection with narrow 37 MHz 3-dB bandwidth and tunable notch central frequencies over 15 GHz. Our demonstration paves the way towards a fully integrated Brillouin-based MWP processor with a reduced footprint and advanced performance.

7.2 Heterogeneous integration of Brillouin As₂S₃ waveguides with foundry-manufactured Si photonics

A schematic of the heterogeneously integrated MWP notch filter with fine spectral resolution is exemplified in Figure 7.1 (a). An RF signal consisting of two tones with frequencies f_1 and f_2 is input into the MWP filter. After filtering, one tone is attenuated, as indicated by the spectrally selective notch response. The key components of the filter, including E-O modulator, As₂S₃ Brillouin waveguide and photodetector, are implemented in the single compact 5mm × 5mm silicon photonic chip presented in Fig. 7.1 (b). We exemplify the future feasibility of achieving a fully-integrated MWP system on a chip by incorporating a design that facilitates lasers through the micro-transfer-print technique [206]. However, the lasers were not populated in the current demonstration. We designed a novel E-O modulator, Fig. 7.1 (c), that transduces the RF signal into the optical domain and enables flexible control of the optical carrier-sideband ratio (CSR) and sideband phases. The chalcogenide Brillouin waveguide that facilitates fine-spectral resolution processing is highlighted in Fig. 7.1 (d).

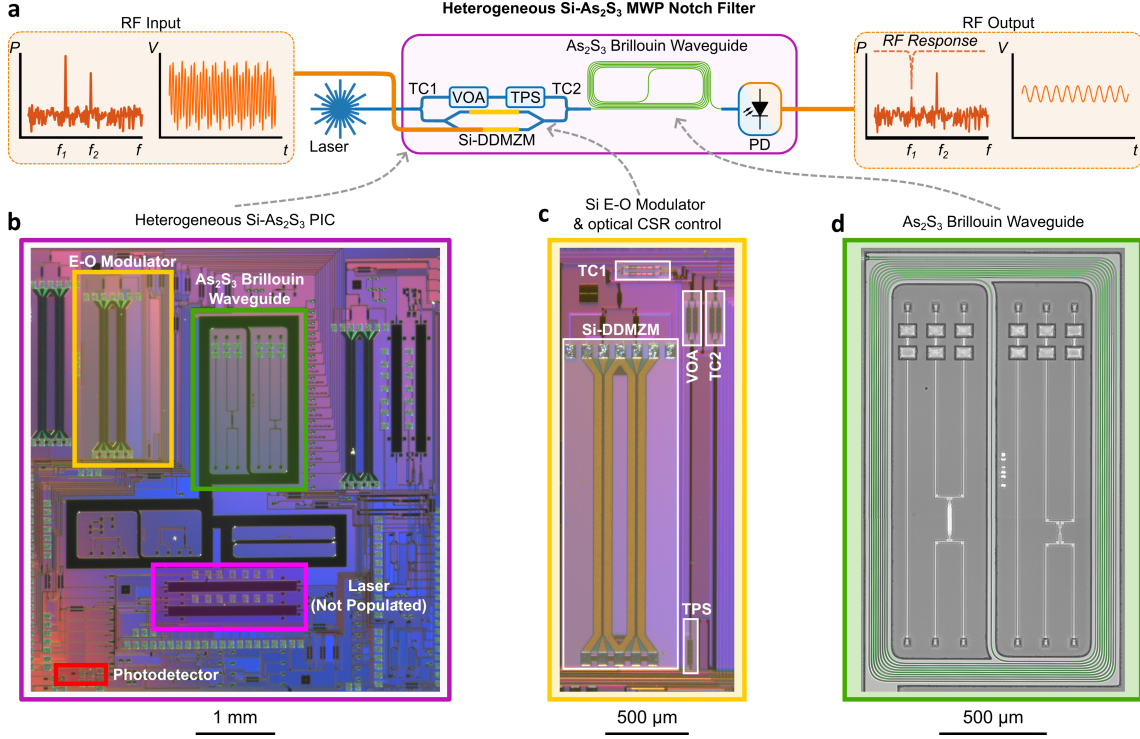


Figure 7.1: Heterogeneous integration of As_2S_3 waveguides and silicon components for an on-chip microwave photonic filter system. (a) High-level schematic of the integrated Si- As_2S_3 MWP notch filter. An RF input signal is modulated onto an optical carrier by a modulator on the Si- As_2S_3 chip. The modulated optical signal is then processed and undergoes photodetection. The notch filter response is visualised in the RF output spectrum where the spectral component at f_1 is attenuated, whereas the signal at f_2 is not. (b) A picture of the Si- As_2S_3 PIC containing (c) a silicon E-O modulator and circuits used for optical carrier control and (d) False-coloured optical microscope image of As_2S_3 spiral waveguides that are used for Brillouin processing. PIC: Photonic integrated circuit, E-O: Electro-optic, CSR: Carrier-sideband-ratio, TC: Tunable coupler, TPS: Tunable phase shifter, VOA: Variable optical attenuator.

7.3 Heterogeneously integrated As_2S_3 waveguides

7.3.1 Fabrication and Design

Figure 7.2 presents the details of the As_2S_3 waveguides that, when integrated into the active silicon photonic platform using custom post-processing steps outlined below, enhance the standard silicon photonic chips with nonlinear Brillouin processing capabilities. The As_2S_3 Brillouin waveguides lie in a trench defined by the IMEC process flow, where the back-end-of-line (BEOL) stack—including metal layers that would otherwise prevent access to silicon waveguides—was removed. Before the As_2S_3 film can be deposited, a 2 μm silica layer covering the silicon waveguides was removed using hydrofluoric acid (HF) etching. We controlled the HF etch depth by performing multiple etching steps and inspected the etch depth after each step using EBL. We used a buffered HF 15:1 (volume ratio of 40% NH_4F in water to 49% HF in water) solution. The etch was performed iteratively. First, we pattern a photoresist (ma-N1420) etch mask and expose the sample

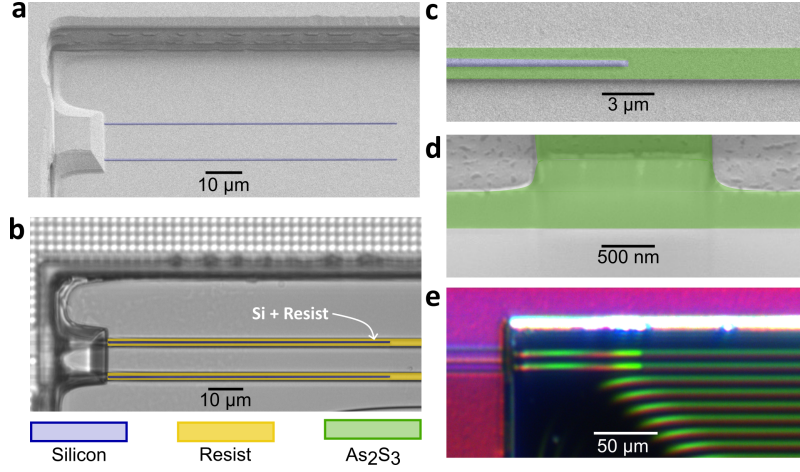


Figure 7.2: As_2S_3 waveguide fabrication and Si- As_2S_3 interface simulation. (a) False-coloured scanning electron microscope (SEM) image of the silicon waveguide taper after the SiO_2 cladding was removed using HF etching. (b) False-coloured optical microscope image of the silicon taper after photoresist is patterned. (c) False-coloured SEM image of the Si waveguide tip and As_2S_3 waveguide. (d) False-coloured SEM image of the As_2S_3 rib waveguide cross-section. (e) True-coloured photograph of the Si- As_2S_3 taper region after fabrication is complete.

to BHF for a duration of approximately 15 minutes, which corresponds to an etch depth of approximately 2000 nm. This was monitored using DekTak profilometer tool. The etch depth was inspected using a scanning electron microscope (SEM), and subsequent etch steps are performed by exposing the sample in BHF for short durations until the required etch depth is achieved. Despite the careful etching process, the HF etching rate was not uniform across the entire trench region. This could lead to localised variations of the As_2S_3 propagation losses and localised scattering. Fig. 7.2 (a) depicts the exposed Si waveguide and trench sidewall after HF etching.

Subsequently, thermal evaporation was used to deposit a 680 nm thick As_2S_3 film and 1800 nm wide waveguides were patterned using electron beam lithography (EBL). The Si- As_2S_3 taper after resist patterning is shown in Fig. 7.2 (b). Finally, a rib waveguide was formed by etching the film to a depth of 340 nm (50%) using inductively coupled plasma-reactive ion etching (RIE). The silicon waveguide tip at the end of the taper is shown in Fig. 7.2 (c) and a cross-section of the As_2S_3 rib waveguide is shown in Fig. 7.2 (d). A photograph of the waveguides after fabrication is shown in Fig. 7.2 (e). The spiral waveguides were designed with Euler bends using an effective radius of 100 μm and minimal bend losses were observed (See Section 7.3.2). Although the As_2S_3 post-processing steps used in this work are similar to the passive Si- As_2S_3 platform reported by Morrison *et al.* [176], the advantage of integrating As_2S_3 into this active platform comes at the expense of increased fabrication complexity associated with the HF etching step.

7.3.2 As_2S_3 propagation losses

The overall As_2S_3 waveguide losses (L_Σ) can be separated into two contributions. The first contribution is from the Si- As_2S_3 interface, and the second is from the propagation losses in the As_2S_3 . We can therefore express the overall loss as

$$L_{\Sigma} = 2L_{\text{Si-As}_2\text{S}_3} + \alpha l_{\text{As}_2\text{S}_3} \quad (7.1)$$

where $L_{\text{Si-As}_2\text{S}_3}$ is the loss per Si- As_2S_3 interface (dB), α is the propagation loss in the As_2S_3 waveguide (dB/cm) and $l_{\text{As}_2\text{S}_3}$ is the length of the As_2S_3 waveguide (cm).

We measured the overall optical losses directly by measuring the input and output optical power of the Si- As_2S_3 chip. A fibre array was used to couple light onto the Si- As_2S_3 chip. A pair of shunt silicon grating couplers allowed for the grating coupler losses to be measured and compensated for in the loss calculation of the As_2S_3 waveguides. Each grating coupler had a loss of ≈ 3.5 dB.

The waveguide propagation losses were extracted using a LUNA OVA5000 optical domain frequency reflectometer (OFDR) and fitting a line to the measured reflected signal. The experimental setup for the measurements of the overall optical losses and the propagation losses are shown in Fig 7.3 (a). A high-level schematic of the heterogeneous Si- As_2S_3 chip with silicon grating couplers is shown in Fig. 7.3 (b) and a schematic of the standalone As_2S_3 chip is shown in Fig. 7.3 (c).

We measured As_2S_3 propagation losses of 2.0 dB/cm in the Brillouin spiral waveguides using optical frequency domain reflectometry, which is significantly higher than the loss of 0.7 dB/cm measured by Morrison *et al.* on a passive Si- As_2S_3 platform [176]. To investigate the source of increased propagation losses, we fabricated standalone As_2S_3 chips (50% etched and fully-etched) with the same waveguide dimensions and spiral layout as fabricated on the Si- As_2S_3 chip. One can note that the only difference in the fabrication process between the standalone As_2S_3 chips and the Si- As_2S_3 chips was that the standalone chips were not exposed to HF before As_2S_3 deposition. The standalone chip exhibited propagation losses of only 0.4 dB/cm for both etch depths. This is much lower than the 2.0 dB/cm for the 50% etched and 3.2 dB/cm for the fully-etched waveguides on the Si- As_2S_3 platform. By considering the round-trip path taken by the signal in the OFDR measurement, we experimentally determined the As_2S_3 waveguides have a group-index of approximately 2.6. This is consistent with the simulated group index of 2.59 that we obtained using Lumerical MODE solver.

Scattering from waveguide sidewalls, as well as surface roughness induced from the

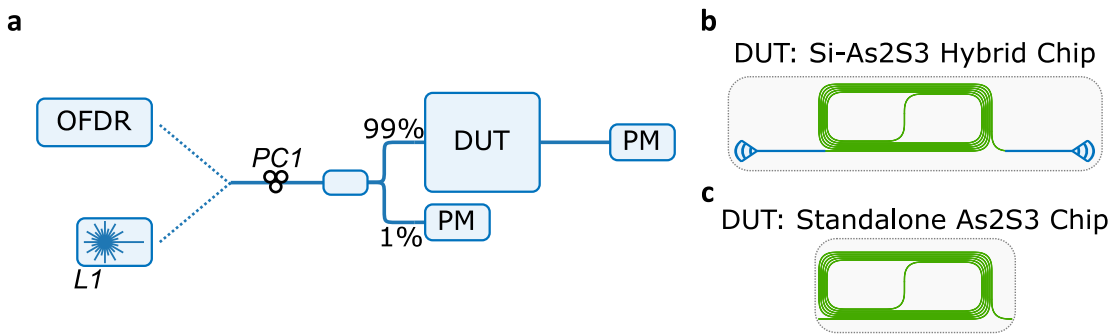


Figure 7.3: Experimental setup for loss measurements of the As_2S_3 waveguides. (a) Setup for OFDR and total insertion loss measurements. (b) Simplified chip schematic of the Si- As_2S_3 chip. (c) Simplified schematic of standalone As_2S_3 chip. OFDR: Optical frequency domain reflectometry, L: Laser PC: Polarisation controller, PM: Power meter DUT: Device under test.

HF etching step could have increased the losses of the As_2S_3 waveguides on the heterogeneous platform. This is evidenced by the reduced As_2S_3 propagation losses on the standalone As_2S_3 platform that was not exposed to HF before chalcogenide deposition, as well as the increased losses of the fully-etched waveguides compared to the 50% etched waveguides on the Si- As_2S_3 platform. We measured the root mean squared (RMS) roughness of the SiO_2 substrate on the heterogeneous platform before and after HF etching using atomic force microscopy (AFM). The RMS roughness increased from 0.3 nm to 3.2 nm after HF etching. The increased SiO_2 roughness, in turn, affects the surface roughness of the As_2S_3 film that is deposited in the Si- As_2S_3 platform. This can lead to increased sidewall roughness once the waveguides undergo RIE etching and can explain why the rib waveguide exhibits significantly reduced propagation losses compared to the fully-etched waveguide on the Si- As_2S_3 platform.

Comparison to losses on a monolithic As_2S_3 platform

Fig. 7.4 shows the measured OFDR traces taken from the Si- As_2S_3 platform. The peak in reflection at $\Delta t = 0$ is due to the index mismatch between the fibre array and the silicon grating coupler. The section from $0 < \Delta t < 0.15$ ns is the reflection from the silicon waveguide. The region corresponding to $0.15 < \Delta t < 0.82$ ns is a reflection from the As_2S_3 waveguide. The reflection from the second Si- As_2S_3 interface is seen at $\Delta t = 0.82$ ns, and the reflection from the second grating coupler can be seen at $\Delta t = 0.9$ ns.

The blue trace in Fig. 7.4 corresponds to the measured signal in the fully-etched As_2S_3 waveguide. The solid line of fit indicates a propagation loss of 3.25 dB/cm. The orange trace represents the signal from the 50% etched rib waveguide. The dashed line of fit indicates a propagation loss of 2.01 dB/cm. The peaks in the rib waveguide measurement are due to localised scattering as the optical signal propagates through the As_2S_3 waveguide. This scattering is likely due to the contamination that occurred during post-processing. Furthermore, the difference in the reflected signal magnitude for the rib waveguide after the first Si- As_2S_3 transition at $\Delta t \approx 0.15$ indicates that the rib waveguide design results in a lower back-reflection from the Si- As_2S_3 interface, as a result of scattering into the As_2S_3 slab, as shown in Fig. 7.8.

We fabricated a standalone As_2S_3 chip with identical waveguide cross-sectional geometry and spiral designs used in the Si- As_2S_3 platform. This chip was fabricated using the same process flow as the heterogeneous Si- As_2S_3 platform, with one exception being that the standalone Si- As_2S_3 chip was not exposed to HF before As_2S_3 deposition. The OFDR results from the standalone As_2S_3 chip are presented in Fig. 7.5. The measured propagation loss for the fully-etched (yellow trace) and 50% etch (purple trace) waveguides were 0.46 dB/cm and 0.40 dB/cm, respectively. These values are much lower than we observed on the Si- As_2S_3 chip and are consistent with previously measured standalone As_2S_3 waveguide losses that are widely reported in the literature, ranging from 0.26-0.5 dB/cm [270], [334]. The propagation losses measured on the As_2S_3 test chip are also consistent with the losses measured using the cut-back method of approximately 0.35 dB/cm. These losses are consistent with the losses extracted using the OFDR technique.

Furthermore, the lack of peaks in the 50% etched waveguide on the standalone chip indicates that there is minimal reflection or mode conversion in the waveguide bends. This confirms that the peaks observed in the 50% etched waveguide in the Si- As_2S_3 chip

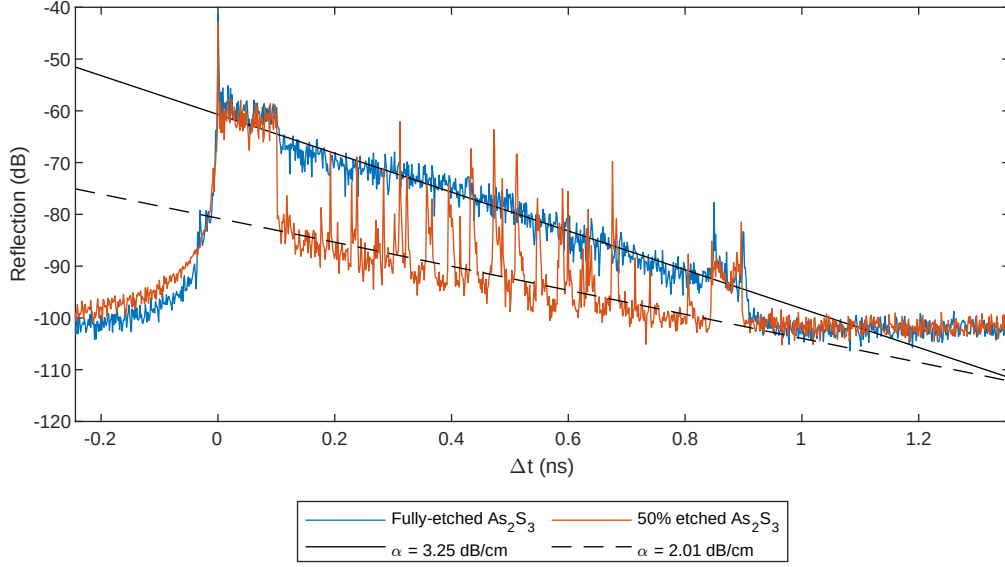


Figure 7.4: OFDR measurement made on on the Si- As_2S_3 platform. The fully-etched (blue) and 50% etched (orange) waveguide losses are shown. The propagation losses are extracted from the linear fits.

are caused by post-processing and not by the waveguide bend design.

Because the propagation losses on the standalone chip are much lower than measured on the Si- As_2S_3 chip, as summarised in Fig. 7.6, we conclude that the source of increased losses is due to the post-processing steps used to fabricate the As_2S_3 waveguides and are not fundamental to the waveguide or spiral design itself. The average propagation loss for variants of fully-etched and rib waveguides on the Si- As_2S_3 and standalone chips are annotated in Fig. 7.6.

Comparing the losses on the Si- As_2S_3 and standalone As_2S_3 chips, we find that there are multiple sources that could contribute to the increased As_2S_3 propagation losses on the Si- As_2S_3 platform. One source of increased loss includes surface roughness induced in the SiO_2 substrate from the HF etching step. We measured the SiO_2 substrate roughness on which the As_2S_3 film was deposited using atomic force microscopy (AFM). We measured a root mean square (RMS) roughness of 0.3 nm before the HF etch step and an RMS roughness of 3.2 nm after HF etching. This could explain why the measured propagation losses for both waveguide designs were higher on the Si- As_2S_3 chip, which was exposed to HF, compared to the overall lower propagation losses on the standalone As_2S_3 platform, which was not exposed to HF. Moreover, the disproportionately larger losses of the fully-etched waveguides on the Si- As_2S_3 chip suggest that the sidewall roughness of the As_2S_3 waveguides on the Si- As_2S_3 platform is higher than the standalone As_2S_3 platform. This can be explained by the rough SiO_2 substrate on the Si- As_2S_3 platform causing roughness on the As_2S_3 film after deposition, which is translated to increased sidewall roughness after etching.

Further investigation is required to fully characterise the sources and relative contributions of the As_2S_3 losses. However, exploring back-side release methods, as demonstrated by Menezo *et al.* [212] and Mai *et al.* [211], could provide a path to reducing As_2S_3 waveguides losses on this active Si- As_2S_3 platform.

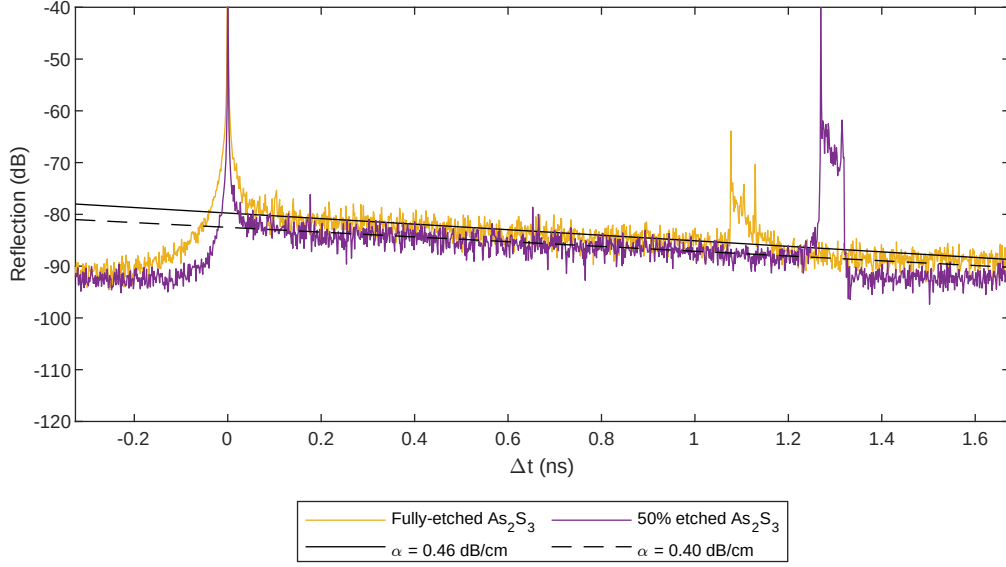


Figure 7.5: OFDR measurement made on the standalone As_2S_3 platform. The fully-etched (yellow) and 50% etched (purple) waveguides losses are shown. The propagation losses are extracted from the linear fits.

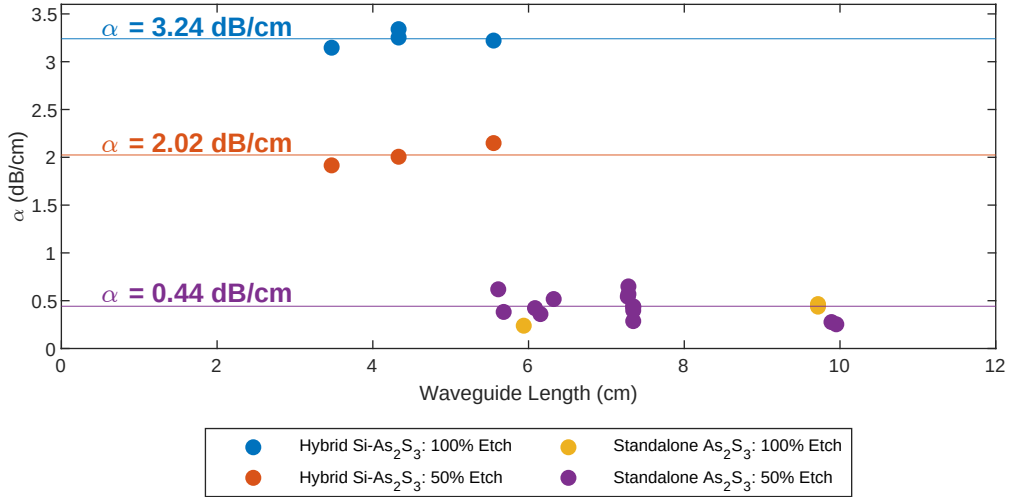


Figure 7.6: Summary of measured As_2S_3 propagation losses. The measured propagation losses of As_2S_3 waveguides on (blue) fully-etched As_2S_3 on the Si- As_2S_3 platform, (orange) 50% etched As_2S_3 on the Si- As_2S_3 platform, (yellow) fully-etched waveguides on the standalone As_2S_3 platform and (purple) 50% etched waveguides on a standalone As_2S_3 platform. Circles represent the measured propagation loss of each individual waveguide. Lines represent the average propagation loss for each combination of device and As_2S_3 etch.

7.3.3 Si- As_2S_3 Interface Losses

Simulations

To adiabatically transition the optical signal into the As_2S_3 waveguide, we taper the width of the silicon waveguide from 450 nm to a minimum tip width of 150 nm whilst maintaining 220 nm height. The TE_{00} mode of the silicon waveguide is input to the taper and the

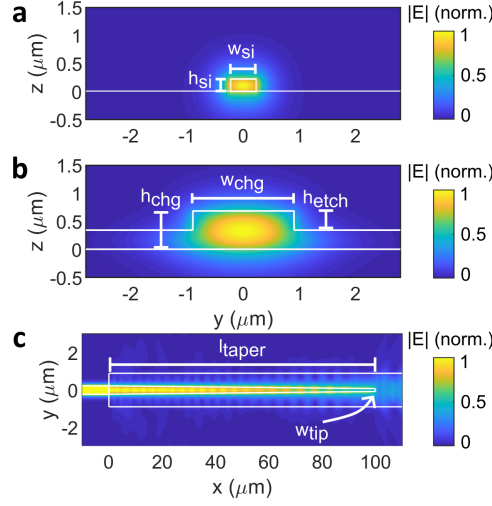


Figure 7.7: (a) Simulated electric field magnitude ($|E|$) of the TE_{00} mode profile for the silicon waveguide. (b) Simulated $|E|$ of the TE_{00} mode profile for the As_2S_3 rib waveguide. The mode profiles were simulated using Lumerical Finite Difference Eigenmode (FDE) solver. (c) Simulated $|E|$ through the Si taper into the As_2S_3 waveguide, simulated using Lumerical Eigenmode Expansion (EME) solver. The labelled critical dimensions have values $w_{\text{Si}} = 450$ nm, $h_{\text{Si}} = 220$ nm, $l_{\text{taper}} = 100$ μm , $w_{\text{tip}} = 150$ nm, $h_{\text{chg}} = 680$ nm, $h_{\text{etch}} = 340$ nm, $w_{\text{chg}} = 1800$ nm.

output of the taper is the TE_{00} mode of the As_2S_3 rib waveguide, which is 1800 nm wide, 680 nm tall with a 340 nm etch. The electric field of the TE_{00} mode of the Si and As_2S_3 waveguides are plotted in Fig. 7.7 (a-b), respectively.

We used Lumerical eigenmode expansion (EME) to simulate the performance of the Si- As_2S_3 interface, which indicated a loss of 0.39 dB for the 50% etch rib waveguide design. The results of the simulation for the rib waveguide are shown in Fig. 7.8. Fig. 7.8 (a-d) shows the TE_{00} mode as the silicon waveguide width changes. Fig. 7.8 (e-f) shows the propagation of the electric field through the taper in the horizontal and vertical planes through the center of the silicon waveguide, which correspond to $z = 110$ nm and $y = 0$ planes, respectively. As the silicon width is reduced, the propagating field transitions from the silicon waveguide to the As_2S_3 waveguide. Scattering into the As_2S_3 slab is observed at the Si- As_2S_3 interface ($x = 0$ μm) and the silicon waveguide tip ($x = 100$ μm). The scattering can be reduced by using a fully-etched As_2S_3 waveguide to increase the confinement of the optical mode. However, the effect that the scattering has on insertion loss is minimal, as simulations indicate that using a fully-etched As_2S_3 waveguide reduces the loss per Si- As_2S_3 interface by only ≈ 0.1 dB. Additionally, we simulated the relationship between the lateral offset, Δy , between As_2S_3 and silicon waveguides in Fig. 7.9. Although we observe significant sensitivity in the Si- As_2S_3 interface losses due to abrupt changes in the HF etch depth, which cause abrupt changes in the silicon waveguide, the simulations indicate that the taper loss is robust to As_2S_3 misalignment as the loss increases by less than 0.05 dB with a 50 nm offset. Although the performance of the taper may be improved by reducing the Si tip width, we selected a tip width to 150 μm to ensure the design was compatible with available foundry processes.

We present the corresponding simulation results for the fully-etched Si- As_2S_3 taper in Fig. 7.10. Although the fully-etched design gives slightly lower interface loss, the

optical propagation losses for the fully-etched design are much higher than the rib design, as shown in Fig. 7.6. The higher propagation losses of the fully-etched design gives a net disadvantage considering overall optical losses and effective length for generating Brillouin gain. Based on the factors discussed above, we used the rib waveguide design presented here to minimise total insertion loss and maximise the effective length of the Brillouin medium.

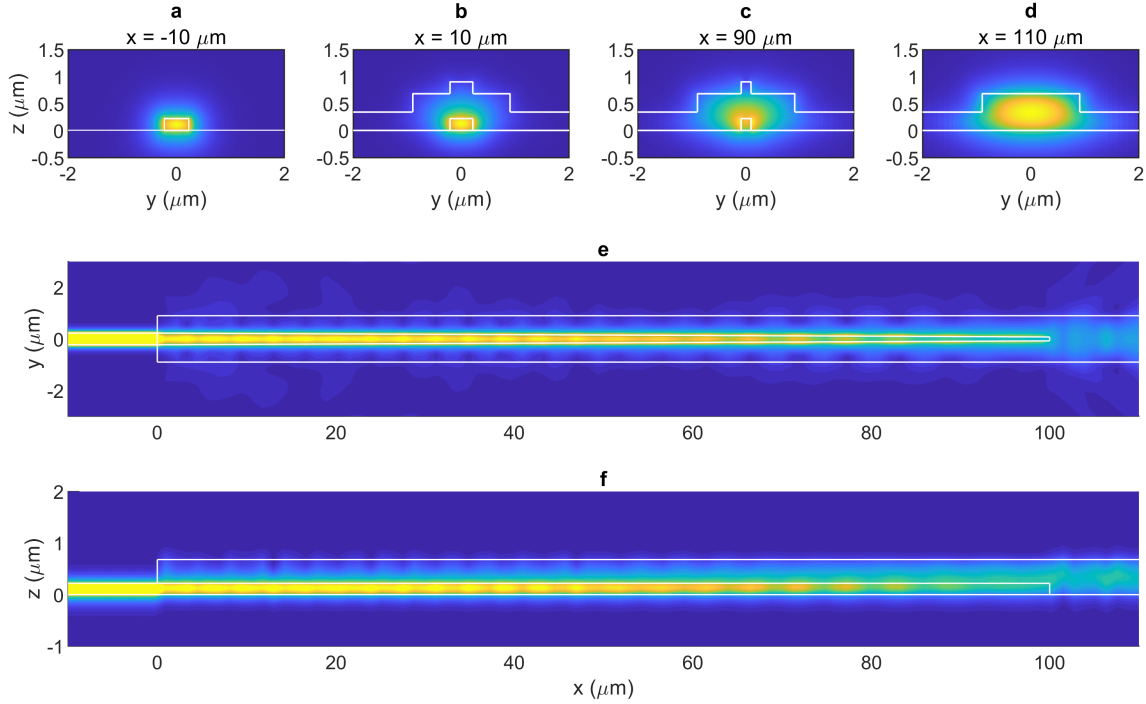


Figure 7.8: Simulation results summary for the Si- As_2S_3 taper with 50% etched As_2S_3 waveguides. (a-d) TE_{00} modes for several cross-sections through the taper. (e) Propagation of the optical electric field in the plane $z = 110 \text{ nm}$. (f) Propagation of the optical electric field in the plane $y = 0$.

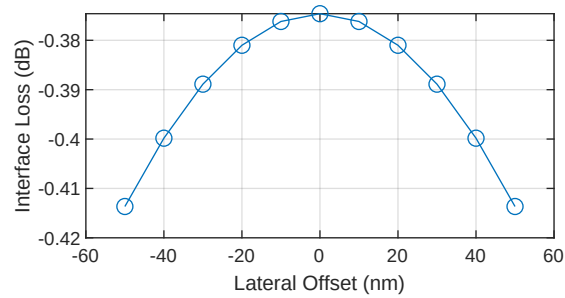


Figure 7.9: Simulation results summary for the Si- As_2S_3 taper as a function of As_2S_3 lateral offset.

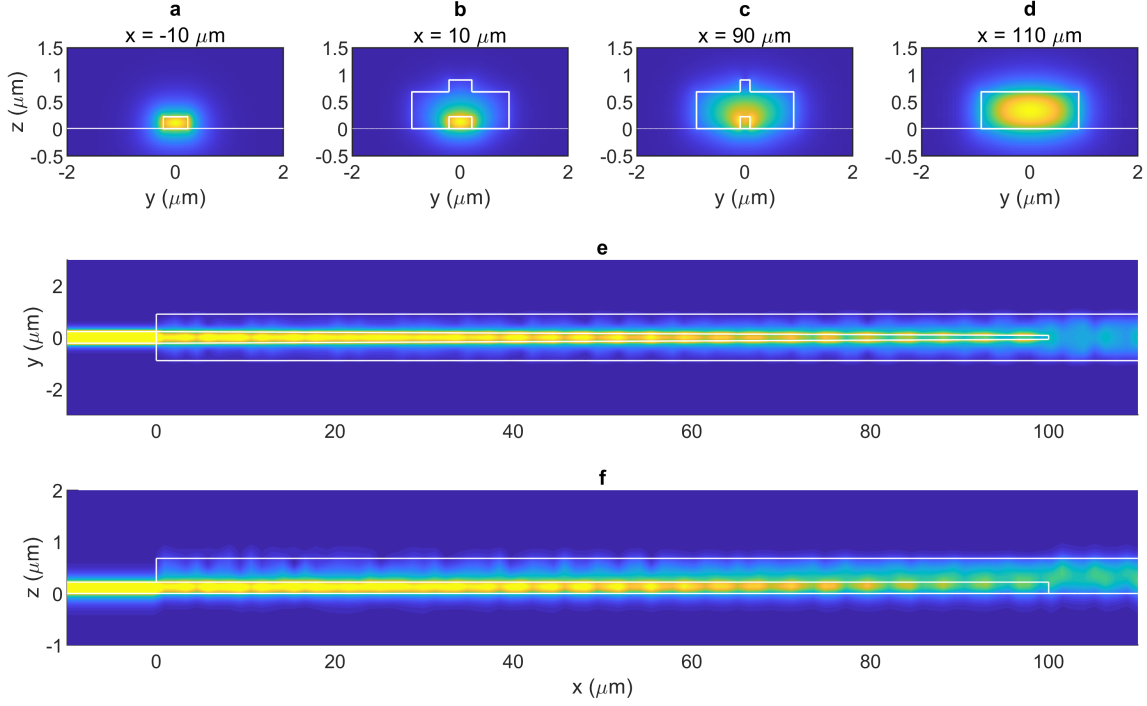


Figure 7.10: Simulation results summary for the Si- As_2S_3 taper with fully-etched As_2S_3 waveguides. (a-d) TE_{00} modes for several cross-sections through the taper. (e) Propagation of the optical electric field in the plane $z = 110 \text{ nm}$. (f) Propagation of the optical electric field in the plane $y = 0$.

Experimental Results

Modifying eq. 7.1, we can calculate the loss per Si- As_2S_3 interface as

$$L_{\text{Si-As}_2\text{S}_3} = \frac{1}{2} (L_{\Sigma} - \alpha l_{\text{As}_2\text{S}_3}) \quad (7.2)$$

Using the total losses measured through the Si- As_2S_3 platform, as well as the propagation losses from the OFDR measurements and compensating for the measured losses of the silicon grating couplers, we calculate the loss per Si- As_2S_3 taper and plot the results for the As_2S_3 rib waveguide design with 1800 nm width, 680 nm thickness and 340 nm etch depth in Fig. 7.11. We ensured that light was coupled to the TE_{00} mode of the silicon waveguide by adjusting the polarisation of the input light into the chip until the transmitted power was maximised.

The calculated Si- As_2S_3 interface losses have considerable variability between 1.8 and 5.5 dB. If we remove the outlier with the largest loss, we calculate an average loss per Si- As_2S_3 interface of 2.1 dB. These values are significantly higher than the EME simulation we performed and the 0.1 dB reported by Morrison *et al.* [176]. The increased losses and variability in the loss of the Si- As_2S_3 transition could also be accounted for due to imperfections introduced in the HF etching step, as outlined above. A source of variability in the measured Si- As_2S_3 interface originates from the HF etching step. The non-uniformity in the etch depth, discussed above in Section 7.3.2, can create abrupt changes in the silicon waveguide, as well as localised over-etching or under-etching, which increases the loss, reflections and variability of the Si- As_2S_3 interface. Although there appears to be

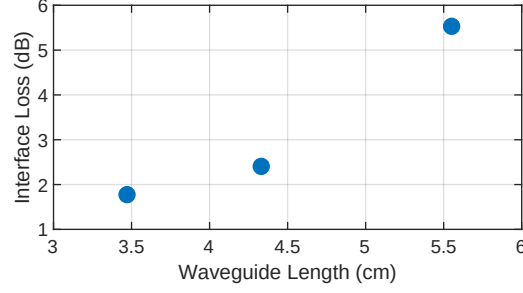


Figure 7.11: Calculated Si- As_2S_3 interface loss. Each data point represents the calculated insertion loss for the length of As_2S_3 waveguide identified.

a trend in the Si- As_2S_3 interface loss as a function of waveguide length in Fig. 7.11, it is independent of the waveguide length and is significantly affected by the HF etch non-uniformity outlined above. On this platform, the Si- As_2S_3 interface loss measurement is also limited by the number of As_2S_3 spiral waveguides that can be implemented on the chip, as the design rules of the multi-project wafer service limited the on-chip trench size. The specific chip we designed in this work was limited to three As_2S_3 spirals. Design improvements that facilitate more As_2S_3 waveguides per Si- As_2S_3 chip, as well as measurements on more chips are required to characterise the losses of the Si- As_2S_3 taper with more accuracy.

7.3.4 As_2S_3 Brillouin Gain Response

The SBS gain, G (linear scale), generated in a waveguide with Brillouin gain factor g_B (m/W), effective optical mode area A_{eff} , effective length L_{eff} (m) and coupled pump power P (W) is given by [325]

$$G = e^{g_B L_{\text{eff}} P / A_{\text{eff}}} \quad (7.3)$$

The effective length relates to the waveguide length L (m) and propagation loss α by

$$L_{\text{eff}} = \frac{1}{\alpha} (1 - e^{-\alpha L}) \quad (7.4)$$

We note that it is common to report the Brillouin gain coefficient (/m/W) of a waveguide, which is given by g_B / A_{eff} .

We measured the magnitude response of the Brillouin resonance in a 5.55 cm As_2S_3 waveguide on the Si- As_2S_3 platform with the setup detailed in Fig. 7.12 (a) using both optical and electronic techniques. For the optical measurement, we used the tracking generator function of an optical spectrum analyser (OSA) to measure the Brillouin response of the As_2S_3 waveguides. The output of the OSA, which constitutes the Brillouin probe, was counter-propagated against the Brillouin pump in the As_2S_3 waveguide and was swept through the Brillouin gain and loss resonances. Off-chip circulators were used to separate the pump and probe, and polarisation controllers were used to ensure the pump and probe were in the TE_{00} polarisation, verified by maximising power transmission through the Si- As_2S_3 chip.

Backwards SBS occurs in As_2S_3 waveguides when an optical signal with large intensity (pump) is counter-propagated against a weaker signal (probe). The pump perturbs the

waveguide through electrostriction and the photoelastic effect [51], forming an acoustic wave. When the pump and probe are separated by the Brillouin frequency shift, Ω_B , a strict phase matching condition is met and pump light is reflected off the acoustic wave. The back-scattered pump light is associated with a Stokes gain resonance and is down-shifted in frequency, coherently amplifying the probe and resonantly enhancing the stimulated process [325]. An anti-Stokes loss resonance also occurs, which is associated with the annihilation of phonons [282]. A Brillouin frequency shift of 7.6 GHz was measured, as annotated in Fig. 7.12 (b). The Brillouin gain and loss resonances are symmetrical around the reflected pump signal.

To characterise the Brillouin gain response of the As_2S_3 waveguides with more precision, we used an RF vector network analyser (VNA), as shown in the simplified schematic presented in Fig. 7.12 (c). We drove the on-chip DDMZM with the output of the VNA. The DDMZM was biased to give single-sideband modulation, ensuring the measured RF transfer function was proportional to the optical transfer function between the modulator and PD. The modulated sideband was swept through the Brillouin gain resonance. We recorded the RF response of the MWP link for multiple pump powers and divided the pump-on response by the pump-off response to determine the SBS gain. A linear slope was then subtracted from the measured response to compensate for roll-off and a Lorentzian curve was fitted to extract the peak Brillouin gain.

The measured SBS response for two different coupled pump powers is shown in Fig. 7.12 (d). The Lorentzian fit for each trace indicates a peak SBS gain of 0.3 dB (blue) and 1 dB (orange). Fig 7.12 (e) shows the peak SBS gain as a function of pump power from which we extracted the SBS gain coefficient. SBS gain saturation is observed for peak SBS gain above ≈ 1 dB, limited by the losses in the As_2S_3 waveguides. Using the measured Si- As_2S_3 interface loss of 5.5 dB, the fit indicates a Brillouin gain coefficient of 703 /m/W. We note that the Brillouin gain coefficient calculation relies on the measured Si- As_2S_3 interface loss that showed great variation across the measured samples.

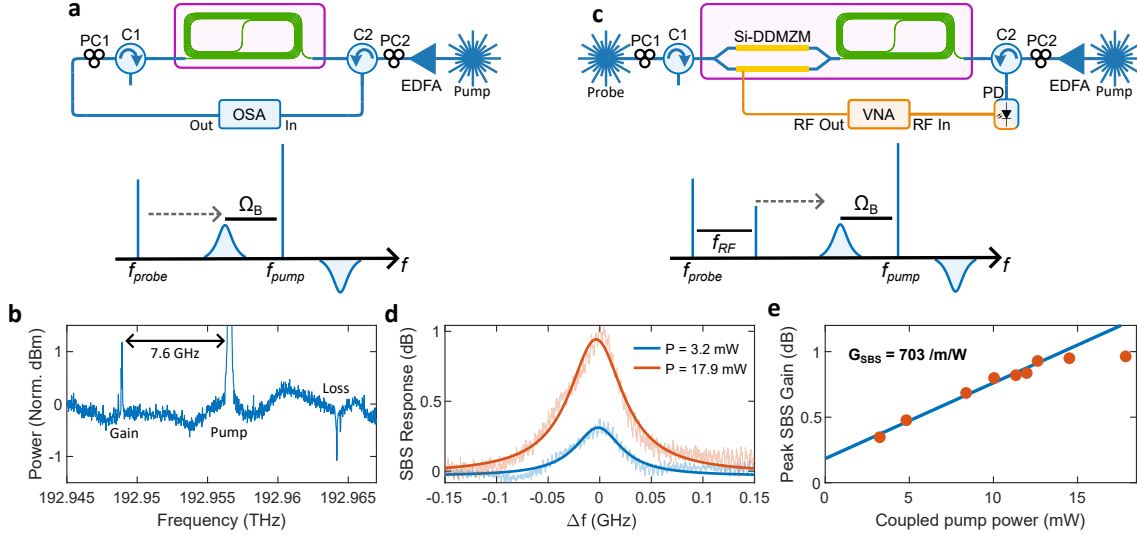


Figure 7.12: Characterisation of the As_2S_3 Brillouin waveguides. (a) Simplified setup for the optical measurement of Brillouin gain. (b) Measured Brillouin gain and loss response. (c) Simplified setup of the electronic measurement of Brillouin gain. (d) The shaded trace represents the measured SBS response in the As_2S_3 waveguides for different coupled pump powers (P). The solid line represents the Lorentzian fit. (e) Measured SBS gain as a function of pump power. OSA: Optical Spectrum Analyser (APEX AP2083A), VNA: Vector Network Analyser (Agilent PNA N5224A), PD: Photodetector, EDFA: Erbium-doped fibre amplifier, DDMZM: Dual-drive Mach-Zehnder modulator. The Brillouin frequency shift is indicated by Ω_B .

7.4 On-chip RF photonic link and novel E-O modulators

We demonstrate the feasibility of the underlying active silicon photonic platform by using the on-chip E-O modulator and PD to construct an on-chip MWP link, as depicted by the schematic in Fig. 7.13 (a). We use a laser around 1550 nm with an on-chip power of only -8.6 dBm. We used RF probes to drive the RF arms of the on-chip DDMZM, using a 180° RF hybrid coupler to induce a phase shift between each arm. The on-chip tunable couplers TC1 and TC2 route the signal to the on-chip DDMZM and bypass upper arm of the larger interferometer. We measure the RF response using a VNA and subtract the measured 0.48 dB/GHz roll-off of the driving RF electronics from the MWP link response. The measurement does not include RF amplification.

The measured response of the MWP link in Fig. 7.13 (b). The on-chip link exhibits a 3-dB bandwidth of approximately 15 GHz, with a measured RF link gain of -52 dB at 1 GHz. We see that the RF link gain (blue) is well above the crosstalk that is measured between the RF probes which drive the modulator and PD (orange).

We implemented a novel on-chip E-O modulator that facilitates control of the amplitude and phase of the optical carrier and sidebands. This allows for the implementation of an MWP filter that is switchable between a bandpass and notch response, while also improving the RF link performance [405], [419]. Although this modulation scheme has been implemented using a bulk lithium niobate-based dual-parallel Mach-Zehnder modulator (DPMZM) in the past [405], [419], its photonic integration here shows a greatly reduced footprint and improved bias stability with an on-chip area of only approximately

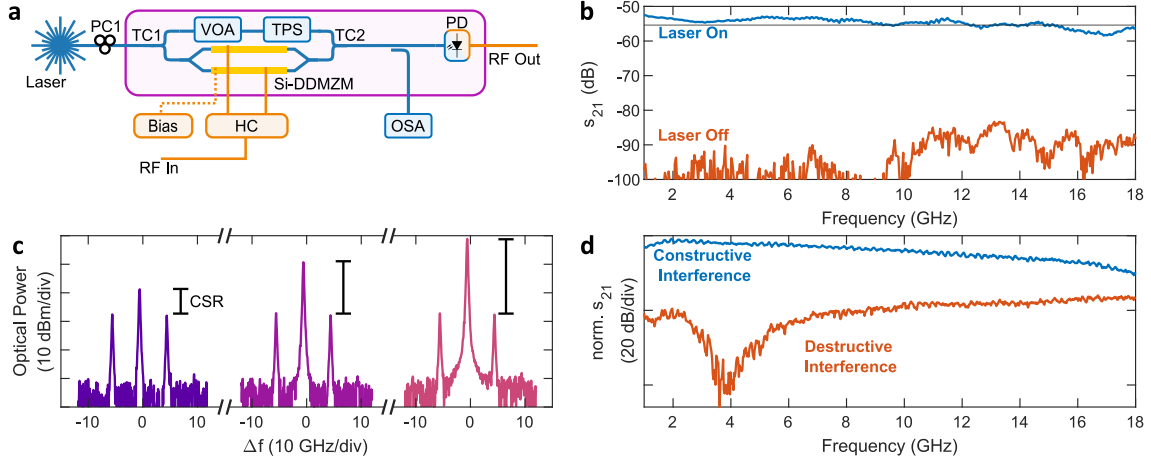


Figure 7.13: On-chip MWP link and E-O modulator characterisation. (a) Simplified schematic of an on-chip MWP link including the custom on-chip modulator. (b) Measured RF gain of the on-chip MWP link using an E-O modulator and PD on the silicon photonic chip. (c) Use of the on-chip E-O modulator to control the optical carrier-sideband ratio. The carrier-sideband ratio is indicated by the black lines. (d) RF response measured using the on-chip modulator to change the carrier phase, such that the RF photocurrents constructively interfere (blue) or destructively interfere (orange). HC: Hybrid coupler, DDMZM: Dual-drive Mach-Zehnder modulator, PD: Photodetector, TC: Tunable coupler, TPS: Tunable phase shifter, VOA: Variable optical attenuator, OSA: Optical spectrum analyser (II-VI WaveAnalyser), CSR: Carrier-sideband ratio.

1 mm $\times$ 2 mm.

The schematic of the novel modulator is depicted in Fig. 7.13 (a). CSR control is achieved by embedding an on-chip dual-drive Mach-Zehnder modulator (DDMZM), supplied by the IMEC iSiPP50G PDK [431], in an interferometer. In this demonstration, a 90° RF hybrid coupler was used to generate the RF drive signal in each arm. First, the input optical signal is split into two arms by a tunable coupler (TC1). High-speed E-O modulation occurs in the lower arm of the interferometer using the on-chip DDMZM. The amplitude and phase of the signal in the upper arm of the interferometer are controlled by a variable optical attenuator (VOA) and tunable phase shifter (TPS). When the upper and lower arms of the interferometer are combined by TC2, the resultant optical carrier at the modulator output is the vector sum of the carrier signals in both arms. As a result, the carrier power can be controlled independently of the sideband amplitude and phase, as shown in Fig. 7.13 (c). Using the novel on-chip modulator, we control the phase of the resultant optical carrier, which in turn controls the phase of the RF photocurrent contributions, such that they constructively (blue), or destructively (orange) interfere, as measured using an off-chip PD in Fig. 7.13 (d).

7.5 Integrated MHz-resolution MWP notch filter

7.5.1 Notch Filter Principle

The RF photonic notch filter principle is outlined in Fig. 7.14 and relies on two core concepts to achieve large RF notch depths with increased RF link gain. The first concept

is destructive interference of RF photocurrents at a given frequency to facilitate a deep RF notch [30], [405], [419], which is achieved through bias control of the DDMZM and Brillouin processing. The second concept is the control of the CSR to increase the RF link gain. In this section, we outline how these two concepts are implemented using the Si-As₂S₃ platform we present in this work.

The implementation of the MWP filter is based on the simplified schematic in Fig. 7.14 (a). The RF input signal, Fig. 7.14 (b), is modulated onto a continuous wave (CW) laser using a DDMZM, Fig. 7.14 (c). The CSR is then controlled using on-chip silicon photonic circuitry, as shown in Fig. 7.14 (d). Brillouin gain is subsequently applied to the upper sideband, shown in Fig. 7.14 (e), and an off-chip EDFA ensures constant power is maintained at the PD, Fig. 7.14 (f).

Upon photodetection, Fig. 7.14 (g), the RF photocurrent consists of two main contributions. One contribution is from the beat between the carrier and upper sideband and another is from the beat between the carrier and lower sideband. For a notch to form, these photocurrent contributions must sum destructively at a given frequency in the RF domain. Therefore, the photocurrent contributions must be equal in magnitude and π out-of-phase. In the passband of the MWP filter, a residual photocurrent is formed because the DDMZM bias voltage ensures that the photocurrent contributions are π out-of-phase, but their magnitudes are not equal. However, when the upper sideband coincides with the Brillouin resonant frequency ($\omega_{RF} = 2\pi f_0$), the photocurrent contributions destructively interfere as they are now equal in magnitude, owing to amplification from Brillouin gain. As a result, the photocurrent contributions cancel and an RF notch response is formed, as shown in Fig. 7.14 (g).

Furthermore, the increase in RF link gain with CSR reduction is exemplified by noting that the RF response with reduced CSR (CSR₂) has larger RF link gain than the RF

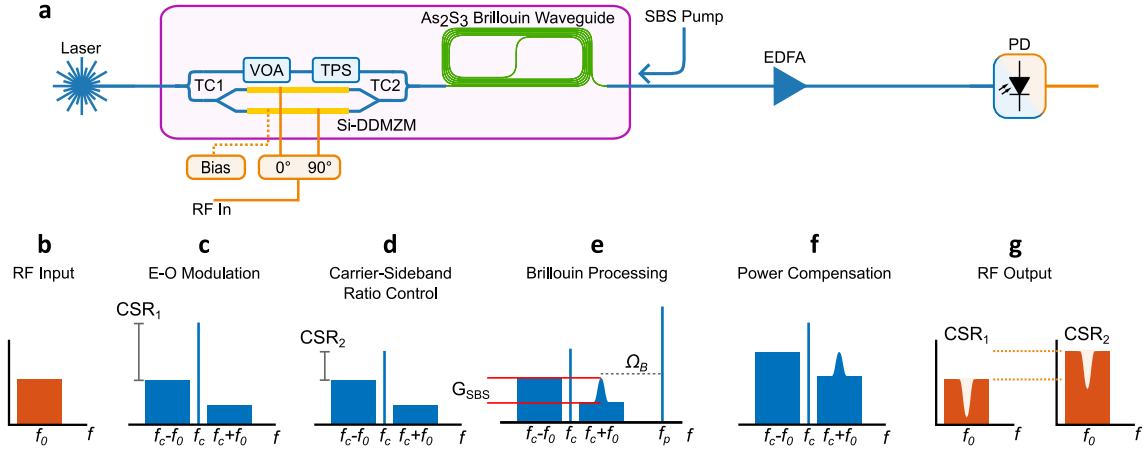


Figure 7.14: Operating principle of the MWP filter. (a) Simplified schematic. (b) The input RF spectrum is (c) modulated onto a continuous-wave laser using on on-chip DDMZM. (d) In parallel, the CSR is modified to increase the RF link gain. (e) The signal then undergoes Brillouin processing. (f) An off-chip EDFA is used to maintain constant optical power at the PD. (g) Upon photodetection, photocurrent contributions from the upper and lower sidebands destructively interfere at a spectrally-localised point whose frequency offset corresponds to the Brillouin resonance. The RF output is a notch response and the RF link gain of the output is increased through optical CSR control.

response with the un-altered CSR (CSR_1), as shown Fig. 7.14 (g). The notch frequency is controlled by changing the relative frequency offset between the signal laser and the Brillouin pump.

The CSR link gain improvement with CSR reduction can be understood by analysing the output of a DDMZM driven under small signal conditions by a single RF tone, with angular frequency ω_{RF} , and a $\pi/2$ phase shift between the RF signal in each arm consists of three main components, which may be expressed as

$$E_{-1} = \frac{E_0}{2} J_1(\beta) (je^{j\phi_B} + 1) e^{j(\omega_c - \omega_{RF})t} \quad (7.5)$$

$$E_{C1} = \frac{E_0}{2} J_0(\beta) (e^{j\phi_B} + 1) e^{j\omega_c t} \quad (7.6)$$

$$E_{+1} = \frac{E_0}{2} J_1(\beta) (je^{j\phi_B} - 1) e^{j(\omega_c + \omega_{RF})t} \quad (7.7)$$

where E_0 is the electric field of the signal laser at the input of the DDMZM (V/m), ω_c is the angular frequency of the optical carrier. J_n is the n^{th} order Bessel function of the first kind, β is the modulation index, defined as $\beta = \frac{\pi V_{RF}}{V_{\pi_{RF}}}$, where V_{RF} is the driving voltage of the DDMZM, $V_{\pi_{RF}}$ is the half-wave voltage of the DDMZM, and ϕ_B is the bias phase angle of the DDMZM. $E_{-1, C1, +1}$, correspond to the electric field of the lower sideband, carrier, and upper sideband, respectively.

If this signal is incident on a photodetector, the RF photocurrent at ω_{RF} can be expressed as [150], [290]

$$i(\omega_{RF}) \propto |E_{C1}| |E_{+1}| e^{j(\phi_{+1} - \phi_{C1})} + |E_{C1}| |E_{-1}| e^{j(\phi_{C1} - \phi_{-1})} \quad (7.8)$$

where $\phi_{-1, C1, +1}$ represents the phase of the corresponding optical electric field.

If we define the optical carrier-to-sideband power ratio (CSR) as

$$CSR = \left(\frac{E_{C1}}{E_{-1}} \right)^2 \quad (7.9)$$

Assuming that $E_{-1} \approx E_1$, we can express eq. 7.8 as

$$i(\omega_{RF}) \propto \frac{E_{C1}}{\sqrt{CSR}} (e^{j(\phi_{+1} - \phi_{C1})} + e^{j(\phi_{C1} - \phi_{-1})}) \quad (7.10)$$

Expressing eq. (7.10) in terms of RF link gain ($s_{21} \propto i^2(\omega_{RF})$) yields the result

$$s_{21} \propto \frac{P_0^2}{CSR} \quad (7.11)$$

Following from eq. 7.11, we see that the RF link gain is inversely proportional to the CSR. Therefore, we can increase the RF link gain by reducing the CSR if constant power is maintained at the PD.

7.5.2 Heterogeneous Si-As₂S₃ MWP Filter Characterisation

The integrated Si-As₂S₃ platform was used to implement an MWP notch filter in the setup depicted in Fig. 7.15 (a). A continuous wave laser (L1) is modulated by a VNA, which drives the on-chip DDMZM and forms the Brillouin probe. An RF hybrid coupler imparts a 90° phase shift between the RF drive of each arm of the on-chip DDMZM. The DDMZM is biased, such that the upper and lower sideband magnitudes differ by the amount of applied Brillouin gain and give π out-of-phase photocurrent contributions.

The probe transitions from the Si waveguide to a 5.55 cm long As₂S₃ waveguide where it undergoes Brillouin processing, which provides narrowband amplification to one sideband, by counter-propagating it against the Brillouin pump, generated by a second laser (L2). Off-chip circulators are used to separate the counter-propagating Brillouin pump and probe. The probe signal subsequently undergoes optical filtering and amplification before photodetection. An RF notch is formed through spectrally-localised destructive interference of RF photocurrents at the PD output, as the magnitude of the Brillouin gain is such that the π out-of-phase photocurrent components are equal in amplitude when the upper sideband coincides with the Brillouin resonant frequency, but unequal otherwise. The novel on-chip E-O modulator controls the CSR at the output of TC2 to improve the RF link gain (See Section 7.5.1).

We did not use the PD on the Si-As₂S₃ chip for the MWP filtering demonstration because the MWP filter exhibited increased RF losses compared to the intensity-modulation direct-detection (IM-DD) link-only measurement without the As₂S₃ waveguide. As a result, the MWP link gain of the filter was below the crosstalk between the RF probes that drive the modulator and PD. The main sources of RF loss are attributed to two main factors. The first factor is the increased insertion losses (20 dB) of the As₂S₃ waveguides, which is linked to the sample preconditioning that removes a 2 μ m silica cladding layer to expose the silicon waveguides. The second source of loss is the partial destructive interference of RF photocurrents that are used to generate the MWP notch filter scheme due to the small 0.8 dB of SBS gain generated on the Si-As₂S₃ chip (See Section 7.6.2). These RF losses can feasibly be overcome by improving the As₂S₃ post-processing fabrication flow, reducing sidewall scattering through further design optimisation of As₂S₃ waveguide dimensions, and avoiding the silica cladding etching step by including dedicated open sections on the initial silicon circuit. Reducing the As₂S₃ waveguide losses will improve the link gain through two mechanisms, a direct improvement from lower optical insertion loss and larger Brillouin gain via the increase in effective length.

Fig. 7.15 (b) presents an RF notch response with narrow 37 MHz 3-dB bandwidth and deep 51 dB RF rejection, generated using only 0.8 dB SBS gain. The broadband response of the MWP filter, with a notch at 8 GHz, is shown in Fig. 7.15 (c). The ripple in the passband is caused by reflections at the Si-As₂S₃ interface (See Section 7.3). Fig. 7.15 (d) shows notch frequency tunability between 2.5 and 15 GHz. The notch frequency was tuned by changing the frequency of the Brillouin pump relative to the probe. The maximum notch frequency is currently limited by the limited magnitude of generated SBS gain and the roll-off of the on-chip E-O modulator response. Additionally, the RF notch depth is sensitive to amplitude and phase perturbations of the optical carrier and sidebands. Such perturbations can arise from changes to the DDMZM bias voltage (See Section 7.6.2). However, with reduced As₂S₃ losses, the magnitude of applied Brillouin gain will increase, enabling a larger notch frequency tuning range and reducing the sensitivity of

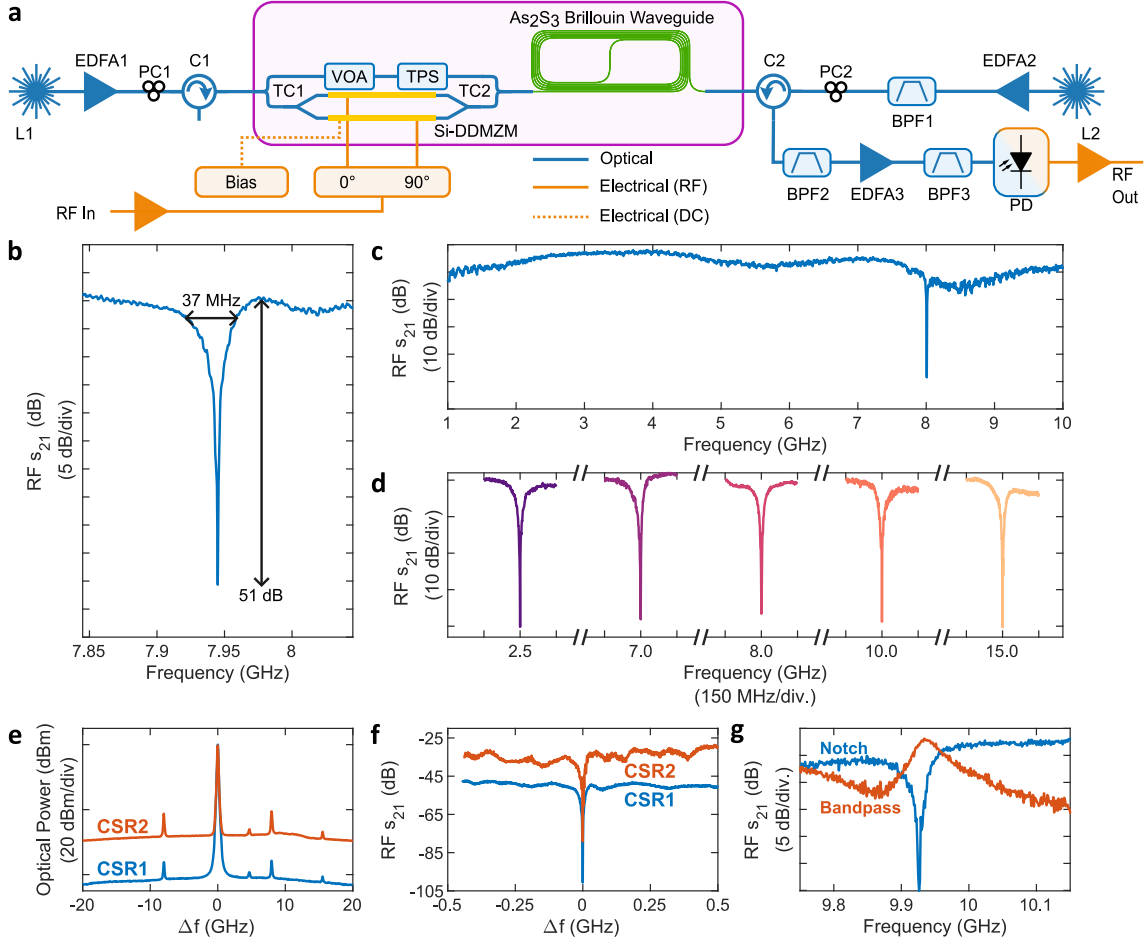


Figure 7.15: Si-As₂S₃ MWP notch filter. (a) Experimental setup. (b) Measured RF notch filter response with the narrow 3-dB bandwidth of 37 MHz and large 51 dB rejection annotated. (c) Measured RF response of the MWP notch filter with a notch at 8 GHz. (d) Notch filter frequency tuning. The notch frequency is changed between 2.5-15 GHz. (e) Measured optical spectra for two different CSRs. Optical processing on the Si-As₂S₃ PIC reduces the CSR from CSR1 (blue) to CSR2 (orange). The modulated sideband is offset by 7 GHz from the optical carrier. (f) Measured RF response for notch filters with CSR1 and CSR2. (g) Measured notch and bandpass filter responses. A bandpass response can be created by changing the DC bias of the on-chip DDMZM, such that the photocurrents destructively interfere outside the Brillouin resonance. L: Laser (Teraxion NLL), EDFA: Erbium-doped fibre amplifier, PC: Polarisation controller, C: Circulator, TC: Tunable coupler, VOA: Variable optical attenuator, TPS: Tunable phase shifter, DDMZM: Dual-drive Mach-Zehnder modulator, BPF: Band-pass filter (BPF1, BPF3: Santec OTF-320. BPF2: Alnair BVF-300CL).

the RF notch depth to environmental factors.

We demonstrate CSR control using the novel on-chip modulator in Fig. 7.15 (e). We present the measured optical spectrum for two different CSRs, using EDFA3 to maintain constant power at the PD. The corresponding RF response for CSR1 and CSR2 is shown in Fig. 7.15 (f). Using the novel on-chip modulator, we demonstrate a 13 dB improvement to the RF link gain through CSR control. The weak seed power into the EDFA causes significant amplified spontaneous emission (ASE) noise in the EDFA output. The excessive

ASE noise is expected to be reduced with lower As_2S_3 losses, such that a net RF link gain and NF improvement is observed with CSR suppression, as observed in Ref. [397].

The flexibility of MWP signal processing is highlighted through our use of the novel on-chip E-O modulator to alternate between bandpass and notch responses in Fig. 7.15 (g). We reconfigure the filter response between notch (blue) and bandpass (orange) filter responses by changing the modulator bias. For a bandpass response, we set the sideband amplitudes to be equal outside the Brillouin resonance, such that the photocurrents destructively interfere. At present, the rejection ratio of the bandpass filter response is limited due to partial broadband destructive interference caused by residual dispersion of the on-chip components.

We further characterise the trend of increasing RF link gain with decreasing CSR in Fig. 7.16. A least squares fit reveals a linear dependence between CSR and RF gain, both in decibel scale, with a gradient of -1.05. This is in good agreement with the theoretical model given by eq. (7.11), which predicts a gradient of -1. An overall RF link gain increase of 16 dB is observed through CSR control. We note that the reported RF link metrics presented in this work include two RF amplifiers, as shown in the setup in Fig. 5 (a), which add a total gain of 47 dB.

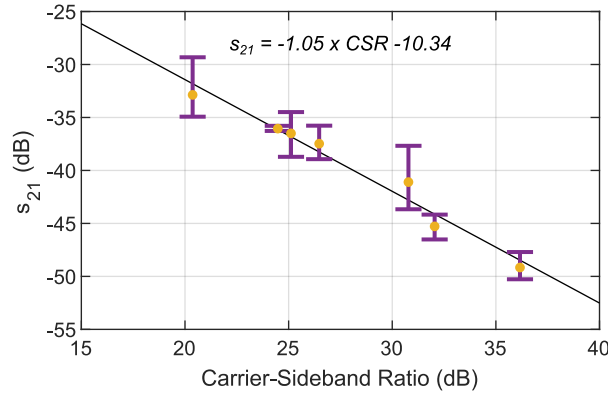


Figure 7.16: RF link improvement using novel on-chip E-O modulator to control the CSR. The measured trend of CSR vs RF link gain. A linear dependency is observed and fit using a least-squares approach.

7.6 Outlook

7.6.1 Increased SBS gain with reduced As_2S_3 losses

We now estimate the expected magnitude of SBS gain with reduced As_2S_3 propagation losses and reduced Si- As_2S_3 interface losses. This models the performance we anticipate when the post-processing of the Si- As_2S_3 chip does not increase the losses of the As_2S_3 waveguides. To calculate the expected magnitude of Brillouin gain, we use the Brillouin gain coefficient that we measured on the Si- As_2S_3 platform and the propagation losses we measured on the standalone As_2S_3 platform. For the coupled power, we used a value of 140 mW, which is 1 dB less than the coupled pump power of 177 mW as recorded by Morrison *et al.* on a passive Si- As_2S_3 platform [176]. The additional 1 dB loss accounts

for on-chip losses from on-chip components required to route the pump in an active Si-As₂S₃ platform. These values are summarised in Table 7.1. We substitute these values in eq.'s 7.3-7.4 and find an expected Brillouin gain of approximately 18 dB.

Table 7.1: Parameters used to model SBS gain increase.

Parameter	Symbol	Value	Unit	Ref
Brillouin gain coefficient	g_B/A_{eff}	703	/m/W	Measured, This Work.
As ₂ S ₃ propagation losses	α	0.44	dB/cm	Measured, This Work.
As ₂ S ₃ waveguide length	L	5.55	cm	Waveguide design, This Work.
Coupled pump power	P	0.140	W	1 dB less than achieved in [176]

7.6.2 MWP link improvements with increased SBS gain

We develop the model for determining the RF link performance of the MWP filter we present in this work as a function of applied SBS gain. The RF link gain of an MWP link using intensity modulation is given by [72]

$$g_{IM} = \left(\frac{\pi r_{pd} R_L P_{opt} \sin(\phi_B)}{4LV_{\pi_{RF}}} \right)^2 \quad (7.12)$$

where r_{PD} is the photodetector responsivity (A/W), R_L is the RF load impedance (typically 50 Ω), ϕ_B is the modulator bias angle, L is the optical loss in the MWP link and $V_{\pi_{RF}}$ is the modulator half-wave voltage.

Given that the MWP filter modulation scheme we present in this work consists of two RF photocurrent contributions, we can express the output power of the MWP link as

$$P = I^2 R = (I_{USB} + I_{LSB})^2 R \quad (7.13)$$

For an MWP link using intensity modulation, the upper and lower sideband photocurrent contributions are equal and in-phase, so we can therefore express the MWP link output power as

$$P_{IM} = (2I_{LSB})^2 R \quad (7.14)$$

For the modulation scheme we present in this work, the photocurrent contributions are out-of-phase with each other and differ in magnitude by a factor of $\sqrt{G}$, where G is the magnitude of the applied Brillouin gain. The RF output power of the MWP link can therefore be expressed as

$$P_{DDMZM} = \left(I_{LSB} - \frac{I_{LSB}}{\sqrt{G}} \right)^2 R \quad (7.15)$$

By simplifying eq. 7.15, the MWP link RF output power for the modulation scheme we present in this work can be related in terms of the RF output power for the IM case as follows

$$P_{DDMZM} = \frac{1}{4} \frac{(\sqrt{G} - 1)^2}{G} \times P_{IM} \quad (7.16)$$

Given that the link gain is proportional to output power, we can therefore express the MWP link gain for the modulation scheme used in terms of the link gain of an IM link using the expression

$$g_{DDMZM} = \frac{1}{4} \frac{(\sqrt{G} - 1)^2}{G} \times g_{IM} \quad (7.17)$$

We note that in the case of large magnitudes of SBS gain, the modulation scheme used in this work approximates single sideband modulation. As a result, the overall RF link gain is reduced by a factor of 4.

The expression for NF and SFDR can be determined the equations below [72]

$$NF = P_N - G_{\text{MWP, dB}} + 174 \quad (7.18)$$

$$SFDR_n = \frac{n-1}{n} (IIP_n - NF + 174) \quad (7.19)$$

$$IIP_3 = \frac{4(V_{\pi, RF})^2}{\pi^2 R} \quad (7.20)$$

where P_N is the RF noise power spectral density, $G_{\text{MWP, dB}}$ is the MWP link gain in decibels and 174 represents the noise power spectral density (PSD) of -174 dBm/Hz.

In the following analysis, we assume that the total optical power and therefore the noise PSD remains constant between the DDMZM and IM case.

We can calculate the expected improvement in RF link performance with the reduced As_2S_3 waveguide losses using the values presented in Table 7.1, and the expression for MWP link gain in eq. 7.17. We identified in the Discussion that reducing the As_2S_3 waveguide losses improved the performance of the MWP filter through two mechanisms.

The first mechanism is from increased optical power at the photodetector. Reducing the loss per Si- As_2S_3 interface from 2.1 dB to 0.1 dB gives a total power increase of 2.0 dB per interface. Reducing the propagation losses from 2.02 dB/cm to 0.44 dB/cm over a 5.55 cm waveguide gives an optical power difference of 8.79 dB. Therefore a total optical power difference of $2 \times 2.0 + 8.79 = 12.79$ dB is to be expected. Because the RF link gain is inversely proportional to the square optical loss in the MWP link, eq. 7.12, the RF link gain is expected to increase by 25.59 dB from the increase in optical power with reduced As_2S_3 losses.

The second mechanism that improves the RF link performance is from the increased Brillouin gain generated in the As_2S_3 waveguides. This is because the upper sideband amplitude, and therefore the corresponding π out-of-phase photocurrent contribution, is reduced in magnitude as shown in Fig. 7.14 (e). In section 7.6.1, we showed that the generated Brillouin gain is expected to increase from 0.8 dB to 18 dB. Substituting these values into eq. 7.17, we expect an RF link gain improvement of 19.97 dB due to the increase in SBS gain.

Therefore, with reduced As_2S_3 waveguide losses, we expect a total RF link gain increase of $25.59 + 19.97 = 45.56$ dB. The increased optical power at the PD does not result in any NF and SFDR_3 improvement, as the RF noise PSD increases at the same rate as the RF link gain. However, if we consider the effect of increasing the SBS gain on the filter performance, the NF will reduce and the SFDR_3 will increase by 19.97 dB. This is

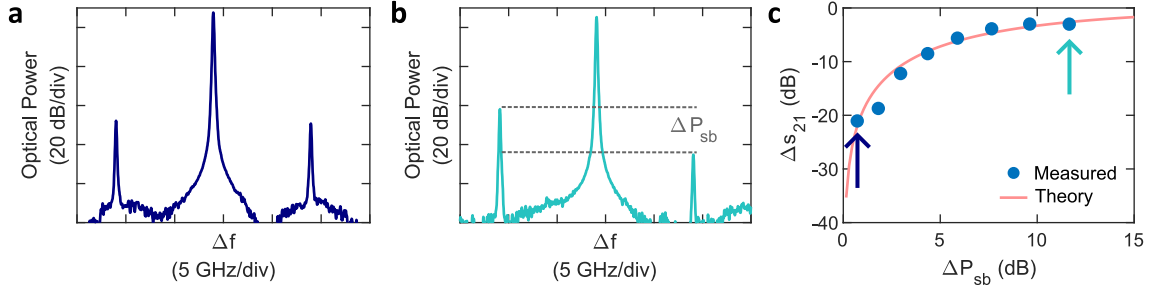


Figure 7.17: Measured MWP link gain as a function of the difference in power between the optical sidebands (ΔP_{SB}). (a) Measured optical spectrum corresponding to $\Delta P_{SB} = 0.7$ dB. (b) Measured optical spectrum corresponding to $\Delta P_{SB} = 11.65$ dB. (c) Measured RF link gain as a function of ΔP_{SB} . The arrows indicate the optical spectra presented in (a-b).

because increasing the SBS gain requires the upper and lower sidebands to have different magnitudes, but the total optical power, and hence RF noise PSD, does not change.

We measure the RF link gain as a function of the difference in power between the upper and lower sidebands (ΔP_{sb}) in Fig. 7.17. We plot the optical spectrum for ΔP_{sb} of 0.7 dB and 11.65 dB in Fig. 7.17 (a-b), respectively. We plot the measured change in RF link gain (blue circles), Δs_{21} , as a function of ΔP_{sb} in Fig. 7.17 (c). The measured values are in agreement with the model presented in eq. 7.17 (solid red line). The arrows indicate the values of ΔP_{sb} presented in Fig. 7.17 (a-b), highlighting that increasing the difference in power between the optical sidebands, and therefore Brillouin gain, increases the RF link gain.

7.6.3 Performance improvement: As_2S_3 platform with state-of-the-art integrated components

In this section, we model the best-case MWP link performance of the filter that we present in this work. We do this by substituting reported parameters of state-of-the-art silicon photonic components, best-case As_2S_3 losses into eq's. 7.17, 7.18, 7.19. The component values that we used to model the link performance are presented in Table 7.2.

We plot the MWP link gain, NF, and SFDR₃ as a function of SBS gain in Fig. 7.18 (a-c), respectively. We summarise the MWP link performance for an equivalent laser RIN at the PD of -165 dBc/Hz in Table 7.3. The results indicate that the link performance deteriorates rapidly for moderate magnitudes of SBS gain (<5 dB), which was a cause of RF performance degradation in this work. Hence, future improvements can be achieved by generating more SBS gain through a reduction in the As_2S_3 waveguide propagation losses.

The performance reported in Table 7.3 indicates that it is feasible to achieve significant improvements to this platform, which is mainly limited by As_2S_3 losses. For example, with 10 dB of Brillouin gain, it is possible to achieve an MWP link gain of -38 dB, NF of 45 dB and SFDR of 86 dB without RF amplification. This is compared to the RF link gain of -35 dB, a NF of 57.5 dB and SFDR₃ of 80 dB that we recorded experimentally, which includes two RF amplifiers providing 47 dB of total gain, as shown in the setup in Fig. 5 (a).

Table 7.2: Summary of parameters used to simulate MWP link performance on a future Si-As₂S₃ platform.

Parameter	Value	Unit	Ref
On-chip Laser Power	115	mW	[242]
Laser RIN	-165	dBc/Hz	[242]
Silicon photonic modulator Loss	6.5	dB	[431]–[433]
Silicon photonic modulator $V_{\pi RF}$	6.5	V	[431]–[433]
PD Responsivity	1	A/W	[296], [297]
As ₂ S ₃ Propagation Losses	0.44	dB/cm	Measured, This Work.
As ₂ S ₃ Interface Loss	0.1	dB	[176]
As ₂ S ₃ Length	5	cm	N/A
Circulator loss	2	dB	[382], [434]

Still, this performance must be further improved to meet the requirements of future RF systems. To understand the key performance trade-offs for the modulation scheme presented in this work, let us consider the case for small and large magnitudes of SBS gain. For small values of gain, the modulation format resembles PM and the two out-of-phase photocurrent contributions are almost equal in amplitude and therefore leave a small residual photocurrent at the PD output which leads to a reduced RF link gain and SFDR, and an increased NF. As the SBS gain increases, the modulation format resembles SSB modulation and a 6 dB link gain penalty is observed compared to IM, as the net photocurrent is halved because it essentially consists of only a single contribution. Further, the relatively large half-wave voltage of silicon modulators limits the RF link performance.

Table 7.3: Simulated MWP link performance for different values of SBS gain. The results correspond to RIN = -165 dBc/Hz

SBS Gain (dB)	RF Link Gain (dB)	NF (dB)	SFDR (dB.Hz ^{2/3})
1.00 dB	-54.27	61.10	75.49
5.00 dB	-42.17	49.01	83.56
10.00 dB	-38.30	45.13	86.14
30.00 dB	-35.27	42.11	88.16

7.6.4 RF Notch Depth Sensitivity Analysis

In this section, we analyse the sensitivity of the RF notch depth to For a notch to perturbations in the DDMZM bias voltage. For a form, the resulting photocurrent contributions from each sideband must be equal in amplitude and π out-of-phase. The DDMZM bias voltage changes the amplitude of each sideband. Therefore, changing the DDMZM bias voltage affects the RF notch depth.

To analyse the dependency of the RF notch depth on DDMZM bias voltage, let us consider the spectral components of the DDMZM output under small-signal modulation, given by eq.'s 7.5-7.7. The RF output is formed after photodetection, which is represented by eq. 7.8.

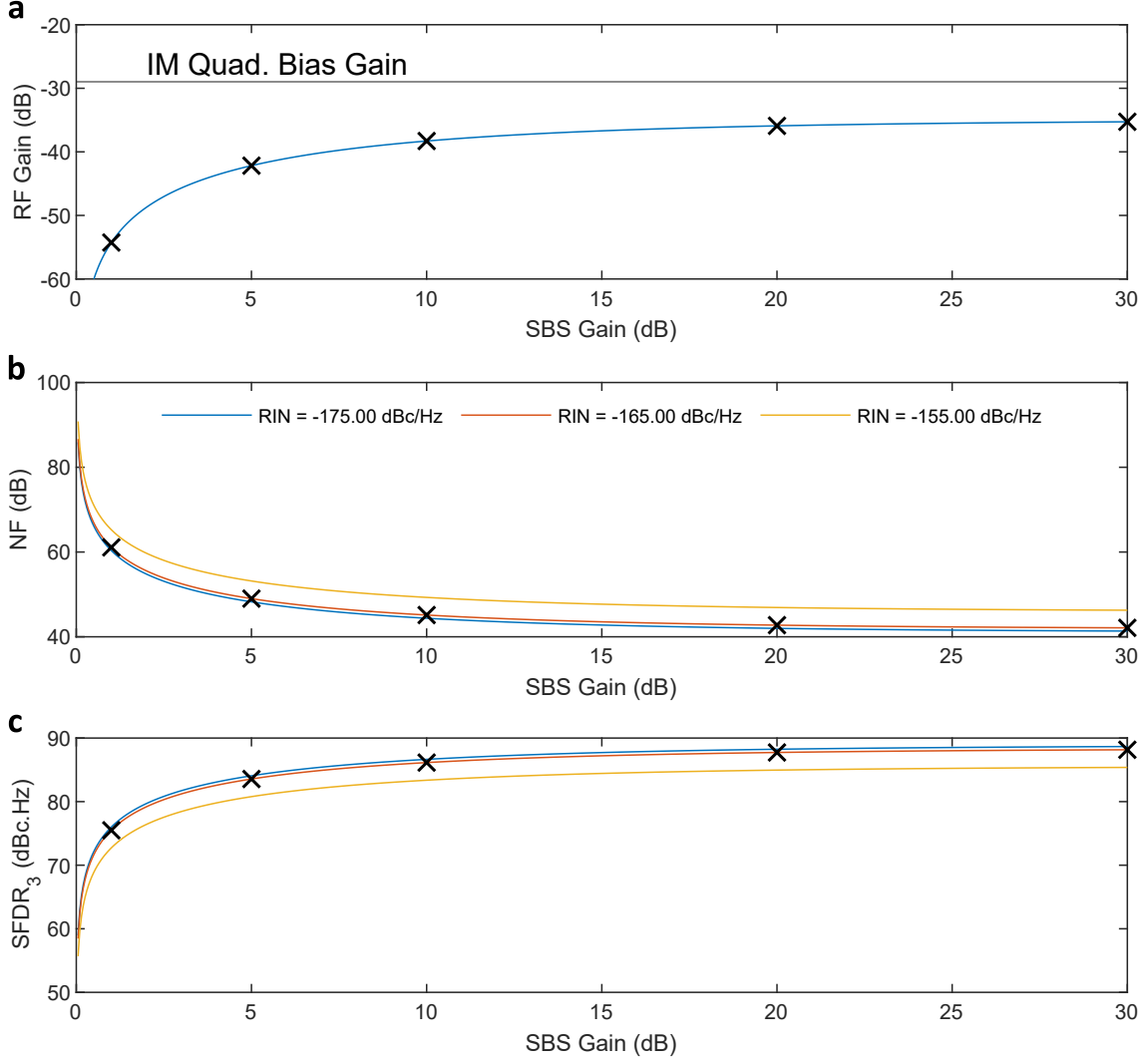


Figure 7.18: Calculated MWP link performance as a function of Brillouin gain.

(a) RF link gain as a function of Brillouin gain. (b) Noise figure as a function of Brillouin gain. (c) Spurious-free dynamic range as a function of Brillouin gain. The crosses represent the values presented in Table 7.3.

In the filter scheme presented in this work, the upper sideband is amplified by SBS gain to form the filter notch. The upper sideband after SBS amplification by a factor of G can be expressed as:

$$E_{+1} = \frac{E_0}{2} \sqrt{G} J_1(\beta) (je^{j\phi_B} - 1) e^{j(\omega_c + \omega_{RF})t} \quad (7.21)$$

In this work, the DDMZM bias was set such that the upper and lower sideband photocurrent contributions were π out of phase and differed in power by the magnitude of applied SBS gain. Assuming 1 dB of applied SBS gain, a DC bias angle of $\phi_B = -0.11$ radians is required. The DDMZM used in this work has a DC half-wave voltage of ≈ 12 V, which corresponds to a voltage of 0.44 V.

Fig. 7.19 (a) plots the net RF photocurrent outside the Brillouin resonance ($G = 1$) in blue, which corresponds to the filter passband, as a function of DC bias offset from

the optimal value of 0.44. Similarly, the net RF photocurrent at the Brillouin resonant frequency ($G>1$) is in orange, which corresponds to the filter stopband. A significant reduction of the RF output power is observed inside the SBS resonance around $\Delta V_{DC} = 0$ due to the interferometric nature of the MWP notch filtering scheme and corresponds to the RF notch.

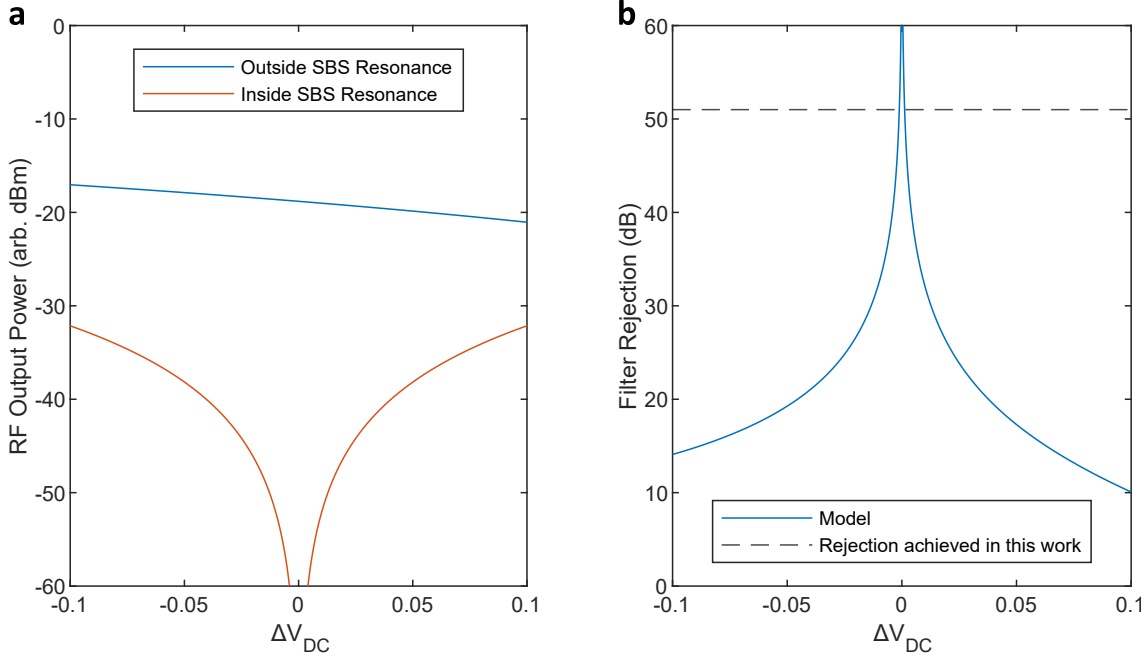


Figure 7.19: Simulated RF rejection ratio dependency as a function of DDMZM DC bias voltage. (a) Simulated RF output of MWP filter as a function of DC bias voltage for frequencies that are outside the SBS resonance (blue) and frequencies that correspond to the SBS central frequency (orange). (b) RF Rejection ratio as a function of DC bias voltage.

Fig. 7.19 (b) plots the RF rejection ratio as a function of ΔV_{DC} , which is calculated by subtracting the RF passband power from the RF stopband power. Large (>60 dB) RF rejection ratio is observed due to the destructively interfering photocurrents at $\Delta V_{DC} = 0$. The black dashed line marks the RF rejection ratio of 51 dB that we experimentally reported in this work. A rejection ratio of 40 dB can be maintained if $|\Delta V_{DC}| < 4.1$ mV. The sensitivity of RF rejection ratio on SBS gain is reduced with larger magnitudes of SBS gain, and hence reduced As_2S_3 waveguide losses. For example, with 5 dB of SBS gain, a rejection ratio of 40 dB can be achieved $|\Delta V_{DC}| < 14.2$ mV.

We note that the sensitivity of RF rejection to bias voltage is not specific to Brillouin-based filters. Many MWP notch filters that use interferometric techniques to enhance RF notch depth have similar sensitivity of RF notch depth to the modulator DC bias voltage.

7.7 Discussion

In this section, we outline improvements that can be made to the Si- As_2S_3 platform regarding optical loss and RF link performance. Reducing the As_2S_3 losses greatly improves RF link performance through two distinct mechanisms, the reduced losses of the As_2S_3

waveguides themselves and from the resulting increase in generated Brillouin gain. The first improvement to the link performance is due to the increased optical power at the PD from the reduced insertion losses. The optical power at the PD would increase by 12.8 dB if the As_2S_3 waveguides on the Si- As_2S_3 platform exhibited the same 0.4 dB/cm propagation loss as measured on the standalone As_2S_3 platform and the same 0.1 dB Si- As_2S_3 interface loss reported by Morrison *et al.* [176]. The second source of RF link performance improvement is from an increase in Brillouin gain, owing to the increased effective length of the As_2S_3 waveguide. This reduces the level of destructive interference in the filter passband. With 0.4 dB/cm propagation loss, the effective length of the 5.55 cm As_2S_3 waveguide would increase from 1.98 to 4.24 cm. As a result, we can feasibly expect the SBS gain to increase from 0.8 dB to 18 dB (See Section 7.6.1). The combined effect of reduced As_2S_3 losses leads to a significant RF link performance improvement compared to the results that we presented here of 46 dB to the link gain, as well as a NF reduction and SFDR₃ increase by 20 dB, respectively (See Section 7.6.2). Furthermore, recent progress of integrated photonic circuits has led to the rapid development of low- $V_{\pi_{RF}}$ modulators that would lead to direct RF link performance improvements [114], [263].

We acknowledge the link performance that we demonstrate here does not yet meet the requirements for real-world systems. However, significant RF link improvements outlined above indicate that the MWP link gain can be significantly improved, providing the opportunity to greatly reduce system SWaP by simultaneously using integrated E-O modulators, PDs and Brillouin waveguides. Additionally, the observed MWP filter 3-dB bandwidth of 37 MHz can be further reduced with lower As_2S_3 waveguide losses, which increases the magnitude of SBS gain and causes an associated gain narrowing [435], with a 3-dB linewidth of 10 MHz having been measured from 52 dB on-chip gain [334]. Further bandwidth reduction can be achieved by exploiting fundamental properties of the Brillouin interaction such as increasing photon lifetime through dynamic gratings [436], or tailoring the response of overlapping Brillouin gain and loss resonances [437]. Similarly, incorporating intermodal SBS [389], forwards SBS [438], or using tunable couplers to separate the counterpropagating pump and probe could enable systems where non-reciprocal devices are not needed, simplifying the overall system design.

Table 7.4 summarises the performance of state-of-the-art MWP filters which exhibit a high degree of integration. It shows that this work is key in unifying the advantages of nonlinear Brillouin signal processing, which offers fine spectral resolution and reconfigurability, with active E-O components. An additional benefit of our work is the flexibility enabled by the novel on-chip E-O modulator, enabling RF link gain improvement through CSR control and simultaneously facilitates the MWP filter to be altered between bandpass and notch responses.

The advantage of this work is further highlighted by comparing the reported 3-dB bandwidth of the MWP filters. Although some MRR and tapped delay line demonstrations have successfully integrated on-chip E-O devices [34], [172], they were generally limited to 3-dB bandwidths that exceeded hundreds of MHz. Additionally, Brillouin processing offers advantages over alternate optical signal processing methods. For example, the FSR of MRRs and tapped delay line MWP filters limit their frequency tunability. However, Brillouin MWP filters do not face such limitations as there is no FSR associated with the SBS process. Furthermore, the thermal stability of the Brillouin frequency shift

(≈ 1 MHz/ $^{\circ}$ C in optical fibres [429]) ensures stable notch operation, whereas the frequency of high-Q MRRs often exhibits a strong dependency on temperature (e.g. ≈ 12 GHz/ $^{\circ}$ C was reported for a silicon MRR with Q of 10^7 in Ref. [427]), which introduces a challenging requirement to either stabilise the MRR, or lock the laser to the MRR resonant frequency. Although the 21 MHz 3-dB bandwidth demonstrated by the MRR-based MWP filter in Ref. [427] is promising for future generations of MRR-based systems, the device did not include an E-O interface and faces challenges from thermal sensitivity and the presence of an FSR in its response, as outlined above.

We have presented an integrated MWP notch filter based on heterogeneously integrated As₂S₃ waveguides in an active silicon photonic platform. This platform enabled an MWP filter with fine spectral resolution of 37 MHz and large RF rejection of 51 dB. Our demonstration addressed the challenge of achieving fine spectral resolution MWP processing and system integration in a foundry-manufactured silicon photonic platform. Furthermore, we improved the RF link performance of the filter through CSR control using a novel on-chip E-O modulator. With future optimisation of this device and integration of a laser, fully integrated Brillouin-based MWP systems can be feasibly achieved. This enables a new class of integrated platforms to implement a range of on-chip MWP processing functionalities, including phase shifters, true time delays, frequency converters and RF sources. Our work represents a significant step toward the ultimate goal of achieving a high-performance fully-integrated MWP processor with fine spectral resolution and broadband tunability.

Table 7.4: Summary of state-of-the-art integrated MWP filters.

Year	Material Platform	Filter Type	Optical Processing	On-chip E-O?	FSR-free operation?	On-chip MWP link improvement?	3-dB Bandwidth (MHz)	Ref
2021	Si + InP	Notch, Bandpass	MRR	Y	N	N	360 - 470	[172]
2022	Si + InP + GaAs	Bandpass	TDL	Y	N	N	900 - 1,100	[34]
2022	Si ₃ N ₄	Notch, Bandpass	MRR	N	N	Y	400	[30]
2022	Si	Bandpass	MRR	N	N	N	20.6	[427]
2020	As ₂ S ₃	Bandpass	SBS	N	Y	N	30-350	[405]
2022	Si	Notch	FSBS	N	Y	N	2.7	[42]
2023	Si + As ₂ S ₃	Notch, Bandpass	SBS	Y	Y	Y	37	This work

MRR - Microring resonator, TDL - Tapped delay line, SBS - Stimulated Brillouin scattering, FSBS - Forward SBS, E-O - Electro-optic

Chapter 8

Summary and Outlook

8.1 Summary

In this thesis, we explored the benefits of integrating As_2S_3 Brillouin waveguides with chip-scale photonic components in Si_3N_4 and silicon PICs. We demonstrated the advantages of using a suppressed carrier phase modulation scheme to create an MWP bandpass filter with low (5 dB) losses with up to three passbands. We also demonstrated the advantages of combining As_2S_3 Brillouin waveguides with low-loss Si_3N_4 MRRs to form the basis of an MWP notch filter with a flexible and reconfigurable response. Finally, we demonstrated the highest degree of photonic integration of a Brillouin device to date by heterogeneously integrating a Brillouin As_2S_3 waveguide in an active silicon photonic platform containing novel E-O modulators and photodetectors. We used this platform as the basis of an MWP notch filter.

In Chapter 2, we provided a historical overview of MWP signal processing. Additionally, we provided an overview of integrated platforms in the context of MWP signal processing, as well as an introduction to common on-chip optical processing components.

Chapter 3 provided a comprehensive overview of the fundamentals of microwave photonics. The expressions for the optical fields for commonly used modulation formats in MWP processing systems were derived. The resulting MWP link response was shown for each modulation scheme. The application of optical processing to enable MWP filtering functionality was introduced. Furthermore, we introduced the expressions used to quantify the key performance metrics of an MWP link, which are extensively used throughout this work.

Chapter 4 provided a detailed discussion of stimulated Brillouin scattering and its applications in MWP signal processing systems. We began by outlining the theoretical foundation of SBS and its advantages for MWP processing, particularly in providing narrowband spectral resolution. We then reviewed the platforms that have been developed to harness on-chip SBS, including chalcogenide-based waveguides, silicon photonic waveguides, silica micro-resonators, and more recently, silicon nitride. We provided a review of on-chip SBS-based MWP signal processing systems, highlighting their key features. Finally, we concluded the chapter by summarising recent developments in on-chip SBS MWP systems and provided a perspective for future research directions.

Chapter 5 presented a novel MWP notch filter that harnessed the flexibility of MWP-based processing which achieved up to three stopbands, each being independently tunable

up to a maximum demonstrated frequency of 20 GHz, with RF rejection ratios surpassing 40 dB. This notch filter is based on optical filtering from an over-coupled Si_3N_4 MRR and on-chip SBS in an As_2S_3 waveguide. By combining Si_3N_4 MRRs with As_2S_3 Brillouin waveguides, we demonstrated RF notches with 500 MHz 3-dB bandwidth. This is a significant increase in spectral resolution compared to previous MWP filters that use As_2S_3 MRRs to implement a similar scheme, which was limited to wide 3-dB bandwidths on the order of 3 GHz, owing to losses in the As_2S_3 MRR. The enhanced functionality of the Si_3N_4 - As_2S_3 MWP filter was highlighted by suppressing unwanted signals in the RF spectrum by up to 50 dB. We also demonstrated the benefits of this filter, using it to maintain a communications channel in the presence of interfering signals.

An advantage of the cascaded MRR-SBS MWP notch filter presented in Chapter 5 is that it is based on IM, enabling favourable RF link performance. However, one drawback of this scheme is that the RF notch bandwidth is limited by the relatively large linewidth of the MRRs. In Chapter 6, we use a suppressed-carrier phase modulation scheme to demonstrate bandpass filtering functionality that maintains fine spectral resolution (≈ 30 MHz) while achieving low-losses (5 dB). By ensuring that destructive interference occurs in the stopband of the MWP filter, we achieved large rejection (40 dB) using only 14 dB of SBS gain. This novel suppressed carrier modulation scheme overcomes challenges many MWP processing systems face in achieving low losses.

One limitation of all Brillouin-based MWP signal processing systems to date is that the only on-chip photonic functionality is the generation of SBS while the rest of the system relies on discrete fibre-based optical components. In Chapter 7, we overcome this limitation by integrating As_2S_3 chalcogenide Brillouin waveguides in a compact platform that contains active silicon photonic components and achieved MWP notch filtering with fine spectral resolution of 37 MHz and large RF rejection of 51 dB. Furthermore, we extended the work presented in Chapter 6 and highlighted the advantages of photonic integration by including optical circuits to enhance the optical gain through carrier-sideband ratio control. By altering the optical CSR, we demonstrated RF link gain improvements of up to 16 dB.

8.2 Outlook

This section provides perspectives on future research to build on the work presented in this thesis. The perspectives fall into two main categories that are summarised below:

- Exploring the integration of As_2S_3 waveguides with high-power handling optical platforms, including Si_3N_4 , LiNbO_3 and InP . A key benefit offered by these platforms over the silicon photonic platform is their ability to handle large optical powers which is critical for MWP and on-chip SBS applications. In addition, these platforms offer a range of passive optical processing components (including MRRs) and active components including modulators, PDs and lasers which can be used to explore further photonic integration.
- Demonstrating a suitable method to achieve a fully-integrated backward Brillouin circuit that does not require on-chip nonreciprocal devices. An architecture that uses on-chip tunable couplers and optical filters to separate and attenuate the counter-propagating pump and probe that could facilitate a proof-of-principle demonstration is presented. The trade-offs introduced by such a scheme are discussed.

Parts of this discussion are based on the work in Ref. [81].

8.2.1 Novel heterogeneous As_2S_3 platforms

There are many advantages to exploring materials beyond the well-recognised use of silicon for developing novel hybrid or heterogeneously integrated Brillouin MWP signal processing platforms. For example, the multilayered TriPleX Si_3N_4 platform [343] offers a mature PDK targeted for MWP applications. Additionally, hybrid integrated lasers, modulators and PDs have been demonstrated when integrated with InP gain sections. The TriPleX platform has formed the basis of a fully-integrated MWP beamformer [16], and more complex designs have been reported which include in-line optical amplifiers to improve the RF link gain [439], [440]. Alternatively, single-layer Si_3N_4 has utilised integration with InP gain sections to create hybrid lasers [249], [250]. Additionally, the platform enables optical amplifiers such as erbium-doped waveguide amplifiers [441] and non-reciprocal components [382], which are all required for a fully-integrated MWP processor.

Furthermore, recent developments in thin-film lithium niobate platforms have led to the rapid development of high-speed and low- $V_{\pi\text{RF}}$ E-O modulators [257], [263], critical in enabling low RF-loss and high-performance MWP processing systems [25]. Integrating As_2S_3 waveguides in thin-film lithium niobate is an exciting future direction to enable high-performance and fine spectral resolution MWP processors.

However, the limited magnitude of SBS gain that has been induced [39], [321], or numerically predicted [442] in the platforms mentioned above highlights the unique role that chalcogenide plays in this context and the priority to integrate high-gain chalcogenide glasses into other integrated MWP processing platforms. In particular, the ability to generate large amounts of Brillouin gain with moderate Brillouin pump power is essential in scenarios where multiple Brillouin pump tones are required, as is the case for MWP filters with multiple notches or passbands that we presented in Chapters 5 and 6, respectively.

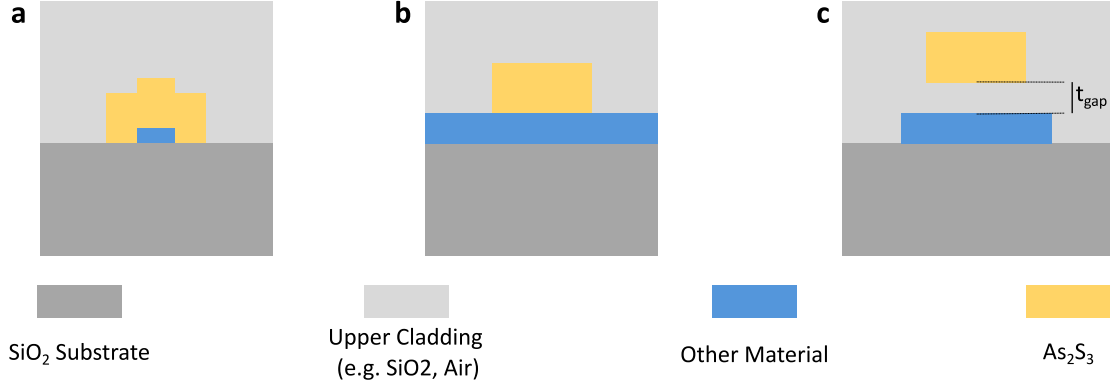


Figure 8.1: Coupling strategies for As_2S_3 Brillouin devices. (a) As_2S_3 waveguide directly over an existing waveguide. (b) As_2S_3 strip-loaded structure. (c) Inter-layer coupling.

In Fig. 8.1 we present a high-level overview that outlines possible geometries for heterogeneous integration of As_2S_3 waveguides with the optical materials mentioned above. In Fig. 8.1 (a), we present an arrangement where the As_2S_3 waveguide surrounds the additional material; this is typically suited to the case where As_2S_3 has a lower refractive index than the other material to be integrated. This coupling geometry was employed in Chapter 7, where we heterogeneously integrate silicon and As_2S_3 waveguides on an active silicon photonic platform, enabling a compact MWP filter. Fig. 8.1 (b) presents a strip-loaded approach, where the As_2S_3 waveguide is patterned directly on the additional material. This approach works best if the index of As_2S_3 is higher than, or similar to the other material that the As_2S_3 is to be integrated with [259], [443]. For example, candidate materials include Si_3N_4 and LiNbO_3 . This arrangement can be used to extend the work presented in Chapter 5 by heterogeneously integrating As_2S_3 with high-quality factor MRRs on low-loss Si_3N_4 , or LiNbO_3 platforms to form the basis of a Brillouin and MRR-based MWP processing platform. Fig 8.1 (c) presents an approach where a thin cladding layer separates the two waveguides. This works for all refractive indexes of the additional materials the As_2S_3 is to be integrated with, provided a suitable coupling geometry can be found. Adiabatic mode conversion between layers is possible, however, accurate control over the vertical gap dimensions is critical to ensure low losses and high reliability.

Fig. 8.2 presents initial simulation results from Lumerical EME which explores integrating As_2S_3 waveguides with a thin-film LiNbO_3 platform. We note that As_2S_3 waveguides have been integrated with titanium-diffused lithium niobate ($\text{Ti}:\text{LiNbO}_3$) platforms in the past [276]–[280], [444]. However, the key benefit of etched TFLN over the Ti-diffused platform for MWP applications is the increased E-O conversion efficiency enabled by the highly-confined etched LiNbO_3 mode and reduced electrode spacing. In the LiNbO_3 - As_2S_3 device that we present, the optical mode transitions from the TE_{00} mode of 1200 nm wide LiNbO_3 waveguide with the 600 nm film etched to a depth of 300 nm, similar to the geometry used in [264] to the TE_{00} mode of a 1000 nm wide, 700 nm high full-etched As_2S_3 waveguide using a linear vertical tapered waveguide over a distance of 200 μm , similar to the work of Lai *et al.* [341]. Although significant effort is required to realise such a platform, it indicates that it is possible to achieve high-performance tran-

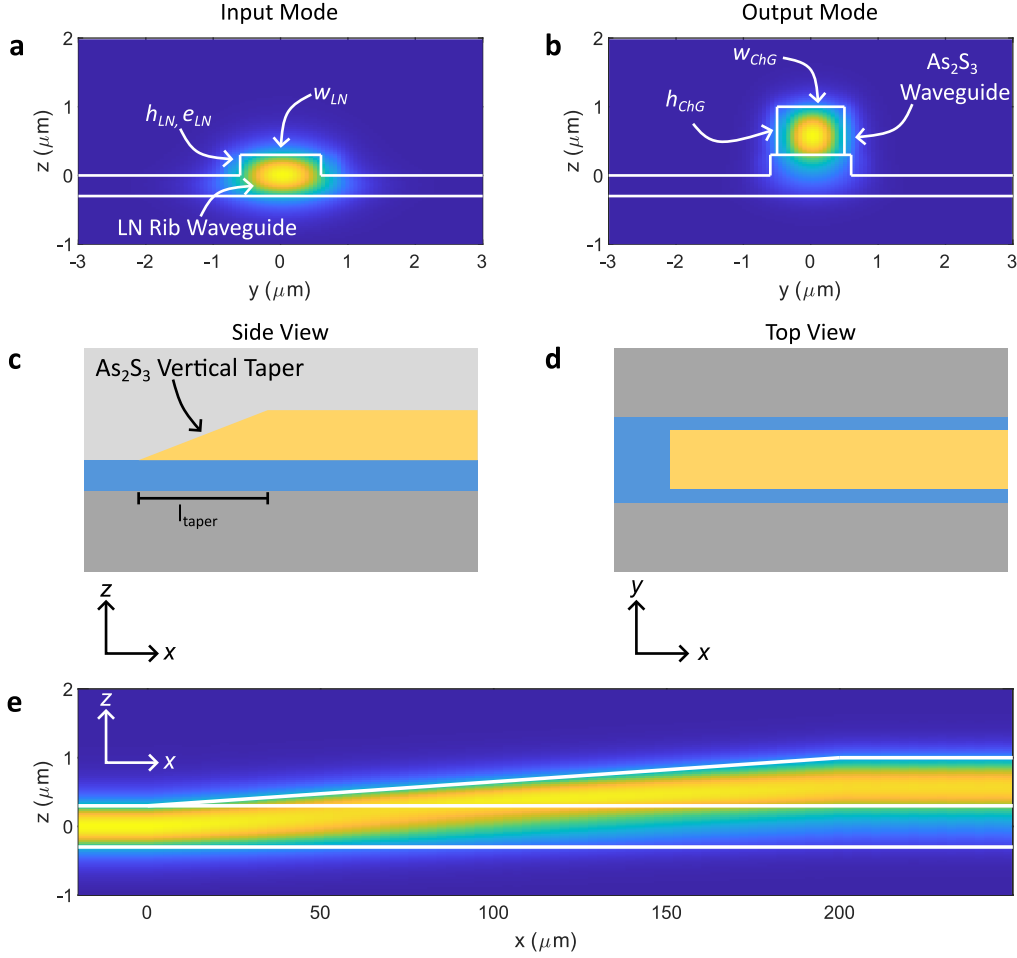


Figure 8.2: Simulation results of As₂S₃ waveguides heterogeneously integrated on a TFLN platform. (a) Simulated TE₀₀ mode of the etched LiNbO₃ waveguide. (b) Simulated mode profile of the TE₀₀ mode in the As₂S₃ waveguide deposited above the LiNbO₃ rib waveguide. (c-d) Side and top views of the taper structure. (e) Simulated propagation through the vertical As₂S₃ taper. The dimensions labelled in the figure are: $h_{LN} = 600$ nm, $e_{LN} = 300$ nm, $w_{LN} = 1200$ nm, $h_{chg} = 700$ nm, $w_{chg} = 1000$ nm, $h_{chg} = 700$ nm.

sitions between As₂S₃ and TFLN platforms. Pursuing this work is attractive for MWP signal processing applications as the low $V_{\pi_{RF}}$ and high power handling capability that is offered by TFLN platforms are key to realising high-performance MWP links [25], [27], [113], [257].

8.2.2 Further photonic integration of As₂S₃ Brillouin processors

The long-term goal of MWP is to achieve a fully-integrated processor with enhanced functionality that enables RF signal processing over wide frequency ranges and must meet the demanding performance specifications of real-world RF systems in terms of noise figure, linearity and insertion loss. Here, we outline the technological advances that are required to realise a high-performance and fully-integrated Brillouin-based MWP processor that incorporates multiple passive and active photonic components to achieve

the ultimate goal of a fully-integrated Brillouin-based MWP signal processor.

We first consider the underlying link performance requirements of the fully-integrated Brillouin based MWP processor, which must exhibit noise figure, linearity, and insertion loss that is competitive with their traditional RF electronic counterparts [25]. Generally, it is desirable to achieve low-loss transmission with noise figure below 10 dB and high linearity with SFDR exceeding $120 \text{ dB}\cdot\text{Hz}^{2/3}$ [25], [118]. However, most integrated microwave photonic systems do not meet these metrics, or do not report them.

To bridge the performance gap, low loss waveguides [343], [445], [446], low $V_{\pi_{RF}}$ modulators ($<1 \text{ V}$) [263], [447], on-chip lasers with high power ($>100 \text{ mW}$) and low RIN noise ($<-165 \text{ dBc/Hz}$) [242], [448], low noise optical amplifiers [441], [448], [449], and high-power handling and high-responsivity ($>1 \text{ A/W}$) PDs [206], [297], [450] must be integrated into a single platform. It is anticipated that the link performance of fully-integrated MWP processors will continue to improve in the future and reach the required performance considering that the performance of the chip-scale components referred to above are starting to meet or even exceed (as is the case for refs. [242], [263]) the performance of their discrete optical counterparts. Despite the continual improvement of integrated optical components, to achieve the required RF link performance it may be essential to incorporate RF link optimisation techniques in the MWP processor. Attractively, it is possible to implement many link optimisation techniques in chip-scale platforms including optical carrier suppression [405], [419], [451], which was demonstrated in chapters 6 and 7, low-biasing [178], and linearisation techniques [30], [408].

An extra consideration for Brillouin-based microwave processors is the integration of circulators, with isolation in the order of 30 dB, to separate the Brillouin pump and probe. However, fabricating low-loss and large isolation non-reciprocal devices such as circulators is challenging and currently limits the demonstration of a fully-integrated Brillouin-based MWP processor [287]. As discussed in Section 4.3, magneto-optical materials are difficult to implement in chip-scale devices, which leads to fabrication complexities [380]. Hence, non-reciprocal devices based on the same materials that are already used in the underlying chip-scale MWP processor are an attractive alternative to using magneto-optical material. Demonstrations of acousto-optic non-reciprocal devices using silicon [388], Si_3N_4 [382], and LiNbO_3 [452] waveguides highlight that non-reciprocal elements can be paired with many existing integrated MWP processing platforms.

Another possible implementation to achieve on-chip isolation to protect the probe laser from the counter-propagating pump laser is to use on-chip phase modulators that rely on phase matching between travelling optical and RF signals, as demonstrated in refs. [453]–[456]. Transmission occurs for propagating optical signals in the forward direction, as the optical signal and RF signal driving the phase modulator travel in opposite directions and are not phase-matched. However, when the optical signal propagates in the reverse direction, it is phase-matched with the RF drive signal and optical carrier power is transferred to the modulated sidebands, as the RF power drive of the modulator is such that the carrier is nulled. This method of producing on-chip isolation is attractive to explore in chip-based Brillouin MWP systems to provide one-way isolation.

It is also attractive for chalcogenide-based Brillouin systems to achieve complete photonic integration without the need for on-chip non-reciprocal components [81]. Liu *et al.* demonstrated backward intermodal SBS in a hybrid $\text{Si-As}_2\text{S}_3$ integrated circuit, which used mode couplers in place of circulators to separate the pump and probe [389]. Al-

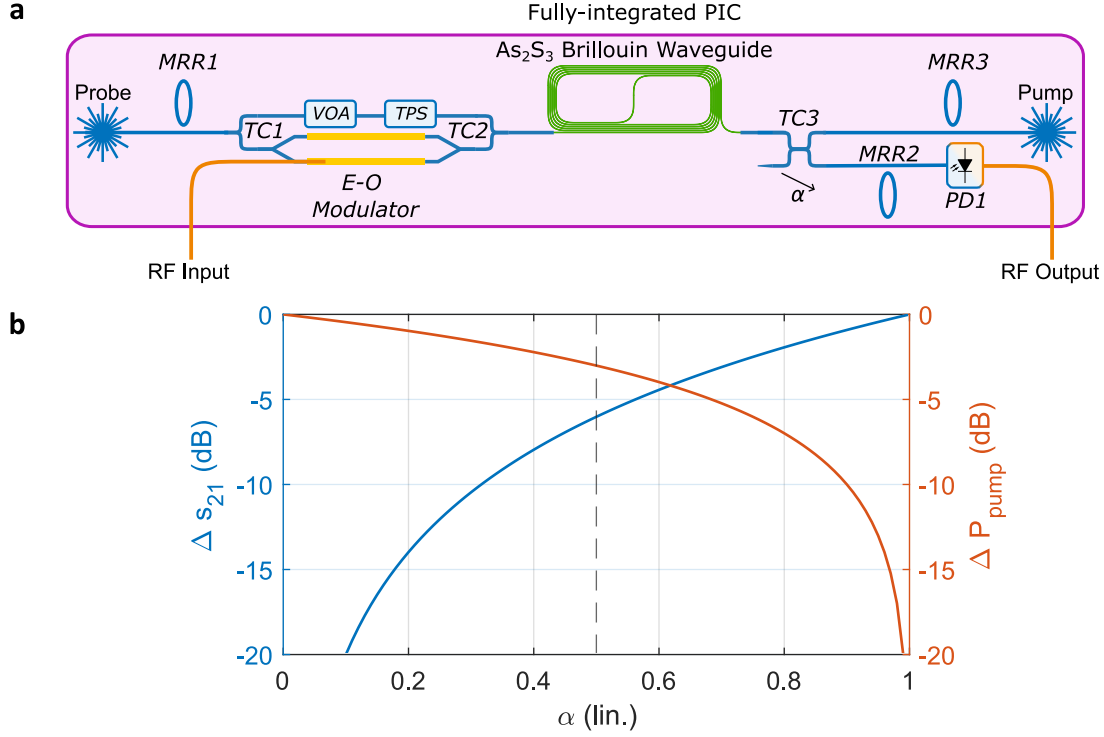


Figure 8.3: (a) Schematic for a fully-integrated and circulator-free Brillouin circuit. This device leverages on-chip optical filters and a tunable coupler in place of non-reciprocal components to provide a more short-term pathway towards a fully-integrated Brillouin MWP processors. (b) A plot of the dependency of the RF link gain (left) and Brillouin pump power (right) as a function of the splitting ratio of TC3, α . MRR: Micro-ring resonator, TC: Tunable coupler, VOA: Variable optical attenuator, TPS: Tunable phase shifter, E-O: Electro-optic, PD: Photodetector.

though this demonstration bypassed the need for on-chip circulators, the magnitude of the generated BIBS is significantly smaller compared to the standard backwards intramodal SBS process. An alternate path towards a circulator-free and fully-integrated backward Brillouin MWP processor is depicted in Fig. 8.3 (a). This architecture circumvents the need for on-chip reciprocal devices by using on-chip tunable couplers to separate the counter-propagating pump and probe. This comes at the expense of requiring on-chip filters to achieve isolation between the two laser sources and to attenuate the pump signal before photodetection. In particular, isolation between the laser sources is required to prevent undesired instabilities [457]. In this example the filters are represented by the MRRs, but can be with a sharper frequency roll-off using a series of Mach-Zehnder interferometers (MZIs), as demonstrated in Refs. [16], [398], [458]. MRR1 and MRR3 attenuate the counterpropagating laser signal to provide isolation between the two lasers. MRR2 is used to attenuate the pump signal before photodetection.

This scheme introduces an inherent trade-off between RF link gain and Brillouin pump power based on the tunable coupler power splitting ratio, α , as shown in Fig. 8.3 (b). To understand the trade-off between RF link gain and coupled Brillouin pump power, let us consider the following two cases. When the tunable coupler (TC3) is set in the cross-state, 100% of the probe light is routed to the on-chip PD and the pump light is routed to the waveguide terminator, giving maximum optical power at the PD but no Brillouin

pump power. When the tunable coupler (TC3) is in the bar state, the probe light is routed away from the PD and 100% of the pump light is directed to the Brillouin waveguide. This gives no RF output from the PD and hence minimal RF link gain, but maximises the Brillouin gain generated. A suitable trade-off may be found by directing equal power into each arm of the tunable coupler. In this configuration, the RF link gain is reduced by 6 dB, and the Brillouin pump power is reduced by 3 dB.

In fact, the Si-As₂S₃ platform that was introduced in Chapter 7 included an on-chip tunable coupler and high-speed PD to separate the counter-propagating pump and probe, as shown in Fig. 8.3 (a). However, due to the significant RF link losses, primarily limited by the small magnitude of the SBS gain generated, only the on-chip modulator was used in the experiment. Replacing off-chip isolators with hybrid couplers is an attractive pathway towards implementing an on-chip Brillouin MWP processor and can be demonstrated in future work.

Furthermore, we identified in Chapter 5 that reflections of the Brillouin pump can increase the spurious content and the noise PSD output of an MWP link. The main source of reflections at the PD in the system presented in Fig. 8.3 (a) would likely originate from the Si-As₂S₃ interface. Although low-reflection interfaces have been demonstrated using vertically tapered As₂S₃ waveguides, it is desirable to further reduce the amplitude of the reflected Brillouin pump to preserve link performance [340]. In the schematic presented in Fig. 8.3, MRR2 represents the optical filter used to implement Brillouin pump filtering. Alternatively, a spectral interleaver can be used to route the pump away from the PD [398], [458].

To summarise, the work presented in this thesis can be extended to facilitate further SWaP reduction and RF performance improvement of Brillouin-based MWP processors. One potential pathway explores novel material combinations of As₂S₃ Brillouin waveguides with materials such as LiNbO₃ and Si₃N₄ that offer high-power handling and high-performance components. Additionally, the work in Chapter 7 demonstrated the first step towards integrating Brillouin devices with key active E-O components for an MWP link. A scheme was presented that could enable a fully-integrated Brillouin-based MWP processing through further integration with active components including lasers and photodetectors, as well as passive filters and couplers for signal conditioning and separation of the pump and probe. Further developments of novel heterogeneous chip-scale platforms and the investigation of fully-integrated MWP signal processing architectures promise many opportunities to enable future generations of integrated Brillouin-based MWP processing systems with fine spectral resolution, compact footprint and high performance.

Appendix A

Integrated Brillouin Phased Array Antenna Systems

This appendix provides a perspective on the deployment of fully-integrated Brillouin devices in a phased array antenna (PAA). This is based on work published in [81]:

- **Matthew Garrett**, Moritz Merklein and Benjamin J. Eggleton, "Chip-Based Brillouin Processing for Microwave Photonic Phased Array Antennas," *IEEE Journal of Selected Topics in Quantum Electronics*, 2023, **29**, 1, 1-20.

A.1 Motivation

Wideband and high frequency microwave systems are attracting significant interest for use with phased array antennas (PAAs) [28], [459], [460]. When these systems are combined with the ability to generate electronically steerable radiation patterns, unique functionalities, including high resolution radars [58], [84] capable of tracking multiple targets [461], and directed high data rate communications to multiple users [462] are enabled. Typically, the emitted radiation pattern of a PAA is controlled by changing the relative phase (or delay) between the radiating elements in a process that is referred to as beamforming [460], [463], [464]. To reduce the presence of side-lobes in the PAA radiation pattern, each antenna element is typically separated by half the operating wavelength [465], [466]. Consequently, the components used for beamforming and analog signal conditioning, which form the interface between subsequent frequency conversion and digitisation stages, must fit in the space between radiating elements; this becomes challenging at higher frequencies, for example above 15 GHz the element spacing is less than a centimeter.

As an example, RF filters, a key component within PAAs, face challenges achieving wideband operation in the required compact form factor. Conventional RF filtering approaches based on surface and bulk acoustic wave processing achieve low losses and large out-of-band rejection [424]. However, they are not suitable for millimeter wave frequencies due to large acoustic losses [67], [424]. Reconfigurable filters based on cascaded banks of high-pass and low-pass filters offer discretely reconfigurable bandwidths and central frequencies [71]. However, many parallel filters are required to enable operation over a wide frequency range, increasing both system complexity and size. Filters based on ferromagnetic and ferrimagnetic materials, including yttrium iron garnet (YIG), achieve

wide frequency tunability using compact cavities [467], [468], but require an external magnetic field to control their resonant frequency, complicating integration. While the above-mentioned devices exhibit impressive performance, there is no clear RF filter technology that can operate over a wide frequency range and be implemented in a compact footprint that is compatible with element-level processing for high frequency PAAs [28].

Microwave photonics (MWP) is well positioned to overcome the challenges associated with the increasing bandwidth and carrier frequencies [25] of next-generation wireless communications [469] and radar [470], [471] systems. Systems that utilise MWP signal processing achieve wideband and high frequency operation by transducing the RF signal into the optical domain using an electro-optic (E-O) modulator. The RF signal is then processed using well established photonic components, that support operation over wide frequency ranges, before conversion back to the RF domain through photodetection [1]. The bandwidth of the MWP link is limited only by the bandwidth of the E-O modulator and photodetector (PD), which convert between RF and optical domains.

At the same time, photonic integrated circuits (PICs), based on established semiconductor fabrication techniques, are enabling significant reductions to the footprint of MWP devices [209], [256], [472]–[475], opening possibilities for processing high frequency and wide bandwidth RF signals at every element of a PAA. Chip-based MWP devices have enabled demonstrations of numerous RF processing functionalities that could form the basis of future element-level processing required for PAAs, including agile RF filters [27], [33], phase shifters [476], [477] and true time delays (TTDs) for beam steering functionality, [15], [16], [478], frequency converters that interface between the operating RF and baseband frequencies [17], [479], and microwave sources [21], [480]–[482]. Initial demonstrations of MWP beamforming networks highlight the potential for implementing chip-based beamforming for PAAs [15], [16], [439], [483]–[485], expanded on below.

On-chip MWP processors, for example RF photonic filters, have been demonstrated using integrated optical micro-ring resonators (MRRs) [16], [178], [408]. Although good performance has been achieved, MRR-based processing has limited spectral flexibility, as the shape of the photonic resonance is determined by the intrinsic waveguide losses and fixed structural properties of the MRR. Alternatively, programmable photonic mesh processors offer flexible RF responses [35], [36], but increasing the number of mesh processors to support additional RF processing functionalities requires longer waveguide lengths, which comes at the expense of increasing RF losses and size [35]. MWP transversal filters can be created using multiple optical carriers propagating through on-chip dispersive delay lines [33], [486], but require many optical carriers to achieve sub-100 MHz processing [292], consequently increasing system complexity.

Integrated MWP beamforming elements highlight the advantages of photonic integration by enabling devices with compact form factors that operate at high frequencies and support wide RF signal bandwidths [16]. Chip-based MWP beamformers have been demonstrated using switchable delay lines [478], [487], but can only produce discrete delays, limiting beam steering flexibility. Continuously tunable delay lines have been implemented using broadband dispersive effects from waveguide gratings [488], [489], by changing the coupling ratios of MRRs [15], [16], [484], or using delay line interferometers [485]. However, these approaches lack the required frequency flexibility, agility and resolution due to the fixed device structural properties and difficulties in scaling the device size, as outlined above.

As we outlined in Section 4.3, stimulated Brillouin scattering (SBS) [51] has enabled microwave photonic signal processors with unmatched frequency selectivity, tunability, and reconfigurability [25]. Inducing SBS in chip-scale platforms greatly reduces the form factor of Brillouin devices, when compared to their fibre-based predecessors [46], [149], [157], [490]. SBS facilitates desirable fine spectral resolution (MHz-level), owing to the intrinsic narrow bandwidth of the Brillouin resonance, which exhibits a Lorentzian gain and loss response with narrow bandwidth of approximately 30 MHz [51], unmatched by other optical processing methods. Because the Brillouin resonance is optically induced, its spectral response can be dynamically reconfigured by tailoring the optical pump spectrum [10], [44], [491], enabling tailored filter shapes that can be tuned over multiple octaves. The narrow bandwidth resonant response of SBS forms the basis of programmable phase and amplitude control used in MWP processors and has been demonstrated in different chip-based platforms, including chalcogenide waveguides [164], [176], suspended silicon on insulator (SOI) waveguides [317]–[319], and high quality factor (Q) Si_3N_4 [227] and SiO_2 resonators [337].

For the remainder of this section, we outline the challenges that must be overcome to enable a fully-integrated Brillouin-based beamforming network suitable for element-level processing in PAAs, and provide perspectives on technical improvements that must be made to achieve the long-term performance goals for MWP processors [27], [118].

A.2 Principle

PAAs are widely used in communications [462] and radar [84], [492] systems, enabling the transmission and reception of RF energy in an electronically controllable manner. Traditionally, PAAs consist of a two-dimensional array of antenna elements spaced by half the operating wavelength, as shown in the artistic impression in Fig. A.1, which depicts a

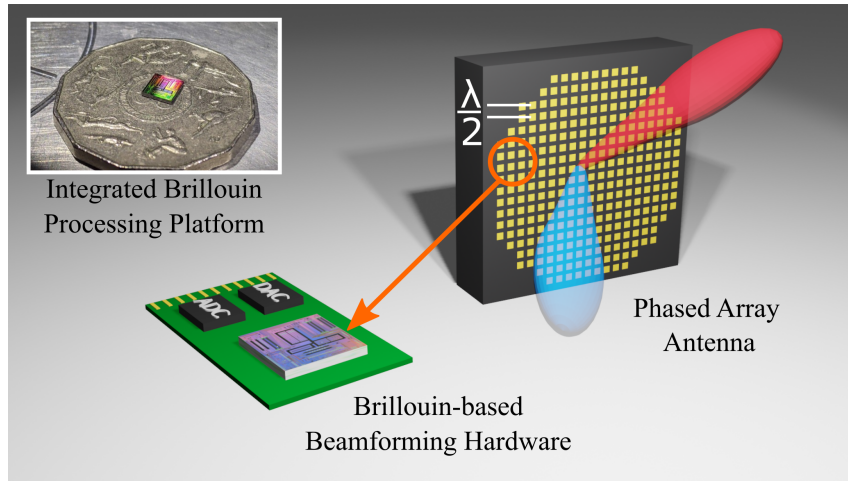


Figure A.1: Artistic impression of a phased array antenna which consists of radiating elements spaced by half the operating wavelength. On-chip Brillouin microwave photonic processing hardware is at each element to implement the required beamforming functionality. The inset shows a silicon photonic integrated circuit, a candidate hybrid platform for integrated Brillouin processing.

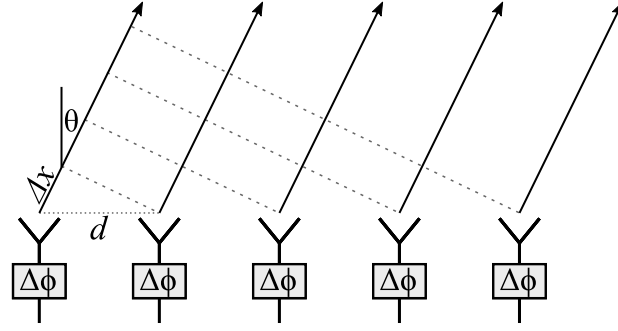


Figure A.2: Phased array antenna where beamforming is enabled by applying a phase shift to consecutive antenna elements.

chip-scale Brillouin processor that implements beamforming at each PAA element. The inset shows a compact silicon-based hybrid photonic circuit, which is a candidate technology for the Brillouin processor that we outline in Section A.2.1.

Beamforming can be implemented in PAAs by introducing a phase shift, $\Delta\phi$, between elements as shown in Fig. A.2. To direct the beam to an angle of θ , each successive wavefront must be delayed an additional path Δx , given by [84], [464], [492],

$$\Delta x = d \sin(\theta), \quad (\text{A.1})$$

where d is the distance between adjacent elements in the PAA. The additional path Δx is caused by the applied phase shift, and is related to the operating wavelength, λ , by,

$$\frac{2\pi}{\Delta\phi} = \frac{\lambda}{\Delta x}. \quad (\text{A.2})$$

Combining (A.1)-(A.2), and expressing the angle of the PAA beam in terms of the operating frequency, f , we get

$$\theta = \sin^{-1} \left(\frac{c\Delta\phi}{2\pi f d} \right), \quad (\text{A.3})$$

where c is the speed of light. As can be seen from (A.3), one drawback of conventional beamforming using phase shifts is that the beam direction changes as a function of the transmitted frequency, referred to as beam squint [84]. Beam squint can be eliminated by introducing a TTD between elements to control the beam angle [82] by applying a linear phase shift to the RF signal of interest. The resultant delay, τ , is the derivative of the applied phase shift with respect to frequency [40], [50]. Substituting the relationship between the applied phase shift and frequency into (A.3), we find that the beam angle no longer directly depends on the PAA operating frequency [84]

$$\theta = \sin^{-1} \left(\frac{\tau c}{d} \right), \quad (\text{A.4})$$

highlighting the advantages of using TTDs to implement beamforming in PAAs with wide signal bandwidths.

To understand the requirements for phase shifters and TTDs for beamforming, we solve for $\Delta\phi$ and τ in (A.3) and (A.4), respectively. Assuming that the PAA consists of $N \times N$

elements spaced by half the operating wavelength, $d = \lambda/2$, we find that the maximum phase shift and delay required is given by:

$$\Delta\phi_{max} = \pi(N - 1) \sin(\theta), \quad (\text{A.5})$$

$$\tau_{max} = \frac{(N - 1) \sin(\theta)}{2f}. \quad (\text{A.6})$$

Using (A.5) and (A.6) we calculate the required phase and time delay to achieve steering over, $\theta = \pm 90^\circ$. For PAAs based on phase shifts, a full 360° is required. For beam steering approaches based on TTDs, the required delay decreases with increasing frequency, and is less than 1 ns for 16×16 arrays above approximately 7.5 GHz, and only 150 ps at 50 GHz.

Fig. A.3 illustrates a typical signal chain of a PAA, which includes beamforming hardware at each antenna element, frequency converters and local oscillators (LO) to interface between the PAA operating frequency and baseband, filters for analog signal conditioning, and baseband digital electronics; more information is provided in Refs [464], [493].

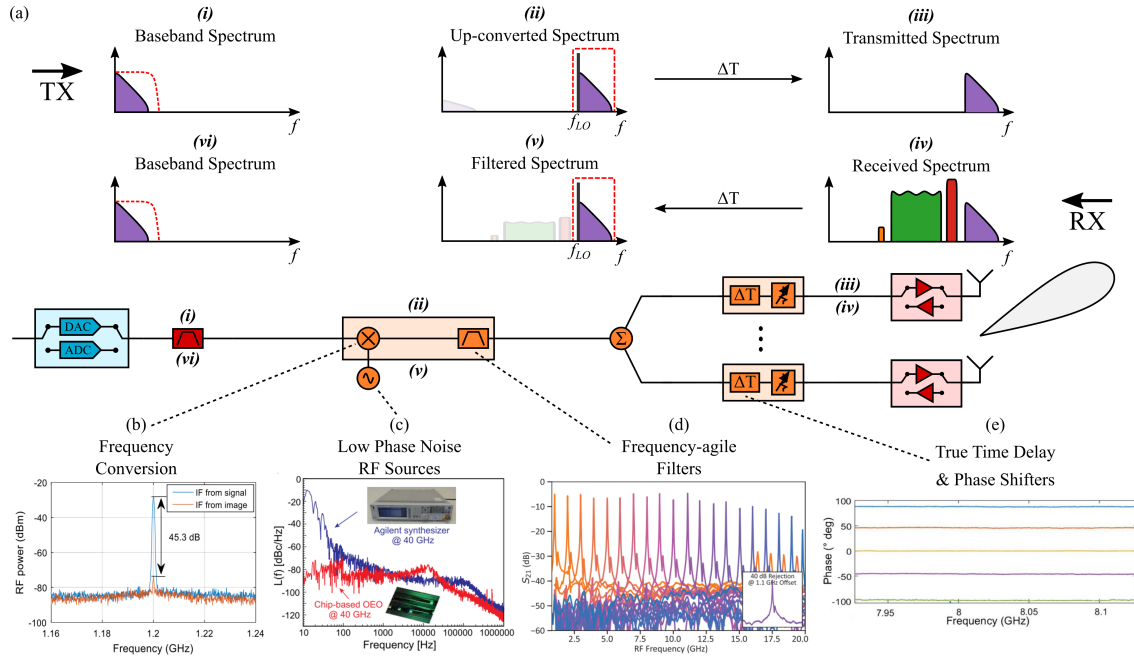


Figure A.3: (a) A typical analog beamforming architecture for a phased array antenna. In transmit mode, (i) a baseband signal from the DAC output is filtered, and (ii) converted to the PAA operating frequency before (iii) a delay and amplitude weighting is applied at each element before it is transmitted. In receive mode, (iv) a signal is received at each antenna element. Delay and amplitude weighting is applied at each element before (v) the spectrum is filtered to select a band of interest and converted to the baseband before (vi) the signal is filtered and undergoes analog to digital conversion. Essential microwave processing in this diagram has been demonstrated using on-chip Brillouin devices. Microwave processing and functional devices are shown in (b-e). (Figure A.3 (b) from [371], © 2020, Optica Publishing Group. Figure A.3 (c) from [22], © 2016, Optica Publishing Group. Figure A.3 (d) from [405], © 2020, Optica Publishing Group. Figure A.3 (e) from [40], © 2017, Optica Publishing Group.)

In transmit mode, a digital-to-analog converter (DAC) converts a digital input to an analog baseband signal at point (i) in Fig. A.3 (a). This signal is filtered by a fixed frequency filter at the baseband to remove spurious content from the DAC. A local oscillator (LO) drives a mixer to convert the generated signal from the baseband to the PAA operating frequency, and is filtered again to remove spurious mixing terms at point (ii). The signal is then distributed to each antenna element of the PAA. Before transmission, delay and amplitude weightings are applied at each element to control the beam direction at point (iii).

The same concept applies for receive beamforming, but the signal chain is reversed, with an analog-to-digital converter (ADC) being used to digitise the received signal. In receive mode, the RF filter serves an additional purpose to select a signal of interest from the RF spectrum. Band-pass filtering may be required to select a particular band of interest, or a notch filter with fine frequency selectivity may be required to reject a particular interfering signal without attenuating adjacent signals at nearby frequencies.

Fig. A.3 (b-e) highlight demonstrations of Brillouin-based MWP processing devices that perform the key functions within a beamforming network, including frequency converters [371], low phase noise RF sources [22], filters [405], phase shifters [40] and TTDs [50]; we expand on these examples and conclude with a perspective on how these distinct functionalities can be combined onto a single photonic chip to enable element-level processing for PAAs.

A.2.1 Proposed Brillouin-based phased array antenna

A potential architecture of a PAA using fully-integrated MWP processors is shown in Fig. A.4 (a). To simplify the discussion without loss of generality, we discuss the PAA operating in receive mode. The proposed architecture consists of a Brillouin beamforming network with a PIC at each row of the PAA, where integrated Brillouin processors implement TTD beamforming at each antenna element. Chip-based pump and probe lasers are routed to each integrated Brillouin processor and the spectrum of each pump is broadened by a DAC which drives MZM1. The signal from each integrated Brillouin processor is combined and converted to the baseband using a Brillouin-based image rejection mixer driven by a low-noise Brillouin-based oscillator. This ensures signals outside of the band of interest are attenuated through destructive interference [20], [371], essentially acting as a MWP filter cascaded with a frequency converter.

The core components of the integrated Brillouin processor are highlighted in Fig. A.4 (b), which works as follows; an input RF signal is modulated onto the optical carrier by MZM2 at (i), a modulation transformer [30], [398] generates the required modulation format at (ii), on-chip optical amplifiers set the pump and probe power as required before the probe undergoes Brillouin processing at (iii), an optical notch filter suppresses reflections of the Brillouin pump at (iv). Reflections are primarily caused by index changes between different materials and can be greatly reduced in an appropriately designed integrated platform, compared to systems that rely on a combination of chip-scale and fibre components [341]. Finally, the signal is converted back to the RF domain. This architecture can be extended to enable phase or delay amplification, as outlined in Section 4.3, by adding additional core components outlined in Fig. A.4 (b).

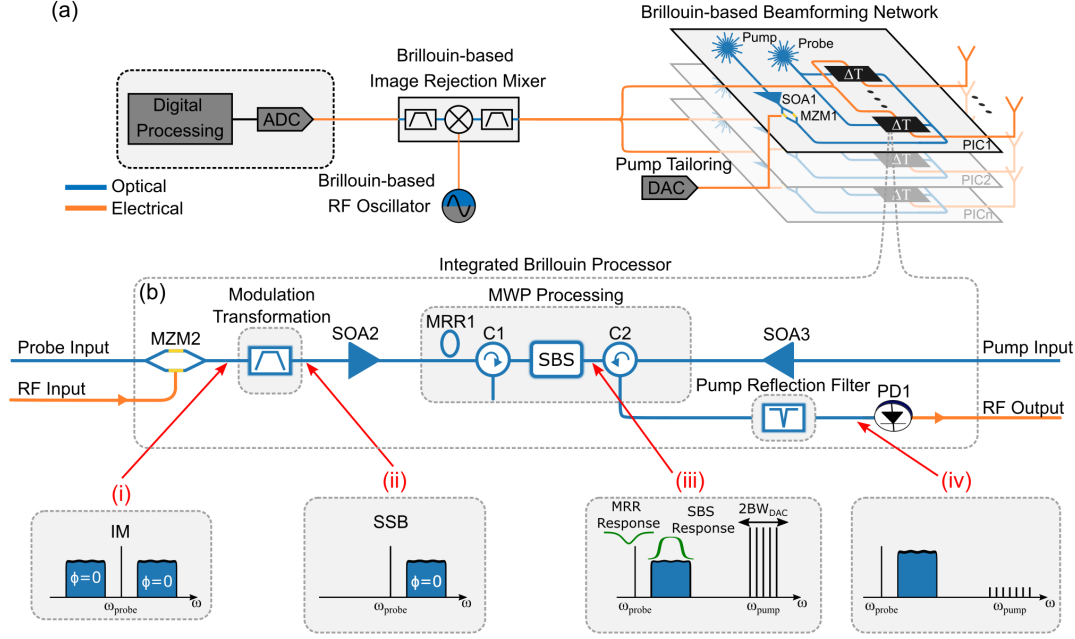


Figure A.4: Phased array antenna enabled by on-chip Brillouin processing. (a) Optical signal distribution scheme for MWP-based PAA processing. (b) Schematic for proposed multi-purpose SBS-enabled MWP beamformer. (i) An RF signal is modulated onto the probe laser. (ii) A modulation transformer generates the required modulation format. (iii) On-chip Brillouin and passive processing. (iv) Processed optical signal with filtered pump reflection. L: laser, MZM: Mach-Zehnder Modulator, LO: local oscillator, BPF: bandpass filter, DAC: digital to analog converter, SOA: semiconductor optical amplifier, SW: switch, MRR: micro-ring resonator, ODL: optical delay line, C: circulator, PD: photodetector

A.2.2 RF performance and SWaP considerations

We provide some additional considerations to the points raised in Section 8.2 regarding the outlook for fully-integrated Brillouin devices in PAA applications.

Care must be taken to minimise reflections of the Brillouin pump, which are caused by refractive index changes between materials. As discussed in Section 4.2, reflections must be minimised in systems that use backward Brillouin processing to avoid degradation to the RF link performance [340]. The vertically tapered interface between Ge:SiO₂ and As₂S₃ that Lai *et al.* demonstrated [341] is an attractive method to achieve low-loss and low reflection interfaces in heterogeneously integrated MWP systems. Demonstrations of similar low-reflection transitions in future integrated Brillouin MWP processing platforms are essential to improve device performance.

To reduce the loss associated with converting between RF and optical domains multiple times, it is desirable to implement all the required functionalities in a single MWP link. Although many MWP processors have been used to demonstrate multiple RF processing functionalities [30], [36], [415], they only demonstrate a single RF processing function at any given time. This motivates future work to implement MWP processors that simultaneously achieve multiple RF processing functionalities in one optimised MWP link. Reza *et al.* performed a theoretical analysis of a fully-integrated MWP beamformer that combines frequency conversion and phase shift beamforming [439]. Through simula-

tions they showed that the overall RF link gain of the multi-functional MWP device was comparable to single function MWP devices, essentially halving the link gain penalty by distributing it between two RF processing functionalities. This highlights the advantages of combining multiple RF processing functionalities into a single MWP link.

It is desirable to reduce the pump power required to generate SBS gain in Brillouin waveguides. Although RF interferometric processing techniques reduce the pump power required by Brillouin-based MWP processors [20], [340], [347], relatively large pump powers are still required to generate Brillouin gain over a wide bandwidth. For example, approximately 350 mW of coupled pump power was used by McKay *et al.* to generate 4 dB of Brillouin gain over a 400 MHz bandwidth [20] in a Ge:SiO₂-As₂S₃ PIC [341]. By analysing (4.6), the magnitude of Brillouin gain generated for a fixed pump power is related to the figure of merit (FOM), $\frac{g_B}{A_{eff}}L_{eff}$, and determines how effective a material platform is at generating Brillouin gain [176]. Increases to both the Brillouin gain coefficient and the effective waveguide length are needed to reduce the required pump power. The former is improved by developing new Brillouin processing platforms and materials, such as fully-etched As₂S₃ waveguides [176], which have larger Brillouin gain coefficients than traditional As₂S₃ rib waveguides. The latter is increased by reducing waveguide losses of the Brillouin medium, which is currently dominated by sidewall scattering [494].

The form factor of each component must be considered to ensure the device is contained in the half-wavelength spacing between PAA elements, which is only 3 mm at 50 GHz. Many components in the design presented in Fig. A.4 are sufficiently compact and introduce no significant form factor limitations. These include MRRs, which have a bend radius of approximately 100 μ m [167], [343], and PDs, which typically have maximum linear dimensions of only tens of microns [206], [297], [450]. Optical gain sections used for SOAs and laser sources are slightly longer with lengths close to 1 mm [206], [249]. While Brillouin As₂S₃ waveguides can require interaction lengths exceeding 6 cm [176], they have been incorporated in compact spirals with dimensions of less than 1 mm² [176], as discussed in Section 4.2. On-chip PPER waveguides typically require shorter interaction lengths, with a recent demonstration using Brillouin waveguides with only 17 mm length [42], which can easily be integrated into a footprint less than 1 mm².

Most work on high-performance modulators to-date has focused on lowering the $V_{\pi_{RF}}$, however, considerations must be made to ensure these devices maintain compact linear dimensions for deployment in space-constrained applications, such as PAAs. For example, the lowest $V_{\pi_{RF}}$ of an E-O modulator is 2.12 V at 50 GHz [263] and has a length of 2 cm, which can be challenging to integrate at each element of a PAA where space is highly constrained. However, folded [495]–[497], or spiral [498] E-O modulators have smaller maximum linear dimension when compared to their straight counterparts, and are attractive solutions for space constrained applications, such as at every element of a PAA. Similarly, acousto-optic non-reciprocal devices have been folded to reduce their on-chip footprint, as demonstrated by Kittlaus *et al.* [388].

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