

MICROWAVE PHOTONICS FOR RANGING AND FILTERING APPLICATIONS

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Microwave Photonics for Ranging and Filtering Applications

by

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Certificate



INDIAN INSTITUTE OF TECHNOLOGY DELHI

I hereby certify that the work which is being presented in the thesis entitled “**Microwave Photonics for Ranging and Filtering Applications**” in the fulfilment of the requirements for the award of the degree of **DOCTOR OF PHILOSOPHY** and submitted in the Department of Electrical Engineering, Indian Institute of Technology Delhi, is an authentic record of my own work carried out during the time period from September 2020 to January 2026 under the supervision of Prof. Amol Choudhary, Associate Professor, Department of Electrical Engineering, Indian Institute of Technology Delhi, India. The matter presented in this thesis has not been submitted by me for the award of any other degree of this or any other institute, and any collaboration with other students with their contributions has been clearly highlighted. Any work that has been published is mentioned under the “List of Publications” on page 186.

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This is to certify that the above statement made by the candidate is correct to the best of my knowledge.

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Abstract

Radio detection and ranging (Radar) is a key enabling technology for sensing, navigation, and situational awareness, facilitating target detection, ranging, and velocity estimation in applications that span air and maritime surveillance, self-driving cars, satellite navigation, remote sensing, and imaging. Modern radars demand high range and velocity resolution, multi-function operation over multiple frequency bands, and higher carrier frequencies. Conventional electronic radar architecture faces bottlenecks in waveform generation and processing of ultra-high time-bandwidth product waveforms. Conventional radars are susceptible to electromagnetic interference and are limited due to bandwidth availability, high jitter, and slow processing speeds. Microwave photonics offer unique characteristics by leveraging the large bandwidth availability by encoding microwave signals into the optical frequency band. MWP enables the generation of large bandwidth waveforms, signal processing such as filtering, phase shifting, and antenna remoting with low-loss optical fiber, immunity from EMI, and compatibility with radio-over-fiber for high-speed data transmission. The primary objective of using this technology is to architect a photonic based, highly reconfigurable radar with tunable frequency and bandwidth, along with highly reconfigurable signal processing, such as filtering in the optical domain, which is difficult to achieve in the RF frequency domain.

The primary focus of this thesis is to address the growing application of microwave photonics (MWP) in radar technology. The thesis examines various aspects of a microwave photonic radar, including the generation of reconfigurable radar waveforms, simultaneous ranging and velocity detection of moving targets, and signal processing for radar applications. A highly frequency-agile dual-chirp waveform generation is demonstrated using a simple, low-bandwidth MZM modulator. A novel multiband complementary waveform for simultaneous range and velocity estimation is

also demonstrated, with experimental validation using a moving toy car in a laboratory environment. Modern radars require operation in multiple frequency bands with different bandwidths to detect targets of varying sizes and velocities. This requires a real-time tunable filter in the microwave frequency band to change the filter's frequency response, bandwidth, or to switch between transmission and rejection filters. A SBS based electromagnetically induced transmission (EIT) analogue and electromagnetically induced absorption (EIA) analogue filters with tunability in frequency and bandwidth are experimentally validated. The thesis also explored chromatic dispersion compensation in a phase-modulated link demodulated with an MZI using stimulated Brillouin scattering. This improved the operational bandwidth of the phase-modulated link for antenna remoting in radar and radio-over-fiber applications. Moreover, the thesis discusses the limitations of the proposed microwave photonic radar and proposes a future research direction for further enhancement.

सार

रेडियो डिटेक्शन एंड रेंजिंग (रेडार) संवेदीकरण, नेविगेशन तथा स्थितिजन्य जागरूकता के लिए एक प्रमुख सक्षम तकनीक है, जो वायु एवं समुद्री निगरानी, स्वचालित वाहन, उपग्रह नेविगेशन, रिमोट सेंसिंग तथा इमेजिंग जैसे अनुप्रयोगों में लक्ष्य पहचान, दूरी मापन और वेग अनुमान को संभव बनाती है। आधुनिक रडार प्रणालियों में उच्च दूरी एवं वेग विभेदन क्षमता, बहु-आवृत्ति बैंडों में बहु-कार्यात्मक संचालन तथा उच्च कैरियर आवृत्तियों की आवश्यकता होती है। परंपरागत इलेक्ट्रॉनिक रडार संरचनाएँ अत्यधिक उच्च टाइम-बैंडविड्थ उत्पाद वाले वेवफॉर्म के निर्माण एवं प्रसंस्करण में बाधाओं का सामना करती हैं। इसके अतिरिक्त, पारंपरिक रडार विद्युतचुंबकीय हस्तक्षेप के प्रति संवेदनशील होते हैं तथा सीमित बैंडविड्थ उपलब्धता, उच्च जितर और धीमी प्रसंस्करण गति के कारण सीमित हो जाते हैं।

माइक्रोवेव फोटोनिक्स (MWP) माइक्रोवेव संकेतों को ऑप्टिकल आवृत्ति बैंड में एन्कोड करके विशाल बैंडविड्थ उपलब्धता का लाभ उठाने वाली विशिष्ट विशेषताएँ प्रदान करता है। MWP के माध्यम से बड़े बैंडविड्थ वाले वेवफॉर्म का निर्माण, फ़िल्टरिंग एवं फेज़ शिफ्टिंग जैसे सिग्नल प्रोसेसिंग कार्य, कम-हानि वाले ऑप्टिकल फाइबर द्वारा एंटीना रिमोटिंग, विद्युतचुंबकीय हस्तक्षेप से प्रतिरक्षा तथा उच्च-गति डेटा संचरण के लिए रेडियो-ओवर-फाइबर के साथ संगतता संभव होती है। इस तकनीक का मुख्य उद्देश्य एक फोटोनिक-आधारित, अत्यधिक पुनर्संरचित रडार संरचना विकसित करना है, जिसमें ट्यून करने योग्य आवृत्ति एवं बैंडविड्थ के साथ-साथ ऑप्टिकल डोमेन में लचीली सिग्नल प्रोसेसिंग संभव हो, जिसे RF आवृत्ति डोमेन में प्राप्त करना कठिन है।

इस शोध-प्रबंध का मुख्य उद्देश्य रडार प्रौद्योगिकी में माइक्रोवेव फोटोनिक्स के बढ़ते अनुप्रयोगों का अध्ययन करना है। शोध-प्रबंध में माइक्रोवेव फोटोनिक रडार के विभिन्न पहलुओं, जैसे पुनर्संरचित रडार वेवफॉर्म का निर्माण, गतिमान लक्ष्यों की एक साथ दूरी एवं वेग पहचान, तथा रडार अनुप्रयोगों के लिए सिग्नल प्रोसेसिंग, का विश्लेषण किया गया है। एक सरल, निम्न-बैंडविड्थ MZM मॉड्युलेटर का उपयोग करते हुए अत्यधिक आवृत्ति-चपल द्वि-चिप वेवफॉर्म निर्माण का प्रदर्शन किया गया है। इसके अतिरिक्त, एक नवीन मल्टीबैंड पूरक वेवफॉर्म को एक साथ दूरी और वेग अनुमान के लिए प्रस्तावित किया गया है, जिसका प्रायोगिक सत्यापन प्रयोगशाला वातावरण में एक गतिमान खिलौना कार के माध्यम से किया गया है।

आधुनिक रडार प्रणालियों को विभिन्न आकार और वेग वाले लक्ष्यों की पहचान के लिए विभिन्न बैंडविड्थ के साथ अनेक आवृत्ति बैंडों में संचालन की आवश्यकता होती है। इसके लिए माइक्रोवेव आवृत्ति बैंड में वास्तविक-समय ट्यून करने योग्य फ़िल्टर आवश्यक होते हैं, जिनके माध्यम से फ़िल्टर की आवृत्ति प्रतिक्रिया, बैंडविड्थ को बदला जा सके अथवा ट्रांसमिशन और रिजेक्शन फ़िल्टरों के बीच स्विच किया जा सके। इस शोध-प्रबंध में स्टिम्युलेटेड ब्रिल्युइन स्कैटरिंग (SBS) आधारित विद्युतचुंबकीय प्रेरित संचरण (EIT) तथा विद्युतचुंबकीय प्रेरित अवशोषण (EIA) समकक्ष फ़िल्टरों का आवृत्ति एवं बैंडविड्थ में ट्यून करने योग्य प्रायोगिक सत्यापन प्रस्तुत किया गया है। साथ ही, स्टिम्युलेटेड ब्रिल्युइन स्कैटरिंग का उपयोग करते हुए MZI द्वारा डीमॉड्युलेट किए गए फेज़-मॉड्युलेटेड लिंक में क्रोमैटिक डिस्पर्सन क्षतिपूर्ति का अध्ययन किया गया है, जिससे रडार एवं रेडियो-ओवर-फाइबर अनुप्रयोगों के लिए एंटीना रिमोटिंग की परिचालन बैंडविड्थ में

सुधार प्राप्त हुआ है। अंततः, शोध-प्रबंध में प्रस्तावित माइक्रोवेव फोटोनिक रडार की सीमाओं पर चर्चा की गई है तथा भविष्य में इसके और अधिक उन्नयन के लिए संभावित अनुसंधान दिशाएँ प्रस्तुत की गई हैं।

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List of Acronyms

A

ADC	Analog-to-Digital Converters
APL	Analog Photonic Link
AF	Ambiguity Function
AWG	Arbitrary Waveform Generator

B

BW	Bandwidth
----	-----------

C

CW	Continuous Wave
----	-----------------

D

DAC	Digital to Analog Converter
dB	Decibel
DDMZM	Dual Drive Mach Zehnder Modulator
DPMZM	Dual-Parallel-Mach-Zehnder-Modulator
DPol-DPMZM	Dual-Polarization Dual parallel Mach-Zehnder Modulator
DFB	Distributed Feedback Laser

E

EIT	Electromagnetic induced Transmission
EIA	Electromagnetic induced Absorption
EMI	Electromagnetic Interference
E/O	Electro-optic Modulators
ESA	Electrical Spectrum Analyzer

F

FBG	Fiber Bragg Gratings
FMCW	Frequency Modulated Continuous Wave
FWHM	Full-Width Half Maxima

G

GHz	Giga Hertz
-----	------------

I

IM	Intensity Modulation
ISAR	Inverse Synthetic Aperture Radar

L

LD	Laser Diode
LFM	Linear Frequency Modulated
LSB	Lower Sideband

M

MaTP	Maximum Transmission Point
miTP	Minimum Transmission Point
MHz	Mega Hertz
MLL	Mode Lock Laser
MZM	Mach Zehnder Modulator
MWP	Microwave Photonics
MTI	Moving Target Indication

N

NF	Noise Figure
----	--------------

O

O/E	Optical to Electrical
OSA	Optical Spectrum Analyser

P

PD	Photodiode
PolM	Polarization Modulator
PM	Power Meter
PS	Power Splitter

Q

Q-Point	Quadrature Point
---------	------------------

R

RADAR

Radio Detection and Ranging

RR	Range Resolution
----	------------------

S

SA Spectrum Analyser

SAR Synthetic Aperture Radar

SBS

Stimulated Brillouin Scattering

SH Spectra hole

SFCW

Stepped Frequency Continuous Wave

T

TBWP Time-Bandwidth Product

THz Tera Hertz

U

USB Upper Sideband

V

VNA Vector Network Analyser

VSG Vector Signal Generator

W

WDM Wavelength Division Multiplexing

Chapter 1

Introduction

1.1 Background

Radio Detection and Ranging (RADAR) is a sensing technology that transmits electromagnetic waves and uses the echo of electromagnetic waves to estimate the range, velocity, and image of target objects such as aircraft, unmanned air vehicles, automobiles, and clouds over land, air, and sea, as illustrated in Fig. 1.1 [1]. Radars are utilized for surveillance, navigation, air traffic control, self-driving cars, fire control, earth exploration, and weather monitoring.

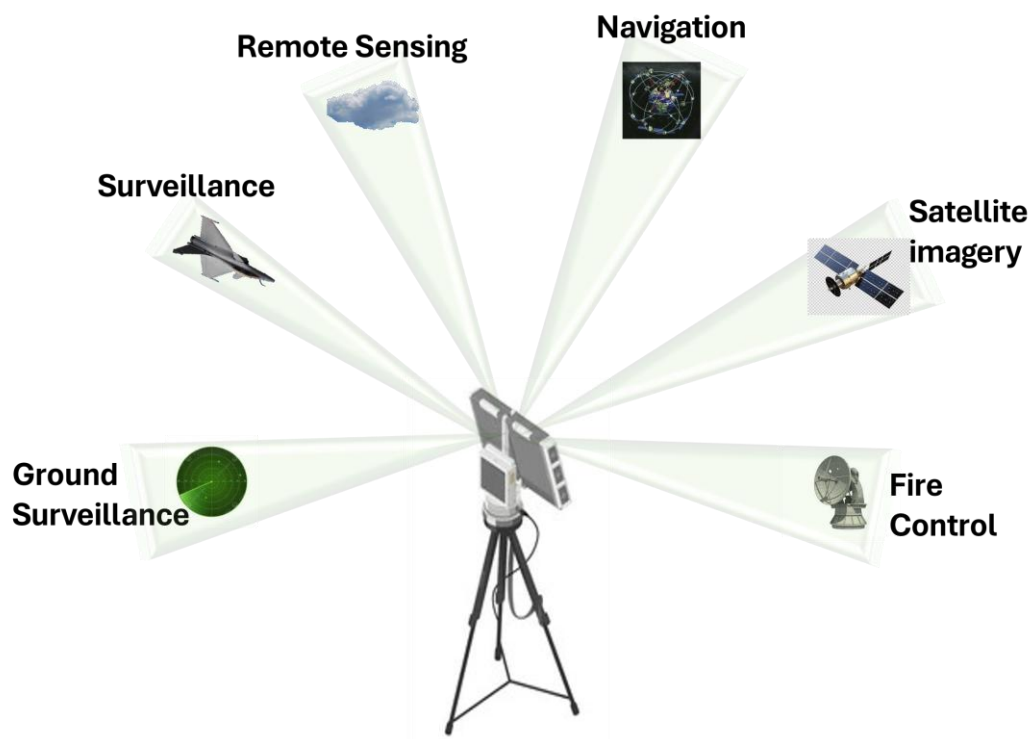


Fig. 1.1: Radar applications

Radar technology has progressed significantly over the last few decades, enhancing automobile technology and leading to the development of driverless cars, which in turn improve navigation accuracy and situational awareness [2]. Radar technology is continuously evolving to meet the unprecedented requirements of detecting high-speed aircraft, small drones, and enhancing medical technology by measuring live breathing patterns.

1.2 Challenges with electrical radar technology

Modern radar technology requires the transmission and reception of radio and microwave frequency signals with unprecedented features compared to traditional microwave technologies [3]. The next generation of radar systems requires characteristics such as waveforms with large time-bandwidth products, low weight, low losses, real-time tunability, immunity to electromagnetic interference (EMI), and flexible bandwidth, all of which are difficult to achieve with traditional electrical methods [4]. Consequently, the detection of low radar cross-section targets, such as small, slow-moving drones and autonomous vehicles, has become increasingly difficult. Moreover, the tunability in frequency and bandwidth of the waveform is challenging using conventional electronics.

The radar's capability to distinguish two closely spaced targets requires high-resolution radars. A linear frequency-modulated (LFM) signal is used to transmit a waveform with a higher time-bandwidth product, thereby improving range and Doppler resolution [5]. The increase in bandwidth improves the time resolution to obtain highly resolved targets, and an increase in the duration of the transmitted waveform enhances the frequency resolution to detect small Doppler shifts for slowly moving targets. The generation of high time-bandwidth product waveforms requires increased complexity in digital-to-analog converters (DACs). This necessitates the exploration of the development of radar with newer technologies, which can enhance radar

capabilities.

1.3 Microwave Photonic Radar

Microwave photonics (MWP) utilizes lightwave technology to transmit, receive, and process microwave signals [6] [7]. Microwave photonics encodes microwave frequency signals onto optical signals and processes them in the optical domain [8]-[9]. The architecture of a photonic radar is illustrated in Fig. 1.2. Waveform generation is performed using optical modulators, where frequency agility is achieved using tunable lasers or other reconfigurable elements. The signal processing techniques, such as highly configurable filters and phase shifters, are performed in the optical domain, reducing bulky electronics-based signal processing and making the MWP highly scalable [10]-[11]. These features enable radars with high processing capabilities for detecting long-range and small-sized targets. This technology holds immense potential across a wide spectrum of applications, including broadband wireless communications, radio-over-fiber (RoF) systems, free-space optical communication links, satellite communications, photonic analog-to-digital conversion, and radar systems.

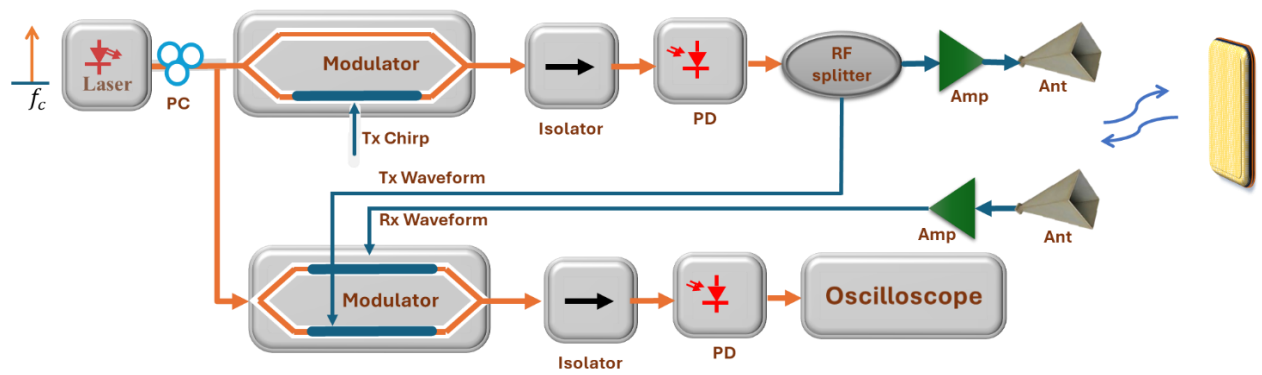


Fig. 1.2: Microwave photonic radar architecture

In a microwave photonic (MWP) radar, the optical carrier is generated by a laser source. The RF signal is then applied to an electro-optic modulator, resulting in up-conversion of the RF

signal to the optical domain. This modulation process generally produces a pair of optical sidebands symmetrically positioned around the carrier frequency. The modulated optical signal is subsequently coupled to a photonic signal processor. Within the photonic signal processor, the modulated optical signal is further processed using various photonic components such as filters, delay lines, or wavelength converters to alter the spectral and temporal characteristics of the optical sidebands. After processing, the optical signal undergoes down-conversion at the photodetector, where the sidebands interact with the optical carrier to reconstruct the processed waveform signal in the electrical domain for transmission in air.

The received echo waveform, along with the reference transmitted waveform, is further coupled to another electro-optic modulator, resulting in the upconversion of both transmitted and received waveform signals to the optical domain. The modulated sidebands beat at the photodetector to generate a low-frequency de-chirped signal, which is further processed using a low-speed ADC to generate the range and velocity of the target object.

1.3.1 Subsystem of Microwave Photonic Radar

In a conventional radar, most operations are performed electrically, including waveform generation, frequency upconversion, distribution to antennas, reception, mixing, filtering, and signal processing. In a microwave photonic radar, some of these functionalities are shifted into the optical domain due to inherent characteristics of photonics, such as large available bandwidth, low propagation loss, and immunity to EMI. The core subsystem of MWP radar consists of a transmitter, a propagation medium, and a receiver system.

The transmitter sub-system consists of waveform generation, transmission of the generated waveform through the antenna, reception of the received echo waveform through the antenna, de-chirping at

the receiver to generate beating products, and signal processing to generate range, velocity, and direction of motion of the target.

1.3.1.1 MWP Transmitter

The transmitter sub-system consists of waveform generation and transmission of the generated waveform through the antenna into the propagation medium, which is air. In an MWP radar, waveform generation governs the achievable range resolution, velocity resolution, ambiguity performance, and operational flexibility, such as frequency and bandwidth tunability. Unlike conventional electronic approaches, where bandwidth and frequency agility are constrained by electronic components, MWP enables the synthesis of ultra-wideband and highly reconfigurable microwave signals by leveraging the large carrier frequency and low loss of optical systems. A narrow linewidth laser acts as an optical carrier, which is modulated using an electro-optical converter driven by a linear frequency modulated signal to generate an LFM waveform at the photodetector.

The type of waveform used is important for parameter estimation, such as range, velocity, and AoA. The LFM waveforms are primarily used in modern radar due to their inherent pulse compression capability, enabling high range resolution. The single LFM [29] [30] has high range-Doppler coupling, resulting in range ambiguity with the target at higher velocity. A dual chirp waveform [31] has been proposed with an up-chirp and a down-chirp at the same starting frequency, resulting in low range-Doppler coupling. However, a dual-chirp waveform can not detect the direction of motion of the target simultaneously. A triangular chirp [32] consisting of an upchirp followed by a down-chirp is proposed to detect range and velocity simultaneously. Various other techniques, such as a multiband LFM waveform [33] and a multiband triangular waveform [34], have been proposed for simultaneous range and velocity estimation.

1.3.1.2 MWP Receiver

The echo waveform reflected from the target object is received by the receiver antenna. The echo waveform and reference transmitted waveform are further coupled to an optical modulator for upconversion into the optical frequency band. The optical modulator is driven by a laser source. The up-converted echo waveform is de-chirped at the photodetector to generate beating frequencies at low frequency, which is further processed after analog to digital conversion by ADC to estimate range and velocity. The de-chirping of the echo waveform is demonstrated using MZM [35] and DDMZM modulator [36].

The MWP technology-based photonic radar system is a promising solution for the next generation of radar technology. The following points outline the advantages of MWP enabled radar systems,

1. **High range resolution:** MWP utilizes lightwave technology to generate and process large bandwidth signals. The range resolution improves with an increase in bandwidth, enabling the detection of small target objects and enhancing the detection of two closely located targets [12].
2. **Reconfigurable waveform generation:** The operational capability of the radar is primarily influenced by the waveform utilized for target detection. MWP offers the capability to generate a highly reconfigurable waveform with tunability in frequency and bandwidth [13]. Furthermore, MWP technology enables the generation of complex waveforms for simultaneous range and velocity estimation, improving the radar efficacy for multifunctional applications.

3. **Low loss:** MWP utilizes optical fiber for signal transportation. The loss in fiber is very low, $\sim 0.2\text{dB/km}$, compared to electrical cables. This enables secure transmission of data over longer distances without any significant attenuation.
4. **Antenna Remoting:** The low loss in optical fibre enables the separation of the antenna system from the signal processing system. This enables easier operation of the radar in difficult terrain, improving the target detection range and minimizing the signal degradation.
5. **Immunity to Electromagnetic Interference:** MWP-based communication systems are inherently immune to electromagnetic interference because light waves, consisting of photons, do not carry any electric charge.
6. **Enhanced immunity to Jamming:** MWP enabled communication systems use optical fibre for transporting information, reducing the interception and jamming. Optical signals are difficult to intercept without access to optical fiber, enhancing security.

Based on the above properties, MWP enabled photonic radar systems address key challenges of conventional radar, such as limited bandwidth availability, system compactness, vulnerability to jamming and interference, enhanced security, and reduced ADC and DAC requirements. These advantages make MWP an ideal technology to enhance the capabilities of radar systems for a wide range of applications.

A majority of microwave photonic radars demonstrated so far utilize MWP for waveform generation, simultaneous range and velocity estimation, angle of arrival estimation, and signal processing, including filtering and phase shifters. The FMCW radar utilizes a linear frequency modulated waveform for target range estimation. A dual-chirp waveform has been demonstrated for simultaneous range and speed estimation. The simultaneous range and velocity are demonstrated using a multi-band LFM waveform. Angle of arrival has been demonstrated by

measuring the phase difference between two antennas. Filters and phase shifters based on Mach-Zehnder interferometers, Bragg gratings, and stimulated Brillouin scattering have been demonstrated.

Recently, some techniques have been demonstrated to generate a tunable waveform by modulating the LFM signal using a dual-polarization dual-parallel Mach-Zehnder modulator (DP-DPMZM) [13] and a DDMZM modulator [17]. However, the bandwidth of the modulator required increases with the increase in frequency of operation of the radar, and a high-frequency tone is required to generate a waveform at higher frequencies. Some techniques using semiconductor lasers are proposed, but the system becomes very complex. The objective is to generate a highly tunable waveform by tuning a CW laser using a low-bandwidth modulator for a low cost highly tunable radar system.

Lately, various techniques have been demonstrated to detect both range and velocity using a Triangular chirp simultaneously; however, this requires complex time-frequency analysis to determine the direction of motion of the target. *One of the objectives of this thesis is to develop a new waveform that can estimate both range and velocity without complex signal processing and exhibit low range-Doppler coupling.*

The new generation of radars requires multi-performance capability, which necessitates tunability in frequency and bandwidth of the waveform to cater to applications such as surveillance, tracking, and imaging using the same radar setup. This required an agile optical filter with tunability in frequency of operation and bandwidth. *Another objective of the thesis is to generate a highly flexible filter which can switch from a bandpass to band-reject filter and is tunable in frequency of operation and transmission or rejection bandwidth.*

The phase modulator has been widely used in optical communication systems for high-speed data transport. The phase modulator also does not require any DC bias. This enables us to use phase modulator for antenna remoting applications, where optical signal is transported to remote antenna using optical fiber. The phase modulator at remote antenna system converts optical to RF during transmission and RF to optical during reception without requiring any DC power. However, the phase-modulated signal has a π phase shift between sidebands, resulting in a null at the photodetector. It requires a phase-to-intensity converter, which decreases the operational bandwidth of the link. *Another objective of the thesis is to use a method to improve the operational bandwidth of the phase-modulated link for radar applications.*

1.4 Key achievements

The recent progress in microwave photonics-based signal processing for microwave applications justifies the need for further exploration of FMCW radar using photonic technology. Microwave photonics enables the generation of highly tunable waveforms and high signal processing capabilities compared to traditional electrical systems. The subsequent statements outline the research goals that have been completed, related to waveform generation, simultaneous range and velocity estimation of the target, and signal processing using stimulated Brillouin scattering.

1.4.1 Photonically generated radar waveforms for simultaneous range and velocity

A straightforward technique for generating a highly tunable dual-chirp waveform through an analog photonic link using a Mach-Zehnder modulator is experimentally demonstrated. The tunability in center frequency of the waveform is achieved by coupling a CW laser before the photodetector. The opto-electronic mixing between chirp-modulated sidebands at MZM with an external laser at PD generates a dual chirp waveform with frequency tuned by tuning the external

CW laser. The dual-chirp waveform, spanning from 5 GHz to 20 GHz, has been experimentally demonstrated. This presents a significant opportunity for multipurpose radar, where the frequency of operation must be tuned for different applications.

The simultaneous range and velocity estimation is important for real-time radar. A multiband complementary dual chirp waveform is presented to estimate both range and velocity, including the direction of motion of the target. The generation of a multiband complementary dual chirp waveform photonically using a single dual-polarization dual-parallel Mach-Zehnder modulator (DPol-DPMZM) is demonstrated with tunable frequency. The waveform consists of a dual chirp in the lower frequency band for high accuracy in range estimation and a complementary dual chirp at a higher frequency for high resolution velocity estimation, due to the higher Doppler frequency shift at higher frequencies. Experiments have been conducted in emulation to validate the concept. The range and velocity have been estimated with an accuracy of 50 cm in range and 0.7 m/s in velocity for a target at 100m moving with a velocity of ± 60 m/s. Furthermore, experiments have been conducted to determine the velocity of a toy car with an accuracy of 0.3 m/s. We have demonstrated waveform generation along with range and velocity estimation for the realization of a reconfigurable microwave photonic radar.

1.4.2 Highly flexible RF filter response using stimulated Brillouin scattering for radar

Radar needs to operate at different frequency bands with waveforms of varying bandwidths to detect targets such as a bird or a slow-moving, small-sized drone. This also requires filters to tune in both the frequency of operation and the transmission or rejection bandwidth. A highly flexible electromagnetically induced transmission (EIT) analogue and electromagnetically induced absorption (EIA) analogue filters in the microwave frequency band are demonstrated using

stimulated Brillouin scattering. In the proposed method, two pump lasers generate an SBS anti-Stokes response on LSB and a SBS anti-Stokes response on USB of an analog photonic link with a phase modulator. The SBS anti-Stokes response on LSB is shifted towards the right or left to switch between electromagnetically induced transparency (EIT) [24]-[27] or electromagnetically induced absorption (EIA) analogue filter response [28] with very low pump power, making the filter highly power-efficient. The tunability in bandwidth for both EIT and EIA analogue filters is achieved by tuning the pump laser frequency. The spectra hole (SH) filter response is also demonstrated by tuning the pump power levels to achieve different rejection ratios in different frequency bands. The generated filter can be used in radar, high-speed optical communication links, satellite communication, and interference rejection.

1.4.3 Null compensation in a phase-modulated link with Stimulated Brillouin scattering for antenna remoting applications

A phase-modulated link demodulated with a Mach-Zehnder interferometer suffers from chromatic dispersion, reducing the operational bandwidth of the link, which is important for high-speed radar applications. The periodic null due to chromatic dispersion created by path length differences in MZI arms is compensated by placing the stimulated Brillouin gain response on the desired null frequency band. We have also demonstrated experimentally multiple null compensation with 3 multiple pump lasers for applications such as multiband radar signal transmission. The phase modulator does not require a DC bias, which improves jitter caused by variations in DC bias due to environmental conditions, making it suitable for antenna remoting with optical signal transported to a remote antenna system using optical fibre, which is further converted back to RF signal for transmission or received RF signal is converted to optical to send the received signal to signal processing location. where DC power is unavailable.

1.5 Thesis Structure

The thesis is structured into six chapters in total. Each chapter provides a brief overview of the work presented, as follows:

Chapter 1 provides a brief overview of radar systems and their applications, followed by an analysis of bottlenecks in traditional electronic-based approaches and highlights the advantages of microwave photonics in addressing these challenges. It also highlights the key accomplishments of this project.

Chapter 2 discusses the relevant technological background for this thesis. This includes radar principles, classification of radar, introduction of MWP technology, overview of stimulated Brillouin scattering, and an SBS based signal processor.

Chapter 3 presents the photonically generated highly tunable dual chirp waveform generation using a tunable reference laser, the waveform generation of a multiband complementary V-shaped waveform for simultaneous range and velocity estimation, including the direction of motion of the target. Experimental validation of a multiband V-shaped waveform for simultaneous range and velocity estimation of a slow-moving target.

Chapter 4 presents the generation of highly tunable electromagnetically induced transparency (EIT) analogue and electromagnetically induced absorption and spectra hole (SH) analogue filters using stimulated Brillouin scattering. The real-time switching between EIT and EIA is also demonstrated by frequency shifting of a pump laser. The variation in bandwidth, frequency of operation, and rejection ratio of the filters is achieved by tuning the frequency shift and pump powers.

Chapter 5 presents the null compensation in a phase modulated (PM) link demodulated with a Mach-Zehnder interferometer (MZI). The periodic nulls generated by the PM+MZI link are compensated by stimulated Brillouin scattering gain response on the specific null frequency band by tuning the pump laser. The multiple null compensation is also demonstrated by generating multiple pumps with an intensity modulator.

Chapter 6 provides a comprehensive summary of the main findings and contributions of this thesis and outlines potential directions for future work.

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Chapter 2

Technical Background

Modern radar technology fused with emerging microwave photonics (MWP) technology enables the transmission and reception of radio and microwave frequency signals with unprecedented features compared to traditional microwave technologies. The next generation of communication systems requires characteristics such as large-time products, low weight, low losses, real-time tunability, immunity to electromagnetic interference (EMI), and flexible bandwidth, all of which are challenging to achieve with traditional electrical methods. Microwave photonics encodes microwave frequency signals onto optical signals and processes them in the optical domain. Signal processing techniques, like highly configurable filters and phase shifters, are performed in the optical domain, reducing bulky electronics-based signal processing and making the MWP highly scalable. These features enable the radar to achieve high processing capability for long-range and small-size target detection, without compromising quality metrics.

This chapter focuses on the principle of radar, radar waveforms for measuring target characteristics, recent progress and advancements in microwave photonic radar technology to achieve significant improvements over conventional radar, including high-resolution target detection along with precise range and velocity detection. Additionally, MWP-enabled optical signal processing is explored to achieve high-bandwidth, reconfigurable filters for FMCW radar applications.

2.1 Overview of Radar

Radar, an abbreviation for radio detection and ranging, transmits an electromagnetic wave and uses the echo electromagnetic wave from target objects to search, track, and identify targets through imaging. The detection of small-sized targets with high resolution requires higher frequency and bandwidth. The bandwidth of the transmitting waveform defines the range resolution of the target, enabling radar to detect close-by targets. The delay between the transmitting and received waveform is harnessed to determine the range of the target, and the Doppler shift in the received echo waveform is utilized to determine the velocity of the target object [1]. The angle of arrival of the target is determined by estimating the phase shift between the transmitted and received waveforms [2]. The shape of the target is determined by synthetic aperture radar imaging, which utilizes the variation in frequency shift as the target approaches or recedes from the radar [3] [4].

The radar capability to distinguish two closely spaced targets requires high-resolution radar [5]. The range resolution is a function of the time-bandwidth product of the transmitted waveform. A linear frequency-modulated signal is used to transmit a waveform with a higher time-bandwidth product, thereby improving the range and Doppler resolution. The increase in bandwidth improves the time resolution to obtain highly resolved targets within range, and an increase in the duration of the transmitted waveform enhances the frequency resolution to detect small Doppler shifts for slowly moving targets. The resolution of the target object's arrival angle is determined by the antenna beamwidth and its aperture.

The operating frequency of the transmitted waveform is also an important parameter of the radar, which varies according to the application, such as surveillance, tracking, or imaging of the target

object. The higher the frequency of the radar, the higher the Doppler shift, which enables the detection of a slow-moving target object by increasing the Doppler shift. The tunability in operating frequency enables radar to detect simultaneously both a slow-moving and a very high-speed target by tuning the frequency [6]. The multiband waveforms can also be utilized to detect different target objects with different band waveforms for different ranges and speeds of the target [7].

The present-day low-speed low radar cross-section (RCS) targets require very high-resolution radar capable of measuring a very small change in frequency due to low Doppler shift. The frequency of operation of the radar needs to be higher to receive a higher Doppler shift with slow-moving targets. The generation of a high-bandwidth waveform at a higher frequency is difficult to generate using a conventional electronics-based approach. The electronic-based approach for radar systems requires a complex, bulky system for waveform generation and is limited by available digital-to-analog converters DACs. MWP based signal processing is very useful in processing the high time bandwidth product signals, decreasing the load on electronics, and improving the size and complexity of the system [8].

Research has been conducted in MWP for signal processing like highly reconfigurable filters [9], phase shifters [10], arbitrary waveform generation [11], high sampling capability, and frequency measurement for applications in phase array antennas, radar [12], mobile communication networks [13], and radio-over-fiber networks [14]. The MWP can be used in radar system for waveform generation, signal processing, and antenna remoting applications significantly improving the radar capabilities.

2.2 Classification of Radar

The radar systems are classified primarily into two categories, continuous wave (CW) and pulse wave radar. The CW radar transmits continuously, which requires a high average transmitted power, whereas pulse radar transmits in bursts, resulting in a low average power. Pulsed radar is used for long-range target surveillance, and CW radar is used for velocity tracking.

2.2.1 Continuous wave Radar

The continuous wave radar is developed to simultaneously transmit and receive radio wave signals, requiring two antennas, one each for transmission and reception. The unmodulated CW radar is primarily utilized to measure the velocity of the target by estimating the Doppler shift between the transmitted and received echo waveforms, as shown in Fig. 2.1. The range of the targets cannot be estimated using CW radar. However, frequency modulated continuous wave radar (FMCW) utilizes a frequency modulated waveform as a transmitting waveform. The FMCW radar estimates range by estimating the delay and velocity by estimating the Doppler shift.

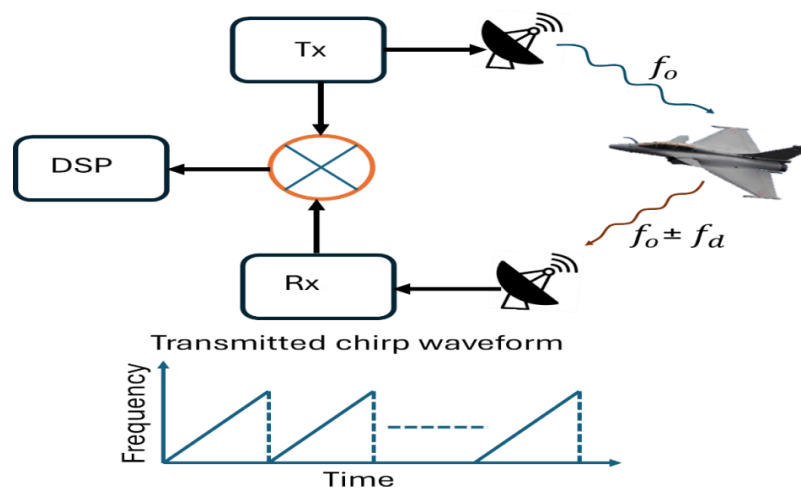


Fig. 2.1: Continuous wave radar, Tx: transmitter, Rx: receiver,

2.2.2 Pulsed Radar

Pulsed radar transmits short RF pulses and then listens for the echo from the target objects. The pulse waveform is transmitted at a fixed pulse repetition interval (PRI) as shown in Fig. 2.2. The target range is estimated from the delay relative to the transmitted pulse as $\left(R = c \frac{\Delta t}{2}\right)$. The target velocity is estimated using pulse-Doppler processing to obtain the Doppler frequency shift, thereby determining the target's velocity. The PRI sets the limits on the unambiguous range and unambiguous Doppler. In pulse radar, during transmission, the radar cannot receive echoes, resulting in a minimum range below which the target cannot be detected.

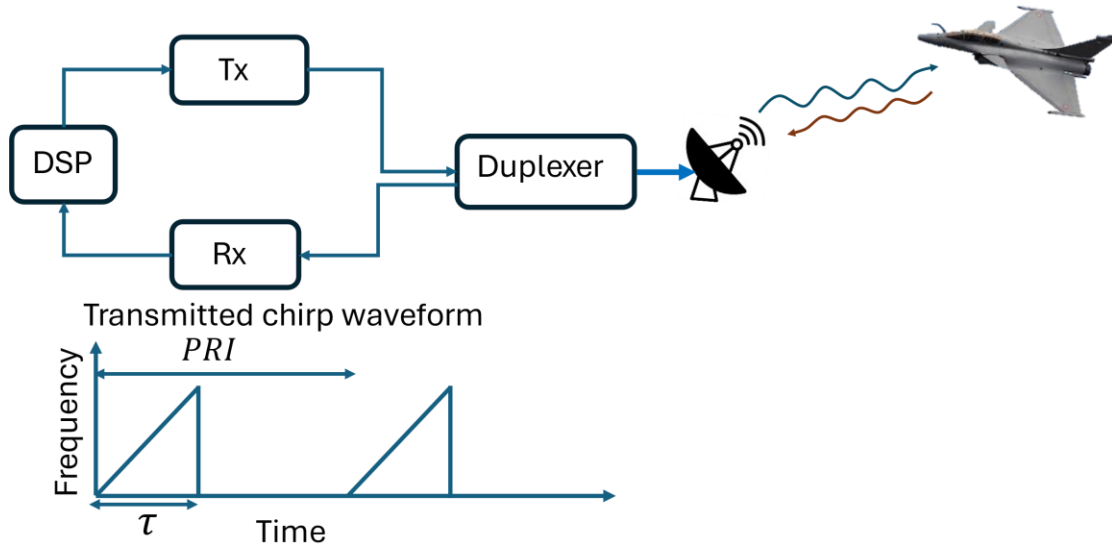


Fig. 2.2: Pulsed Radar, Tx: transmitter, Rx: receiver, PRI: pulse repetition interval

The pulsed waveform-based radar can have blind spots as it is not transmitted while waiting for an echo, whereas CW radar does not have blind spots and can be used for both short range and long-range target estimation, tracking, imaging, and early warning radars. The FMCW radar measures both range and velocity simultaneously [14], further enhancing the real-time

performance of the radar.

The current research work is focused on the development of an FMCW radar due to its inherent capability to measure range and velocity for diverse applications such as short-range navigation, aircraft detection, vehicle speed detection, automobile cruise control, battlefield monitoring, and target imaging. CW radar is a valuable resource for very precise and real-time monitoring and situational awareness.

The unmodulated CW radar transmits a single tone to measure the velocity of the target by estimating the Doppler between the transmitted and received frequencies. The range of the targets cannot be estimated as no time information is transmitted to estimate the delay. The frequency modulated CW radar transmits a linear frequency modulated (LFM) waveform with a temporal width (T) and chirp rate (k) determining the bandwidth of the waveform. To achieve long-range target detection, the temporal width needs to be increased, whereas high range resolution requires a small temporal width. LFM waveform resolves this issue due to its inherent pulse compression capability, significantly improving the capability to resolve nearby targets by increasing the range resolution through an increase in the bandwidth of the waveform ($RR = R = c/2B$). The range resolution of FMCW radar enables it to excel in many applications, such as search and rescue operations, aerospace, radar altimetry, imaging, security, surveillance, traffic monitoring, weather monitoring, and many more. While FMCW radar has numerous advantages, there are some limitations compared to other radar technologies, such as pulse-Doppler radar. FMCW radar needs to increase the temporal width of the transmitted waveform to extend range, which limits its usage for long-range target detection, while pulse-Doppler radar can detect long-range targets by transmitting a shorter pulse. The FMCW radar is less power efficient due to continuous transmission, limiting its usage in remote areas. However, FMCW radar technology's capability to

simultaneously measure both range and velocity with high range resolution, along with no blind spots, makes it a reliable technology for a wide range of applications from surveillance to target imaging.

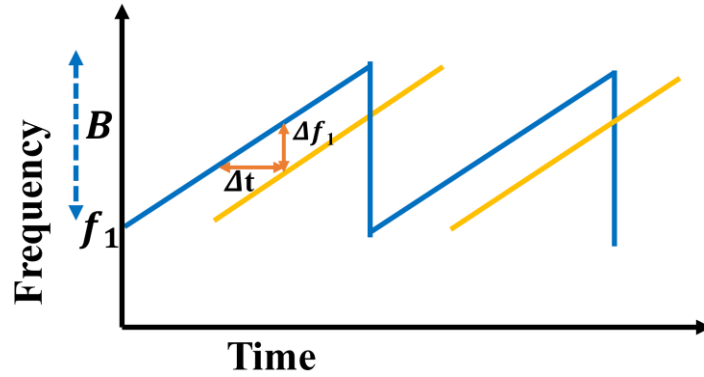


Fig. 2.3: Time-Frequency relationship of LFM waveform in an FMCW radar

The time-frequency relationship of the beating of the transmitted and received echo waveform in an FMCW radar is shown in Fig. 2.3. Here, an LFM waveform with bandwidth (B) and starting frequency f_1 is transmitted, shown in blue, and the received echo waveform is shown in yellow, mixed at the receiver to generate Δf_1 and Δt . The Doppler shift (f_d) is estimated from Δf_1 to estimate the velocity and delay Δt is used to estimate the target range.

$$f_d = 2 \frac{v}{c} \quad (2.1)$$

$$R = c \frac{\Delta t}{2} \quad (2.2)$$

The angle of arrival (AoA) of the target object is estimated by spatial estimation using multiple spatially separated receiver antennae by measuring the phase differences of the beat signal across the array. This is highly useful for self-driving cars to track multiple vehicle types and estimate its position and direction of motion relative to radar sensors. AoA for a uniform linear array is

represented as,

$$\theta \approx \sin^{-1} \left(\frac{\lambda}{2\pi d} \Delta\phi \right) \quad (2.3)$$

Here, λ is the wavelength of the signal, d is the distance between the receive antenna and the $\Delta\phi$ is the phase shift between the echo signal received at different antennas.

The capability of the radar system to detect nearby targets is determined by its range resolution. The range resolution of FMCW radar is inversely proportional to the bandwidth of the waveform. The velocity of the multiple targets with nearly the same speed is distinguished by Doppler resolution. The Doppler resolution depends on the temporal width of the transmitted chirp waveform. The increase in temporal width of the transmitted waveform decreases the chirp slope, which in turn decreases the range beat, resulting in a higher range error. The range and velocity are coupled, referred to as range-Doppler coupling, resulting in an increase in range error with high velocity resolution or an increase in velocity error with higher range accuracy. The range error due to Doppler coupling in an FMCW radar is represented as,

$$\Delta R = \frac{cT}{2B} f_d \quad (2.4)$$

Here, ΔR is the range error, c is the velocity of light, T is the temporal width of the chirp signal, B is the bandwidth, and f_d is the Doppler frequency shift.

The range-Doppler coupling estimation is necessary to estimate the range resolution, Doppler resolution, and range-Doppler coupling. The ambiguity function (AF), which utilizes autocorrelation between transmitted and echo signals, is primarily used to graphically estimate range resolution, Doppler resolution, and range-Doppler coupling [15].

The ambiguity function is represented as a two-dimensional Fourier transform of the autocorrelation function of the transmitted waveform. It demonstrates the correlation between the transmitted and received waveform as a function of delay and Doppler frequency shift. The mathematical expression of the autocorrelation function is represented as,

$$\chi(\tau, \nu) = \int_{-\infty}^{\infty} x(t)x^*(t - \tau)e^{-j2\pi\nu t} dt \quad (2.5)$$

Here, $\chi(\tau, \nu)$ is the ambiguity function, $x(t)$ represents the transmitted chirp signal, τ is the delay, and ν is the velocity of the target. The ambiguity in range is represented by the range spectrum $\chi(\tau, 0)$ and in velocity is represented by the Doppler spectrum $\chi(0, \nu)$. $\chi(\tau, 0)$ denotes the range resolution and $\chi(0, \nu)$ represents the Doppler resolution. The narrower the main lobe in the range(delay) spectrum, the better the range resolution, while a narrower main lobe in the Doppler spectrum represents better velocity resolution. The sidelobes can also be analyzed using the ambiguity function, which can impact the accuracy of target detection. The ambiguity function enables the waveform performance over range and velocity, which can be used to enhance the performance of the radar system.

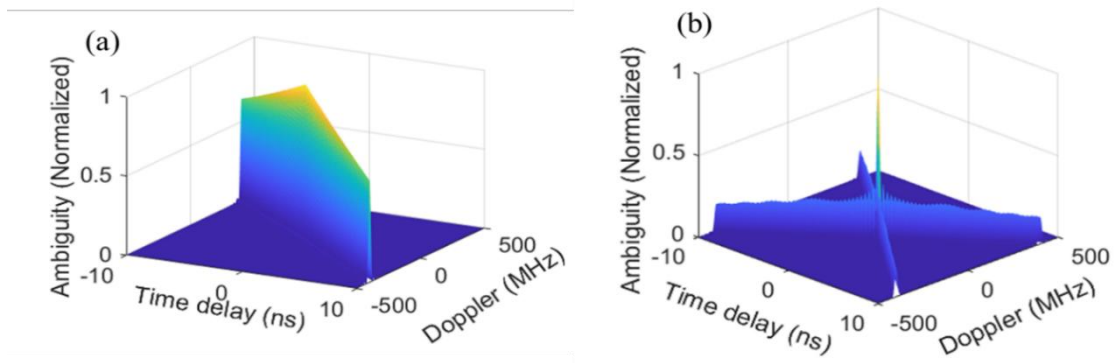


Fig. 2.4 : (a) Ambiguity function to estimate delay (range) and Doppler coupling using a single LFM waveform, (b) Ambiguity function to estimate the delay(range) and Doppler (velocity) coupling using a dual-chirp waveform

Fig. 2.4(a) represents the ambiguity function for a single LFM signal. The knife-edge ambiguity function represents high range-Doppler coupling, resulting in higher range error with an increase in Doppler frequency shift. The ambiguity function of a dual chirp is shown in Fig. 2.4(b). The ambiguity function of a dual-chirp waveform has a small wedge, resulting in low range-Doppler coupling. The low main lobe width represents a high range and Doppler resolution. The dual-chirp waveform can be utilized for high-accuracy range and velocity estimation.

The conventional radar system utilizes electronic-based solutions for waveform generation, detection, and signal processing techniques, such as pulse compression, frequency modulation or coding techniques to improve the performance of the radar. These techniques use electronics-based techniques for waveform generation. Moreover, waveform transmission and reception are limited by carrier frequency tunability, bandwidth availability, and high jitter using electronics-based systems. The signal processing capability is also limited by the availability of analog-to-digital converters and digital-to-analog converters, which restrict the radar's reconfigurability for high range and velocity resolution requirements.

In Contrast, MWP based systems enable wideband signal generation, frequency-agile operation, and low-jitter photonic mixing, effectively eliminating many of the limitations inherent in conventional electronic architectures. The integration of microwave photonic (MWP) technology has significantly enhanced radar performance by providing a broader bandwidth, low-loss optical signal transmission, immunity to electromagnetic interference (EMI), and a compact size and weight. Moreover, MWP systems offer improved security and the potential for high levels of photonic integration, resulting in compact, lightweight, and energy-efficient radar designs. The following section elaborates on the fundamental principles of MWP systems, establishing a foundation for understanding their role and the mechanisms by which MWP technology enhances

the overall functionality and versatility of modern radar architecture.

2.3 Introduction to Microwave Photonics Technology

Microwave Photonics (MWP) uses light wave technology to transmit, receive, and process microwave signals. This technology holds immense potential across a wide spectrum of applications, including broadband wireless communications, radio-over-fiber (RoF) systems, free-space optical communication links, satellite communications, photonic analog-to-digital conversion, and radar systems.

The core principles of MWP are based on inherent characteristics of integration, functionality, and performance. Integration refers to the seamless incorporation of electronic and photonic components within a compact, chip-scale platform, enabling enhanced system efficiency and miniaturization [16]. Functionality denotes the capability of MWP systems to overcome the inherent limitations of conventional electronic approaches, such as restricted bandwidth and electromagnetic interference. Finally, performance encompasses the overall system efficiency, speed, reliability, and responsiveness. Collectively, these pillars aim to realize integrated, high-performance, and energy-efficient microwave photonic systems that enable the next generation of high-speed communication and sensing technologies.

The primary objective of the microwave photonic approach for developing next-generation radar systems is to leverage the advantages of photonic technologies for waveform generation, signal processing, and the tunability of radar waveforms, thereby enabling complex functionalities that are often infeasible within purely electronic systems.

As illustrated in Fig. 2.5, the fundamental concept of a microwave photonic (MWP) link involves seamless integration between optical and microwave signals on a single platform. A conventional

MWP system typically consists of a laser light source, an optical modulator, an optical signal processing unit, and a photodetector [17]. The electrical-to-optical (E/O) modulator modulates the microwave signal using an optical laser. This modulated optical signal is processed and then driven through an optical fiber before being converted back to the electrical domain via optical-to-electrical (O/E) conversion using a photodiode at the receiver. Various signal processing techniques, such as the generation of large-bandwidth waveforms, highly tunable filters, phase shifters, and instantaneous frequency measurement, can be performed by a photonic signal processor.

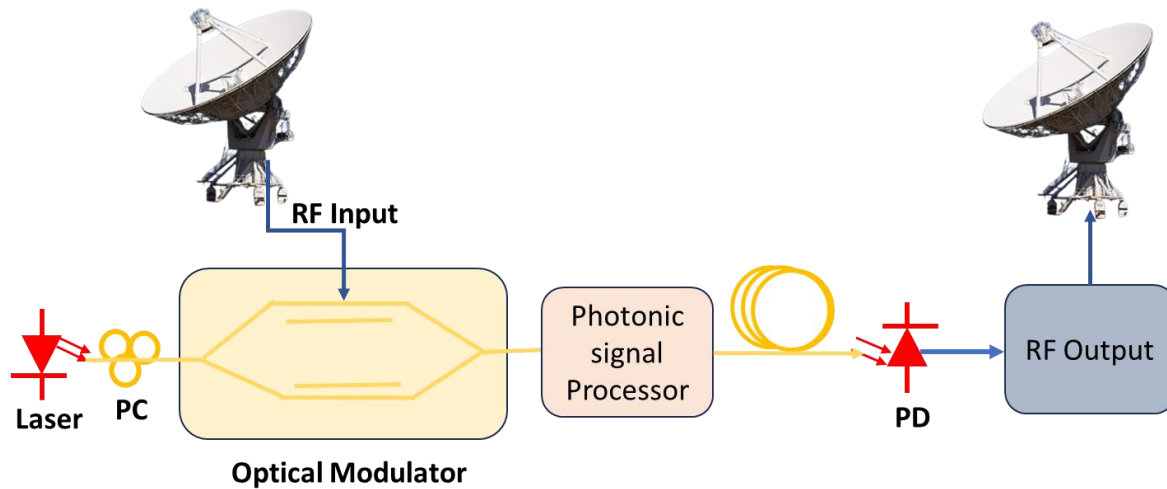


Fig. 2.5: Schematic of a Microwave photonic link, PC: polarization coupler, PD: photodiode

In a microwave photonic (MWP) link, the optical carrier is generated by a laser source, while the radio frequency (RF) signal originates from an electronic oscillator. The RF signal is then applied to an electro-optic modulator, resulting in the upconversion of the RF signal to the optical domain. This modulation process generally produces a pair of optical sidebands symmetrically positioned around the carrier frequency. The modulated optical signal is subsequently coupled to an optical signal processor through an optical fiber.

Within the optical signal processor, the modulated optical signal is further manipulated using various photonic components such as filters, delay lines, or wavelength converters to alter the spectral and temporal characteristics of the optical sidebands. After processing, the optical signal undergoes down-conversion at the photodetector, where the sidebands interact with the optical carrier to reconstruct the processed RF signal in the electrical domain. This complete process establishes a microwave photonic link, offering several distinct advantages that collectively enhance the efficiency, performance, and versatility of modern radar systems as described in the following section.

❖ Enhanced Bandwidth

MWP-based systems operate at THz frequencies, offering very high bandwidth availability over RF and microwave frequency bands, resulting in high-speed data rate transport over optical fiber. High bandwidth availability enables the generation and processing of a large bandwidth signal in the optical domain, reducing the high DSP requirement [18].

❖ Low-Loss Optical Transmission and Antenna Remoting

By facilitating the modulation of RF signals in the optical domain, MWP allows low-loss signal transmission (~ 0.2 dB/km) over optical fibers [19]. This capability minimizes signal degradation and supports antenna remoting, where remote antenna units are connected to secure base stations through optical links, thereby improving performance and enhancing system security.

❖ Advanced Photonic Signal Processing

MWP systems integrate photonic components such as modulators, filters, phase shifters, and delay lines to perform essential signal-processing tasks, including modulation, filtering, and time-delay operations [20]. These photonic functionalities effectively replace traditional RF multipliers, mixers, and filters, reducing electronic bottlenecks and improving signal quality. As a result, MWP

enables the generation of frequency chirps and their selective filtering across multiple bands, allowing a single system to support diverse radar functions simultaneously.

❖ Superior Timing Stability and Low Jitter

Electronic systems often suffer from timing jitters, which degrade synchronization and signal accuracy. MWP systems inherently exhibit ultra-low timing jitters, ensuring precise frequency control and de-chirping accuracy, both of which are essential for achieving high-resolution range and velocity estimation in radar, optical communication, and Radio-over-fiber applications [21].

❖ Miniaturization and System Simplification

The replacement of bulky microwave hardware with compact, efficient photonic devices substantially reduces system size and complexity. This miniaturization enables lightweight, space-efficient radar architectures suitable for integration into mobile and aerial platforms.

❖ Seamless Integration with Electronic Infrastructure

MWP technology can be readily interfaced with existing electronic systems, such as antenna elements or transceiver units, facilitating incremental upgrades and hybrid implementations. This compatibility mitigates bottlenecks associated with technology transitions, allowing smooth migration from traditional electronic to hybrid photonic-electronic architectures [22].

❖ Improved Security and Electromagnetic Immunity

MWP exhibits high immunity to electromagnetic interference (EMI) and low susceptibility to interception or eavesdropping due to operation in optical frequency bands [23]. However, after photodetection and conversion to radio frequency, the signal behaves as conventional EM wave and can be jammed and interfered. However, reduced RF path significantly enhances communication security and signal integrity, addressing one of the key limitations of purely

electronic radar systems.

2.4 Overview of MWP link subsystems

The core of MWP is based on an analog photonic link (APL) that consists primarily of a laser, a modulator, transmission fiber, and a photodetector. The modulator is used to convert the electrical input into the optical domain, which is transmitted over optical fibers and converted back to an electrical signal at the receiver. In the MWP link, input and output signals operate in the microwave frequency domain, while transportation over long distances and processing are performed in the optical domain, thereby reducing digital to analog converter / analog to digital converter complexity and increasing speed.

2.4.1 Laser source

The continuous-wave (CW) laser is primarily used as a light source in MWP based systems. Various other light sources, such as broadband light sources [24], laser arrays [25], and frequency combs [26], are utilized according to MWP link requirements. The output of the laser is used as an optical carrier in an MWP system. The laser employed in the MWP link should have very low intensity noise, a high output power level, and a low spatial linewidth to exhibit a high signal-to-noise ratio in the MWP link. The photo detector converts the intensity fluctuation into photocurrent to recover the microwave signal, so a small intensity fluctuation in the laser will get transported to the output electrical signal and distort the output microwave signal.

2.4.2 Electrical-to-Optical conversion: Modulator

The conversion of RF information from the MHz/GHz frequency band to an optical frequency band is achieved using an optical modulator. The upconversion is achieved by modulation of an optical laser source with RF information. The fundamental concept underlying modulation is that

the RF signals modulate the properties of the light source, such as amplitude, phase, polarization, or frequency. In this work, we have used two kinds of modulators, intensity modulators and phase modulators.

There are two types of modulators, direct modulators and external E/O modulators. The direct modulator modulates the intensity of the laser carrier signal by directly modulating the optical power through modulation of its injection current [27]. This modulator has a limited operational bandwidth and suffers from chromatic dispersion over long fiber lengths due to strong frequency chirp caused by direct modulation, which negatively impacts the RF signal. The distributed feedback laser (DFB) with integrated modulator is used for high-speed data transmission. DFB lasers have low output powers and limited tunability in wavelength, which limits their application where wavelength tunability is required, such as in the wavelength division multiplexing technique [28].

In contrast to direct modulation techniques, external modulation has gained significant prominence due to its high modulation bandwidth, superior speed, and low chirp characteristics, which ensure minimal variation in the optical carrier's frequency or wavelength over time. These attributes enable precise spectral and temporal control, thereby ensuring high-quality signal transmission and accurate measurements in microwave photonic systems.

External modulation employs an electro-optic modulator (EOM), which operates based on the electro-optic effect in nonlinear materials. Commonly used electro-optic crystals include lithium niobate (LiNbO_3) [29], potassium di-deuterium phosphate (DKDP) [30], potassium titanyl phosphate (KTP) [31], beta-barium borate (BBO) [32], lithium tantalate (LiTaO_3) [33], and ammonium dihydrogen phosphate (ADP) [34]. In addition to these inorganic crystals, poled

electro-optic polymers have also emerged as promising candidates for modulator fabrication, leveraging the Pockels effect to achieve efficient and tunable modulation.

The Pockels effect refers to the phenomenon in which the refractive index of a material varies linearly with the strength of an applied electric field, resulting in linear electro-optic modulation. The Mach–Zehnder Modulator (MZM) or intensity modulator is used when the amplitude of the optical carrier is used to encode the information. The MZM enables high-speed intensity modulation through either a single-arm drive or a push–pull configuration, which depicts the device's functional structure and modulation mechanism. The phase modulator changes the phase of the optical carrier over the applied RF field. In the next section, the phase modulator and intensity modulator will be discussed, as these modulators are utilized in Chapters 3, 4, and 5.

2.4.2.1 Phase modulator

The optical phase modulator changes the phase of the coupled light wave over the applied RF or microwave fields, as shown in Fig. 2.6. The phase modulator does not require any DC bias. However, the phase modulator has an inherently π phase difference between the two sidebands, resulting in perfect cancellation at the photodetector. The phase modulator cannot be used directly for data transmission and requires a mechanism to convert phase information to intensity at the photodetector. A Mach-Zehnder interferometer can be used in combination with a phase modulator

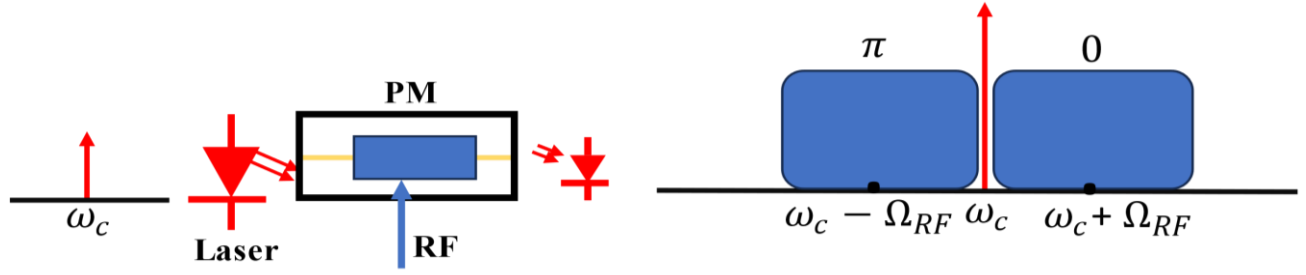


Fig. 2.6: Optical modulation of RF signal using a Phase modulator

for detection at a photodetector. A phase modulator does not require a DC bias, making it highly suitable for high-speed data communication and radar applications.

The Mach–Zehnder Interferometer (MZI) is a fundamental optical device widely utilized in photonic and microwave photonic systems for its capability to manipulate and analyze light through interference. Structurally, the MZI consists of two beam splitters that divide an incident optical beam into two separate arms. An additional path length difference is introduced between the two arms by adding an extra fiber length between them. Both arms combine to produce constructive or destructive interference depending on the relative phase difference between the arms. The schematic of MZI is shown in Fig. 2.7.

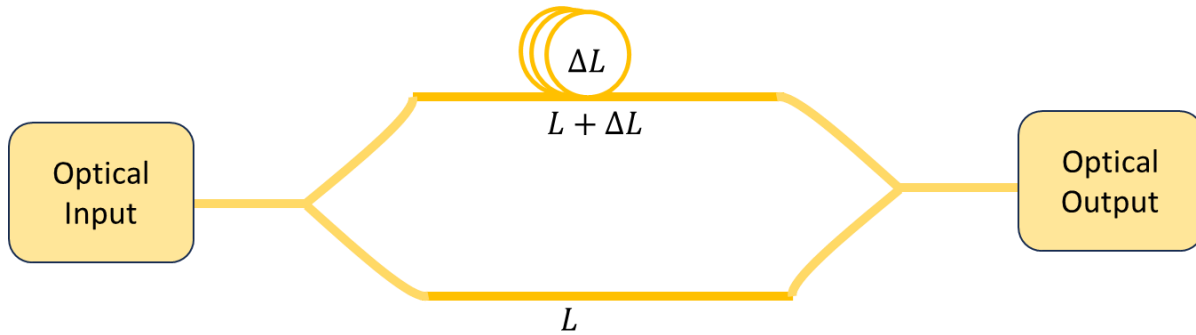


Fig. 2.7: Mach-Zehnder Interferometer

The inherent advantages of high linearity, broad bandwidth, and tunability make the MZI indispensable in advanced optical communication systems, signal processing circuits, and

photonic radar architectures. Its interferometric nature provides both phase and amplitude modulation capabilities, forming the operational foundation of numerous optical signal manipulation techniques in modern photonic technology.

The phase modulated modulator is integrated with MZI at the receiver to get phase to intensity conversion at the photodetector to get photocurrent at the photodetector.

The Electric field at the output of the phase modulator, coupled with the probe laser of frequency ω_c and RF frequency Ω with amplitude V_{RF} is represented as,

$$E_0(t) = A \exp(j(\omega_c t + \phi(t))) \quad (2.6)$$

ω_c is defined as the optical carrier frequency of the laser, and $\phi(t)$ is defined as the phase shift added by the phase modulator and is specified as

$$\phi(t) = \phi \sin(\omega t) \quad (2.7)$$

Here, ϕ is specified as $\phi = \pi \frac{V_{RF}}{V_\pi}$, where V_π is the voltage required by the phase modulator for introducing a 180° phase shift to the signal.

The optical electric field at the output of the MZI is expressed as [27],

$$\begin{bmatrix} E_1(t) \\ E_2(t) \end{bmatrix} = \frac{\sqrt{l_{MZM} g_o}}{2} \begin{bmatrix} 1 & i \\ i & 1 \end{bmatrix} \begin{bmatrix} \Gamma(\tau) & 0 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & i \\ i & 1 \end{bmatrix} \begin{bmatrix} E_0(t) \\ 0 \end{bmatrix} \quad (2.8)$$

Here g_o is the net optical gain, and $\Gamma(\tau)$ is the delay function proportional to the path length difference between the two arms of the MZI. Solving the transfer matrix $E_1(t)$ can be expressed as,

$$E_1(t) = L(E_0(t - \tau) - E_0(t)) \quad (2.9)$$

Where, L includes the losses in the link. The beating of the electric field $E_1(t)$ at the photodetector with a responsivity, $\Re$ results in the generation of RF signals with periodic null response determined by MZI delay τ .

Applying $E_0(t)$ from (6) in (8) and solving the optical electric field at the output of MZI is represented as,

$$E_1(t) = L \{ \exp(j(\omega_c(t - \tau) + \phi(t - \tau))) - \exp(j(\omega_c(t) + \phi(t))) \} \quad (2.10)$$

The MZI output is coupled to a photodetector for optical to electrical conversion, and the current detected at the photodetector is expressed as,

$$I(t) = \Re E_1(t) \cdot E_1(t)^* \quad (2.11)$$

Applying $E_1(t)$ from (2.10) in (2.11),

$$I_1(t) = I_{dcq} - \cos(\omega\tau) \cos(\Delta\phi) + \sin(\omega\tau) \sin(\Delta\phi) \quad (2.12)$$

Here I_{dcq} is the DC photocurrent at the quadrature point of the MZI. The phase $\Delta\phi = \phi(t) - \phi(t - \tau)$. Applying the Anger series, the photocurrent can be expressed as,

$$\begin{aligned} I_1(t) = I_{dcq} - I_{dcq} \cos(\omega_c\tau) J_0(x) - I_{dcq} \sin(\omega_c\tau) \sum_{j=0}^{\infty} (-1)^j J_{2j+1}(x) \cos \left\{ (2j+1) \left(\Omega t - \frac{\Omega\tau}{2} \right) \right\} \\ + I_{dcq} \cos(\omega_c\tau) \sum_{m=0}^{\infty} (-1)^m J_{2m}(x) \cos \left\{ (2m) \left(\Omega t - \frac{\Omega\tau}{2} \right) \right\} \end{aligned} \quad (2.13)$$

The initial two terms are DC photocurrents, the third and fourth terms denote the odd and even terms.

Here, $x=2\phi\sin(\pi f\tau)$, the link gain considering the fundamental terms is expressed as [35],

$$g = \left(\frac{2\pi I_{dcq} \sin(\omega_c \tau) \sin\left(\frac{\Omega \tau}{2}\right) |H_{pd}|}{V_\pi} \right)^2 R_o R_i |H_{pd}|^2 \quad (2.14)$$

where, R_i is the input resistance and R_o is the output resistance. $|H_{pd}|$ is the frequency response of the photodiode itself and the circuit between the p-i-n junction and the load. $|H_{pd}| = 1/2$, for a matching load. $|H_{pd}| = \frac{1}{2\sqrt{2}}$ for a balanced photodiode. As it can be observed from equation (2.14), to achieve a higher link gain, it is required to have a modulator with a low V_π value, low insertion loss, high optical power at the photodiode, and high responsivity of the photodiode.

The integration of a phase modulator with MZI has significant advantages, however, a phase modulated link detected by MZI has a periodic null response at the photodetector, resulting in a reduction of the operational bandwidth of the link as shown in Fig. 2.8. This motivated the development of a mechanism to improve the operational bandwidth of the link using stimulated Brillouin scattering, explain in detail in Section **Error! Reference source not found..** SBS gain response can be tailored using a pump laser in frequency, resulting in reconfigurability of the phase-modulated link demodulated by MZI, making the system highly agile.

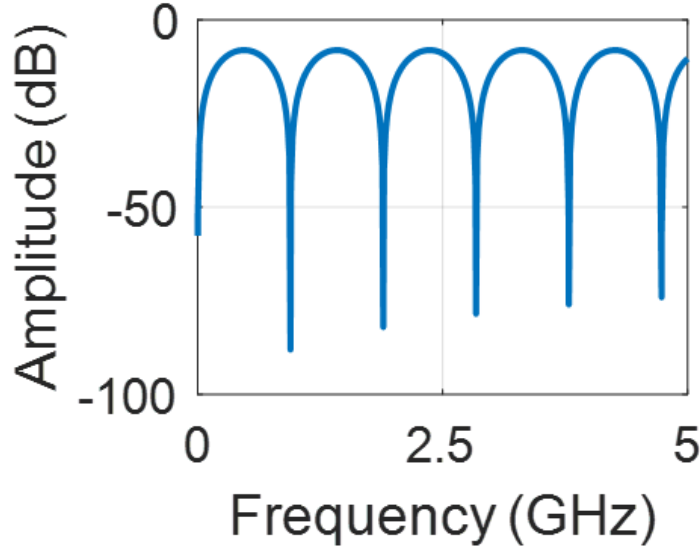


Fig. 2.8: Periodic null response with a phase modulated link demodulated with MZI

2.4.2.2 Mach-Zehnder modulator

The Mach-Zehnder modulator consists of a Y-branch coupler to split the optical signal into both arms of the MZM. The schematic of a photonic link with MZM is shown in Fig. 2.9. The electrical signal is applied to one of the arms of the MZM, which changes the crystal's refractive index, resulting in phase modulation in one arm. The phase-modulated sidebands beat with the unmodulated carrier from the second arm after a Y-branch coupler, combining both arm signals for phase-to-intensity conversion at the photodetector. A DC bias voltage is applied to set the MZM at quadrature, and a linear optical intensity modulation will result when an RF signal is applied.

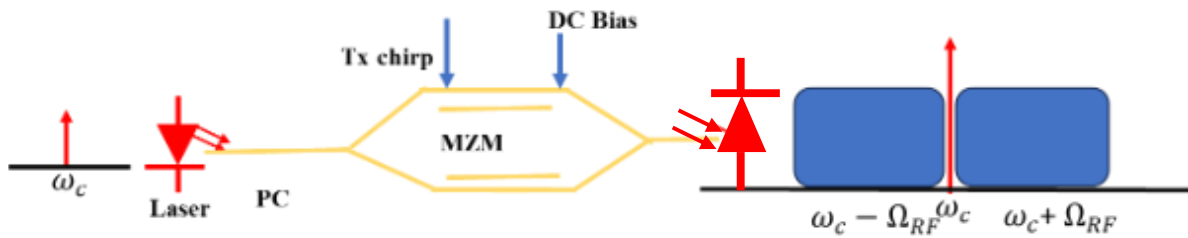


Fig. 2.9: Mach Zehnder modulator

The Mach–Zehnder Modulator (MZM) is driven by a continuous-wave (CW) laser, followed by a polarization controller that optimizes the polarization state of the light to align with the LiNbO₃ crystal, ensuring maximum modulation efficiency. When the phase-modulated and unmodulated optical fields combine in phase, they produce complete constructive interference, corresponding to the maximum transmission point (MaTP). Conversely, when the two optical components combine out of phase, destructive interference occurs, resulting in the minimum transmission point (miTP). Between these two extremes lies the quadrature point (Q-point), positioned midway between the peak and null of the transfer curve. The MZM exhibits a linear response at this bias point, offering maximum modulation efficiency and the highest gain.

The electric field at the output of the modulator, consisting of two Y-branch couplers an input electric field $E_0(t)$ is expressed as [35],

$$\begin{bmatrix} E_1(t) \\ E_2(t) \end{bmatrix} = \frac{\sqrt{l_{MZM}}}{2} \begin{bmatrix} 1 & i \\ i & 1 \end{bmatrix} \begin{bmatrix} e^{i\varphi(t)} & 0 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & i \\ i & 1 \end{bmatrix} \begin{bmatrix} E_0(t) \\ 0 \end{bmatrix} \quad (2.15)$$

Here $\varphi(t)$ is the phase shift due to DC biasing and RF input to the modulating arm of the MZM and is expressed as,

$$\varphi(t) = \varphi_{DC} + \varphi_{RF} \cos(\Omega t)$$

Here, $\varphi_{DC} = \pi \frac{V_{DC}}{V_{\pi,DC}}$ and $\varphi_{RF} = \pi \frac{V_{RF}}{V_{\pi,RF}}$, φ_{DC} and φ_{RF} are the half-wave voltage of the MZM modulator. φ_{DC} controls the DC biasing and φ_{RF} controls the modulation index of the modulator. l_{MZM} is the power loss factor of the MZM.

$$\begin{bmatrix} E_1(t) \\ E_2(t) \end{bmatrix} = \frac{\gamma}{2} e^{i\omega_0 t} \begin{bmatrix} e^{i\varphi(t)} - 1 \\ e^{i\varphi(t)} + 1 \end{bmatrix} \quad (2.16)$$

$$E_1(t) = \frac{\gamma}{2} e^{i\omega_0 t} [e^{i\varphi(t)} - 1]$$

$$E_1(t) = \frac{\gamma}{2} e^{i\omega_0 t} [e^{i(\varphi_{DC} + \varphi_{RF} \cos(\Omega t))} - 1] \quad (2.17)$$

$$E_1(t) = \frac{\gamma}{2} e^{i\omega_0 t} [e^{i(\varphi_{DC})} e^{i(\varphi_{RF} \cos(\Omega t))} - 1] \quad (2.18)$$

Expanding the electric field using the Jacobi-Anger expression,

$$\begin{aligned} E_1(t) = & -\frac{\gamma}{2} \cos(\omega_0 t) + \frac{\gamma}{2} \cos(\varphi_{DC}) \sum_{i=-\infty}^{\infty} J_i(\varphi_{RF}) \cos(\omega_0 t + i\Omega t) \\ & -\frac{\gamma}{2} \sin(\varphi_{DC}) \sum_{k=-\infty}^{\infty} J_k(\varphi_{RF}) \sin(\omega_0 t + k\Omega t) \end{aligned} \quad (2.19)$$

Considering the fundamental, the electric field can be expressed as,

$$\begin{aligned} E_1(t) = & -\frac{\gamma}{2} \cos(\omega_0 t) + \frac{\gamma}{2} \cos(\varphi_{DC}) J_0(\varphi_{RF}) \cos(\omega_0 t) + \frac{\gamma}{2} \cos(\varphi_{DC}) J_1(\varphi_{RF}) \cos(\omega_0 t + \Omega t) \\ & -\frac{\gamma}{2} \sin(\varphi_{DC}) J_1(\varphi_{RF}) \sin(\omega_0 t + \Omega t) \end{aligned} \quad (2.20)$$

The carrier can be suppressed by setting $\varphi_{DC} = 2n\pi$ at MiTP and the sideband power can be controlled by φ_{RF} .

The optical to electrical conversion is done at the photodetector. The current detected at PD is proportional to the sideband mixing and responsivity $\mathfrak{R}$ is expressed as,

$$I(t) = \mathfrak{R} P_{out} \quad (2.21)$$

Here, the average output power can be represented as,

$$P_{out} = \frac{A}{2} \left(\frac{\epsilon}{\mu} \right)^{1/2} E_1(t) E_1(t)^* \quad (2.22)$$

Here, A is the cross-sectional area of the fiber core, ϵ is the permittivity of the fiber core, μ is the permeability of the fiber core, P_0 is the average optical power, and ω is the angular optical frequency.

The photocurrent term at the fundamental drive frequency is of importance for the analysis, which can be written at MiTP as,

$$I(t) = I_{DC} J_1(\varphi_{RF}) \cos(\Omega t) \quad (2.23)$$

Here, $I_{DC} = \Re \frac{g_0 l_{mzm} P_0}{A}$, and $\gamma = \sqrt{\frac{2 l_{mzm} P_0}{A}} \left(\frac{\mu}{\epsilon} \right)^{1/4}$, $J_1(\varphi_{RF}) \approx \frac{\varphi_{RF}}{2}$ for $\varphi_{RF} \ll 1$ and

The average output power at the fundamental frequency is represented as,

$$P_{\Omega} = \langle I(t)^2 \rangle R_o |H_{pd}|^2$$

The input power to the RF port of the MZM modulator can be represented as,

$$P_{IN} = \frac{V_{RF}^2}{R_i}$$

Here, R_o is the output resistance, R_i is the input resistance and $|H_{pd}|$ is the frequency response of the photodetector.

The link gain is the output power detected at the photodetector over the input power applied to the MZM modulator.

$$g = \frac{P_{\Omega}}{P_{IN}} = \frac{I_{DC}^2 (\varphi_{RF})^2 R_o R_i}{V_{RF}^2} |H_{pd}|^2 \quad (2.24)$$

$$g = \frac{I_{DC}^2 \pi^2 R_o R_i}{V_{\pi, RF}^2} |H_{pd}|^2 \quad (2.25)$$

The link gain is estimated using a vector network analyzer and by measuring the scattering parameter $|S_{21}|^2$.

The MWP link utilizes various modulators like phase modulator, Dual parallel Mach Zehnder modulator (DPMZM) [36], Dual drive Mach Zehnder modulator (DDMZM) [37], Dual polarization dual parallel Mach Zehnder modulator (DP-DPMZM) [38] for different applications as per the number of RF inputs required or separate polarization between modulated sidebands.

The research conducted in this thesis used MZM, DDMZM, and DP-DPMZM modulators. The higher end modulator DDMZM houses two MZMs, each driven by individual RF inputs and using a differential DC bias. The DP-DPMZM modulator houses two DPMZM modulators, both in different polarizations, to have no mixing between the modulated sidebands of individual DPMZM modulators at PD.

The intensity modulator, like MZM, requires a DC bias to modulate the microwave signal using an optical carrier, which limits the operation in some scenarios where direct DC power is not available. The phase modulator can be utilized in these scenarios as it does not require any DC bias. However, the phase modulator produces π out-of-phase sidebands, resulting in a null response with direct detection. We need a dispersive device, such as a Mach-Zehnder interferometer, to convert phase information to intensity information.

2.4.3 Photodetector

The photodetector (PD) is used to convert the electrical signal back from an optically modulated signal. The PD can only detect the variation in intensity, which is then converted back into photocurrent variations. The quality of PD is determined by the responsivity ($\mathfrak{R}$), bandwidth, and linearity. The bandwidth of the photo detector is important for detecting high-frequency signals.

The current detected at PD is determined by the PD's optical input power and responsivity.

2.5 Performance of the MWP link system

In this section, the performance of the MWP link is discussed. The metrics include various figures of merit (FOM) for quantifying the link performance, such as link gain (G), Noise figure (NF), compression dynamic range (CDR), and spurious-free dynamic range (SFDR) [35]. The photocurrent detected at the photodetector and the input power applied to the modulator are used to quantify the performance of MWP link.

2.5.1 Link Gain

Link gain (G) determines the electrical output power over the input power applied to the input in an MWP link. It is the ratio of the RF signal power delivered to the load to the input RF signal power applied at the source. The link gain needs to be maximized to prevent any RF signal power loss. The losses in the electrical and optical components used in the MWP link need to be minimized to have a high link gain for better signal power at the receiver.

$$G = 10 \log_{10} \left(\frac{P_{out_{RF}}}{P_{in_{RF}}} \right) \quad (2.26)$$

Here, G is the link gain in dB and $P_{out_{RF}}$ is the output RF power detected at the photodiode and $P_{in_{RF}}$ is the input RF power applied to the optical modulator.

2.5.2 Noise Figure

The noise factor is a typical metric to describe the performance of an RF component. The noise factor is defined as the ratio of the input signal-to-noise (SNR) to the output signal-to-noise ratio of the RF signal. Noise factor can also be defined at a given input frequency as the ratio of the total

noise power per unit bandwidth available at the output at the corresponding output frequency, with the input terminated at the standard temperature of 290 K, to the output noise contribution arising only from the 290 K input termination noise at that input frequency.

The MEP link is primarily limited by noise sources such as thermal, shot, RIN, and SBS-induced noise [35], represented as,

$$N_{thermal} = (G + 1)kTB$$

$$N_{shot} = 2qI_{DC}R_oB|H_{PD}|^2$$

$$N_{RIN,probe} = RIN_{probe} \cdot I_{probe}^2 R_oB|H_{PD}|^2$$

$$N_{SBS} = 2R_{PD}I_{probe}P_{n,sbs}R_oB_1|H_{PD}|^2$$

Here, k is the Boltzmann constant, T is the temperature, and B is the bandwidth RIN is the relative intensity noise of the laser, q is the elementary charge constant, I_{DC} is the photocurrent and I_{probe} is the probe current. $P_{n,sbs}$ is the SBS noise optical power.

Noise figure in a thermal-limited regime is represented as,

$$NF(dB) = 10 \log(F) = 174 + N_{out} \left(\frac{dBm}{Hz} \right) - G(dB) \quad (2.27)$$

2.5.3 Compression dynamic range (CDR)

The dynamic range of an MWP system is a crucial metric for determining the performance of a nonlinear system. As no RF system is perfectly linear, non-linearities arise in terms of compression, harmonic distortion, and intermodulation distortion (IMD).

CDR_{ndB} is defined as the input power at which the output signal power is reduced by n dB

compared to the linear response at the output [34]. CDR_{ndB} is determined using a single tone test and is represented as,

$$CDR_{ndB} = \frac{P_{ndB} 10^{n/10}}{N_{out} B} \quad (2.28)$$

Here, P_{ndB} is the output power at the n-dB compression point. The 1dB compression point is represented as,

$$CDR_{1dB} = \frac{P_{1dB} 10^{0.1}}{N_{out} B} \quad (2.29)$$

2.5.4 Spurious free dynamic range (SFDR)

The SFDR metric determines the metric to quantify the new frequencies generated due to nonlinearity in the system. SFDR can be defined as the range of input or output power over which the output signal is above the output noise floor, and all spurious signals are less than or equal to the output noise floor. The metric used to define the n th order SFDR is the n th-order output intercept point (OIP_n). The (OIP_n) is the output power at the intersection of the extrapolated small-signal responses of the fundamental and the n th-order distortion. The n th order $SFDR_n$ in terms of the n th order output intercept point (OIP_n) is represented as,

$$SFDR_{ndB} = \left(\frac{OIP_n}{N_{out} B} \right)^{\frac{n-1}{n}} \quad (2.30)$$

Here, $SFDR_{ndB}$ is n th order spurious free dynamic range, OIP_n is the n th order intercept point, N_{out} is the output noise power, and B is the bandwidth.

2.6 Recent work on microwave photonic radar

Recent advances in microwave photonics have significantly expanded the capabilities of modern radar systems, enabling higher operational bandwidths, improved functionality, and seamless integration with emerging technologies. These developments have opened radar applications across diverse fields, including defense, telecommunications, aerospace, and remote sensing applications.

Table 2.1 summarizes a selection of research contributions that provide important insights into the progress of microwave photonics-based FMCW radar systems. The table highlights a comparison of recent state-of-the-art microwave photonic radar for estimation of high-resolution range and velocity, inverse synthetic aperture radar (ISAR) imaging, and deployment of fully integrated photonic chip-based highly reconfigurable radar platforms.

Table 2.1: Recent work on microwave photonic radar systems

Ref.	Radar Type	Freq. Band	BW	Range	Velocity	Remarks
[96]	FMCW	K	8 GHz	~3.5m	NA	Generation of high bandwidth LFM waveforms and target range detection with range resolution of ~1.875 cm
[101]	FMCW	W	2 GHz	Up to 3m	NA	Handheld operation has applications for health monitoring and inspections of ancient arts.
[88]	FMCW	K	8GHz	NA	NA	ISAR imaging of 3 metal plates having the size of 6

						cm × 7 cm
[89]	FMCW	C	1GHz	150cm-210cm	NA	Detection of a static target. The system is emulated to detect multiple targets (up to four), showing a range resolution of <15 cm.
[91]	FMCW	Ku	6GHz	NA	NA	The proposed system is suitable for miniaturized multifunctional radar signal processing systems with continuous operation of transmitting and receiving antennas in unmanned aerial vehicles (UAVs), automotive radar,
[92]	FMCW	HF to Ku	Varying 6-15GHz	6m	NA	LFM waveforms with ultrawide bandwidths. Detection of two metal targets with high resolution
[93]	FMCW	V	10 GHz	Multiple	Varying	A range resolution of 1.5cm and velocity resolution of 0.067m/s is demonstrated. ISAR imaging is also demonstrated
[97]	FMCW	X	10GHz	~3.5M	NA	Distance measurement ranging from 0.50 m to 0.78 m has been implemented, and the absolute errors are less than 3 mm. Both range and velocity can be measured.
[106]	Pulse	K	4 GHz	160.9 cm	NA	Detection of range. Multiple orthogonal LFM signal combined using WDM.
[114]	FMCW	Ku	5 GHz	3 & 5cm	NA	ISAR image of the target
[115]	FMCW	W	600 MHz	500 m		Target range and velocity for an automobile

						application are demonstrated.
[107]	FMCW	S & X		3 Km		A MIMO radar system is demonstrated.
[116]	FMCW	W/	600 MHz	6 km	41.66m/s	Weather conditions' impact on automobile range and velocity estimation
[104]	FMCW	X	4 GHz	2m	Rotating target	Range, velocity estimation, along with ISAR imaging for a target with 3 cuboids and a cylinder in the center.
[108]	FMCW	X	5GHz & 2.5GHz	2.7 km	Max 72 km/h	Field demonstration of Drone range and velocity detection. ISAR Imaging of a drone is also demonstrated.
[109]	FMCW	X	100 MHz	3m	NA	MIMO radar for a coherent centralized network
[110]	FMCW	W	600 MHz	390m	80 km/h	Car, Truck, Motorbike, Bus, Pedestrian detection
[95]	SFCW	Ku & Ka	18.2 GHz	1 m	NA	CM level range resolution with target as two metallic planes (2 cm × 2 cm)
[113]	SFCW	Ka	11.52 GHz	1.5m	NA	CM level range resolution with a target consisting of three cylindrical objects with a radius of 3 cm and a height of 4 cm
[87]	SFCW	HF-Ku	18GHz	8.3MM	NA	Distinguished two simulated point targets with a distance of 8.3 mm, and achieved the ranging error within $\pm 225 \mu\text{m}$

[111]	FMCW	K	8 GHz	17m	NA	Real-time tracking and high-resolution ISAR imaging of a drone are demonstrated
[96]	Pulse	Ka	8 GHz	1.6m	NA	ISAR imaging of two pyramids
[105]	Pulse	X	300 MHz	5.4 km/NA	NA	Range Detection
[98]	Pulse	S&X	200 MHz	3.6km	160 km/h	Detection of both range and velocity. X band results in better velocity resolution compared to S band
[12]	Pulse	S&X	200 MHz	8 nmi	55m/s at 4.6 km	Detection of both range and velocity of a boat.

2.7 Challenges in Conventional Radar Technology

The modern generation of radar systems requires high bandwidth and frequency reconfigurability, with the capability to tune the time-bandwidth product in real-time to detect highly stealthy targets with very low cross-sections. The radar systems require high resolution to detect small targets and resolve nearby targets. It also needs to have high resilience against jamming and avoid detection by modern electronic detection systems, which require real-time bandwidth tunability along with frequency agility. Linear frequency-modulated waveforms are widely used in radar technology due to their high pulse compression, which improves range resolution without increasing peak power.

The advancement of microwave photonic technology has enabled highly augmented radar systems with the capability to reconfigure bandwidth, frequency, and power, resulting in highly tunable radar systems that are also immune to detection by enemy radar systems. Conventional electronic systems require received waveforms, as well as reference transmitted and received waveforms, to

be converted to digital signals using high-speed analog-to-digital converters before processing to generate the range and velocity of targets. The limitations of high-end processing systems, along with the cost and availability of high-speed ADCs, severely limit the radar capabilities for high-bandwidth and high-frequency operation in detecting small, slow-moving targets.

MWP technology enables real-time tunability in frequency, making it highly adaptable for different target sizes and positions, as well as the detection of slow-moving objects. The capability to tune bandwidth enables tuning the time bandwidth product to resolve a small-sized object. The LFM signal is predominantly used in radar systems, as it brings certain inherent characteristics that significantly improve the radar systems. These characteristics are discussed in subsequent sections for further exploration to improve radar technology.

2.7.1 Large time-bandwidth product

The bandwidth of linear frequency-modulated signals is critical for achieving high time-bandwidth products, which in turn enable high range and Doppler resolution [39]. The range resolution is inversely proportional to the bandwidth of the transmitted waveform, enabling the detection of multiple targets flying very close to each other as specified in equation (2.31).

$$\text{Range resolution} = \frac{c}{2B} \quad (2.31)$$

Here, c is the velocity of light and B is the bandwidth of the transmitted waveform.

The MWP signal processing uses a terahertz laser frequency signal for signal processing. The bandwidth of the microwave waveform, for example, $\sim 10\text{GHz}$, is very low as compared to $\sim 193\text{THz}$ frequency level, which enables easy integration of a large bandwidth LFM signal with photonic signal processing. The large bandwidth of the waveform also enables the detection of

very small targets, such as small particles, making it highly suitable for medical applications.

The time of transmitting waveform is an essential part to improve the detection of frequency resolution of Doppler shifts for the detection of very slow-moving targets. The increase in time transmitted waveform with a specific bandwidth increases the integration time and, hence, reduces the Doppler main lobe of the ambiguity function, thereby improving the velocity resolution [40].

2.7.2 Waveform generation for high performance Radar

A radar system operates over a broad range of frequencies from a few megahertz to several gigahertz [41]. The low-frequency bands are primarily used for long-range surveillance, navigation over the sea, early warning radar, and astronomy applications. Radars operating in higher frequency bands are primarily used for tracking, imaging, autonomous vehicles, and air traffic control, [42]. Conventional radar based on electronics can be tuned only within a limited frequency band, which restricts the radar's functionality. The MWP-based radar waveform generation can be used to generate highly tunable waveforms, spanning a low-frequency bandwidth of a few megahertz to a several-gigahertz bandwidth [43] [44]. The MWP-based radar system enables highly reconfigurable radar operation for various applications. MWP can also be used as a remote antenna system, separating the transceiver system and antenna at two different locations using optical fiber, which makes the system highly responsive in mission-critical applications [45].

MWP based chirp waveforms have been proposed using frequency-to-time mapping [46], frequency modulation using electro-optic modulators [47] – [51], an optoelectronic oscillator [52] [53], optical injection of light using a semiconductor laser [54], and optical frequency multiplication [55]. However, these techniques require a complex modulator with multiple DC bias control through a dual parallel Mach Zehnder modulator DPMZM, dual polarization dual

parallel Mach Zehnder modulator (DPol-DPMZM), or Dual drive Mach Zehnder modulator (DDMZM), which compromises the linearity over a long period of time. Most of these techniques use RF signals to tune the frequency of the waveform, which limits the frequency tunability range and complicates the design with a minimum of two electrical input requirements.

This motivates the development of a new methodology to generate different waveforms using a simple MZM modulator with a single DC bias control requirement and frequency tunability by tuning a CW laser frequency. This will improve the system linearity, and frequency tunability will not be limited by an electrical oscillator circuit with limited frequency tunability.

2.7.3 Simultaneous range and velocity Detection

MWP based frequency modulated continuous wave (FMCW) radar utilizes beating between transmitted and received echo waveforms for the detection of target range and velocity. The range of the target is a function of delay and is measured as $\left(R = \frac{c\Delta t}{2}\right)$. The accuracy of the estimation of the range of targets is determined by the bandwidth of the transmitted pulse. The frequency-time graph of the transmitted and received echo pulse for a single LFM is shown in Fig. 2.10. A LFM signal with a starting frequency f_1 and chirp rate $k_1 \left(\frac{B}{T}\right)$ is transmitted. The received echo waveform with a delay Δt beats with the transmitted waveform to generate a de-chirped frequency Δf_1 .

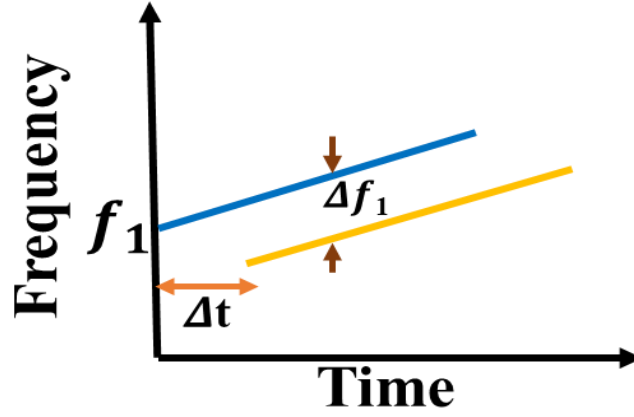


Fig. 2.10: De-chirped frequency generation from transmitted and echo received waveform

$$\Delta f_1 = k_1 \Delta t - f_{d1} \quad (2.32)$$

Here Δf_1 is the de-chirped frequency and f_{d1} is the Doppler frequency shift in the echo received signal. Considering the f_{d1} very small compared to $k_1 \Delta t$, the range of the target is estimated as,

$$\Delta f_1 = k_1 \Delta t \quad (2.33)$$

$$\Delta t = \frac{\Delta f_1}{k_1} \quad (2.34)$$

The estimated range of the target is $R = \left(\frac{c \Delta t}{2}\right)$, putting Δt from equation (2.34), the target range is represented as,

$$R = \left(\frac{c \Delta f_1}{2 k_1}\right) \quad (2.35)$$

With an increase in target velocity, the Doppler shift f_{d1} increases. This increases range-Doppler coupling, resulting in a decrease in range estimation accuracy. With a single LFM as the transmitted waveform, only the range can be estimated from a single echo, and it also has a very

high range of Doppler coupling due to knife edge ambiguity function of a single LFM [41]. To decrease the range-Doppler coupling, a dual chirp waveform is proposed, consisting of an up-chirp and a down-chirp with the same chirp rate and at the same starting frequency [42]. The dual chirp waveform has low range-Doppler coupling due to the low slope of the wedge in the ambiguity function. This waveform also enables simultaneous estimation of both range and speed.

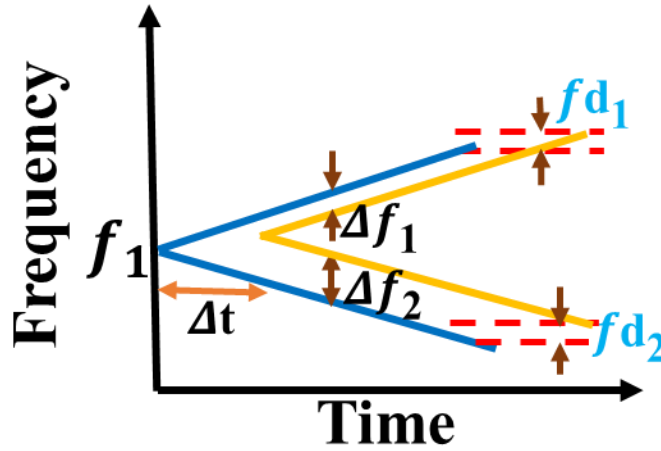


Fig. 2.11: De-chirped frequency spectrum for a dual chirp transmitted and echo received waveform.

The time-frequency plot of the transmitted and received echo waveform for a receding target is shown in Fig. 2.11. Here Δf_1 and Δf_2 are de-chirped frequencies through beating between the transmitted and received echo waveforms. Here Δt is the delay between the transmitted and received echo waveforms. f_{d1} and f_{d2} are the Doppler shifts on up-chirp and down-chirp, respectively. The Doppler shift f_{d1} and f_{d2} are the same due to the same starting frequency and are represented as,

$$f_{d1} = f_{d2} = \frac{2v}{c} f_1 \quad (2.36)$$

$$\Delta f_1 = k_1 \Delta t - f_{d1} \quad (2.37)$$

$$\Delta f_2 = k_1 \Delta t + f_{d1} \quad (2.38)$$

Here, the delay Δt and the Doppler shift f_{d1} can be estimated from equations (2.37) and (2.38) and is represented as,

$$\Delta t = \frac{\Delta f_1 + \Delta f_2}{2k_1} \quad (2.39)$$

$$f_{d1} = \frac{\Delta f_2 - \Delta f_1}{2} \quad (2.40)$$

The range and speed estimated from equations (2.39) and (2.40) can be represented as,

$$R = c \frac{(\Delta f_1 + \Delta f_2)}{4k_1} \quad (2.41)$$

$$v = c \frac{(\Delta f_2 - \Delta f_1)}{4} \quad (2.42)$$

The dual chirp waveform can be used to estimate both range and speed simultaneously. However, the direction of motion cannot be estimated using this waveform. Several techniques have been presented, like using a triangular chirp waveform consisting of an up-chirp followed by a down-chirp [42]. However, this waveform requires double the time for the same bandwidth as down-chirp comes after up-chirp, but the swept bandwidth is equal to either up-chirp or down-chirp bandwidth, limiting the range resolution ($\Delta R = c/2B$) to the bandwidth of either up-chirp or down-chirp, reducing the range-resolution, compared to a single chirp or a V-shaped chirp as shown in Fig. 2.12(a) and Fig. 2.12(b). The main lobe width of the dual chirp is smaller than that of the triangular chirp, representing high range resolution in the dual chirp compared to the triangular chirp. Other techniques like multiband LFM chirp use three different LFM [43],

however, this technique has higher range-Doppler coupling due to different starting frequencies of individual LFM, resulting in different Doppler shifts for all three individual LFM, increasing the range ambiguity and hence, reducing the range accuracy at higher velocities of the target.

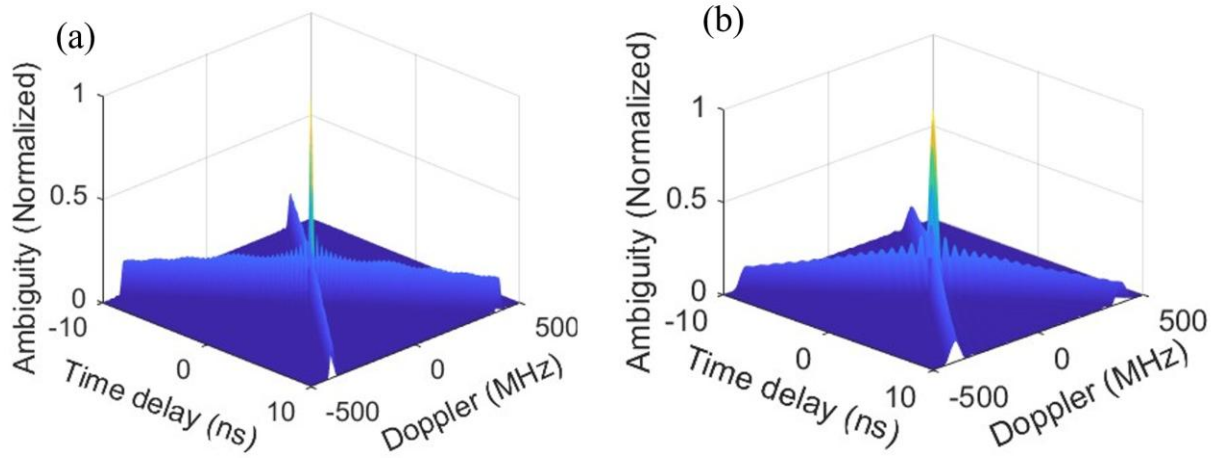


Fig. 2.12:(a) delay(range) and Doppler(velocity) coupling for a V-shaped dual chirp, (b) the delay(range) and Doppler(velocity) coupling for a Triangular waveform

This motivates the development of a new waveform that can be used for simultaneous range and velocity estimation with high accuracy and resolution. The development of highly tunable waveforms with the capability to change the operational frequency for different applications like surveillance, tracking, and imaging requires highly tunable filters.

Filters are an essential component in all RF and microwave systems for communication systems, radar, and radio-over-fiber networks [57]. The modern radar technology requires the waveform to be tuned in real time to filter the unwanted frequencies generated during waveform generation, changing the filter bandwidth for a change in waveform bandwidth, and to reject the interference signal or jammers. This requires a highly reconfigurable filter with real-time tunability in frequency of operation and bandwidth. The conventional electronic based microwave frequency filters have limited frequency tunability and low bandwidth tunability, requiring different filters

for different frequencies of operation with different bandwidths, limiting the radar reconfigurability.

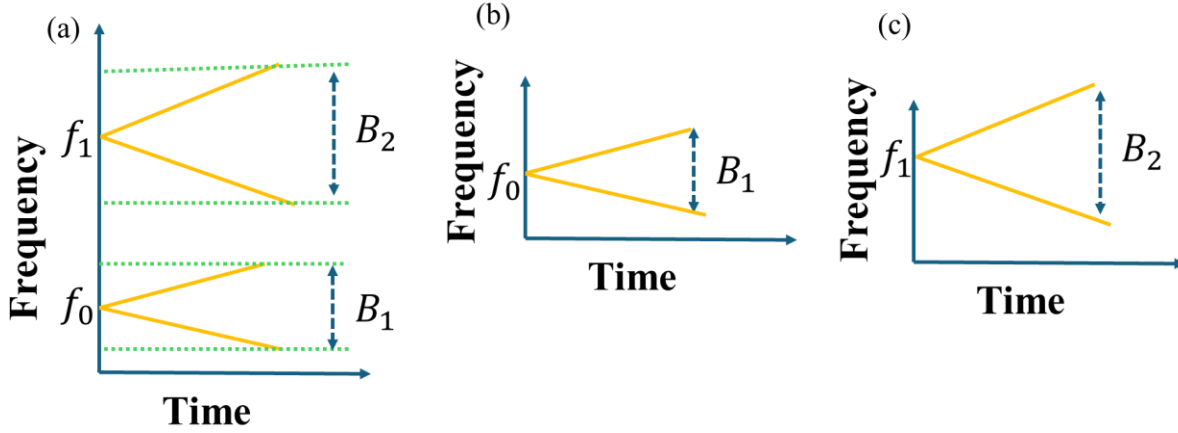


Fig. 2.13: (a) Multiband waveform with different bandwidths, (b) Filtering the lower band waveform, (c) Filtering the upper band waveform

Fig. 2.13(a) illustrates a multiband dual-chirp waveform with varying bandwidths for multiband radar operation, featuring a distinct bandwidth profile. To switch from a waveform with frequency f_0 and bandwidth B_1 as shown in Fig. 2.13(b) to waveform at frequency f_1 and bandwidth B_2 as shown in Fig. 2.13(c), the filter frequency and bandwidth both need to be tuned. The generation of such highly reconfigurable filters is difficult to achieve using electronics.

MWP with high frequency and bandwidth availability in THz frequency ranges is highly suitable to generate a highly reconfigurable filter with tunable frequency and bandwidth operation. Various microwave filters have been demonstrated based on Mach-Zehnder interferometers [58], fiber Bragg gratings [59], ring resonators [60], and stimulated Brillouin scattering [61]- [66]. However, switching from a very narrowband to a wideband filter response is difficult. The bandwidth of the SBS frequency response is very narrow, making it suitable for narrowband filtering. The SBS narrowband filtering capability and tunability in filter response over the pump laser frequency

motivated us to focus on SBS-based signal processing for radar applications.

2.8 FMCW Radar signal processing with Stimulated Brillouin scattering

Frequency-modulated continuous-wave radar uses the frequency difference between the transmitted and echo-received signals to estimate the range and velocity of the target. MWP enables a radar waveform with a large bandwidth, a multiband waveform with different chirp rates, and high frequency tunability. However, waveform generation using optical modulators, such as intensity modulators, phase modulators, or higher-order modulators like DPMZM or DP-DPMZM, generates higher-order sidebands, which need to be filtered before being sent to the receiver [67] - [70]. An optical filter is required to filter the undesired frequency band before coupling the modulated transmitted and received echo signals for beating at the photodetector. The new generation of radar requires multi-performance capability, which necessitates tunability in frequency and bandwidth of the waveform to cater to applications such as surveillance, tracking, and imaging using the same radar setup. This required an agile optical filter with tunability in frequency of operation and bandwidth. An optical filter based on stimulated Brillouin can be used as a highly reconfigurable filter to select a band of interest by tuning the pump laser frequency.

Our group has demonstrated SBS based frequency agile filter for band selection of the waveform by tuning the pump laser frequency [50]. Here, a photonics-based approach for the generation of multi-carrier linear frequency-modulated (LFM) waveforms and agile selection of individual LFM bands through a stimulated Brillouin scattering (SBS) based tunable filter . In this approach, a DDMZM modulator is used to generate chirp-modulated sidebands by using higher-order mixing between the RF tone and chirp signal. The generated multiband modulated LFM waveform is coupled to a phase modulator, and a pump laser is coupled using a circulator to generate an SBS gain response to filter the desired frequency band and reject

other bands. The frequency of the SBS filter response is tuned by changing the pump laser frequency, making the waveform adaptable in real time. Fig. 2.14(a) shows the LFM waveform generated at 2,4,6, and 8GHz frequencies. Fig. 2.14(b) to Fig. 2.14(e) show the filtered LFM signals at 2 GHz, 4 GHz, 6 GHz, and 8 GHz, respectively, with the SBS gain response. The SBS gain response used to filter the chirp signal is shown in Fig. 2.14(f). However, the bandwidth of the SBS gain response used as a filter is very low, ~ 30 MHz.

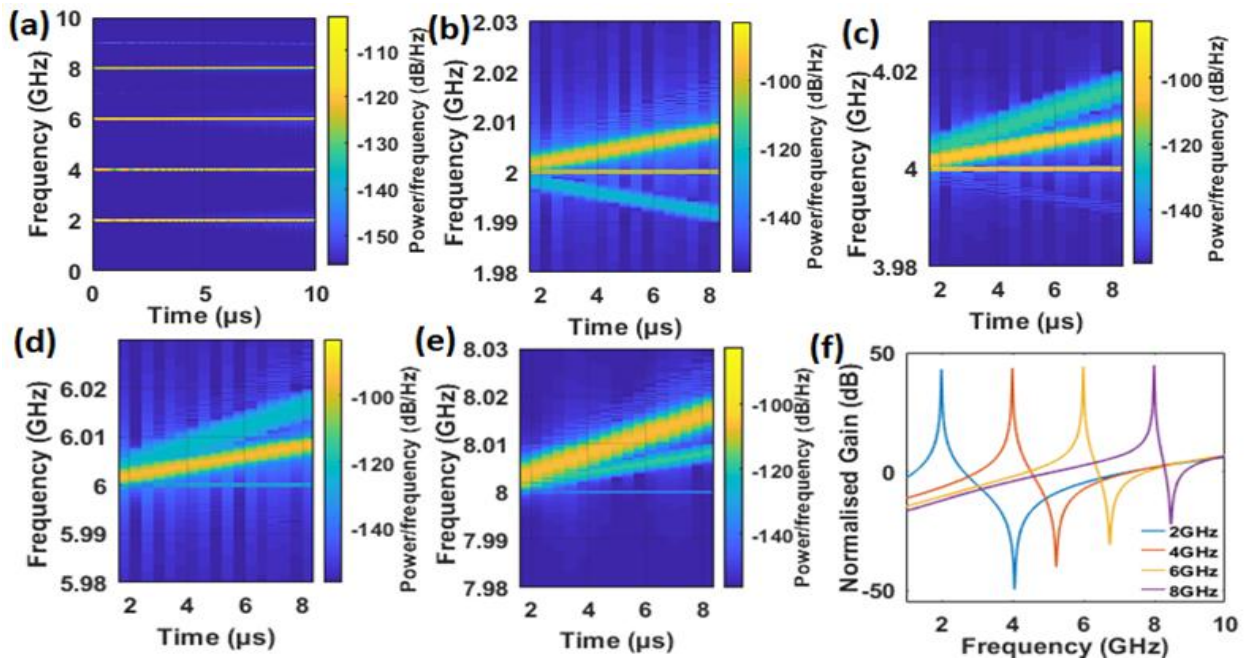


Fig. 2.14: (a) Spectrogram plot of the generated tupled LFM at 2 GHz, 4 GHz, 6 GHz, and 8 GHz, (b, c, d, e) zoomed-in view of the spectrogram plot showing all up-chirps at 2 GHz (BW 10 MHz), 4 GHz (BW 10 MHz), 6 GHz (BW 10 MHz), and 8 GHz (BW 20 MHz), respectively, (f) represents the simulated SBS filter response at 2 GHz, 4 GHz, 6 GHz, and 8 GHz [86].

This restricted the bandwidth of the waveform to be less than 30 MHz. The low bandwidth availability reduces the radar capability to detect small targets moving at slow speeds due to low range resolution. The rejection ratio of the SBS gain response is also low and requires very high pump power to increase it. This decreases the signal-to-noise ratio, as higher amplification of noise occurs at higher pump power, thereby reducing sensitivity and, consequently, range and velocity

detection capability.

This motivated the design of a highly flexible filter based on stimulated Brillouin scattering with tunability in frequency of operation, bandwidth, and rejection ratio to tune the radar. The complex waveform generation based on higher modulators, such as DPMZM and DPol-DPMZM, also requires filtering to remove the unwanted sidebands generated by these modulators before transmission, as discussed in Chapter 3.

2.9 Overview of Stimulated Brillouin Scattering

The process of stimulated Brillouin scattering (SBS) can be described as a nonlinear interaction among the pump and Stokes fields and an acoustic wave through the process of electrostriction in an optical medium such as fiber [71]. The acoustic wave is initially generated thermally inside the optical medium. The acoustic wave produces density modulations, which in turn modulate the refractive index of the medium. The refractive index grating produced by the pump wave scatters the pump light through Bragg diffraction. The scattered light is downshifted in frequency due to the Doppler shift associated with a moving grating at the acoustic velocity. The scattering process can be explained as the annihilation of a pump photon, creating a Stokes photon and an acoustic phonon simultaneously. As momentum must be conserved during this annihilation process, the frequencies and the wave vectors of the pump wave, Stokes wave, and acoustic wave are represented as,

$$\Omega_B = \omega_p - \omega_S$$

$$k_A = k_p - k_S$$

Here, Ω_B is the Brillouin frequency shift and ω_p and ω_s are the pump and Stoke wave frequencies. k_A , k_p and k_s are the wave vectors of the acoustic wave, pump wave and Stokes wave, respectively.

The Brillouin frequency shift can be represented in terms of wave vectors as,

$$\Omega_B = v_A k_A \approx v_A k_p \sin(\theta/2)$$

The frequency of acoustic waves is maximum at a scattering angle of π , confirming that the acoustic wave travels in the opposite direction to the direction of propagation of the pump wave. The spontaneous Brillouin scattering occurs in the forward direction, resulting in the creation of an anti-Stoke's wave. The anti-Stoke's wave is weak and is in the direction of propagation of the pump wave.

Fig. 2.15 describes the electrostriction process. When a pump wave propagates through a dielectric medium, the electric field exerts an electrostrictive force that periodically compresses the medium. This interaction gives rise to an acoustic wave that travels along the direction of light propagation, establishing a dynamic modulation of the refractive index. The presence of this acoustic wave results in the formation of a moving density grating within the medium. The optical pump light is scattered from this grating, producing a backward-propagating Stokes wave, arising from the motion of the acoustic grating relative to the propagating optical wave.

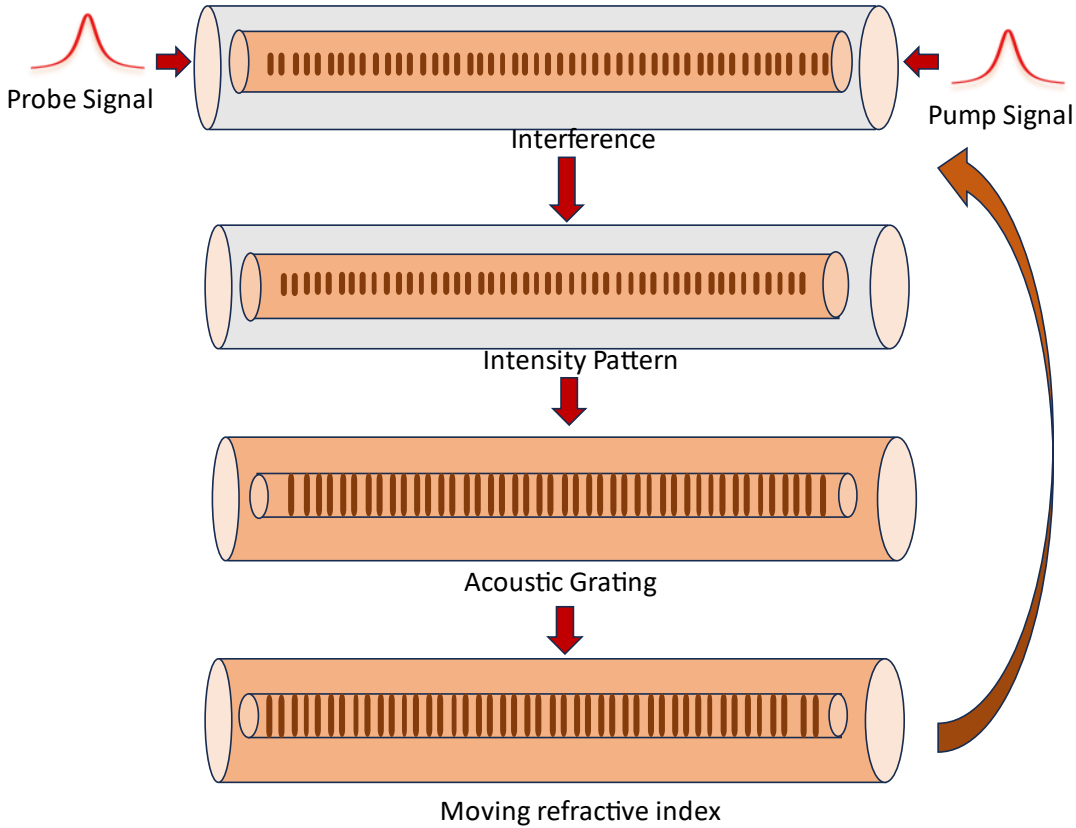


Fig. 2.155: Refractive index variation generation in an optical fiber with a counterpropagating probe and pump laser signals

As the Stokes wave builds up, it interferes with the pump wave, reinforcing the spatial intensity pattern within the medium. This enhanced interference pattern further strengthens the acoustic wave through electrostriction, thereby increasing the modulation of the refractive index. The resulting acoustic grating acts as a Bragg reflector, scattering additional light in the backward direction and amplifying the Stokes component. This sequence of interactions constitutes a positive feedback mechanism, the growth of the Stokes wave intensifies the acoustic wave, which in turn increases the backscattering efficiency. The process continues until energy transfer from the pump to the Stokes wave reaches equilibrium, at which point SBS saturation occurs. Thus, the stimulated Brillouin scattering process can be viewed as a coherent coupling between optical and acoustic waves, mediated by the electrostrictive response of the medium.

Over the past decade, various microwave functionalities have been demonstrated using stimulated Brillouin scattering (SBS). The SBS in an optical medium results in a narrow bandwidth gain response with a bandwidth of $\sim 30\text{MHz}$. The acoustic wave generated results in modulation of the refractive index of the optical medium, resulting in backscattering generating Stoke's and anti-Stoke signal. The backscattered wave in the direction of the pump laser results in forward SBS and a frequency shift Ω_B away from the pump laser frequency. The backscattered wave travelling in the opposite direction of the pump laser results in backward SBS or Stoke's wave. The forward wave traveling in the direction of the pump laser results in forward SBS or anti-Stoke's wave. The SBS gain and loss response are shown in Fig. 2.17.

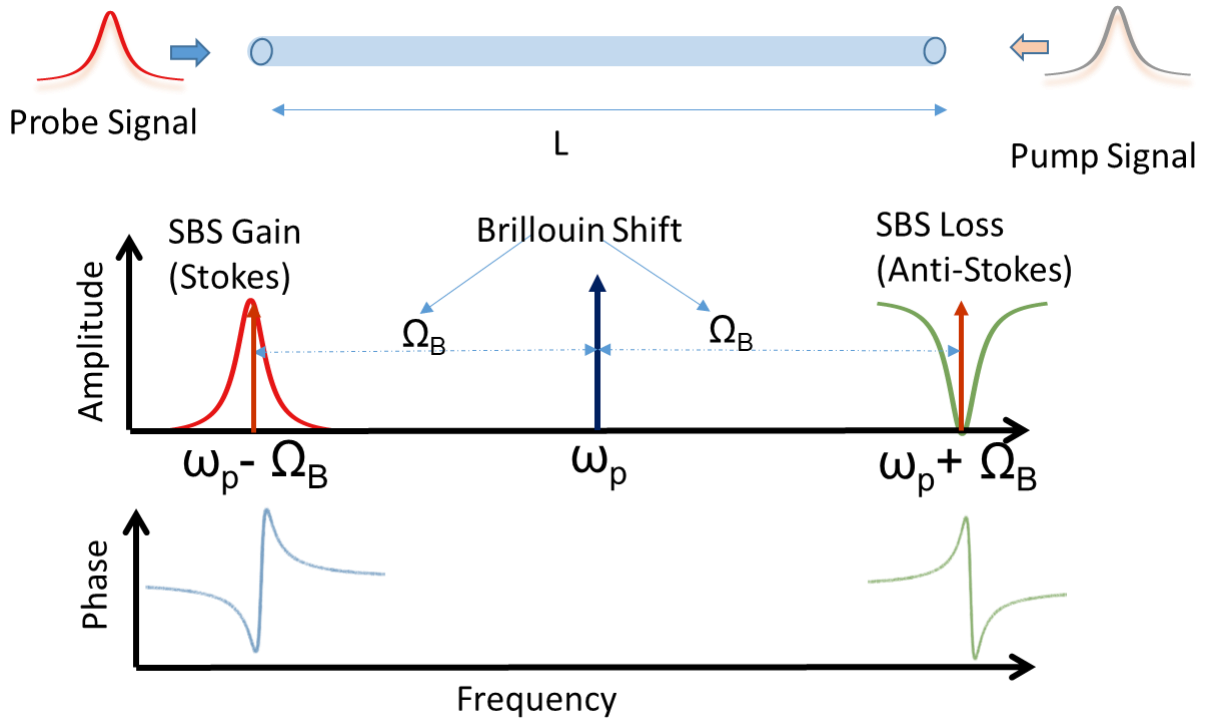


Fig. 2.166: Stimulated Brillouin scattering gain and loss response generation in optical fiber

The evolution equation in terms of Brillouin gain corresponding to the m^{th} acoustic mode can be

represented as [72],

$$\frac{dP_s}{dz} = \alpha P_s - \frac{g_B}{A_{eff}} L(v) P_s P_p \quad (2.45)$$

$$\frac{dP_p}{dz} = -\alpha P_s - \frac{g_B}{A_{eff}} L(v) P_s P_p \quad (2.46)$$

Here peak Brillouin gain corresponds to m^{th} acoustic wave is represented as,

$$g_B = \frac{2n^8 p_{12}^2 \omega^3}{\Omega_m \rho_0 c^4 \Gamma} \quad (2.47)$$

Here, $n^4 p_{12} = \gamma_e$ is the electrostrictive constant of the optical media [71]

The effective acoustic optic area determines the strength of the acousto-optic interaction in optical fibers with different profiles of core doping. is represented by,

$$A_{eff} = \frac{\langle f^2(r) \rangle}{\langle \xi_m(r) f^2(r) \rangle} \langle \xi_m(r)^2 \rangle \quad (2.48)$$

The SBS Lorentzian profile of the Brillouin gain can be represented as,

$$L_{gain}(\Omega) = \frac{(\Gamma/2)^2}{(\Omega - \Omega_m)^2 + (\Gamma/2)^2} \quad (2.49)$$

The dispersive Lorentzian of the Brillouin phase can be as,

$$L_{phase}(\Omega) = \frac{(\Omega - \Omega_m)}{(\Omega - \Omega_m)^2 + (\Gamma/2)^2} \quad (2.50)$$

The SBS gain response can be represented as,

$$G_{sbs} = g_B P_p L_{eff} L_{gain}(\Omega) \quad (2.51)$$

The SBS phase response can be represented as,

$$\phi_{sbs} = g_B P_p L_{eff} L_{phase}(\Omega) \quad (2.52)$$

Here, g_B is the Brillouin gain coefficient, L_{eff} is the effective interaction length.

2.9.1 SBS gain and Phase response

SBS is a promising microwave photonic signal processing technique due to its narrowband gain and loss response, which can be tailored using the pump laser frequency and power. The numerical expression of SBS gain and the phase response is given by equations (2.53) and (2.54) [73],

$$G_{sbs} = \frac{\gamma P_p}{1 + 4 \left(\frac{\Delta\omega}{\Gamma_B} \right)^2} L_{eff} \quad (2.53)$$

$$\phi_{sbs} = \frac{\gamma P_p \left(\frac{\Delta\omega}{\Gamma_B} \right)}{1 + 4 \left(\frac{\Delta\omega}{\Gamma_B} \right)^2} L_{eff} c \quad (2.54)$$

where $\gamma = \left(\frac{g_B}{A_{eff}} \right)$ is the nonlinear SBS parameter, defined as the ratio of the SBS gain coefficient g_B and effective area A_{eff} . $\Delta\omega$ is the detuning of the frequency from the peak SBS gain and Γ_B is the Brillouin linewidth. P_p is the pump power, L_{eff} is the effective length of the optical fiber. The gain experienced by the signal is given by $\exp(G_{sbs})$.

2.9.2 SBS Stoke's frequency response

The SBS gain and phase response for a 3 km single mode fiber over different pump power levels

is shown in Fig. 2.17(a) and 2.17(b). The gain response increases with an increase in pump power level. With an increase in pump power level, the slope of the SBS gain response increases, resulting in a sharp increase in gain response.

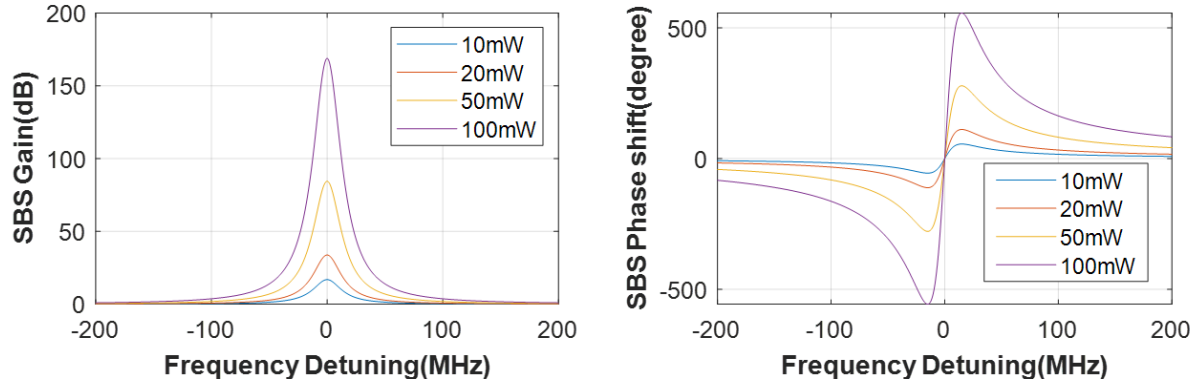


Fig. 2.177: (a) simulated SBS gain response for pump power variation from 10mW to 100mW, (b) simulated SBS phase response for pump power variation from 10mW to 100mW.

2.9.3 SBS anti-Stoke's frequency response

The SBS anti-Stoke loss and phase response for a 3 km single mode fiber over different pump power levels is shown in Fig. 2.18(a) and Fig. 2.18(b). The loss response increases with an increase in pump power level. With an increase in pump power level, the slope of the SBS gain response increases, resulting in a sharp increase in loss response.

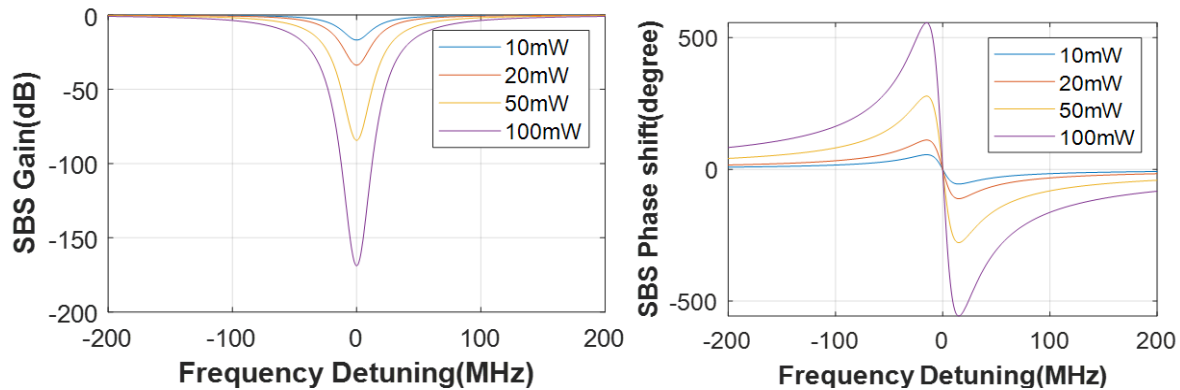


Fig. 2.188: (a) simulated SBS loss response for pump power variation from 10mW to 100mW, (b) simulated SBS phase response for anti-Stoke pump power variation from 10mW to 100mW.

2.10 SBS based signal processor

The radio frequency (RF) and microwave filter constitutes a fundamental element within the front end of both the RF transmitter and RF receiver, where it performs critical functions, such as suppressing self-interference originating from the transmitter, mitigating jamming signals to prevent amplifier saturation, and selectively isolating the signal of interest. The realization of RF and microwave filters through photonic implementation has garnered considerable academic and industrial interest, as photonic technologies inherently provide broader operational bandwidths and superior tunability compared to their electronic counterparts, thereby addressing the limitations associated with conventional electronic filter architectures.

Stimulated Brillouin scattering based signal processing enables the generation of highly tunable filters. Over the past decade, several techniques to generate highly reconfigurable filters have been proposed. The following section discusses the research articles and publications for the generation of reconfigurable filter using SBS.

2.8.1 SBS based wideband filter

A SBS based photonic processor for wideband filtering is demonstrated in Fig. 2.20(a) [73]. Here, the RF signal is modulated with an optical modulator and passed through the SBS processing unit to filter the required frequency band, tunable by tuning the pump laser signals. Here, a series of pump lasers has been used to generate multiple SBS gain responses, which, when combined, generate a wideband bandpass filter with a bandwidth tunable from 30MHz to 440MHz.

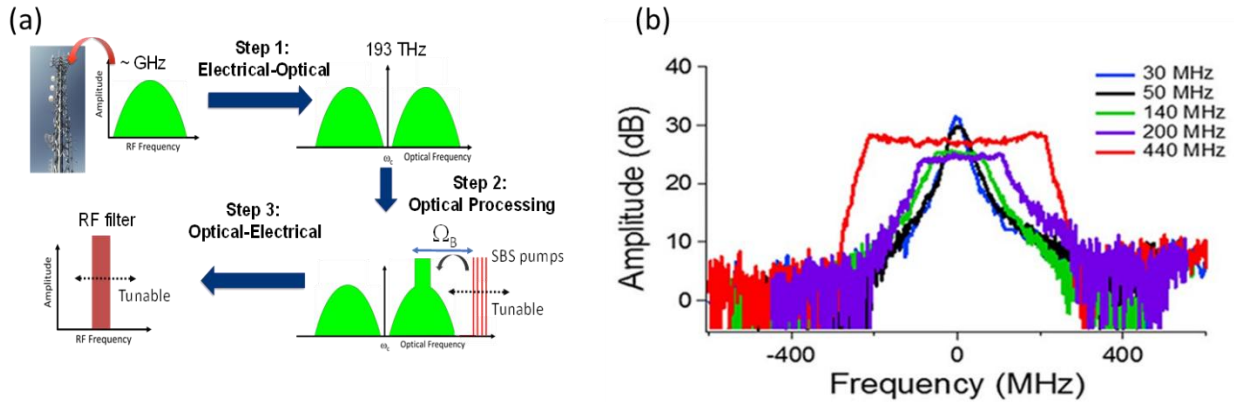


Fig. 2.199: (a) Tunable wideband filter generation using stimulated Brillouin scattering with multiple pumps, (b) Wideband filter with bandwidth variation from 30 MHz to 440 MHz [85]

Table 2.2 lists out the other articles and publications that offer valuable perspectives on the generation of microwave frequency band filters using stimulated Brillouin scattering for practical applications in FMCW microwave photonic radar systems.

Table 2.2: Microwave photonic filter

RF filter using microwave photonics for radar application		
[74]	Tunable photonic microwave single bandpass filter based on phase shift introduced in the periodic structure of a fiber Bragg grating	Microwave bandpass filter obtained exhibits a FWHM of 130 MHz and an extinction ratio (ER) close to 15 dB. The tunability of the filter is achieved by changing the DC voltage
[75]	GHz-bandwidth optical filters based on high-order silicon ring resonators	The tunable filters having a bandwidth of 1-2 GHz, FSRs of 50 GHz, and out-of-band extinction ratios of 50 dB at the drop port.
[76]	Single-Passband Microwave Photonic Filter Based on Stimulated Brillouin Scattering	A microwave filter with frequency variation from 1 to 20 GHz, featuring high resolution with a -3-dB bandwidth of only 20 MHz and a high out-of-band rejection of 31 dB, is demonstrated.

2.8.2 Electromagnetically induced Transparency/absorption analogue Filter

The EIT analogue filter has transparent spectra between absorption spectra, and the EIA has absorption spectra in between transmission spectra, allowing tight control over the transmission and reception of the signal of interest while attenuating interfering signals. EIT and EIA have been demonstrated in multiple media, however SBS based EIA and EIT analogues are of primary interest due to narrowband gain or loss response and high frequency tunability. Table 2.2 lists the articles and publications which offer valuable perspectives on the generation of highly reconfigurable EIA and EIT analogue filters for FMCW radar applications.

Table 2.3: Electromagnetically induced Transparency and electromagnetically induced absorption

Ref.	Electromagnetically induced transparency	Key Achievements
[77]	Electromagnetically induced transparency (EIT) filter using SBS	EIT filter with a BW of approx. 25MHz and frequency tunability up to 40GHz using two SBS gain responses and a loss response.
[78]	Fano resonances and induced transparency by coherent interactions between Brillouin resonances	The Fano and EIT-like resonance is demonstrated up to 43 GHz. The asymmetry and polarity are electrically controlled by varying the bias to the intensity modulator, whereas the strength is varied by tuning the Brillouin pump power and/or the bias. The depth and 3 dB linewidth of the transparency window in the EIT-like resonance are controlled using pump and probe parameters.
[79]	Electromagnetically induced Transparency	have shown how an opaque optical medium may be rendered transparent by means of a destructive interference, which is caused by applying an

		additional electromagnetic field
[80]	Electromagnetically induced transparency and slow light with optomechanics	This work has demonstrated EIT and tunable optical delays in a nanoscale optomechanical crystal, utilizing the optomechanical nonlinearity to control the velocity of light through engineered photon–phonon interactions. At low temperature (8.7 kelvin), optically tunable delay of 50 nanoseconds with near-unity optical transparency, and superluminal light with a 1.4 microsecond signal advance is demonstrated.
	Electromagnetically induced absorption	Key Achievements
[81]	Electromagnetically induced absorption using SBS	An EIA/EIT filter using a single laser is demonstrated, featuring frequency tunability of up to 40 GHz.
[82]	Electromagnetically induced absorption	Demonstrated the possibility of a large resonant increase in atomic absorption due to coherent interaction with optical radiation. The phenomenon of EIA is complementary to EIT.
[83]	Analog of Electromagnetically Induced Absorption in Plasmonics	The electromagnetically induced absorption is demonstrated for classical oscillators by coupling a plasmonic dipole nanowire to a plasmonic quadrupole wire pair. Manipulating the coupling phase of the oscillators obtains constructive interference of the two excitations, leading to enhanced absorption.
[84]	SBS loss based electromagnetically induced absorption in a single sideband modulated signal	The analogue of EIA within Brillouin loss is harnessed to demonstrate an efficient MPNF with an ultrahigh rejection of >75 dB.

The tunability of the EIT filter allows transmission within a specified frequency band, and EIA to reject in the desired frequency band is highly useful in a jamming environment where the filter transmission or rejection band need to be tuned. This motivated the development of highly flexible EIT analogue and EIA analogue filter responses with high tunability in bandwidth, frequency, and rejection ratio. In many applications, it is necessary to switch from a bandpass filter to a bandstop filter. Therefore, an investigation has been conducted to transition between the EIT analogue filter response and the EIA analogue filter response.

The primary focus of the thesis is on the development of an FMCW radar with a waveform capable of simultaneously estimating both range and velocity with high accuracy and resolution. This includes the generation of the waveform photonically with high tunability in frequency, enabling real-time tuning for multipurpose radar applications. The study also includes the development of a highly flexible RF filter using stimulated Brillouin scattering to filter the desired frequency band in both the transmitter and receiver systems. The investigation further extended to utilizing stimulated Brillouin scattering to mitigate chromatic dispersion in phase-modulated links demodulated with an MZI, which expands the scope of using a phase modulator for FMCW radar applications.

2.11 Conclusion

Microwave photonics-based communication links possess several advantages, including large bandwidth waveform generation, low transmission loss, high speed, very low jitter, real-time analog signal processing, and immunity from EMI. These capabilities bring significant attention from researchers to investigate microwave photonic technology-based radar systems for future

generations. Some techniques for waveform generation, radar signal processing for simultaneous range and velocity estimation, and signal processing, such as highly tunable filters, have been demonstrated with significant advantage over the electronic-based approach.

With significant improvement in microwave photonic technology in recent years, there is scope for innovative techniques to address the challenges of photonic radar. The enhanced demand for high-frequency agility, multiband, and highly tunable radar to cater to multiple applications has become apparent. This signifies that photonics-based solutions are of great interest for future-generation radars and communication systems, facilitating novel techniques and approaches for future research. I will discuss the research conducted in this thesis in the coming chapters.

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Chapter 3

Photonicallly Generated Radar Waveforms for Range and Velocity Estimation

The ability to simultaneously detect the range and velocity of a target using radar systems is a fundamental requirement in numerous applications, including navigation and autonomous vehicles. High-speed, high-resolution radars are crucial for detecting challenging targets, such as low-radar cross-section (RCS) aircraft and unmanned aerial vehicles (UAVs). Linear chirped waveforms are widely used to enhance range resolution for detecting low RCS targets; however, single linearly chirped waveforms exhibit high range-Doppler coupling. The chirped waveforms at higher frequencies lead to a higher Doppler shift, which is useful for detecting slow-moving targets, such as UAVs. We propose a straightforward method to generate a frequency-agile dual-chirp waveform using a single MZM modulator and a tunable reference laser, where the center frequency is tuned by varying the reference laser frequency. The dual-chirp waveform cannot simultaneously detect the direction of motion of the target. We have presented a multiband complementary V-shaped chirp waveform for simultaneously estimating the target's range, speed, and direction of motion with higher accuracy. A V-shaped chirp is used in the lower frequency band to estimate range with very high accuracy due to the low Doppler shift at lower RF

frequencies, and a complementary V-shaped chirp is employed in the higher frequency band with a higher bandwidth to achieve a higher Doppler shift, thereby enabling high velocity resolution to detect even slow-moving targets. A theoretical analysis is performed, along with experimental verification, of the generation of a multiband complementary waveform using a dual-polarization dual-parallel Mach-Zehnder modulator (DPol-DPMZM). The range and velocity of the target have been estimated over an emulated 100m target range. Furthermore, the range and velocity of a slow-moving toy car are also experimentally validated.

The remainder of this chapter is organized as: the introduction, and challenges are addressed in Section 3.1, followed by generation of frequency agile dual chirp waveform generation using low bandwidth Mach Zehnder modulator in Section 3.2, generation of multiband complementary V-shape waveform in Section **Error! Reference source not found.**, simultaneous range and velocity estimation in Section **Error! Reference source not found.**, simulation of waveform generation and receiver setup in Section **Error! Reference source not found.** and Section **Error! Reference source not found.**, followed by the experimental setup for waveform generation in Section **Error! Reference source not found.**, emulation of receiver setup in Section **Error! Reference source not found.**, simultaneous range and velocity estimation of toy car in Section **Error! Reference source not found.**, with conclusion and future scope in Section **Error! Reference source not found.**

This chapter presents results from the journal and conference contributions listed below.

- Mukund Jha, Lakshay Kumar, Rajveer Dhawan, Atul Khanna, Anu Sharma, and Amol Choudhary, “Multiband complementary dual-chirp waveform for the simultaneous range and velocity detection,” in *Appl. Opt.* 64, 6746-6756 (2025)

- M. Jha, A. Khanna, L. Kumar, R. Dhawan, and A. Choudhary, “Frequency Agile Dual-Chirp Microwave Waveform Generation with MZM,” in *Conference on Lasers and Electro-Optics/Europe (CLEO/Europe 2025)*
- M. Jha, R. Dhawan, A. Sharma, and A. Choudhary, “Photonically-generated Multiband Complementary Dual Chirp Waveforms for Range and Velocity Detection,” in *Conference on Lasers and Electro-Optics Pacific Rim (CLEO-PR 2024)*.

This chapter includes results from a patent filed given below.

- Amol Choudhary, Mukund Jha, R. Dhawan, Anu Sharma, “System And Method For Velocity Detection of a Radar Target” *Indian Patent No. 202411039577, May 2024 (filed)*.

3.1 Introduction

Radar sensors are a primary technology for detecting, ranging, classifying, and imaging long-distance targets in all weather conditions. The radio frequency signal is transmitted in the air, and the reflected echo signal is computed to estimate the range, speed, and direction of motion of target objects [1] - [4]. Traditional radar systems have low bandwidth availability, a large size, low resolution, and high DSP requirements for processing, making it difficult to detect and track small, slow-moving targets.

Frequency-modulated continuous-wave radar utilizes linear frequency-modulated signals, inheriting high pulse compression capability for high-resolution target detection. Traditionally, the LFM waveforms are generated using conventional electronics, which restricts the system functionality due to limited tunability in bandwidth and frequency. To overcome this bottleneck, a microwave photonics-based approach offers the generation of high-bandwidth waveforms, tunability in frequency and bandwidth, immunity to electromagnetic interference (EMI), and low signal degradation over long distances [5]-[12]. It also provides a direct interface with a radio frequency signal, making it suitable for radar applications. Owing to the above advantages, MWP-based signal processing [13]-[19], waveform generation [20]-[24], and antenna remoting technology [25] are utilized for ultra-high-resolution, real-time imaging radars [26]-[29].

Over the past decade, various approaches have been demonstrated to realise radar using microwave photonics technologies. MWP radar has been demonstrated using a single LFM waveform [30], however, this generated high range-Doppler ambiguity, resulting in the estimation of either range or velocity with high accuracy. This significantly reduces the range accuracy for high-velocity targets. To overcome this, a dual-chirp-based waveform has been proposed, consisting of an up-

chirp and a down-chirp that start at a common frequency for simultaneous range and speed estimation [31] - [32]. However, the generation of these waveforms requires complex modulators and precise DC bias control to produce a dual-linear chirped waveform, which limits system robustness in the face of environmental challenges. These require a tunable RF tone to tune the frequency of the dual-chirp waveform, which necessitates a high-speed modulator for generating higher-frequency waveforms. Some photonic techniques utilize optical injection techniques to generate dual chirps [4], but this approach comes with challenges related to system complexity, where precise control of the master and slave lasers is required. In this chapter, we present a straightforward technique for generating a frequency-agile dual-chirp waveform, where the frequency of the dual chirp is tuned by tuning the laser frequency. The frequency tunability is achieved using an external laser, resulting in the bandwidth of the modulator becoming independent of high-frequency dual chirp generation.

A single dual chirp waveform can simultaneously estimate only range and speed. The direction of motion cannot be estimated using a single dual chirp waveform. To estimate the direction of motion of the target, a triangular chirp [33] or a multiband triangular chirp [34] is proposed, consisting of an upchirp followed by a downchirp. This waveform requires double the time period to achieve the same range resolution as a normal dual chirp waveform and also requires complex time-frequency analysis to estimate the direction of motion. The time-frequency analysis requires a high-speed ADC, thus increasing the complexity of the radar receiver. The simultaneous range and velocity are also demonstrated using asymmetric dual chirp [34] and multiband LFM [35] techniques; however, these methods have moderate range-Doppler coupling, which increases the range errors for high-velocity moving targets. Table 3.1 presents the various state-of-the-art articles and publications on waveform generation, range, and velocity estimation for an FMCW radar.

Table 3.1 List of state-of-the-art articles for waveform generation & detection technique in a photonic radar

Ref	Microwave waveform generation Technique	Key Achievements
[35]	Photonic Generation of Frequency and Bandwidth Multiplying Dual-Chirp Microwave Waveform	Dual-chirp microwave waveforms at 16 GHz with a 2048 MHz bandwidth and at 32 GHz with a 4096 MHz bandwidth are generated through an 8 GHz RF carrier and a 512 MHz bandwidth baseband signal.
[36]	Bandwidth-Doubled dual-chirp waveform with freely tunable central frequency	The dual-chirp microwave signals centered at 1.5 GHz and with bandwidths of 100 MHz and 200 MHz are generated by using a DPMZM modulator and a frequency-shifted optical carrier generated by an MZM
[37]	Transmission of dual-chirp microwave waveform over fiber with compensation of dispersion-induced power fading	The generation of a dual chirp is demonstrated using a DPMZM modulator. Chromatic dispersion compensation is achieved by using a polarization controller (PC) and verified for a 25km SMF fiber.
[38]	Dual band dual-chirp waveform generation	The generation of dual-band dual-chirp microwave signals at C-band and X-band (centered at 7 GHz and 10 GHz) with tunable bandwidth.
[23]	Waveform generation of switchable down-, up-, and dual-chirped linear frequency-modulated microwave signals	The LFM signal, switchable between up-chirp, down-chirp, and dual-chirp, is demonstrated using a DP-DPMZM modulator. The switching is achieved by tuning the DC bias.
[39]	Generation of Linear Frequency-Modulated Waveforms by a Frequency-Sweeping Optoelectronic Oscillator	The LFM waveform is generated by applying an injection strength controller in the OEO for rapid frequency sweeping..
[40]	Generation of Broadband Reconfigurable LFM Waveforms Via Heterodyne-Beating	A dual-laser synchronization system is established, and an LFM waveform with reconfigurable bandwidth, central frequency,

	Synchronized Lasers	and pulse duration is produced
[22]	Linearly chirped microwave waveform generation with large time-bandwidth product by optically injected semiconductor laser	LCMW with TBWPs as large as 1.2×10^5 (bandwidth 12 GHz; temporal duration 10 μ s) are successfully generated. The flexibility for tuning the center frequency, bandwidth and temporal duration is also demonstrated.
Ref	Microwave Photonic Radar	Key Achievements
[29]	Quadrupled LFM signal generation and implementation of high range resolution radar	Accurate distance measurement with an error of 4 mm within a range of ~ 3.5 meters. Detection of two targets is demonstrated with a range resolution of 1.875 cm.
[41]	FMCW Radar with multi target detection	Demonstrated detection of multiple targets at different distances ranging from 150 cm to 210 cm. The range resolution of < 15 cm is also shown
[31]	Dual-chirp microwave photonic radar based on compressive sensing for distance and velocity measurement.	The distance for a stationary target is estimated with a measurement error of about 1.560 cm. With moving targets, the simulation results present that the maximum distance error is 1.2 cm, and the velocity error is below 0.140 m/s
[42]	Range and velocity estimation using multifomat waveform	The radar is demonstrated to have a distance error of less than 2 cm and a velocity error of less than 0.180 m/s. With the switch of the single-chirp signal and the dual-chirp signal, the direction of the moving target can be distinguished.
[33]	Range and velocity estimation using a triangular chirp	Photonic radar in X-band with a bandwidth of 2 GHz is demonstrated. A distance measurement ranging from 0.50 m to 0.78 m with absolute errors are less than 3 mm is shown.

In this chapter, we have also demonstrated the simultaneous estimation of range and velocity using a multiband complementary V-shaped chirp waveform. This waveform has a V-shaped chirp at a lower frequency band with a chirp rate (k_1) and a complementary V-shaped chirp waveform at a higher frequency band with chirp rate ($k_2=2k_1$). A dual chirp at a lower frequency is used to estimate the range due to low Doppler shifts, and a complementary V-shaped chirp at a higher frequency band is used to estimate velocity with higher accuracy due to high Doppler shifts at higher frequency, increasing the velocity resolution estimation. The direction of motion is estimated by estimating a reference frequency ($k_1\Delta t$) and comparing with the de-chirped frequencies of the complementary V-shaped chirp.

The generation of a multiband complementary V-shaped chirp waveform is also demonstrated using a single dual-polarization dual-parallel Mach-Zehnder modulator (DPol-DPMZM), as shown in Fig. 3.4. A V-shaped chirp at the lower frequency band is generated by applying a baseband up-chirp and an RF tone to the upper DPMZM of DPol-DPMZM. A complementary V-shaped chirp is generated by applying a downchirp LFM and an RF tone at a higher frequency to lower DPMZM of the DPol-DPMZM modulator. The frequency of the lower band V-shaped chirp and the upper band V-shaped chirp is tuned by tuning the RF tone frequency, making the waveform highly tunable.

The purpose of having a multiband complementary V-shaped chirp is to use low low-frequency band V-shaped chirp to have a lower Doppler shift to reduce the range-Doppler coupling and improve the range estimation accuracy, and use an upper-frequency band V-shaped chirp to have a high Doppler shift to have better estimation of velocity with higher resolution to detect slow-moving targets. The direction of motion of the target is estimated by harnessing the difference in Doppler shift of V-shaped chirps at different frequency bands, using low-bandwidth spectrum

analysis with low-speed analog-to-digital converters (ADCs) to achieve a simpler receiver system.

The receiver setup for estimating simultaneous range and velocity using a multiband complementary V-shaped chirp waveform consists of a dual-drive Mach-Zehnder modulator (DDMZM), as shown in Fig. 3.6. The transmitted and received waveform is coupled to two RF ports of the DDMZM modulator. The beating between the modulated transmitted and received echo waveforms at the photodetector generates de-chirped frequencies. The two de-chirped frequencies ($\Delta f_1, \Delta f_2$) generated by beating of lower band V-shaped chirps have been used to estimate range. The de-chirped frequencies ($\Delta f_3, \Delta f_4$) generated by beating the upper band V-shaped transmitted and received echo waveform generated by beating of the upper band has been used to estimate velocity. The combination of V-shaped chirps at different frequency bands has been used to estimate both range and velocity simultaneously with higher accuracy.

3.2 Frequency agile dual chirp waveform generation using a low bandwidth single modulator for range and speed estimation

A dual chirp waveform consists of an upchirp and a downchirp signal with a common starting frequency as shown in the spectrogram plot in Fig 3.1(c). The range Doppler coupling in dual chirp waveform is low due to opposite Doppler shift in upchirp and downchirp. The schematic diagram of frequency agile dual chirp waveform generation using a MZM modulator is shown in Fig 3.1. The optical carrier with frequency ω_0 is coupled to optical port of MZM modulaotr. The upchirp signal with chirp rate k_1 is fed to RF port of MZM modulator. An external laser with frequency ω_p is coupled using a circulator. The modulated chirp signal with external laser beats at photodetector to generate a dual chirp signal with center frequency of the dual-chrip as $(f_p - f_c)$.

The electric field at the output of the MZM consisting of two Y-branch coupler and an upchirp

signal applied to RF port can be represented as,

$$\begin{bmatrix} E_1(t) \\ E_2(t) \end{bmatrix} = \frac{\sqrt{l_{MZM}}}{2} \begin{bmatrix} 1 & i \\ i & 1 \end{bmatrix} \begin{bmatrix} e^{i\phi(t)} & 0 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & i \\ i & 1 \end{bmatrix} \begin{bmatrix} E_0(t) \\ 0 \end{bmatrix} \quad (3.1)$$

Here l_{MZM} is the optical loss and $\phi(t)$ is the relative phase shift between the two arms of the MZM modulator.

$$\phi(t) = \phi_{DC} + \phi_{upchirp} \cos(kt^2) \quad (3.2)$$

Here, $\phi_{DC} = \pi \frac{V_{DC}}{V_{\pi,DC}}$, $\phi_{upchirp} = \pi \frac{V_{upchirp}}{V_{\pi,upchirp}}$, with $V_{\pi,DC}$ is the half-wave voltage of MZM, $V_{\pi,upchirp}$ is the upchirp signal voltage value, respectively, and V_{DC} is the DC voltage applied to the MZM through a DC power supply. ϕ_{DC} controls the DC bias of the modulator and $\phi_{upchirp}$ is the modulation index of MZM.

Solving equations (3.1) and 3.2, the electric field at the output of the MZM modulator considering the real part can be represented as,

$$E_1(t) \propto e^{i\omega_0 t} [e^{i\phi(t)} - 1] \quad (3.3)$$

Applying the Jacobi-Anger expression $e^{i\phi(t)}$ can be represented as,

$$e^{i\phi(t)} = e^{i\phi_{DC}} \sum_{n=-\infty}^{\infty} J_n(\phi_{upchirp}) e^{inkt^2} \quad (3.4)$$

The electric field at the output of MZM can expressed as,

$$E_1(t) \propto \frac{e^{i\omega_0 t}}{2} \left[e^{i\phi_{DC}} \sum_{n=-\infty}^{\infty} J_n(\phi_{upchirp}) e^{inkt^2} - 1 \right] \quad (3.5)$$

Considering the null point biasing of the modulator, $\phi_{DC} = 2n\pi$, the electric field considering only the fundamental terms can be represented as,

$$E_1(t) \propto \frac{1}{2} \left[J_0(\phi_{upchirp})e^{i\omega_0 t} + J_1(\phi_{upchirp})e^{ikt^2} + J_{-1}(\phi_{upchirp})e^{-ikt^2} - e^{i\omega_0 t} \right] \quad (3.6)$$

Considering $\phi_{upchirp} \ll 1$, the carrier term cancels and a baseband upchirp and a downchirp is generated at the output of MZM modulator,

$$E_1(t) \propto \frac{\phi_{upchirp}}{4} \left[e^{ikt^2} - e^{-ikt^2} \right] \quad (3.7)$$

An external laser with carrier frequency ω_p is coupled using a circulator, the modified electric field can be represented as,

$$E_1(t) \propto \frac{\phi_{upchirp}}{4} \left[e^{i(\omega_0 t + kt^2)} - e^{i(\omega_0 t - kt^2)} \right] + e^{i\omega_p t} \quad (3.8)$$

The beating at the photodetector with responsivity $\Re$ generates current represented as,

$$I(t) = \Re E_1(t) E_1(t)^* \quad (3.9)$$

Putting the $E_1(t)$ in the above expression, $I(t)$ can be represented as,

$$I(t) \propto \Re \frac{\phi_{upchirp}}{2} \left[e^{i((\omega_p - \omega_0)t + kt^2)} - e^{i((\omega_p - \omega_0)t - kt^2)} \right] \quad (3.10)$$

The photocurrent $I(t)$, represents an upchirp and a downchirp at center frequency determined by $(\omega_p - \omega_0)$ in equation (3.10). The center frequency of the generated dual-chirp can be tuned by tuning the external laser frequency ω_p making the waveform generation highly agile. The MZM modulator is used to modulate only the baseband chirp signal, the bandwidth of modulator required

for high frequency dual-chirp generation is same as bandwidth of chirp significantly reduces the cost of the radar system.

The experimental setup used to generate the frequency agile dual chirp waveform is shown in Fig. 3.1. The carrier frequency of 193THz with 8dBm of power is coupled to optical port of MZM modulator using a polarization coupler. The up-chirp signal of 1GHz bandwidth and temporal width of 10 μ s is coupled to RF port of MZM modulator. An external laser with tunable frequency and 11dBm of power is coupled using an optical coupler. The beating of modulated chirp signal with external laser generates a dual-chirp waveform which is recorded at oscilloscope. The recorded data is processed in MATLAB to generate spectrogram of dual chirp waveform.

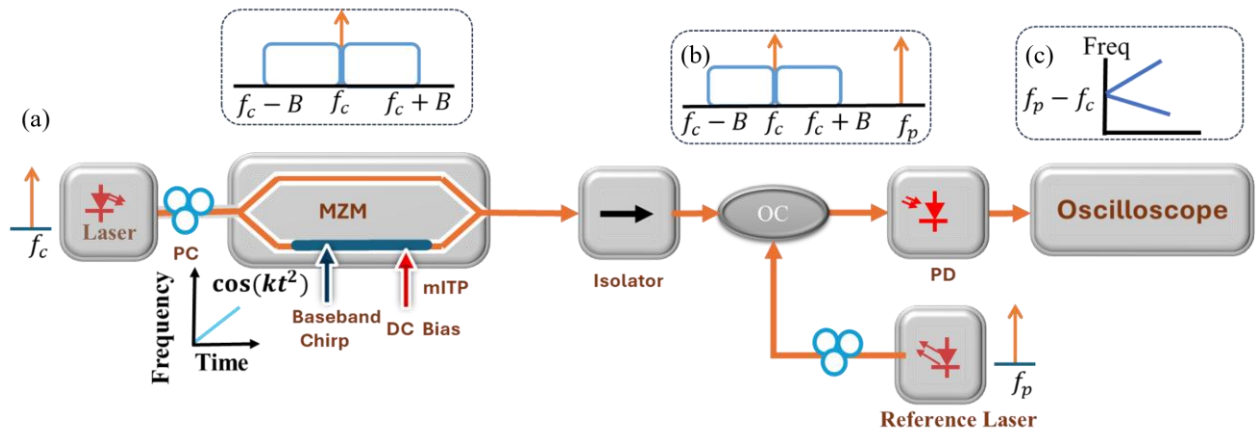


Fig. 3.1: Schematic of frequency agile dual chirp waveform generation using a low bandwidth MZM modulator.

The mixing of the external laser signal with the modulated signal at the PD generates a dual-chirp waveform, as shown in the amplitude-time plot in Fig.3.2(b) and the respective frequency spectrum in Fig.3.1(c). The frequency-time plot of the data recorded on oscilloscope confirms the generation of a dual-chirp waveform centered at 5 GHz having the up- and down-chirp bandwidth of 2 GHz with temporal width of 10 μ s as shown in Fig 3.2(a). The system's agile nature where the generation of dual chirp at different carrier frequencies, such as at 10 GHz, 15 GHz, and 20

GHz, is achieved by only tuning the reference laser frequency from 193.410 THz to 193.420 THz with a step size of 5 GHz as shown in Fig. 3.2(b) to Fig. 3.2(d). The bandwidth of the chirp is tuned by changing the input chirp bandwidth. The technique can be used for generation of other baseband waveforms like triangular waveform by tuning the external laser around the center of the modulated baseband downchirp signal.

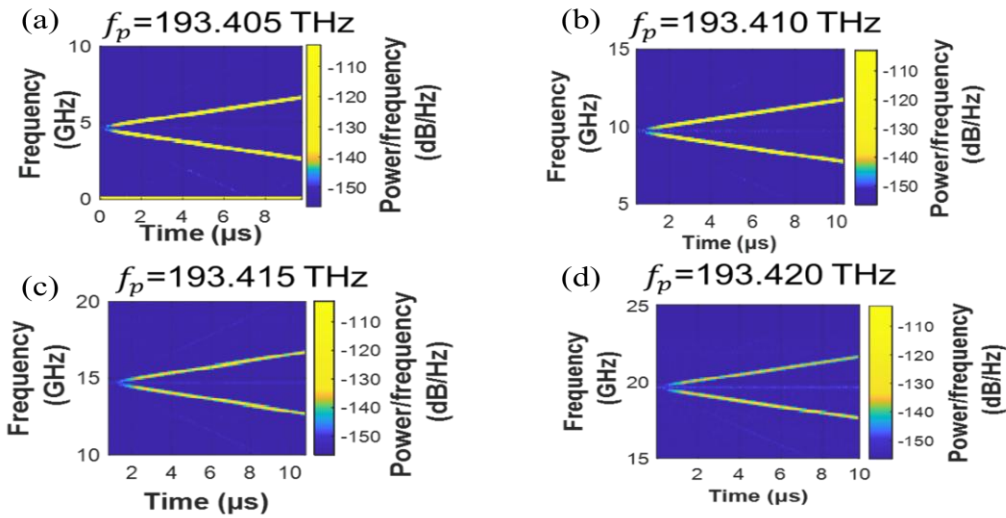


Fig. 3.2: Dual-chirp waveform generated at 5 GHz, 10 GHz, 15 GHz, and 20 GHz by tuning the external laser frequency from 193.405 THz to 193.420 THz as shown in (a) 5GHz, (b) 10GHz, (c) 15GHz, (d) 20GHz.

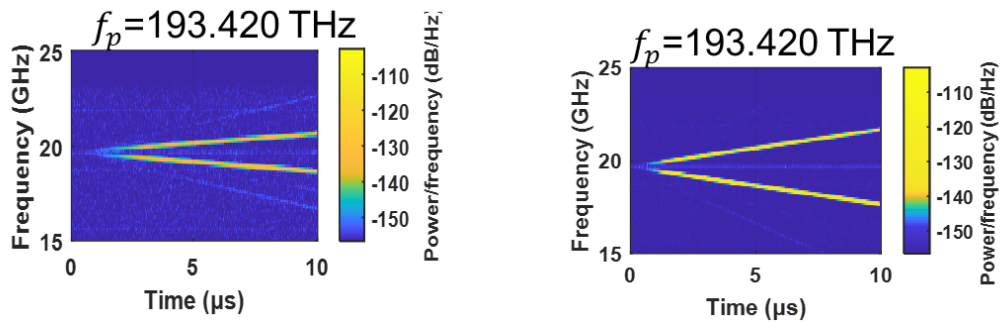


Fig. 3.3: The bandwidth tunability of dual chirp by tuning the input chirp bandwidth (a) 1GHz (b) 2GHz.

Hence this method will enable generation of highly tunable dual-chirp waveform with a low bandwidth and single MZM modulator for radar operation in all frequency bands like L, S, X, Ku, Ka for both long range and short-range target range and speed detection. However, the dual-chirp

waveform can simultaneously estimate range and speed. The direction of motion of target cannot be detected by a single dual chirp waveform. We have presented a multiband complementary dual chirp waveform for simultaneous range and velocity estimation along with direction of motion of the target.

3.3 Generation of Multiband Complementary V-shaped chirp waveform for simultaneous range and velocity estimation

The multiband complementary V-shaped chirp waveform for simultaneous range and velocity estimation is generated using a dual-polarization, dual-parallel Mach-Zehnder modulator (DPol-DPMZM), as shown in Fig. 3.4. The DPol-DPMZM has two DPMZM modulators with different polarizations. An upchirp LFM signal along with an RF tone is applied to RF port 1 and RF port 2 of the x-polarization DPMZM. A downchirp LFM along with a second RF tone is applied to RF port 3 and RF port 4 of the lower DPMZM. The RF tone 2 frequency is kept higher than RF tone 1 to have a complementary V-shaped chirp at a higher frequency band. The mixing of the x-polarization DPMZM modulated signal generates a V-shaped chirp at a frequency determined by RF tone 1. The mixing of the y-polarization modulated signal generates a complementary V-shaped chirp at a frequency determined by RF tone 2.

Error! Reference source not found.5 demonstrates the schematic of DPol-DPMZM utilized for the development of a multiband complementary V-shaped chirp. The DPol-DPMZM consists of a polarization beam splitter, two sub-DPMZMs, and a polarization beam combiner. The upper DPMZM comprises two MZM modulators, MZM_a and MZM_b , whereas the lower DPMZM consists of MZM_c and MZM_d . A

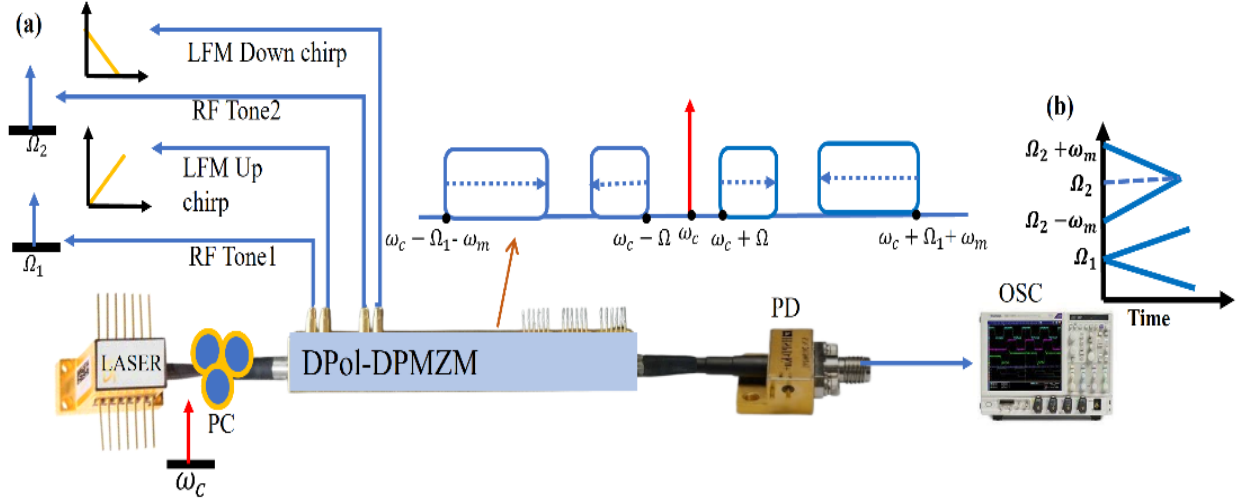


Fig. 3.4: (a) Architecture of multiband complementary V-shaped waveform generation using DPOL-DPMZM modulator; (b) Multiband complementary V-shaped chirp waveform. AWG: arbitrary waveform generator; RF: radio frequency, DPOL-DPMZM: Dual polarization dual parallel Mach Zehnder modulator; PD: photodetector; PC: polarization coupler.

continuous-wave laser with a carrier frequency ω_c is connected into a dual-polarization, dual-parallel Mach-Zehnder modulator via a polarization controller. A baseband up-chirp LFM waveform with a chirp rate k_1 is applied to MZM_a , while an RF tone with frequency Ω_1 is applied to MZM_b of the x-polarization DPMZM modulator. The RF tone of frequency Ω_2 is applied to

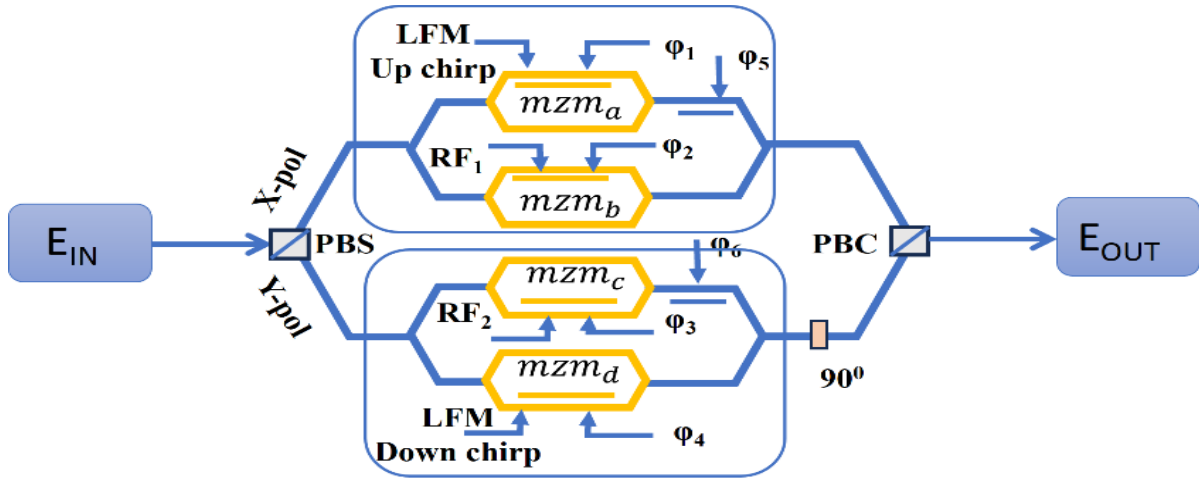


Fig. 3.5: Schematic of dual polarization dual parallel Mach-Zehnder modulator for complementary V-shaped chirp generation. E_{IN} is the input signal, and E_{OUT} is the output modulated signal, ϕ_1 to ϕ_6 : DC bias, PBS: polarization beam splitter, and PBC: polarization beam combiner.

MZM_c while a down-chirp LFM waveform with a chirp rate k_2 and an initial frequency of downchirp ω_m is applied to MZM_d of the y-polarization DPMZM modulator.

The optical field at the output of the x-pol DPMZM modulator is represented as,

$$E_{DPMZM(xpol)} = E_{MZM(a)} + E_{MZM(b)} \exp(i\phi_5) \quad (3.11)$$

Here, the electric field at the output of $E_{MZM(a)}$ can be represented as,

$$E1_{mzm,a}(t) \propto e^{i(\omega_c t)} [e^{i(\phi_1 + \phi_{upchirp} \cos(\pi k_1 t^2))} - 1] \quad (3.12)$$

Applying Jacobi-Anger expansion and biasing the modulator at MiTP to suppress the carrier, the output electric field of MZM_a can be represented as,

$$E1_{mzm,a}(t) \propto J_1(\phi_{upchirp}) \cos[(\omega_c + \pi k_1 t)t] + J_{-1}(\phi_{upchirp}) \cos[(\omega_c - \pi k_1 t)t] \quad (3.13)$$

where $\phi_1 = \pi \frac{V_{DC}}{V_{\pi(DC)}}$ and, $\phi_{upchirp} = \pi \frac{V_{upchirp}}{V_{\pi(RF)}}$, Here $V_{\pi} \left(\frac{DC}{RF} \right)$ is the half-wave voltage of MZM.

Similarly, the Electric field for $MZM_{b,c,d}$ can be represented as,

$$E1_{mzm,b}(t) \propto J_1(\phi_{rf1}) \cos[(\omega_c + \Omega_1)t] + J_{-1}(\phi_{rf1}) \cos[(\omega_c - \Omega_1)t] \quad (3.14)$$

$$\begin{aligned} E1_{mzm,c}(t) \propto & J_1(\phi_{downchirp}) \cos[(\omega_c + \omega_m - \pi k_2 t)t] \\ & + J_{-1}(\phi_{upchirp}) \cos[(\omega_c - \omega_m + \pi k_2 t)t] \end{aligned} \quad (3.15)$$

$$E1_{mzm,d}(t) \propto J_1(\phi_{rf2}) \cos[(\omega_c + \Omega_2)t] + J_{-1}(\phi_{rf2}) \cos[(\omega_c - \Omega_2)t] \quad (3.16)$$

Here, $\phi_{rf1} = \pi \frac{V_{rf1}}{V_{\pi(RF)}}$, $\phi_{rf2} = \pi \frac{V_{rf2}}{V_{\pi(RF)}}$, $\phi_{downchirp} = \pi \frac{V_{downchirp}}{V_{\pi(RF)}}$,

The output electric field of DPMZM ($xpol$) consists of MZM_a and MZM_b can be represented as,

$$E_{DPMZM(xpol)} \propto J_1(\phi_{upchirp}) \cos[(\omega_c + \pi k_1 t)t] + J_{-1}(\phi_{upchirp}) \cos[(\omega_c - \pi k_1 t)t] \\ - \{J_1(\phi_{rf1}) \cos[(\omega_c + \Omega_1)t] + J_{-1}(\phi_{rf1}) \cos[(\omega_c - \Omega_1)t]\} \exp(i\phi_5) \quad (3.17)$$

The output electric field of DPMZM ($ypol$) consists of MZM_c and MZM_d can be represented as,

$$E_{DPMZM(ypol)} \propto J_1(\phi_{downchirp}) \cos[(\omega_c + \omega_m - \pi k_2 t)t] \\ + J_{-1}(\phi_{downchirp}) \cos[(\omega_c - \omega_m + \pi k_2 t)t] \\ + J_1(\phi_{rf2}) \cos[(\omega_c + \Omega_2)t] \exp(i\phi_6) \\ + J_{-1}(\phi_{rf2}) \cos[(\omega_c - \Omega_2)t] \exp(i\phi_6) \quad (3.18)$$

The combined electric field at the output of DPol-DPMZM modulator consisting of x polarized DPMZM and y polarized DPMZM can be represented in vector form as,

$$E_{DPol-DPMZM} = E_{DPMZM(xpol)} + E_{DPMZM(ypol)}$$

The current detected at PD for both X-pol and Y-pol modulators can be represented by,

$$I(t) = \Re E_{DPol-DPMZM} \cdot E_{DPol-DPMZM}^*$$

$$I(t) = E_{DPMZM(xpol)} \cdot E_{DPMZM(xpol)}^* + E_{DPMZM(ypol)} \cdot E_{DPMZM(ypol)}^*$$

Applying $E_{DPMZM(xpol)}$ and $E_{DPMZM(ypol)}$ and considering the fundamental, $I(t)$ can be represented as,

$$\begin{aligned}
I(t) \propto & J_1(\phi_{upchirp})J_1(\phi_{rf1})\cos[(\Omega_1 + \pi k_1 t)t] \\
& + J_1(\phi_{upchirp})J_1(\phi_{rf1})\cos[(\Omega_1 - \pi k_1 t)t] \\
& + J_1(\phi_{downchirp})J_1(\phi_{rf2})\cos[(\Omega_2 + \omega_m - \pi k_2 t)t] \\
& + J_1(\phi_{downchirp})J_1(\phi_{rf2})\cos[(\Omega_2 - \omega_m + \pi k_2 t)t]
\end{aligned} \tag{3.19}$$

The mixing of the x-polarization product at PD generates a V-shaped chirp centered at Ω_1 and mixing of y-polarization products at PD generates a complementary dual chirp centered at Ω_2 . The RF tone Ω_2 is kept higher than Ω_1 to have a complementary dual chirp at the upper frequency band to have higher Doppler shifts. The bandwidth of the downchirp is kept double the bandwidth of the upchirp to have wide frequency separation between de-chirped frequencies due to V-shaped and complementary V-shaped chirps. Here, the starting frequency of both V-shaped chirps is tuned by changing the RF tone inputs to DPOL-DPMZM, making the waveform highly tunable.

3.4 Simultaneous Range and Velocity estimation using multiband complementary V-shaped waveform

FMCW radar based on microwave photonic technology utilizes beating between the transmitted waveform and the received echo waveform to generate de-chirped frequency components for range and velocity estimation. The multiband complementary V-shaped waveform has been used to estimate both range and velocity simultaneously. This waveform comprises a V-shaped chirp at a lower frequency band and a complementary V-shaped chirp at a higher frequency band with the same temporal width, as shown in **Error! Reference source not found.** (a). The complementary V-shaped chirped (depicted in green) waveform has a higher chirp rate $k_2 (B_2/T)$ than the V-shaped waveform (depicted in blue) with a chirp rate of $k_1 (B_1/T)$. This leads to a higher Doppler shift difference between the received up-chirp and down-chirp components of the complementary V-shaped chirp for direction estimation. Here, $B_{1/2}$ is the chirp bandwidth over a temporal width

of T. The range of the target has been estimated using de-chirped frequencies generated by the beating product of the transmitted and the reflected V-shaped chirped signal centered at a lower frequency band (f_1), which has a low range-Doppler coupling. Meanwhile, the velocity has been estimated from the de-chirped frequencies generated by the beating of the transmitted and reflected complementary V-shaped chirps at higher frequency bands, with higher Doppler frequency shifts, to achieve high velocity resolution. The combination of V-shaped chirps at different frequency bands results in simultaneous range and velocity detection with very high accuracy.

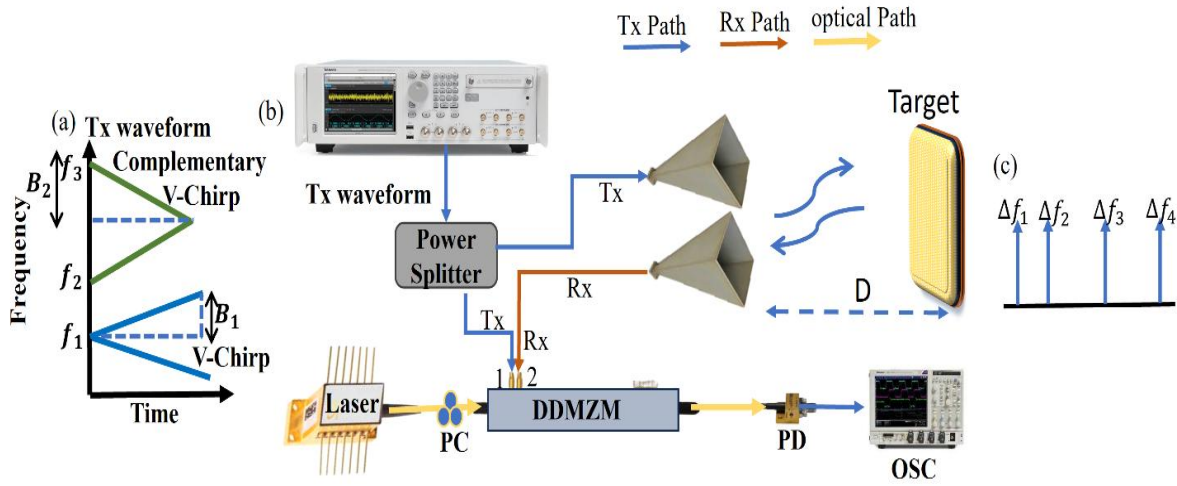


Fig. 3.6: Schematic of the multiband complementary V-shaped chirp based FMCW Radar receiver model for simultaneous range and velocity detection, (a) Tx waveform, (b) Radar receiver model, (c) De-chirped frequencies, DDMZM: Dual drive Mach Zehnder modulator; PD: photodetector; OSC: Oscilloscope, Tx: transmitted chirp, Δf_{1-4} : de-chirped frequencies, B_1, B_2 : bandwidth of V-shaped chirps at lower and upper frequency bands.

The FMCW radar receiver model, utilizing a dual-drive Mach-Zehnder modulator, is demonstrated in Fig. 3.6(b). Here a CW laser with frequency, ω_c is coupled into a dual-drive Mach Zehnder modulator (DDMZM). The transmitting waveform comprises a multiband dual-chirp (V-shaped and complementary V-shaped), as shown in **Error! Reference source not found.**(a), and is coupled to RF port 1 through an RF splitter. The reflected echo waveform is then coupled into RF port 2 of the DDMZM. The dual chirp starting at frequency f_1 has a chirp rate k_1 with bandwidth

B_1 and the complementary V-shaped chirp starting from f_2 and f_3 has a chirp rate k_2 with bandwidth $B_2(2B_1)$. The de-chirped frequencies Δf_{1-4} are shown in Fig. 3.6(c). The range and velocity of the target are estimated from the de-chirped frequencies Δf_{1-4} .

The transmitted multiband V-shaped waveform frequency can be expressed as,

$$f_{rx} = \left\{ \begin{array}{ll} f_1 + k_1 t & (\text{upchirp}) \\ f_1 - k_1 t & (\text{downchirp}) \end{array} \right\} \{V \text{ Chirp}\} \left\{ \begin{array}{ll} f_2 + k_2 t & (\text{upchirp}) \\ f_3 - k_2 t & (\text{downchirp}) \end{array} \right\} \{Complementary V \text{ Chirp}\} \quad 0 \leq t \leq T \quad (3.20)$$

The received echo waveform is delayed, reflecting the target range, and frequency-shifted, indicating the Doppler effect due to the target's velocity, which is fed into the second RF port of the DDMZM modulator. The frequency of the received echo waveform reflected from the target can be represented as,

$$f_{rx} = \left\{ \begin{array}{ll} f_1 + k_1(t - \Delta t) + f_{d1} & (\text{upchirp}) \\ f_1 - k_1(t - \Delta t) + f_{d2} & (\text{downchirp}) \end{array} \right\} \{V \text{ Chirp}\} \left\{ \begin{array}{ll} f_2 + k_2(t - \Delta t) + f_{d3} & (\text{upchirp}) \\ f_3 - k_2(t - \Delta t) + f_{d4} & (\text{downchirp}) \end{array} \right\} \{Complementary V \text{ Chirp}\} \quad 0 \leq t \leq T \quad (3.21)$$

The Doppler shifts for the up- and down-chirps of a V-shaped chirp with initial frequency f_1 are denoted by f_{d1} , f_{d2} , respectively. The Doppler shifts f_{d3} and f_{d4} correspond to the complementary V-shaped chirp, initiating frequencies f_2 and f_3 for the up-chirp and down-chirp, respectively. In a complementary V-shaped chirped waveform, the frequency content of the down chirp commencing at frequency f_3 exceeds that of the up-chirped waveform beginning at frequency f_2 resulting in f_{d4} being higher than f_{d3} . The complementary V-shaped chirp has a greater bandwidth than the V-shaped chirp and has a larger separation between f_{d3} and f_{d4}

compared to f_{d1} and f_{d2} . The difference in Doppler shifts between the V-shaped chirp and its complementary V-shaped chirp is harnessed to ascertain the target's direction of motion. The beating between the transmitted and received echo signal frequencies (f_{tx}), (f_{rx}), respectively, at the photodetector produces four de-chirped frequencies Δf_1 to Δf_4 , expressed as follows:

$$\begin{aligned}\Delta f_1 &= k_1 \Delta t - f_{d1} \\ \Delta f_2 &= k_1 \Delta t + f_{d2} \\ \Delta f_3 &= k_2 \Delta t - f_{d3} \\ \Delta f_4 &= k_2 \Delta t + f_{d4}\end{aligned}\tag{3.22}$$

Here, the Doppler shifts are $f_{d1} = \left(\frac{2v}{c}\right) f_1$, $f_{d2} = \left(\frac{2v}{c}\right) f_1$, $f_{d3} = \left(\frac{2v}{c}\right) f_2$, $f_{d4} = \left(\frac{2v}{c}\right) f_3$. The Doppler shift f_{d1} is equal to f_{d2} as the starting frequency is the same, and the Doppler shift is symmetrical with a change in frequency of both the up-chirp and down-chirp of the V-shaped chirp. The delay Δt is estimated using de-chirped frequencies (Δf_1 and Δf_2) with ($f_{d1} = f_{d2}$) is represented as,

$$\Delta t = \left(\frac{(\Delta f_1 + \Delta f_2)}{2k_1} \right)\tag{3.23}$$

The estimated range of the target is $R = \left(\frac{c \Delta t}{2}\right)$, putting Δt from (4), the target range is represented as,

$$R = \left(\frac{c (\Delta f_1 + \Delta f_2)}{4k_1} \right)\tag{3.24}$$

The target speed is estimated from Δf_3 and Δf_4 from (3) and is represented as,

$$v = \left| \frac{c(\Delta f_4 - \Delta f_3)}{2(f_2 + f_3)} \right|\tag{3.25}$$

The target receding away from the radar has positive velocity, while the target approaching the radar has negative velocity. The Doppler shift, f_{d4} , is greater than f_{d3} due to the higher initial frequency, (f_3), of the complementary dual chirp. Consequently, the de-chirped frequency (Δf_4) resulting from the mixing of the down-chirp components of the transmitted and echo complementary V-shaped signal is greater than the de-chirped frequency (Δf_3) arising from the beating of the up-chirps, as indicated in **Error! Reference source not found.**(a). An approaching target has a negative Doppler shift f_{d4} , which exhibits a greater magnitude than f_{d3} according to the fundamental frequency Doppler relationship. The combination of the down-chirp components from the transmission signal and the echo signal of the complementary V-shaped chirp produces a detuned frequency (Δf_4) that is lower than the detuned frequency (Δf_3) of the up-chirp components, as indicated in **Error! Reference source not found.**3.7 (b).

The target's direction of motion is determined by assessing the frequency difference between the estimated reference frequency $k_2\Delta t$ and the de-chirped frequencies Δf_3 and Δf_4 of the complementary dual chirp. The target is receding when the reference frequency $k_2\Delta t$ is nearer to the beating frequency Δf_3 than to Δf_4 , as shown in **Error! Reference source not found.**3.7 (c). Conversely, the target is approaching when $k_2\Delta t$ is closer to the beating frequency Δf_3 than to Δf_4 , as illustrated in **Error! Reference source not found.**7 (d). The Δf_1 , Δf_2 values interchange positions in the de-chirped spectrum for approaching and receding targets, while Δf_3 and Δf_4 exhibit a similar interchange for approaching and receding targets. For clarity, de-chirped frequencies are utilized in ascending order as Δf_1 to Δf_4 from the de- chirped spectrum. The direction of motion of the target is estimated from arranged de-chirped frequencies as,

$$|\Delta f_4 - k_2\Delta t| > |(\Delta f_3 - k_2\Delta t)| \quad \text{target is receding} \quad (3.26)$$

$$|\Delta f_4 - k_2 \Delta t| < |(\Delta f_3 - k_2 \Delta t)| \quad \text{target is approaching} \quad (3.27)$$

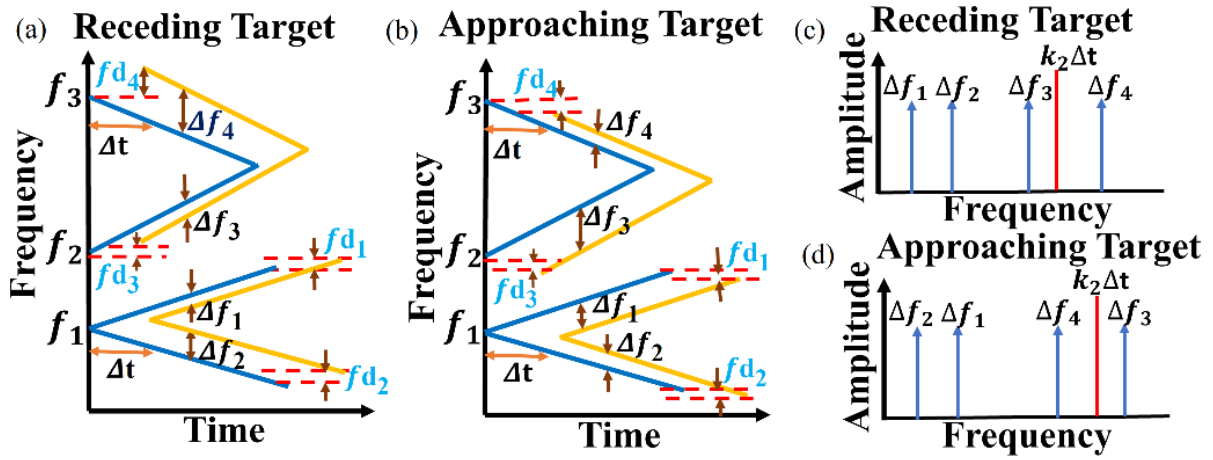


Fig. 3.7: The frequency vs time plot of multiband complementary V-shaped chirp for (a) receding, and (b) approaching target, (c) amplitude-frequency plot for de-chirped frequencies for receding target, and (d) approaching target, respectively. f_1 is the starting frequency of the V-shaped chirp f_2, f_3 , are starting frequency of the up-chirp and down-chirp of the complementary V-shaped chirp, Δf_{1-4} are de-chirped frequencies, f_{d1-d4} are Doppler shifts and Δt is the delay between transmitted and received echo signals.

The V-shaped chirp at a lower frequency (f_1) of the multiband complementary V-shaped chirp waveform has been used to precisely estimate the delay (Δt) as both the up-chirp and the down-chirp share a common starting frequency and identical chirp rates. The target range is estimated from equation (3.24). The speed of the target is estimated from the complementary V-shaped dual-chirp at the upper-frequency band from equation (3.25). The target's direction of motion is estimated from equations (3.26) and (3.27). In previous works, range accuracy decreases due to different starting frequencies of all chirps in multiband LFM [43], resulting in different Doppler shifts, which led to higher range-Doppler coupling and hence higher range and velocity error or in asymmetric dual chirp [44], the Doppler shifts are asymmetric for up-chirp and down-chirp, which increases range-Doppler coupling resulting in higher range and velocity error.

3.5 Simulation Model for the generation of multiband complementary V-shaped chirp using DPOL-DPMZM

Simulations in MATLAB were carried out to verify the feasibility of the specified approach for the generation of a multiband complementary V-shaped chirp waveform, as discussed in Section 4.2. The multiband complementary V-shaped chirp waveform generation simulation model includes a dual-polarization dual-parallel Mach-Zehnder modulator (DPOL-DPMZM), a laser, and a photodetector. The DPOL-DPMZM modulator comprises four RF ports, with two RF ports of x-polarization DPMZM receiving a baseband up-chirp LFM signal characterized by a bandwidth of 1 GHz and an RF tone signal at 3 GHz. The two RF ports of the y-polarization DPMZM receive a down-chirp signal with a bandwidth of 2 GHz and an RF tone signal at 8 GHz.

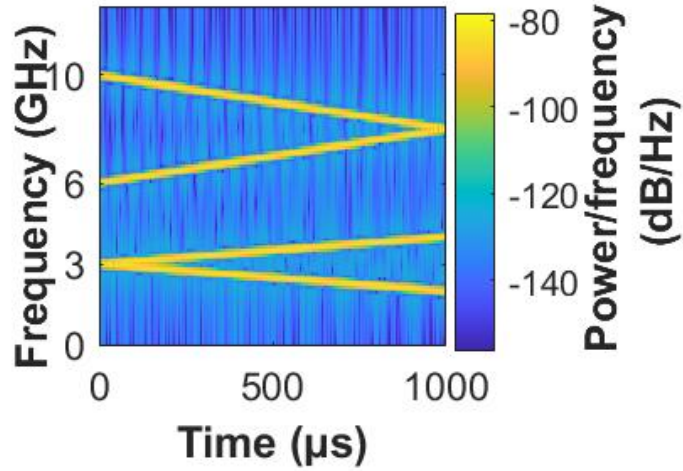


Fig. 3.8: frequency-time graph of multiband complementary V chirp with 1 GHz BW V chirp centered at 3 GHz and 2 GHz BW complementary V chirp centered at 8 GHz.

The beating of the first-order sideband of two MZMs in the x-polarization DPMZM modulator at the photodetector produces a V-shaped chirp centered at 3 GHz. The complementary V-

shaped chirp is produced by the beating of first-order sidebands from the two Mach-Zehnder modulators (MZMs) of the y-polarization DPMZM modulator at a central frequency of 8 GHz, as illustrated in Fig. 3.8. The precision of velocity estimation improves with an increase in the RF tone input of the y-polarized DPMZM modulator, which generates a complementary V-shaped chirp at a higher frequency. This leads to a greater Doppler shift and enhanced velocity resolution. The multiband complementary waveform generated can be tuned in frequency by tuning the two RF tones, making the waveform generation highly agile.

3.6 Simulation model of MWP radar receiver

To verify the concept of microwave photonic receiver discussed in section 3.4, a simulation in MATLAB is carried out. The simulation model of a radar receiver based on dual drive Mach Zehnder modulator is created, in which the transmission waveform and reception waveform are modulated using the DDMZM and subsequently de-chirped at the photodetector to validate the photonic receiver. The Tx waveform comprises a dual-complementary V-shaped chirp, featuring a V-shaped chirp centered at 3 GHz and a complementary V-shaped chirp centered at 8 GHz. The Rx echo waveform is produced by incorporating a delay corresponding to a target distance of 100 m and a Doppler shift ranging from -60 m/s to +60 m/s. The beating of the Tx waveform and Rx waveform at the photodetector produces four de-chirped frequencies, denoted as Δf_1 to Δf_4 . Fig. 3.9(a) illustrates the de-chirped frequencies (Δf_1 to Δf_4) for a target receding at a velocity of 60 m/s. Fig. 3.9 (b) presents the de-chirped frequencies (Δf_1 to Δf_4) for a target approaching at a velocity of -60 m/s. The target range is estimated from equation (3.24), and the target's speed is estimated from equation (3.25). The direction of the target is estimated from equations (3.26) and (3.27). The actual velocity vs the estimated velocity is shown in Fig. 3.9(c), and the actual range

vs the estimated range is shown in Fig. 3.9(d). The target is detected along with the direction of motion with a maximum range error of 14 cm and velocity error of 0.89 m/s over the range of 50 m to 100 m and velocity variation of -60 m/s to 60 m/s.

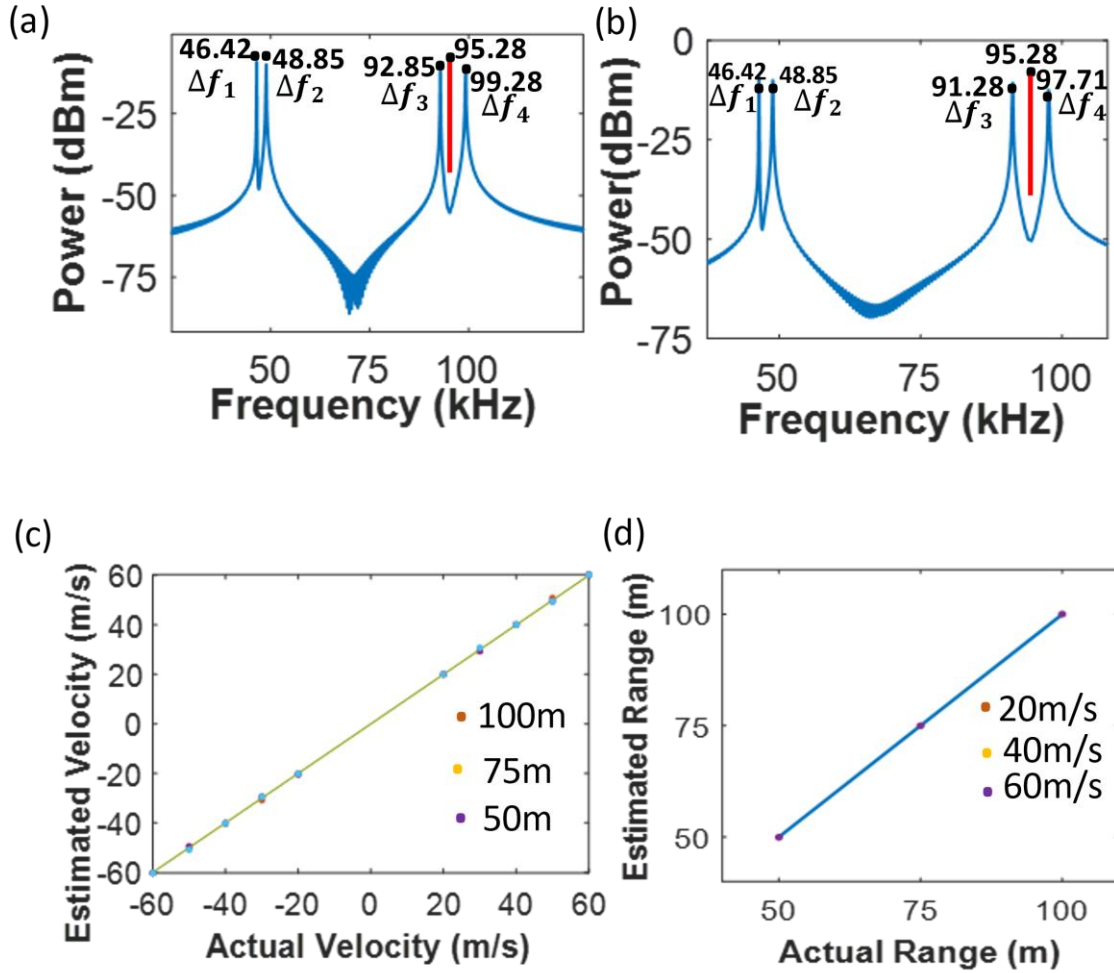


Fig. 3.9: (a) De-chirped frequencies for the target at 100 m receding with 60 m/s, (b) De-chirped frequencies for the target at 100 m approaching with 60 m/s, (c) Estimated velocity vs Actual velocity, (d) Estimated Range vs Actual Range,

3.7 Experimental verification of multiband complementary V-shaped chirp waveform

An experiment based on the schematic shown in Fig. 3.10(a) is set up and is carried out to validate the simulated results. Fig. 3.10(b) demonstrates the experimental setup utilized for the generation of a multiband complementary V-shaped waveform. A laser emitting a carrier signal at 1550 nm

with a power of 13.5 dBm is coupled into the DPol-DPMZM modulator via a polarization coupler to ensure polarization alignment with the modulator. Signal generators were utilized to produce RF tones, while a two-port arbitrary waveform generator (AWG) was employed to create the baseband up-chirp and down-chirp. The lab setup used to implement this experiment is shown in Fig. 3.10(b). The DPol-DPMZM consists of two distinct DPMZM architectures. The upper DPMZM receives an RF tone of 3 GHz along with a baseband up-chirp signal characterized by a bandwidth of 1 GHz and a temporal width of 1 ms. In contrast, the lower DPMZM is supplied with an RF tone of 8 GHz and a baseband down-chirp signal that has a bandwidth of 2 GHz and a temporal width of 1 ms.

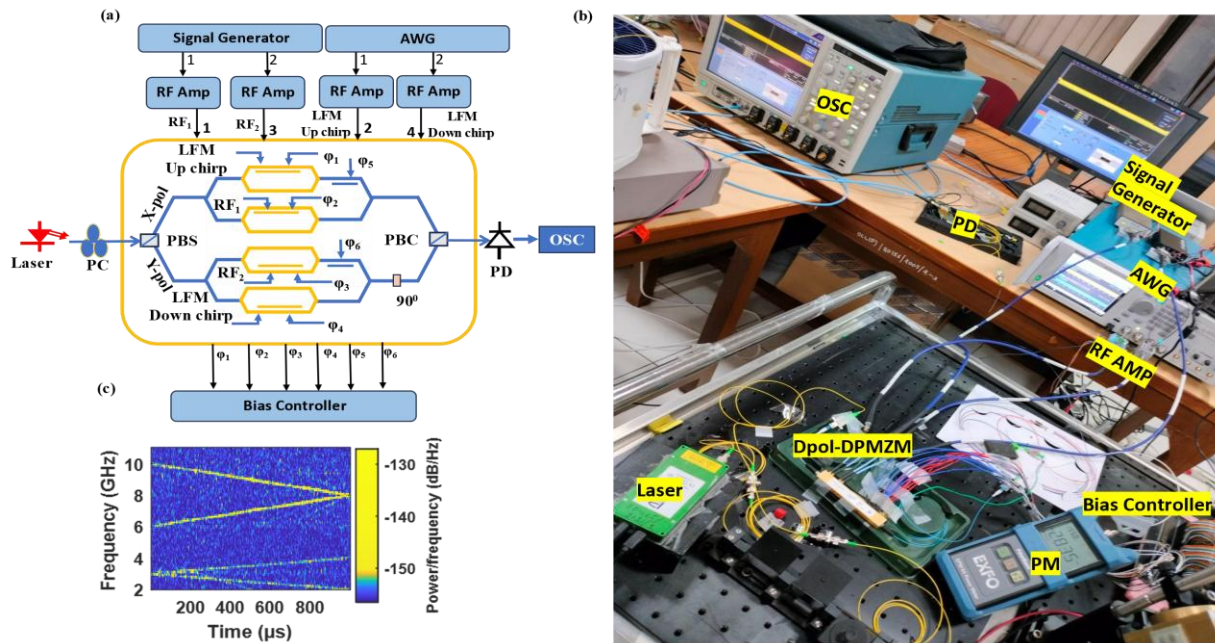


Fig. 3.10: (a) Schematic of Experimental setup of generation of multiband complementary V-shaped chirp, (b) Actual setup of waveform generation, (c) multiband complementary waveform generated using DPol-DPMZM modulator. PC: polarization controller; LFM: linear frequency modulated signal, PBS: polarization beam splitter, PBC: polarization beam coupler, PD: photodetector. OSC: oscilloscope

The RF tone signals are amplified with RF amplifiers and subsequently coupled to the RF ports of the modulator. The two DPMZMs exhibit bias at the minimum transmission point to produce

carrier-suppressed first-order sidebands. A bias controller regulates all six biases of the DPol-DPMZM modulator, ensuring the voltage aligns with the minimum transmission point. An oscilloscope captures the waveform produced at PD, and the FFT transform is applied to the recorded data in MATLAB to display the frequency-time plot of the generated dual complementary chirp waveform, as shown in Fig. 3.10 (c).

The modulated signals from DPol-DPMZM, which include the X-Pol signal (a 3 GHz tone with a 1 GHz bandwidth baseband up-chirp) and the Y-Pol signal (an 8 GHz tone with a 2 GHz bandwidth baseband down-chirp), are coupled to a photodetector. The mixing at the photodetector results in X-Pol DPMZM generating a V-shaped chirp centered at 3 GHz with a bandwidth of 1 GHz. In contrast, the mixing of the Y-Pol DPMZM modulated signal generates a complementary V-shaped chirp centered at 8 GHz, exhibiting a bandwidth of 2 GHz, as shown in Fig. 4.6(c). The center frequency of both dual chirps is adjusted by varying the RF tone, resulting in a highly tunable waveform. The scheme demonstrates the generation of tunable multiband complementary V-shaped chirp via a DPol-DPMZM modulator. The suppression ratio of the generated waveform can be enhanced by utilizing a higher-gain RF amplifier. The signal-to-noise ratio of the lower band V-shaped chirp is diminished because the DDMZM modulator is biased at MiTP, leading to a reduced signal-to-noise ratio in the lower frequency band. Nonetheless, enhancement is possible through link optimization techniques [45].

3.8 Emulation of Simultaneous Range and velocity detection using DDMZM

To validate the radar receiver model simulation, an experiment is set up with both the transmitted and echo waveforms emulating using AWG for different ranges and velocities. The model of the microwave photonic receiver is demonstrated through a dual-drive Mach-Zehnder modulator

(DDMZM) featuring two RF ports, which aims to estimate simultaneous range and velocity via a multiband complementary V-shaped waveform. The setup for the radar receiver model is shown in Fig. 3.11(a). The experimental configuration of the photonic receiver features a continuous-wave laser with a wavelength of 1550 nm, which is coupled to a dual-drive Mach-Zehnder modulator. The Tx-chirp waveform comprises multiband complementary dual-chirp waveforms, featuring a V-shaped chirp centered at 3 GHz with a chirp bandwidth of 1 GHz, alongside a complementary V-shaped chirp centered at 8 GHz with a bandwidth of 2 GHz, and a temporal width of 7 ms. The Tx-chirp waveform produced by the AWG is connected to the RF ports of the DDMZM.

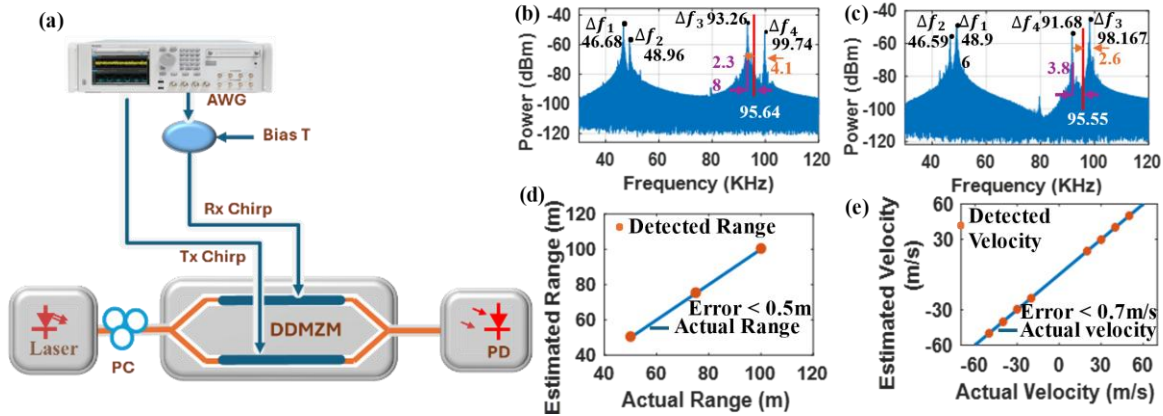


Fig. 3.11 : (a) Experimental setup of photonics receiver model with DDMZM modulator, (b) de-chirped frequencies of approaching target, and (c) de-chirped frequencies of a receding target. (d) Actual vs estimated range, (e) actual vs estimated velocity error. PC: polarization controller, PD: photodetector, AWG: arbitrary waveform generator.

The mimicked Rx-chirp waveform is generated by delaying the Tx-chirp waveform over distances of 50 m, 75 m, and 100 m, and applying a frequency shift corresponding to velocities ranging from -60 m/s to +60 m/s. This process is sequentially executed using an AWG and coupled to RF Port 2 of the DDMZM via a bias tee. A bias tee has been employed to couple the DC bias to the modulator. The output signals of the photonic modulated Tx-chirp and Rx-chirp waveforms mixing at the photodetector are captured on an oscilloscope and subsequently analysed using

MATLAB to produce the frequency spectrum of detuned frequencies Δf_1 to Δf_4 .

The target range is derived from equation (3.24), while the velocity magnitude is obtained from equation (3.25). The direction of the target's motion is estimated using equations (3.26) and (3.27). The detuned frequencies Δf_1 and Δf_2 are 46.68 kHz and 48.96 kHz, respectively, while Δf_3 , and Δf_4 are 93.26 kHz and 99.74 kHz for the target emulated at a distance of 100 m and a velocity of 60 m/s. The reference frequency $k_2\Delta t$, as estimated from equation (3.23), is indicated by the red line in the spectrum of de-chirped frequencies, illustrated in Fig. 3.11 (b). The estimated range is 100.4 meters, and the velocity is 60.7 meters per second. The error in the range of 0.4m and velocity of 0.7m/s is achieved. The direction is estimated from equations (3.26) and (3.27) as $|(\Delta f_4 - k_2\Delta t)|$, which is 4.1 kHz, exceeding $|(\Delta f_3 - k_2\Delta t)|$ at 2.38 kHz, as shown in **Error! Reference source not found.** (b). Here $|(\Delta f_4 - k_2\Delta t)| > |(\Delta f_3 - k_2\Delta t)|$ confirming that the velocity is positive, and the target is receding away from the radar.

Fig. 3.11 (c) demonstrates the detuned frequencies for the approaching target at 100 m and a velocity of -60 m/s. The detuned frequencies are Δf_1 , and Δf_2 , measured at 46.59 kHz and 48.96 kHz, respectively, while Δf_3 , and Δf_4 are recorded at 91.68 kHz and 98.16 kHz, respectively. The reference frequency $k_2\Delta t$ is indicated by the red line. The estimated range is 100.3 meters, and the speed is 60.7 meters per second which resulted in range error of 0.3m and speed error of 0.7m/s. The direction is estimated from equations (3.26) and (3.27). The value $|(\Delta f_4 - k_2\Delta t)|$ is (2.6 kHz) and $|(\Delta f_3 - k_2\Delta t)|$ is (3.8 kHz), resulting in $|(\Delta f_4 - k_2\Delta t)| < |(\Delta f_3 - k_2\Delta t)|$ confirming that the velocity is negative and the target is approaching the radar at -60.7 m/s. The maximum error in range estimation is 0.5 m, while the maximum error in velocity estimation is 0.7 m/s for a target measured at distances of 50 m, 75 m, and 100 m, with velocity varying from -60 m/s to 60 m/s, as shown in Fig. 3.11(d) and Fig. 3.11(e), respectively.

3.9 Experimental verification of simultaneous range and velocity detection of a toy car V-shaped chirp waveform

This section demonstrates the simultaneous detection of range and velocity for a moving toy car utilizing a multiband complementary V-shaped waveform, as depicted in Fig. 3.12(a). The configuration includes a continuous wave laser operating at 1550 nm with an output power of 12 dBm, coupled to a dual-drive Mach-Zehnder modulator. The Tx-chirp consists of a V-shaped chirp at 3 GHz and a complementary V-shaped chirp at 8 GHz, each with a temporal width of 2 ms. This signal is directed to an RF splitter, where one output is connected to an antenna via an RF amplifier, while the second output is linked to a DDMZM modulator to serve as a reference waveform. The Rx-chirp, after reflection from the toy car, is connected to the second RF port of the DDMZM

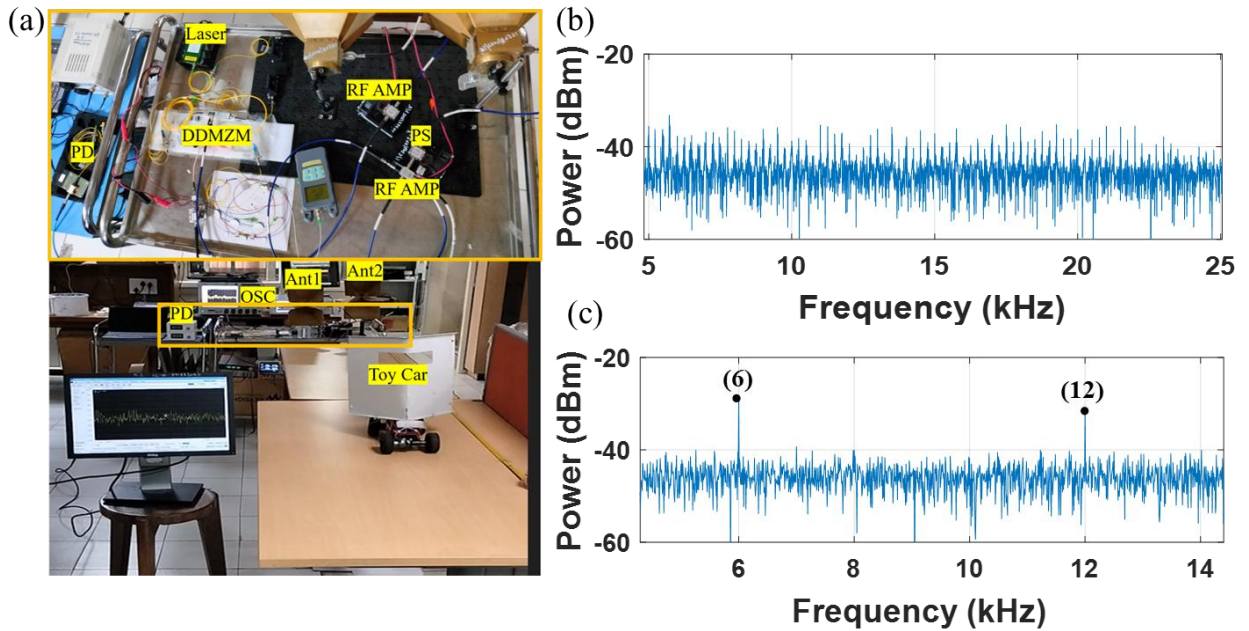


Fig. 3.12: (a) Experimental setup of a photonic FMCW Radar with multiband complementary V-shaped chirp. (b) spectrum of de-chirped frequencies for no target on the table (c) Spectrum of de-chirped frequencies for a static target (toy car) with clutter peaks filtering.

modulator via the Rx-antenna, following amplification by an RF amplifier. The modulated Tx-chirp and Rx-chirp signals are input to the photodetector. The mixing of the signals at the photodetector produces four de-chirped frequencies (Δf_1 to Δf_4), which are recorded using an oscilloscope.

Fig. 3.12 (b) displays the de-chirped frequencies in the absence of a target on the table. Continuous residual frequencies are observed because of reflection from the table. Fig. 3.12 (c) presents the de-chirped frequencies recorded with the toy car stationary on the table at 180 cm from the antenna, with clutter peaks from the table removed by filtering in MATLAB. The power of the clutter peak is lower than the reflection from the toy car, which is then filtered by setting a threshold value. The two de-chirped peaks occur at frequencies of 6 kHz and 12 kHz. The de-chirped frequency of 6 kHz results from the interference of a V-shaped chirp in a lower frequency band, while the de-chirped frequency of 12 kHz arises from the interference of a complementary V-shaped chirp in a higher frequency band.

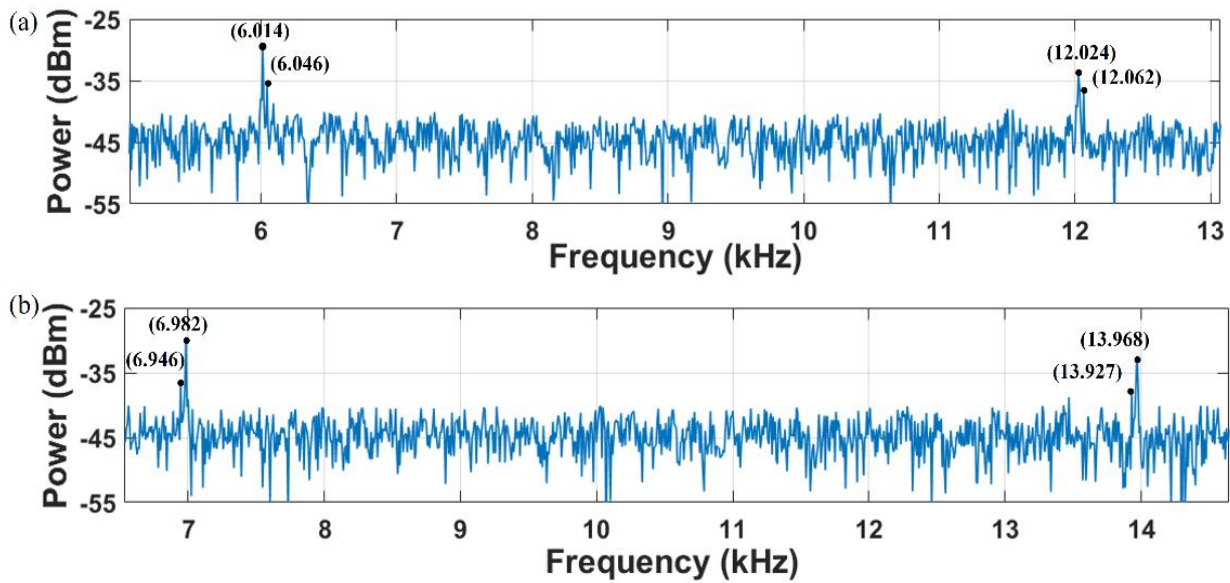


Fig. 3.13: (a) The spectrum of de-chirped frequencies for a toy car approaching the radar. (b) The spectrum of de-chirped frequencies for a toy car receding away from the radar.

Fig. 3.13 (a) shows the de-chirped frequencies resulting from the beating between the reference transmission waveform and the reception waveform for a toy car approaching the radar at a velocity of 0.7 m/s. The measurements were conducted with the toy car traveling at a speed of 0.7 m/s, covering a distance from 1500 cm to 2500 cm in a back-and-forth motion relative to the antenna. The data is filtered and processed using MATLAB DSP to eliminate unwanted peak frequencies resulting from reflections of static targets, including the table and car track.

The de-chirped frequencies of 6.014 kHz and 6.046 kHz result from the beating of the lower frequency V-shaped chirp, while the frequencies of 12.024 kHz and 12.062 kHz arise from the beating of the higher frequency complementary V-shaped chirp, as observed between the reflected waveform from the toy car and the reference Tx-waveform, and are represented below.

$$\Delta f_1 = 6.014 \text{ kHz}, \Delta f_2 = 6.046 \text{ kHz}$$

$$\Delta f_3 = 12.024 \text{ kHz}, \Delta f_4 = 12.062 \text{ kHz}$$

The range, velocity, and direction of approaching toy car are estimated from equations (3.24), (3.25), (3.26), and (3.27) by using de-chirped frequencies $\Delta f_1, \Delta f_2, \Delta f_3, \Delta f_4$, and chirp rate ($k_1 = 5 \times 10^{11}$) for the lower band V-shaped chirp and ($k_2 = 10 \times 10^{11}$) for upper band V-shaped chirp for temporal width of 2ms as,

$$R = \left(\frac{c (\Delta f_1 + \Delta f_2)}{4k_1} \right) = 1.81 \text{ m}$$

$$v = \left| \frac{c(\Delta f_4 - \Delta f_3)}{2(f_2 + f_3)} \right| = 0.36 \text{ m/s}$$

$$|(\Delta f_4 - k_2 \Delta t)| = 2 \text{ Hz}$$

$$(\Delta f_3 - k_2 \Delta t) = 36 \text{ Hz}$$

$$|(\Delta f_4 - k_2 \Delta t)| (2Hz) < (\Delta f_3 - k_2 \Delta t)(36Hz) , \text{ target is approaching}$$

The range for the toy car measured is 1.81 m, while the estimated speed is 0.4 m/s. The estimated reference frequency ($k_2 \Delta t$) is 12.06 kHz. The target is approaching the radar, which is confirmed by validating the equation $|(\Delta f_4 - k_2 \Delta t)| (2Hz) < (\Delta f_3 - k_2 \Delta t)(36Hz)$ above. This confirms that the target is approaching, indicating a negative velocity of -0.4 m/s.

The de-chirped frequencies due to the beating of the reflected waveform from the toy car receding at 0.7 m/s are shown in Fig. 3.13 (b). The de-chirped frequency spectrum is filtered to remove unwanted peaks resulting from reflections from the table. The de-chirped frequencies 6.946 kHz and 6.982 kHz are due to the beating of the reflected lower band V-shaped chirp with the lower band V-shaped chirp of the reference Tx-waveform. The de-chirped frequencies 13.927 kHz and 13.968 kHz are due to the beating of the upper band of the received echo from the toy car with the upper band complementary V-shaped chirp of the reference waveform, as shown below,

$$\Delta f_1 = 6.014 \text{ kHz} , \Delta f_2 = 6.046 \text{ kHz}$$

$$\Delta f_3 = 12.024 \text{ kHz} , \Delta f_4 = 12.062 \text{ kHz}$$

The range, velocity, and direction of a receding toy car are estimated from equations (3.24), (3.25), (3.26), and (3.27) by using de-chirped frequencies Δf_1 , Δf_2 , Δf_3 , Δf_4 , and chirp rate ($k_1 = 5 \times 10^{11}$) for the lower band V-shaped chirp and ($k_2 = 10 \times 10^{11}$) for upper band V-shaped chirp for temporal width of 2ms as,

$$R = \left(\frac{c (\Delta f_1 + \Delta f_2)}{4k_1} \right) = 2.08m$$

$$v = \left| \frac{c(\Delta f_4 - \Delta f_3)}{2(f_2 + f_3)} \right| = 0.39m/s$$

$$|(\Delta f_4 - k_2 \Delta t)| = 38 \text{ Hz}$$

$$(\Delta f_3 - k_2 \Delta t) = 3 \text{ Hz}$$

$$|(\Delta f_4 - k_2 \Delta t)| (2 \text{ Hz}) > (\Delta f_3 - k_2 \Delta t)(36 \text{ Hz}) , \text{ target is receding}$$

The estimated range of a toy car is 2.08 m, and the speed of a toy car is 0.4 m/s. The reference frequency ($k_2 \Delta t$) estimated using Δt from (4) is 13.929 kHz. The target is receding away from the radar.

This technique determines the target range and velocity directly from de-chirped peaks with high accuracy. This technique does not require a highly complex frequency-time analysis to determine velocity, as suggested in previous works that used triangular chirps. This technique also does not compromise the range error due to asymmetry in chirps. Hence, this technique proposes a simple method to estimate simultaneous range and velocity with high accuracy. Comparison of the proposed radar system with the previous reported method is shown in Table 3.2

Table 3.2: Comparison of the proposed radar system with previous reported methods

Ref	Range Resolution/Error	Velocity Resolution	Estimation of the direction of Motion	DSP requirement

[42]	Range error of 2cm	Velocity error of 0.180m/s	Yes, requires switching between single LFM and dual-chirp	High
[45]	Range error of 0.93 cm	0.16 m/s	Yes, requires complex time-frequency analysis	High
[33]	Range error of 3mm		Yes, time-frequency analysis is required.	High
[47]	Range resolution 15.98 cm	NO	NO	Low
[43]	Range error of 0.005%	Velocity error of 0.59%	Yes, Not explicit requires to physically check beating frequencies on the spectrum analyser	High
Our work	Range error of 8 cm	Velocity error of 0.3m/s	Yes, directly computed from beat frequencies	Low

The method presented has some cons as well, as the waveform required to estimate range and velocity is complex and requires a higher bandwidth power amplifier and antenna to support multiband transmission and reception.

3.10 Conclusion and Future Outlook

In this chapter, a simple and straightforward method for generating a dual-chirp waveform using a single low-bandwidth modulator is successfully demonstrated. The tunability in frequency is

achieved using an external tunable laser. The frequency tunability from 5 GHz to 20 GHz is experimentally validated using a low-bandwidth modulator. In this method, the bandwidth of the modulator required is limited to the baseband chirp bandwidth only, reducing the overall cost of the system. This waveform can be used for highly reconfigurable radars where the frequency of the radar needs to be tuned based on the surveillance requirements of a slow-speed target or a high-speed target.

A novel approach for simultaneously estimating range and velocity is also demonstrated using a multiband complementary V-shaped chirp waveform. The core of this work lies in the frequency-agile generation of this unique waveform, which was realized using a single dual-polarization dual-parallel Mach-Zehnder modulator. This method enables high-accuracy range estimation using a lower-band V-shaped chirp with low range-Doppler coupling. The high resolution of velocity estimation is achieved by a complementary V-shaped chirp at a higher frequency band to have a higher Doppler shift. The waveform generation allows tuning of the V-shaped chirp center frequency by simply adjusting the RF tones, offering high flexibility.

The experiment was conducted to validate the MWP-based FMCW radar using a multiband V-shaped chirp waveform in an emulated environment for the simultaneous detection of range and velocity of high-speed targets. We achieved a range accuracy of 50 cm and a velocity accuracy of 0.7 m/s, demonstrating the system's robustness across a wide range of target distances (50 to 100 m) and velocities (-60 to 60 m/s). The target's direction of motion is detected directly from the spectrum of the de-chirped frequencies. This technique significantly reduces receiver complexity by minimizing the need for complex digital signal processing (DSP) and high-speed analog-to-digital converters (ADCs). With the current advancements in lasers and modulators on chips using integrated photonics, this technique could be harnessed for on-chip photonic radar with real-time

imaging, low data storage requirements, high frequency agility, and accuracy, paving the way for next-generation sensing technologies in autonomous vehicles and industrial robotics.

The radar system, which utilizes a complex waveform for transmission and reception, requires highly reconfigurable microwave frequency band filters to reject the unwanted frequencies generated during mixing at the photodetector. The radar waveform for surveillance of small-size, slow-speed targets to high-speed targets requires the bandwidth of the waveform to be tuned to increase the range resolution, which also necessitates real-time filter reconfigurability. Stimulated Brillouin scattering-based filtering with real-time tunability in frequency can be utilized for this purpose. The generation of highly flexible filters for multiband radar applications is discussed in the next chapter.

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Chapter 4

Highly Flexible Electromagnetically Induced Absorption and Transmission Analogue Filters using Stimulated Brillouin Scattering

Radars are used for various applications, including self-driving cars, medical diagnostics, surveillance, tracking, imaging, and astronomy. These diverse applications require radar to operate in different frequency bands, ranging from low frequencies in MHz to high frequencies in the hundreds of GHz. Low-frequency bands are primarily used for early warning radar, and navigation. The mid-frequency range in VHF and UHF, as well as the L and S bands, is used for surveillance, air traffic management, and radio communication. Higher frequency bands in C, Ku, Ka, V, and W are used in autonomous vehicles, satellite communication, healthcare, and precision aircraft landing. The radar frequency of operation and bandwidth needs to be reconfigured for different applications, which require a highly reconfigurable filter with the capability to tune the filtering frequency and bandwidth. Microwave photonics has been extensively used for signal processing to filter out unwanted frequency bands and pass the desired frequency bands.

This chapter presents the generation of a highly reconfigurable electromagnetically induced transparency (EIT) analogue filter with transparent frequency spectra between two absorption

spectra, and an electromagnetically induced absorption (EIA) analogue filter with absorption spectra between two transparent spectra, utilizing stimulated Brillouin scattering. The generation of the spectra hole (SH) analogue filter response is also demonstrated with agility in frequency, bandwidth, and rejection ratio. The switching from the EIT analogue filter response to the EIA filter response has also been demonstrated. This highly programmable frequency response can be utilized for photonic radar applications that require precise control over transmission and rejection bandwidths with very low power.

The remainder of this chapter is organized as follows: A brief introduction is provided, and the challenges are addressed in Section 4.1, followed by the principle of operation and theory in Sections 4.2 and 4.3. Section 4.4 explains the simulations, followed by the experimental setup in Section 4.5, with discussions & results in Section 4.6 and the conclusion and future scope in Section 4.7.

This chapter presents results from the journal and conference contributions listed below.

- Mukund Jha, Rajveer Dhawan, Reena Parihar, and Amol Choudhary, "Highly Flexible RF Filter Responses Using Stimulated Brillouin Scattering with Bandwidth up to 1 GHz," *J. Lightwave Technol.* 41, 6675-6683 (2023)
- M. Jha, R. Parihar, R. Dhawan, and A. Choudhary, "Switchable Analogue of Electromagnetic Induced Absorption and Transmission-like Filter Responses using Stimulated Brillouin Scattering," in *Conference on Lasers and Electro-Optics/Europe (CLEO/Europe 2023)*, paper cd_12_3.

4.1 Introduction

Microwave photonics is widely used for millimeter-wave signal processing in radar applications

[1] [2]. Modern radar systems require operation over multiple frequency bands with reconfigurable bandwidth to fuse data from various bands, thereby achieving a greater detection range and higher range resolution to detect even small targets [3] [4]. This requires a highly agile and reconfigurable filter with real-time tunability over multiple frequency bands to filter the interfering signal. Over the past decade, various microwave photonic filters have been proposed based on Mach-Zehnder interferometers [5], fiber Bragg gratings [6]– [7], ring resonators [8]-[9], and stimulated Brillouin scattering [10]-[18]. Recently, multiband radar waveform generation has been demonstrated using a DDMZM modulator with filtering using stimulated Brillouin scattering [19]. However, the SBS bandwidth is very narrow, ~ 30 MHz, limiting the radar's capability to operate over a higher bandwidth. In practical applications, highly flexible filters with precise control over transmission/rejection bandwidth, stopband attenuation/passband gain, and rejection levels in various frequency bands, as well as high selectivity, are required. SBS with narrow gain and loss response enables the development of highly programmable and tunable filters.

Stimulated Brillouin scattering based MWP systems have been utilized to demonstrate various functionalities, including a single band-pass filter [18], a bandstop filter [19], null compensation of an analog photonic link [20], and non-reciprocal light storage [21]. SBS has also been utilized to replicate many responses in the RF domain within microwave photonics, including electromagnetically induced transparency (EIT), electromagnetically induced absorption (EIA), and spectra hole (SH) burning.

The EIT analogue filter has a transparent spectrum between absorption spectra, and EIA has an absorption spectrum between transmission spectra, allowing for tight control over the transmission and reception of the signal of interest while attenuating interfering signals. EIA/EIT has recently been reported in systems such as atomic systems [22-23], opto-electromechanical systems [24-25],

resonators [26], and plasmonic systems [27]. EIA/EIT analogues in the millimeter frequency band have been demonstrated using SBS [28]. An EIT analogue has also been shown using three pumps [29], but the filter's bandwidth was limited, and bandwidth tuning required high pump power, thereby reducing the filter's power efficiency. The spectral shapes of the EIT, EIA, and SH analogue filter responses are shown in Figs. 4.1 (a) -4.1 (c), respectively.

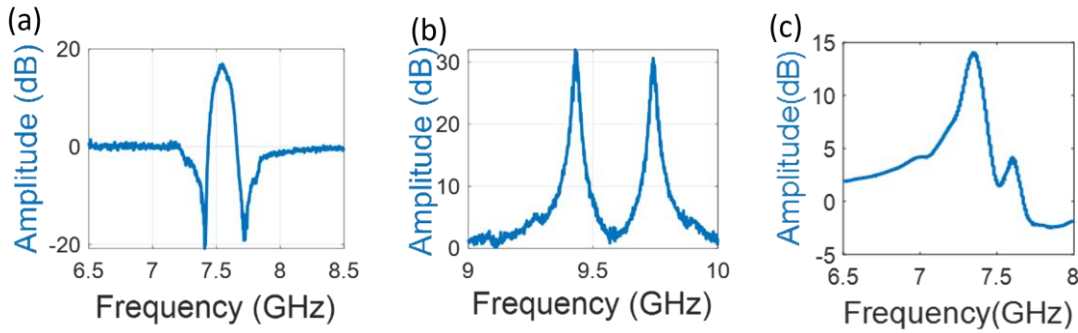


Fig. 4.1: EIA analogue, EIT analogue, and SH-burning analogue filter frequency response

The spectra hole burning analogue response has dual transmission spectra with a variable extinction ratio. SH response can be generated by selective absorption in a specific band of frequencies, resulting in increased transmission within a particular spectral band. SH burning-like Fano resonance [30] frequency response has been demonstrated using SBS. SH burning responses have been previously demonstrated in PMF and SMF [31], but they lack bandwidth tunability and frequency variation. SH frequency responses can be utilized for picosecond optical data processing [32]. SH filter shapes can also be utilized for controlled transmission/rejection of signals in the frequency band of interest.

In recent years, highly flexible filters with an inherent capability to have a transparent transmission window followed by rejection frequency windows on both sides have been proposed for the next generation of communication systems. The tunability in transparent/rejection bandwidth and

transmission/rejection ratio enables the mitigation of complex interfering signals, thereby improving the selectivity of the desired signal.

The control of EIT and EIA analogues frequency response depth and width can be harnessed to generate tunable band-pass and bandstop filters. EIT/EIA and SH frequency response have several applications in sensing, highly selective filtering, and signal processing. The EIT filter response has been previously demonstrated by interfering with the SBS Stokes response with a broad SBS anti-Stokes response [28]. A single laser-based EIA and EIT analogue response using an anti-Stokes response with low tunability and extinction ratio has been demonstrated [29]. Another approach to demonstrating an EIT analogue filter is using a single SBS Stokes response between two SBS anti-Stokes responses, with tunable bandwidth and rejection ratio. However, analogues of EIA and EIT with criteria such as seamless transition from EIA to EIT, wideband operation, and dynamic control of bandwidth, frequency, and rejection ratio with high power efficiency have yet to be demonstrated.

This chapter will discuss a method of generating highly tunable EIT analogue RF filter responses, EIA analogue RF filter responses, and an SH-burning analogue filter response using two SBS anti-Stokes responses. An EIA analogue with two SBS Stoke responses is also demonstrated, with a high rejection ratio. Switching between EIT and EIA, or vice versa, is achieved simply by tuning one of the pump laser wavelengths, making it highly suitable for applications that require live changes in frequency response. The technique to demonstrate these filters utilizes a phase-modulated analog photonic link leveraging RF phase interferometry between two SBS anti-Stokes/Stokes responses. We have significantly improved the power efficiency of these filters, as a high rejection ratio is achieved by utilizing the phase interferometry with low pump power.

Electromagnetically induced transparency and absorption response have been demonstrated with high bandwidth and rejection ratio tunability. The bandwidth tunability of over 1 GHz has been achieved for EIA, along with absorption depth tuning from 22dB to 50dB with high power efficiency. The EIT with transmission bandwidth variation of up to 1.2 GHz, along with transmission depth variation of 25dB to 51dB, is achieved. The frequency variations of the EIA and EIT filters are also demonstrated up to 15 GHz, limited by the available modulator bandwidth. The switching from EIA analogue filter frequency response to EIT analogue filter frequency response is achieved by tuning the pump laser frequency, making the filter highly reconfigurable without requiring a change to the setup.

A spectra hole burning analogue filter response has been generated using the same technique as the EIA analogue by tuning one of the two pump power levels. Finally, spectra hole burning frequency responses have also been demonstrated, with control over hole depth variation ranging from 14 dB to 36 dB. An improved EIA analogue with an absorption depth of up to 63 dB has been demonstrated using two Stokes responses. The use of RF phase interferometry to achieve bandwidths over 1 GHz demonstrates a significant improvement in power efficiency compared to previous approaches.

4.2 Principle of Operation

In this section, the working principle of the proposed EIA, EIT, and SH analogue filters for the microwave frequency band is discussed. These filters are demonstrated by harnessing RF phase interferometry in a phase-modulated link with two SBS anti-Stokes responses, as shown in Fig. 4.2. The probe signal of frequency ω_c is fed to a phase modulator (PM) to generate π out-of-phase sidebands. Two tunable pump lasers counter-propagate a probe laser in a fiber medium, generating

an SBS anti-Stokes response on the lower sideband (LSB) and upper sideband (USB) of a phase-modulated signal. The RF phase interferometry of PM sidebands with an SBS anti-Stokes response at a PD generates switchable EIA and EIT analogue frequency responses by tuning the wavelength of a pump laser.

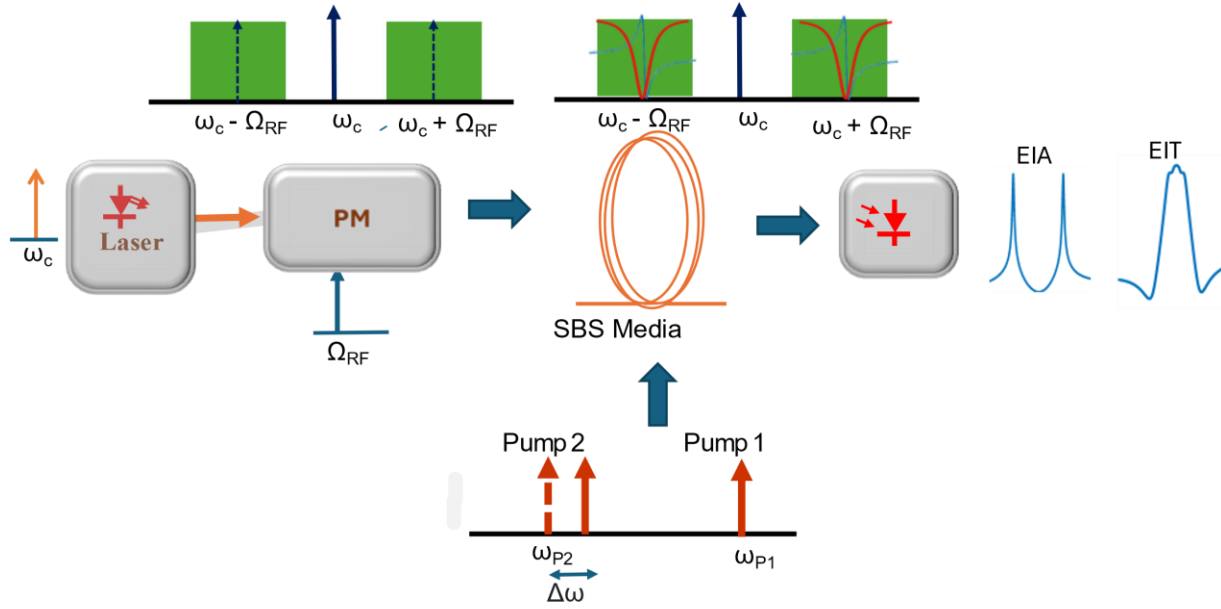


Fig. 4.2: Schematic of EIA and EIT analogue filter response generation using an analog photonic link with a phase modulator and two pump lasers to generate SBS anti-Stokes response. PM: phase modulator, PC: polarization coupler, EIA: electromagnetically induced absorption, EIT: electromagnetically induced transparency.

The proposed concept of a switchable analogue of EIA and EIT filters in the microwave domain using two SBS anti-Stokes responses has a probe signal of frequency ω_c is fed to a PM to generate π out-of-phase sidebands. When an RF signal of frequency Ω_{RF} is swept in a particular frequency band and is applied to a PM driven by a probe laser of carrier frequency ω_c , it generates two sidebands at $\omega_c \pm \Omega_{RF}$. Direct detection of this PM signal results in no RF signal at PD due to the cancellation of out-of-phase sidebands. The pump signal with the frequency ω_P generated by a tunable pump laser counter-propagates in the fiber with the probe laser. The resulting interaction between the probe, the pump, and an acoustic grating results in the nonlinear process known as

stimulated Brillouin scattering, which results in the generation of Stokes waves and anti-Stokes waves at $\omega_p - \Omega_b$, and $\omega_p + \Omega_b$, respectively, where, Ω_b is the Brillouin shift. The Stoke's and anti-Stoke's responses are also referred to as gain and loss responses, respectively, and they have associated phase responses. The slope of the SBS phase response is a function of the pump power level; the slope increases with an increase in pump power and decreases with a decrease in pump power.

4.2.1 Electromagnetically induced absorption analogue filter

The EIA filter frequency response has been generated using two pump lasers to produce an SBS anti-Stokes response on the LSB and a SBS anti-Stokes response on the USB of a phase-modulated link. The pump1 laser generates an SBS loss response on the upper sideband at frequency $(\omega_c + \Omega_{RF})$, and pump2 laser generates an SBS loss response on the lower sideband at frequency $(\omega_c - \Omega_{RF} + \Delta\omega)$. EIA filter frequency response has been generated by adjusting the pump2 laser frequency, which imparts a loss response on the LSB of the PM signal. The SBS anti-Stoke response on LSB is shifted toward the right, towards the probe frequency ω_c by $\Delta\omega$ as shown in Fig 4.3(a). The two sidebands combine at the PD with additional phase shifts provided by dispersion in the fiber. The combined phase shift results are close to or equal to $\pm 180^\circ$ at the center frequency, bringing both sidebands almost entirely out of phase. This results in enhanced cancellation of both sidebands at the photodetector, hence a null response in the center. On both sides of the center frequency, the phase response deviates from 180° , with the sidebands tending towards being in-phase, resulting in an increase in amplitude on both sides. Then, the combined phase difference increased on both sides, resulting in a rejection frequency band between the two transmission frequency bands, which produces an EIA analogue filter response.

The bandwidth of the EIA analogue filter increases with the frequency shift $\Delta\omega$, as shown in

Fig. 4.3 (b), representing a significant improvement over traditional Brillouin bandwidth tailoring, which requires additional pump power to increase bandwidths. Furthermore, with an increase in pump power, the slope increases, resulting in a larger spectral region with a 180° phase difference. This leads to more cancellation bandwidths and hence more EIA analogue absorption bandwidth.

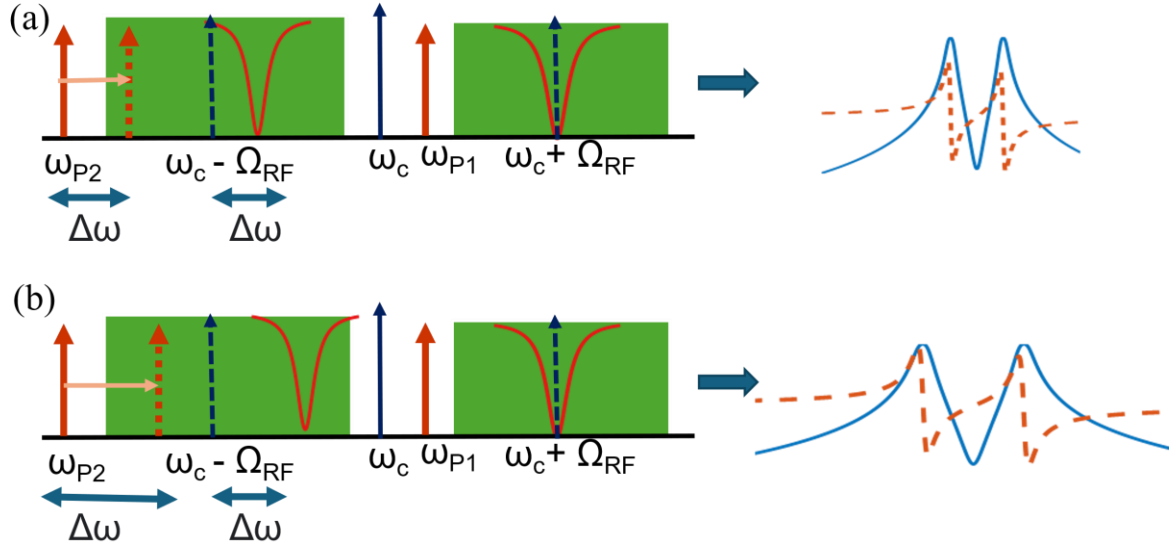


Fig. 4.3: (a) EIA analogue filter response generation by tuning pump2 laser frequency away from carrier frequency, (b) EIA analogue filter bandwidth tailoring by tuning the frequency shift of pump2 laser.

The concept diagrams for the EIA filter generation using two SBS loss responses are shown in Fig. 4.4. Two SBS loss responses on each sideband, with a frequency shift in the pump2 laser frequency on the LSB, are shown in Fig. 4.4(a). When these sidebands combine at the PD, the phase response of one sideband reverses, as shown in Fig. 4.4(b). The dispersion introduced by the fiber enhances the phase response, resulting in a combined phase and amplitude response, as shown in Fig. 4.4(c), which manifests an EIA analogue filter response.

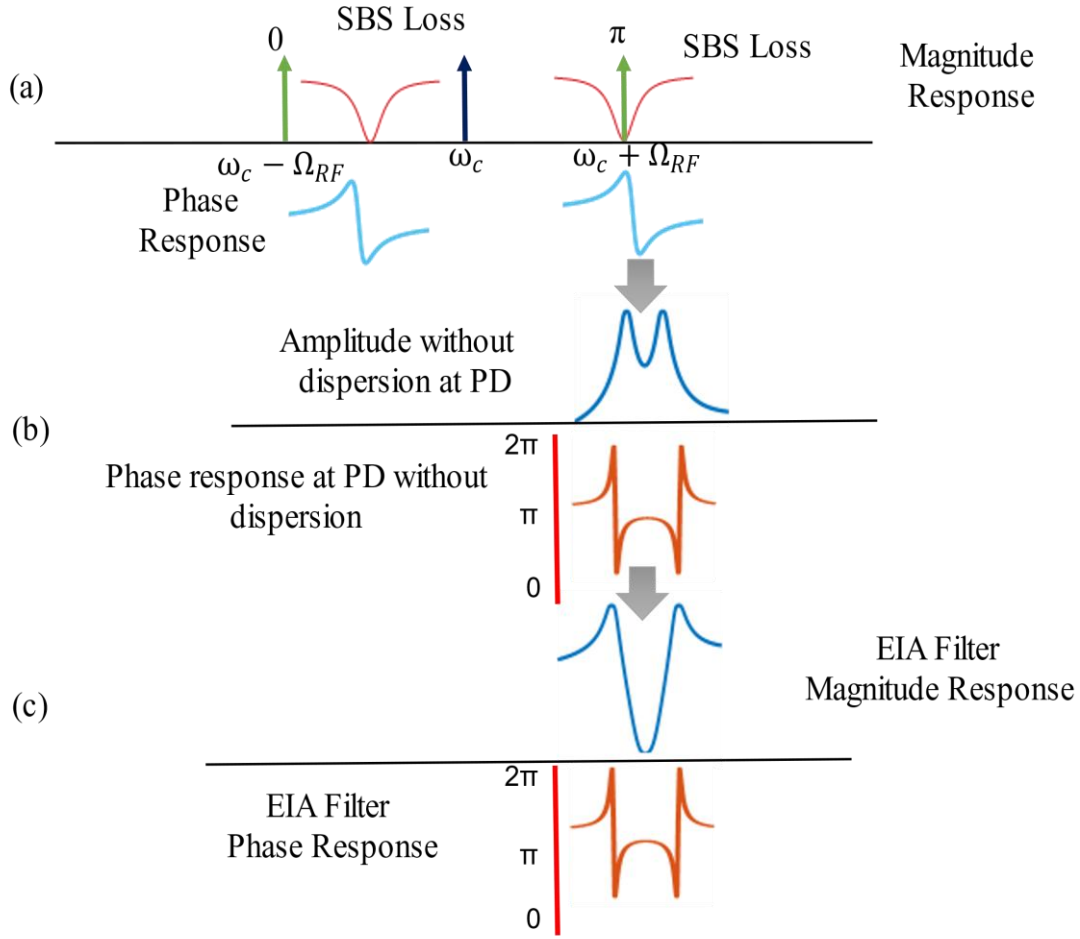


Fig. 4.4: Concept explanation of EIA filter (a) Amplitude and phase response of SBS loss response on each sideband of PM with anti-Stokes response at LSB shifted towards carrier (b) Amplitude and phase response at PD without dispersion effect (c) Filter magnitude and phase response with RF interference between two SBS anti-Stokes responses and dispersion

An Improved EIA analogue filter with sharp gain responses, tunable rejection ratio, and bandwidth is also demonstrated using two SBS gain responses as opposed to loss responses. Here, two pump laser frequencies have been tuned to have two gain responses, one each at the lower sideband (LSB) and upper sideband (USB) of the PM link. The pump2 laser is shifted away from the carrier frequency by a frequency shift $\Delta\omega$, as shown in Fig. 4.5(a). The bandwidth is tuned by tuning the frequency shift $\Delta\omega$ away from the carrier frequency, as shown in Fig. 4.5(b). The rejection ratio is

primarily a function of phase interferometry, resulting in high power efficiency. A rejection ratio of up to 63dB is achieved with a bandwidth variation of more than 1GHz.

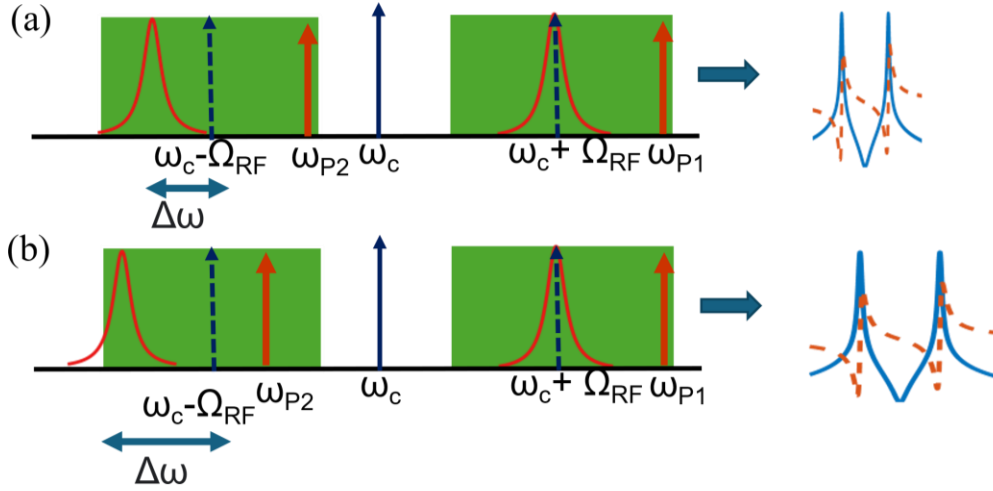


Fig. 4.5: (a) EIA analogue filter response with two SBS gain responses, one each on LSB and USB of the phase modulated link, (b) The EIA analogue bandwidth tailoring by tuning the pump2 laser frequency.

The EIA analogue filter using two SBS gain responses harnesses the same method as using two SBS loss responses. The frequency of pump1 laser is tuned to have SBS gain response at $(\omega_c + \Omega_{RF})$, and the pump2 laser frequency is tuned to have SBS gain response at $(\omega_c - \Omega_{RF})$. The pump2 laser is further shifted by $\Delta\omega$ to have SBS gain response at $(\omega_c - \Omega_{RF} - \Delta\omega)$. The combined effect of phase interferometry, resulting from the combination of both PM sidebands with SBS gain responses and phase shift due to dispersion in the fiber at the PD, yields an EIA analogue filter response. Here, a near 180° phase difference in the center between the two sidebands results in enhanced cancellation of both sidebands in the center. An increase in the phase shift at the edges brings the sidebands into phase. This results in a passband response at the edges, manifesting the EIA analogue filter. The bandwidth of the filter is tuned by tuning the frequency shift in the pump2 laser. The EIA filter depth and bandwidth have also been configured by tuning the pump power levels. Here, the depth of the absorption window increases with an increase in the pump power

level due to an increase in the SBS gain amplitude. Since the slope of the SBS phase response also increases with an increase in pump power, the bandwidth of the EIA increases with increasing pump power levels. The increase in EIA filter rejection bandwidth due to variations in pump power level is low compared to the frequency shift.

4.2.2 Electromagnetically induced transparency analogue filter

An EIT analogue filter has a transparent frequency spectrum between two absorption spectra. The pump1 laser with frequency ω_{p1} is set to have a loss response on the USB at frequency $(\omega_c + \Omega_{RF})$, and the pump2 laser with frequency ω_{p2} is set to have a loss response on the LSB at frequency $(\omega_c - \Omega_{RF} - \Delta\omega)$ as shown in Fig. 4.6(a). The pump2 laser is shifted away from the carrier frequency to increase the bandwidth of the EIT analogue filter. This also results in switching from an EIA analogue to an EIT analogue filter response by tuning a pump laser towards or away from the carrier frequency. Here, shifting the SBS anti-Stoke response towards the carrier laser frequency ($\Delta\omega$ positive) generates EIA and, away from the carrier laser frequency ($\Delta\omega$ negative), generates EIT.

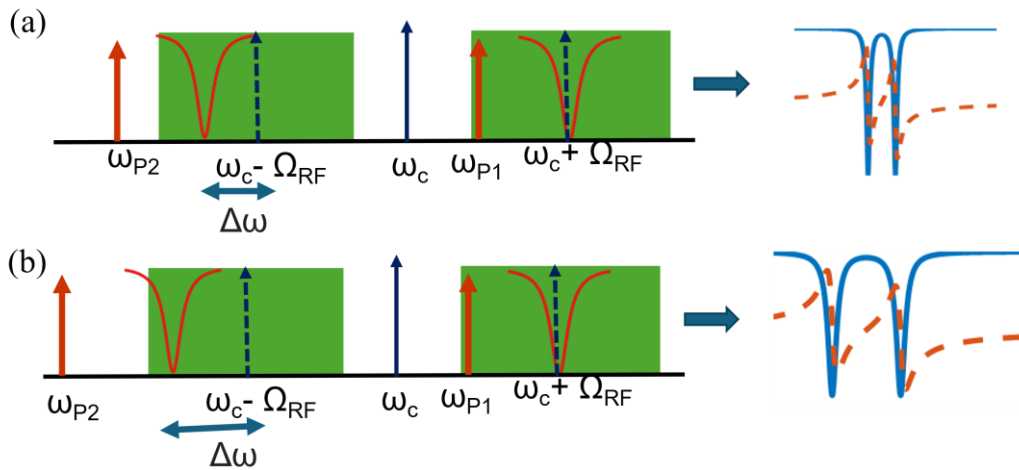


Fig. 4.6: (a) EIT analogue filter response using two SBS anti-Stoke's response, one each at LSB and USB of the phase-modulated link. (b) EIT analogue filter bandwidth tailoring by tuning the frequency shift of the pump laser away from the carrier frequency.

Here, the shifting of the pump2 laser frequency ω_{p2} results in a combined near-zero phase shift of both sidebands, resulting in maximum amplitude response at the center frequencies at the PD. The phase added by dispersion in the fiber enables the combined phase shift to approach a 0° phase shift. As we move away from the center frequency on both sides, the combined phase shifts move away from 0° , resulting in a decrease in amplitude response, followed by a combined phase shift crossing 0° , exhibiting an analogue of the EIT filter response. The amplitude and phase shift of the EIT analogue filter are shown in Fig. 4.6(a). The bandwidth of the EIT filter can be tuned by adjusting the frequency shift of the pump2 laser, i.e., increasing the frequency shift of the pump2 laser, $\Delta\omega$, as shown in Fig. 4.6(b). With an increase in $\Delta\omega$, the interference region between two SBS responses increases, increasing the transmission bandwidth. The pump power level is also used to alter the bandwidth by changing the slope of the SBS phase response.

The concept diagram for the EIT filter is shown in Fig. 4.7. The amplitude and phase responses of the SBS loss resonance with frequency shift in a phase-modulated link are shown in Fig. 4.7(a). At the PD, the SBS loss response of one sideband of PM reverses to the SBS gain response, as shown in Fig. 4.7(b), without the dispersion. The dispersion added by the fiber raises the phase responses above 180° bringing the two sidebands in phase, resulting in a transmission response at the center, followed by nulls on both sides. The final amplitude and phase response, which manifests as EIT, is shown in Fig. 4.7(c). The EIT filter response exhibits a small asymmetry in magnitude due to a slight variation in phase shift caused by dispersion over the frequency band. The switching between EIA and EIT occurs when the LSB anti-Stokes phase response precedes or follows the USB anti-Stokes phase response during the combination of LSB and USB of PM at PD.

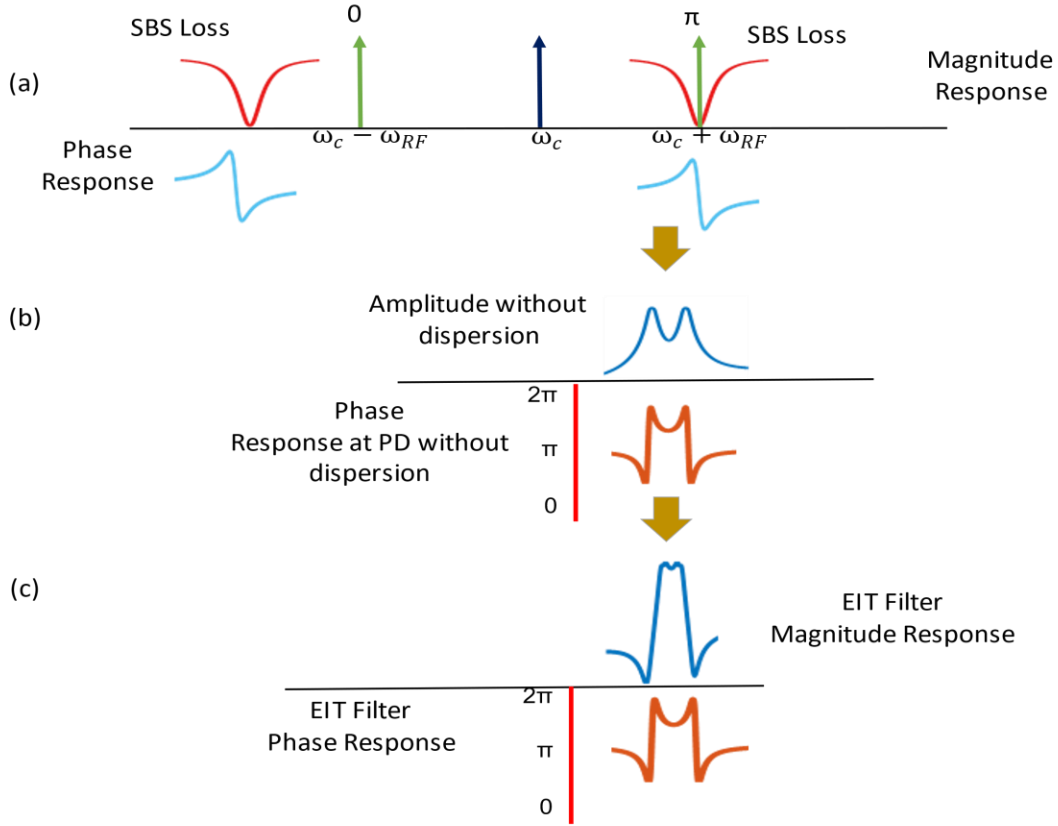


Fig. 4.7: Concept explanation of EIT filter (a) Amplitude and phase response of SBS loss response on each sideband of PM with anti-Stokes response at LSB shifted away from carrier (b) Amplitude and phase response at PD without dispersion effect (c) Filter magnitude and phase response with RF interference between two SBS anti-Stokes responses and dispersion of the fiber.

4.2.3 Spectra hole burning analogue filter response

SH analogue frequency response in the microwave domain was generated using the same setup as the EIA analogue filter response, with different pump power levels. The pump1 laser frequency ω_{p1} is tuned to have SBS loss response at $(\omega_c + \Omega_{RF})$, and the pump2 laser frequency ω_{p2} is tuned to have SBS loss response at $(\omega_c - \Omega_{RF})$. The pump2 laser is further shifted by $\Delta\omega$ to have SBS gain response at $(\omega_c - \Omega_{RF} + \Delta\omega)$ to generate an EIA analogue frequency response. The pump1 laser power level is tuned below the pump2 power level to generate SBS phase responses with different slopes. The frequency of the pump2 laser is shifted towards the carrier frequency by a frequency shift $\Delta\omega$ to create an interference region, resulting in a near-180° phase difference at the

center, which in turn produces absorption spectra. As we move away from the center, the phase shift increases, generating transmission spectra on both sides. The pump 1 power level is low compared to pump 2, resulting in a spectrum hole in the analogue filter response. The amplitude and phase responses of the SH burning filter responses are shown in Fig. 4.8(a). The SH analogue frequency response can be used to attenuate a particular frequency spectrum selectively.

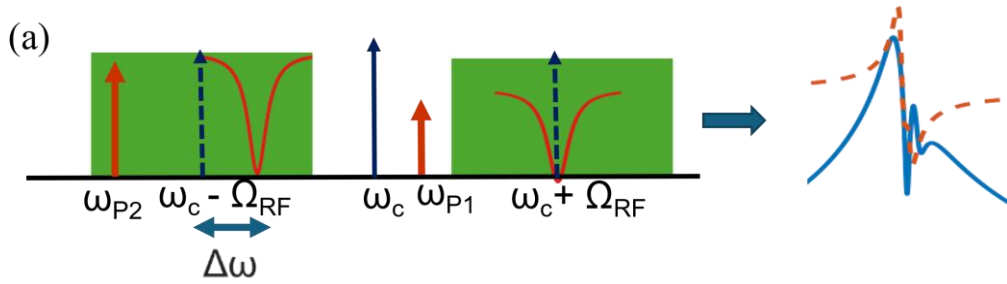


Fig. 4.8: SH-burning filter response with pump1 power level lower than pump2 power level

4.3 Theory

In this section, a mathematical model is derived to characterize RF phase interferometry for generating switchable EIA, EIT, and SH analogue filter responses using a phase modulator with two SBS anti-Stokes responses.

The numerical expression of stimulated Brillouin-induced loss and gain response is represented as [33],

$$G_{SBSl}(\omega) = \frac{g_B P_P L_{eff}}{A_{eff}} \frac{1}{\left(1 + 2j \left(\frac{\Delta\omega}{T_B}\right)\right)} \quad (4.1)$$

$$G_{SBSg}(\omega) = \frac{g_B P_P L_{eff}}{A_{eff}} \frac{1}{\left(1 - 2j \left(\frac{\Delta\omega}{\Gamma_B}\right)\right)} \quad (4.2)$$

Here g_B , is the Brillouin gain coefficient in the fiber. P_P , is the pump power, L_{eff} and A_{eff} , are the effective length and effective area. $\Delta\omega$ is the detuning of the frequency from the peak SBS gain and Γ_B is the Brillouin linewidth. Here, the real part of G_{SBSl} denotes the SBS loss amplitude response, and the imaginary part denotes the SBS loss phase response. SBS loss response at frequency shift $\Delta\omega$ is given as,

$$G_{SBS}(\omega) = G_{SBSl}(\omega + \Delta\omega) \quad (4.3)$$

The electric field at the output of the phase modulator modulates the carrier signal of frequency driven by an RF signal of frequency Ω_{RF} , and considering small signal analysis and ignoring higher order terms,

$$E_{PM} = E_0((J_0(m) \exp(j\omega_c t) - J_1(m) \exp(j(\omega_c - \Omega_{RF})t) + J_1(m) \exp(j(\omega_c + \Omega_{RF})t)) \quad (4.4)$$

Where, E_0 is the electric field of the probe laser signal, Ω_{RF} is the angular frequency of the RF signal input to the PM and $J_n(m)$ defines the n^{th} order Bessel function of the first kind, where m is the modulation index of the PM.

SBS loss response on the lower sideband and on the upper sideband can be expressed in the form of the transfer function as,

$$H_{LSB}(\omega) = \exp(G_{SBS}(\omega)) \quad (4.5)$$

$$H_{USB}(\omega) = \exp(G_{SBS}(\omega)) \quad (4.6)$$

The electric field signal with SBS loss response applied on the lower sideband and upper sideband with frequency shift $\Delta\omega$ applied on the lower sideband is represented as,

$$E_{out}(t) = E_0 \left(J_0(m) \exp(j\omega_c t) - J_1(m) \exp(j(\omega_c - \Omega_{RF})t) H_{LSB}(\omega_c - \Omega_{RF} \pm \Delta\omega) H_D + J_1(m) \exp(j(\omega_c + \Omega_{RF})t) H_{USB}(\omega_c + \Omega_{RF}) H_D \right) \quad (4.7)$$

Here, $H_{LSB}(\omega_c - \Omega_{RF} \pm \Delta\omega)$, $H_{USB}(\omega_c + \Omega_{RF})$ are the SBS anti-Stokes responses at the lower sideband and the upper sideband of the PM signal, respectively, and H_D , is the transfer function of dispersion in the fiber.

When the above electric field signal is detected at a photodetector with responsivity " R_0 ", the current detected at PD is,

$$I_{PD} = R_0(t) E_{out}(t) E_{out}(t)^* \quad (4.8)$$

Considering only the fundamental terms and rejecting the higher order terms, the current detected at PD can be represented as,

$$I(t) \propto 2 J_0(m) J_1(m) \text{Re} \{ H_{USB}(\omega_c + \Omega_{RF}) - H_{LSB}(\omega_c - \Omega_{RF} + \Delta\omega) \} H_D \quad (4.9)$$

The current detected at PD with an EIA analogue response is represented as,

$$I(t) \propto 2 J_0(m) J_1(m) \{ \exp(G_{SBS}(\omega_c + \Omega_{RF})) \exp(\emptyset_D) - \exp(G_{SBS}(\omega_c - \Omega_{RF} + \Delta\omega)) \exp(\emptyset_D) \} \quad (4.10)$$

$$I(t) \propto 2 J_0(m) J_1(m) \{ G_{ueia} \cos(\emptyset_{ueia}) - G_{leia} \cos(\emptyset_{leia}) \} \quad (4.11)$$

where G_{ueia} , $\emptyset_{ueia}$, G_{leia} , $\emptyset_{leia}$ are expressed as,

$$G_{ueia} = \exp \{ \text{Re}(G_{SBS}(\omega_c + \Omega_{RF})) \} \quad (4.12)$$

$$\emptyset_{ueia} = \exp \{ \text{Img}(G_{SBS}(\omega_c + \Omega_{RF})) + \emptyset_D \} \quad (4.13)$$

$$G_{leia} = \exp \{ \text{Re}(G_{SBS}(\omega_c - \Omega_{RF} \pm \Delta\omega)) \} \quad (4.14)$$

$$\emptyset_{leia} = \exp \{ \text{Img}(G_{SBS}(\omega_c - \Omega_{RF} \pm \Delta\omega)) + \emptyset_D \} \quad (4.15)$$

From Eq. (11), we can deduce that the EIA filter operates such that the combined phase due to anti-Stokes response at USB and LSB of PM, along with the phase added by dispersion in fiber, approaches 180° at the center point of the interference zone. This leads to the current $I(t)$ detected by PD being very low, resulting in an absorption response at the center frequency. Transmission response on both sides of the absorption response occurs at the VNA due to the combined phase moving away from 180° and then approach 180° on both sides of the center frequency, respectively.

The current detected at PD with an EIT analogue response is represented as,

$$I(t) \propto 2 J_0(m) J_1(m) \{ \exp(G_{SBS}(\omega_c + \Omega_{RF})) \exp(\emptyset_D) - \exp(G_{SBS}(\omega_c - \Omega_{RF} - \Delta\omega)) \exp(\emptyset_D) \} \quad (4.16)$$

$$I(t) \propto 2 J_0(m) J_1(m) \{ G_{ueit} \cos(\emptyset_{ueit}) - G_{leit} \cos(\emptyset_{leit}) \} \quad (4.17)$$

where, $G_{ueit}, \emptyset_{ueit}, G_{leit}, \emptyset_{leit}$ are expressed as, $G_{ueit} = \exp \{ \text{Re}(G_{SBS}(\omega_c + \Omega_{RF})) \}$, $\emptyset_{ueit} = \exp \{ \text{Img}(G_{SBS}(\omega_c + \Omega_{RF})) + \emptyset_D \}$, $G_{leit} = \exp \{ \text{Re}(G_{SBS}(\omega_c - \Omega_{RF} - \Delta\omega)) \}$, $\emptyset_{leit} = \exp \{ \text{Img}(G_{SBS}(\omega_c - \Omega_{RF} - \Delta\omega)) + \emptyset_D \}$

where, $\emptyset_D$ is the phase added by dispersion in the fiber and is represented as,

$$\emptyset_D = (\pi^2 \lambda^2 \left(\frac{\Omega_{RF}}{2\pi} \right)^2 DL) / c \quad (4.18)$$

Here, D is the dispersion in the fiber, and L is the length of the fiber.

From Eq. (18), when the pump2 laser frequency is shifted away from the carrier frequency with frequency shift ($-\Delta\omega$), the combined phase at the center of the interference zone due to anti-Stokes response at USB and the LSB, and the phase added by dispersion in the fiber in a PM link, is close to 0° . This results in the detection of a higher current and, consequently, a more pronounced transmission response around the center frequency. The combined phase comes closer to 180° , while moving away from the center frequency, this leads to absorption spectra on both sides of the transmission spectra, generating an EIT analogue filter response.

The mathematical model presented is simulated in MATLAB to verify the generation of the EIA analogue and EIT analogue filter response. The variation in bandwidth of the EIA and EIT filter is simulated over the frequency shifting of a pump laser frequency. The switching between EIA and EIT, or vice versa, over a frequency shift is also simulated and verified. Moreover, the spectra hole analogue filter response over tuning of a pump power level is also simulated. Furthermore, the generation of an EIA analogue filter response with a higher rejection ratio is verified using two SBS gain responses.

4.4 Simulation

Based on the mathematical model presented to achieve EIA, EIT, and SH analogue response, a simulation in MATLAB has been performed. As part of the simulation, a CW laser, a phase modulator, two tunable CW pump lasers, a 3.2 km fiber for generating SBS Stokes or anti-Stokes responses, and a PD are used to generate an EIA, EIT, and SH analogue filter response. The dispersion of 17 ps/nm/km is considered for a fiber of length 3.2km.

To simulate an EIA analogue filter response, an RF input is applied to a phase modulator to

generate π out-of-phase sidebands. Two pump lasers are tuned to generate SBS loss responses at the LSB and the USB of PM. One of the pump lasers with an SBS response on the LSB is shifted towards the carrier, with a frequency shift varying from 200 MHz to 800 MHz. The beating of PM sidebands with SBS loss response, along with the phase contribution from the fiber, generates an EIA analogue filter response, as shown in Fig. 4.9(a). The amplitude and phase responses of the EIA filter bandwidth variation over frequency shifts from 200 MHz to 800 MHz are shown in Fig. 4.9 (b) and 4.9 (c), respectively.

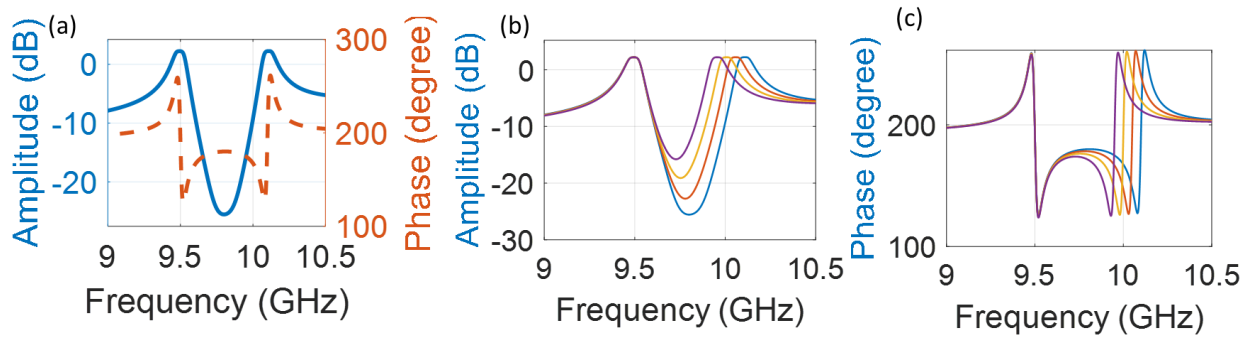


Fig. 4.9: (a) The amplitude and phase response of an EIA analogue filter; (b) Amplitude response of EIA filter bandwidth tailoring by tuning frequency shift of pump2 laser; (c) Phase response of EIA filter bandwidth tailoring by tuning frequency shift of pump2 laser.

To simulate an EIT analogue filter response, an RF input is applied to a phase modulator to generate π out-of-phase sidebands. Two pump lasers are tuned to generate SBS loss responses at the LSB and the USB of PM. The beating of PM sidebands at PD, accompanied by an SBS loss response and a phase contribution from the fiber, generates an EIT analogue filter response, as shown in Fig. 4.10(a). The amplitude and phase responses of the EIT filter, with bandwidth variation over frequency shifts from 200 MHz to 500 MHz, are shown in Fig. 4.10 (b) and 4.10 (c), respectively.

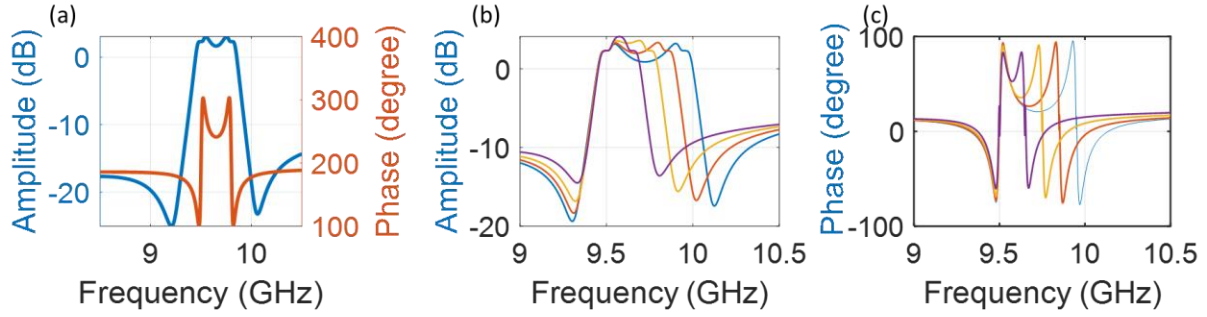


Fig. 4.10: (a) The amplitude and phase response of an EIT analogue filter response, (b) Amplitude response of EIA filter bandwidth tailoring by tuning frequency shift of pump2 laser from 200 MHz to 500 MHz, (c) Phase response of EIA filter bandwidth tailoring by tuning frequency shift of pump2 laser from 200 MHz to 500 MHz

To simulate a spectra hole analogue filter response, an RF input is applied to a phase modulator to generate π out-of-phase sidebands. Two pump lasers are tuned to generate SBS loss responses at the LSB and the USB of PM. The beating of PM sidebands with SBS loss response, along with the phase contribution from the fiber at the PD, generates a spectrum hole-burning analogue filter response, as shown in Fig. 4.11(a). The amplitude and phase response variations of the SH analogue filter over pump power variations from 5 dBm to 9 dBm are shown in Fig. 4.11(b) and Fig. 4.11(c), respectively.

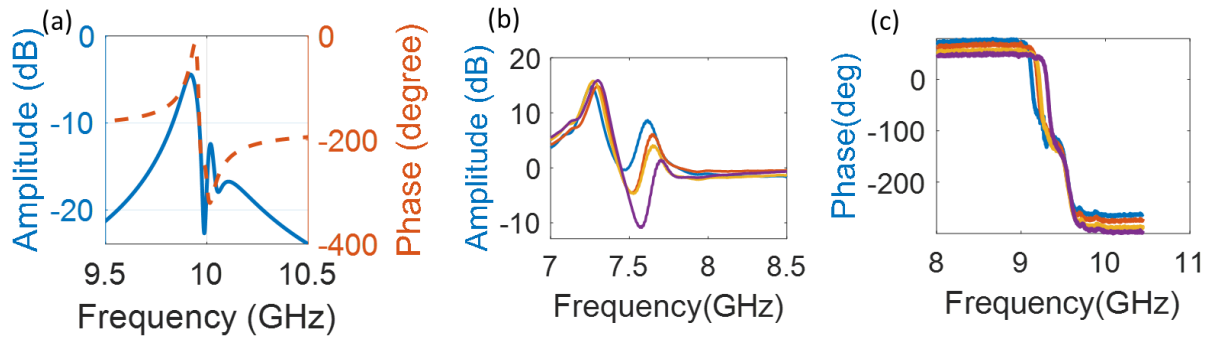


Fig. 4.11: (a) The amplitude and phase response of an EIT analogue filter, (b) Amplitude response of EIA filter bandwidth tailoring by tuning frequency shift of pump2, (c) Phase response of EIA filter bandwidth tailoring by tuning frequency shift of pump2 laser.

To simulate an EIA analogue filter response using two Stoke's responses, an RF input is applied to a phase modulator to generate π out-of-phase sidebands. Two pump lasers are tuned to generate a

SBS gain response with a frequency shift towards the carrier at the LSB and a SBS gain response at the USB of PM. The beating of PM sidebands with the SBS gain response, along with the phase contribution from the fiber at the PD, generates an EIA analogue filter response with a higher rejection ratio and a sharp gain response on the edges of the rejection window, as shown in Fig. 4.12(a). The amplitude and phase response variations of the EIA analogue filter over pump frequency shifts from 200 MHz to 800 MHz are shown in Fig. 4.12(b) and Fig. 4.12(c), respectively.

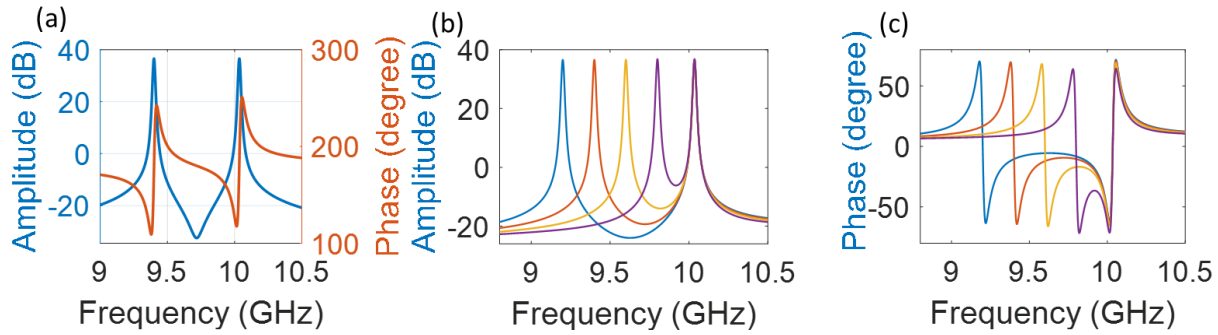


Fig. 4.12: (a) Amplitude and phase response of EIA filter response using two Stokes's. (b) EIA bandwidth variation over frequency shift from 200 MHz to 800 MHz, (c) Phase response of EIA over frequency shift from 200 MHz to 800 MHz.

4.5 Experimental Setup

The proposed EIA/EIT and SH analogue filters are achieved by the interaction of two SBS anti-Stokes responses applied in a phase-modulated analog photonic link. SBS anti-Stokes response is applied to both LSB and USB of the phase-modulated signal. The experimental setup for the

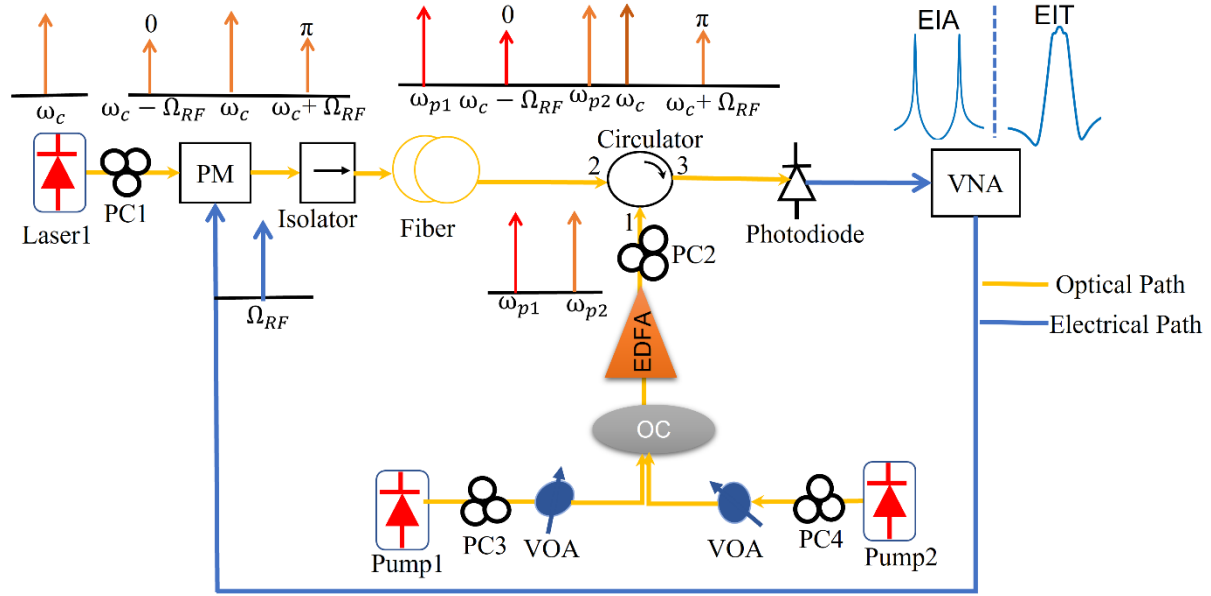


Fig. 4.13: Experimental setup for the realization of a switchable EIA and EIT filter. PC: Polarization controller; PM: Phase modulator; EDFA: Erbium doped fibre amplifier; VOA: variable optical attenuator; OC: Optical Combiner; VNA: Vector network analyser; SMF: single-mode fibre.

the generation of EIA, EIT, and SH analogue filter response is shown in Fig. 4.13. The probe signal generated by a laser with a frequency ω_c is modulated by a phase modulator to generate two 180° out-of-phase sidebands at the VNA. The phase-modulated signal is passed through an isolator and fed into an optical fiber of 3.2km length. The PM signal output after the SBS fiber is connected to Port 2 of the circulator.

Two tunable pump lasers were used in the pump arm to achieve the RF phase interferometry, followed by an isolator and a polarisation controller to generate two SBS anti-Stoke's resonances. The output of the polarization controller in the pump arm is fed to a variable optical attenuator, which controls the output power level of individual pump lasers. An optical coupler combines the pump laser signal, which is then amplified by an Erbium-doped fiber amplifier (EDFA). The gain of the EDFA is controlled to vary the combined pump power of both pump lasers. The combined pump signals from the EDFA are connected to the polarisation controller PC2, followed by the

circulator port 1. The interaction of the counter-propagating pump laser signals and the phase-modulated signal in the SBS fiber processes the signal exiting the circulator at Port 3. The filtered RF signal is detected by a photodetector, followed by a vector network analyzer to visualize the gain and phase response of the RF signal.

To demonstrate the EIA analogue frequency response, the pump frequency is tuned to $\omega_c + \Omega_{RF} - \Omega_B$ to have an SBS anti-Stoke response at the frequency $\omega_c + \Omega_{RF}$, in the upper sideband of the PM signal. The SBS frequency shift Ω_B of the fiber is 10.8 GHz. Pump2 laser frequency is tuned to $\omega_c - \Omega_{RF} - \Omega_B + \Delta\omega$ to have an SBS absorption response at the frequency $\omega_c - \Omega_{RF} + \Delta\omega$ in the lower sideband of the PM signal.

The switching from EIA to EIT and vice versa is achieved by having $\Delta\omega$ as positive or negative. The bandwidth variation is achieved by tuning the value of $\Delta\omega$. When the $\Delta\omega$ is positive, the EIA filter is detected at the VNA, and with an increase in $\Delta\omega$, the bandwidth of the EIA filter increases. When the $\Delta\omega$ is negative, the EIT analogue filter is detected at the VNA, and the bandwidth increases with the increase in the negative $\Delta\omega$ value. This separation between the center frequency of the PM signal and the SBS loss resonance frequency determines the bandwidth of the EIA/EIT filter. The normal SBS loss resonance has a bandwidth of ~30 MHz; however, the interaction of the two SBS loss resonances, one on each upper and lower sideband, results in an EIA/EIT filter with a tunable bandwidth of more than 1 GHz.

4.6 Results and Discussions

The switching between EIA and EIT analogue frequency response is achieved by shifting one of the pump lasers in a PM link with two SBS loss responses. Here, the probe laser ω_c is set at 194.1212 THz, and two counter-propagating pumps, pump1 is set at $\omega_c + \Omega_{RF} - \Omega_B$, and pump2

at $\omega_c - \Omega_{RF} - \Omega_B + \Delta\omega$ with Ω_{RF} , set at 9.5 GHz. The propagation fiber, with a length of 3.2km, is used, featuring an attenuation of 0.2dB/km and a linewidth of ~40 MHz for SBS. The frequency shift $\Delta\omega$ in the pump2 laser was set to -200 MHz to create an EIT filter with a 3dB bandwidth of 84 MHz, as shown in Fig. 4.14(a). The EIA analogue frequency response is achieved by setting the frequency shift $\Delta\omega$ to 380MHz. When these modified SBS loss responses are combined at the PD, an EIA filter response with a 3 dB bandwidth of 500 MHz is detected as shown in Fig. 4.14(b). The absorption bandwidth of the EIA filter is tuned by tuning the frequency shift $\Delta\omega$. The bandwidth increases as the frequency shift increases. The phase modulator link has a 180° phase difference between LSB and USB. The SBS loss response is applied to both LSB and USB of the PM link, with a frequency shift applied to the LSB pump laser to shift the SBS loss response towards the carrier.

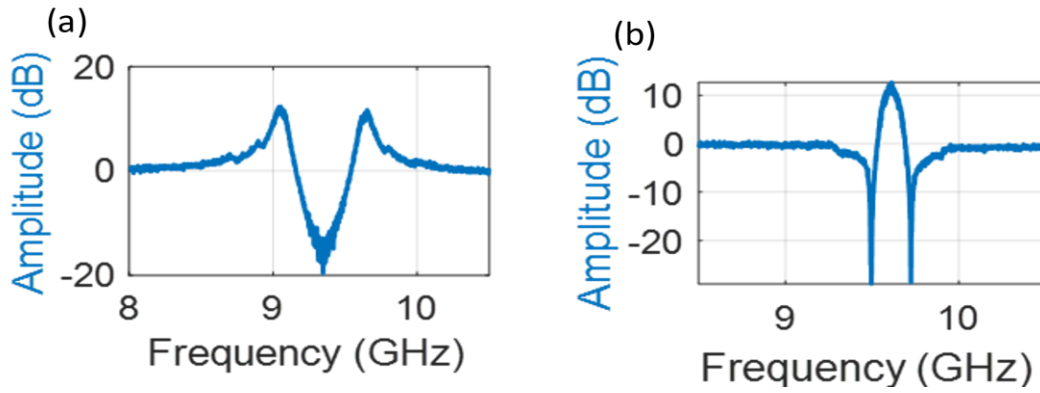


Fig. 4.14: Switching from EIT to EIA filter response by shifting the pump laser 2 frequency by 300MHz. (a) EIA filter; and (b) EIT filter

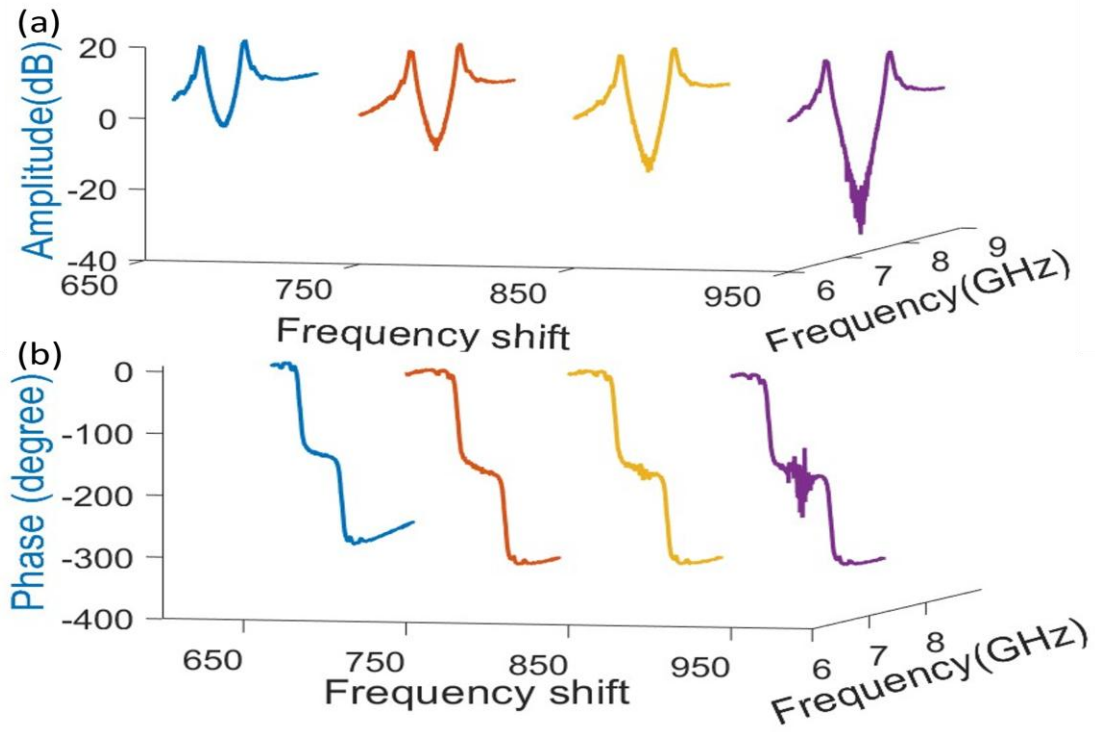


Fig. 4.15: (a) EIA filter over frequency shifts of pump2 laser from 650MHz to 950MHz (b) Phase response of EIA filter variation over frequency shifts of pump2 laser from 650MHz to 950MHz.

The combination of LSB and USB with SBS gain and phase response generates an interference region with a 180° phase shift around the center, which moves towards 0° on both sides. The absorption spectra due to a 180° phase shift around the center increase with increasing frequency shift, resulting in a larger frequency band with a phase shift of close to 180° around the center, which increases the EIA bandwidth. The EIA filter bandwidth variation, from 700 MHz to 1 GHz, over a frequency shift $\Delta\omega$ from 650 MHz to 950 MHz, is shown in Fig. 4.15(a), with the pump

power levels kept constant. The corresponding phase response is shown in Fig.4.15(b).

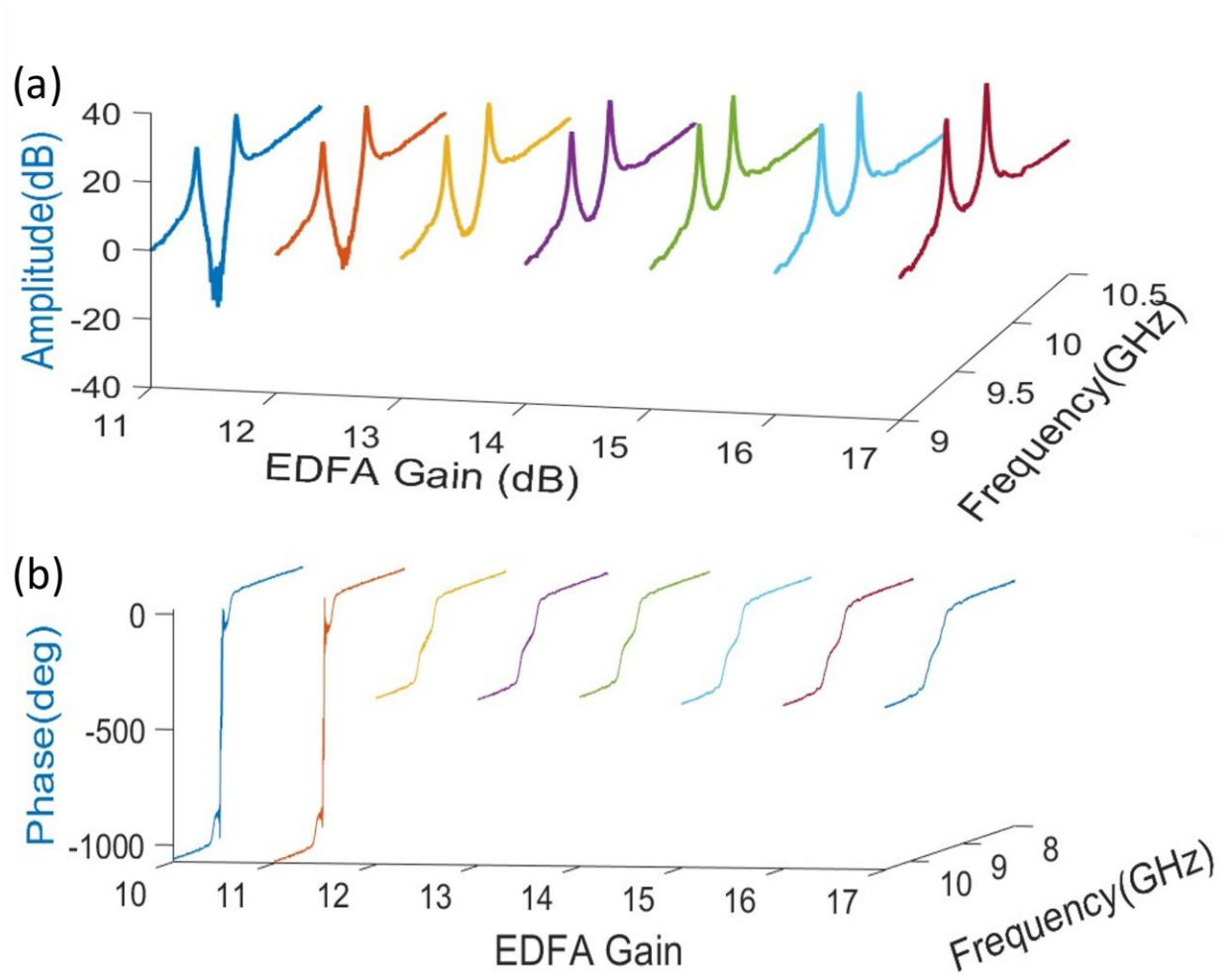


Fig. 4.16(a) EIA filter bandwidth and rejection ratio variation over different pump power levels by varying the EDFA gain. (b) Phase response of the EIA filter for various pump power levels.

The rejection ratio of the EIA filter also increases with an increase in frequency shift due to enhanced cancellation around the center frequency. With a further increase in $\Delta\omega$, a further increase in absorption bandwidth can be achieved with a slight decrease in the rejection ratio around the center frequency due to a decrease in the phase shift's contribution from two anti-Stokes responses at the center. As we move the frequency shift to 950 MHz, the rejection bandwidth of the EIA filter increases. The depth of the absorption window in the EIA filter also increases with

an increase in the frequency shift. Here, the depth of EIA increases from 24 dB to 50 dB. We can see that the phase is close to 180° within the absorption window, as shown in Fig. 4.16(b), resulting in enhanced cancellation of the sidebands and an increase in the depth of absorption. We can maintain the center frequency by increasing the frequency shift in both pumps, with a positive frequency shift for both the upper sideband and the lower sideband. The depth of the EIA filter also increases with a decrease in pump powers as shown in Fig. 4.16(a). The depth of EIA increases from 13 dB to 40 dB for both pump power levels, ranging from 15 dBm to 7 dBm.

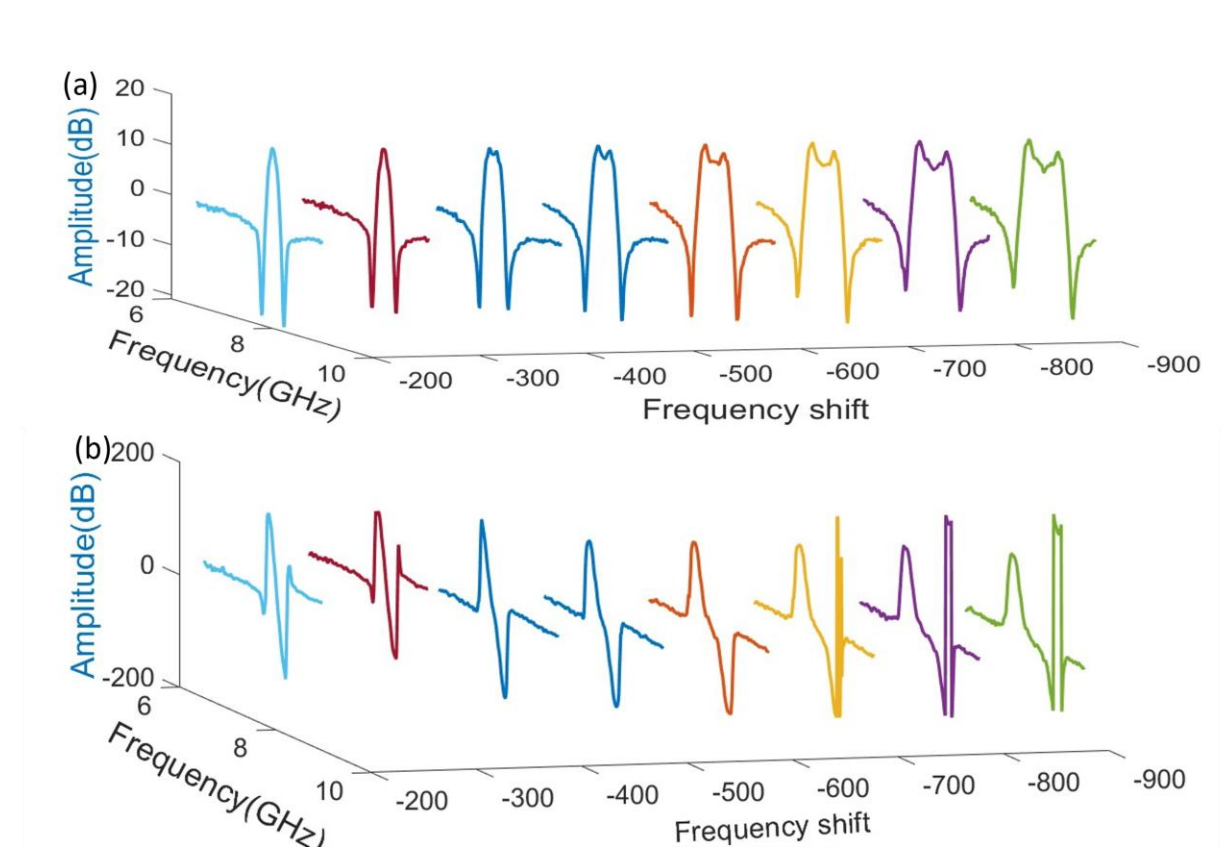


Fig. 4.17: EIT filter bandwidth variation over frequency shift in pump2 laser from -200 MHz to -900 MHz (b) Phase response of EIT filter over frequency shift from -200 MHz to -900 MHz.

The transmission bandwidth of the EIT filter can also be varied by modifying the frequency shift $\Delta\omega$ of the pump2 laser. Here frequency shift $\Delta\omega$ is negative, which means SBS loss response moves away from carrier frequency. As we increase the frequency shift, the transmission bandwidth of

the EIT filter increases due to larger interference zone with close to 0° phase shift. The transmission bandwidth variation from 80 MHz to 1.1 GHz is achieved over a frequency shift from 200 MHz to 900 MHz of the pump₂ laser frequency, as shown in Fig. 4.17(a). The phase response of the EIT filter over the frequency shift variation is shown in Fig. 4.17(b). With a further increase in $\Delta\omega$, a further increase in transmission bandwidth can be achieved with a slight decrease in the rejection ratio around the center frequency due to a decrease in the phase shift's contribution from two anti-Stokes responses at the centre.

The transmission bandwidth of the EIT filter also varies with tuning the pump power, due to a change in the slope of the SBS loss response. Since pump power determines the slope of phase shift, which in turn determines the slope of the SBS loss response. The slope of the SBS loss response increases as the pump power increases, leading to a larger interference region around the center with a phase shift approaching 0° to achieve a higher transmission bandwidth. Since the amplitude of SBS increases with the pump power level, the EIT depth also increases with the increase in the pump power level. Variation of the EIT bandwidth from 84 MHz to 130 MHz and depth from 26dB to 35dB over pump power levels from 8 dBm to 14 dBm is shown in Fig. 4.18(a). The phase response of the EIT filter over pump power variation is shown in Fig. 4.18(b).

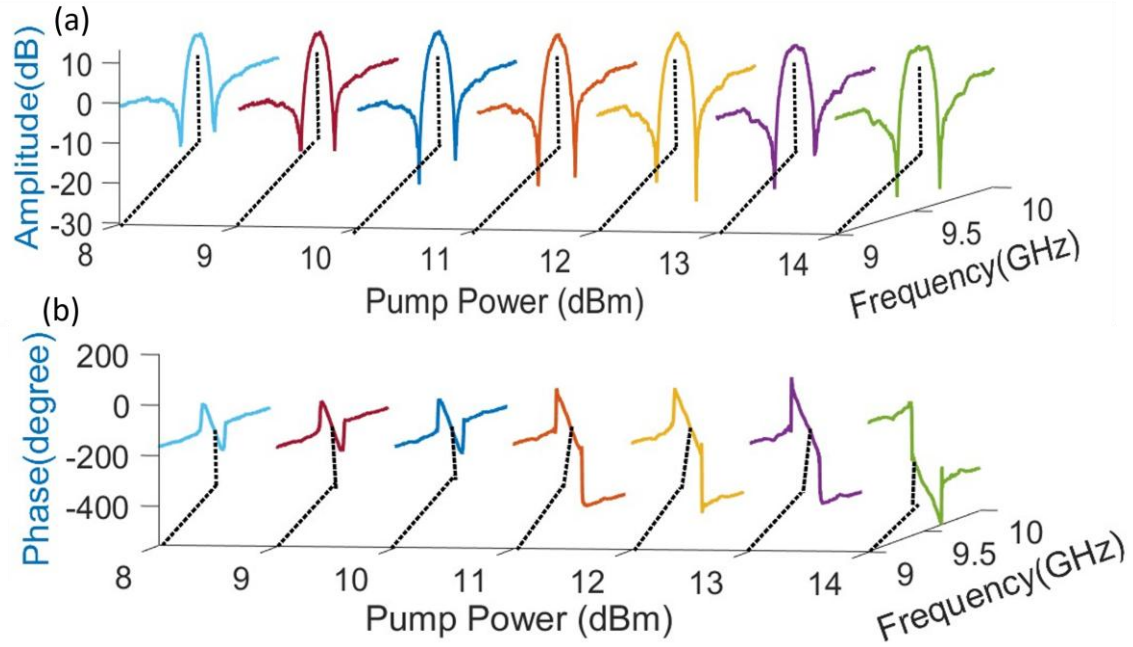


Fig. 4.18: (a) EIT bandwidth and depth variation over pump power variation from 8 dBm to 14 dBm, (d) Phase response of EIT filter over pump power variation.

EIA and EIT analogue filters are also demonstrated at different frequencies by tuning the pump laser frequencies and RF input to the phase modulator. The frequency of the pump1 laser frequency

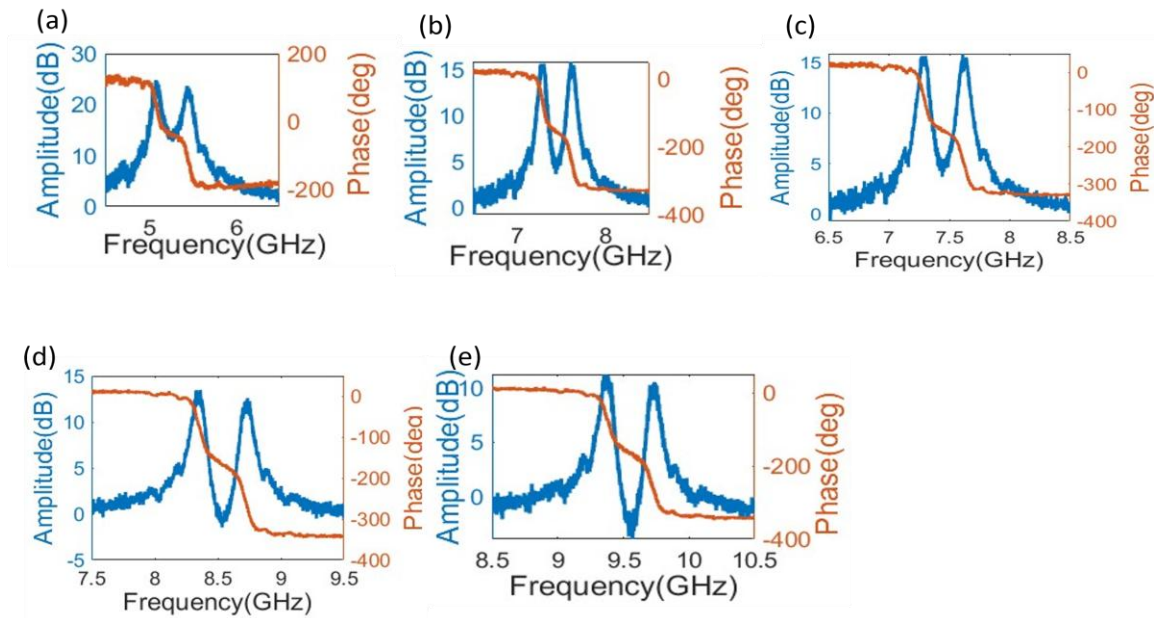


Fig. 4.19: EIA filter amplitude and phase response variation over frequency ranges from 5.5GHz to 10GHz.

is tuned to have SBS loss responses at $\omega_c + \Omega_{RF}$, and the pump2 frequency is tuned to place the SBS loss response at $\omega_c - \Omega_{RF} + \Delta\omega$ on the upper and lower sidebands of the PM-modulated signal, respectively. This generates the EIA/EIT responses with a central frequency of Ω_{RF} . The frequency shift $\Delta\omega$ is tuned to be either positive or negative, allowing for a transition from the EIT analogue to the EIA analogue and vice versa. EIA analogue filter response from 5 GHz to 10 GHz is shown in Fig. 4.19 (a) to Fig. 4.19 (e). The EIT analogue filter response from 3 GHz to 10 GHz is shown in Figs. 4.20(a) to 4.20(e).

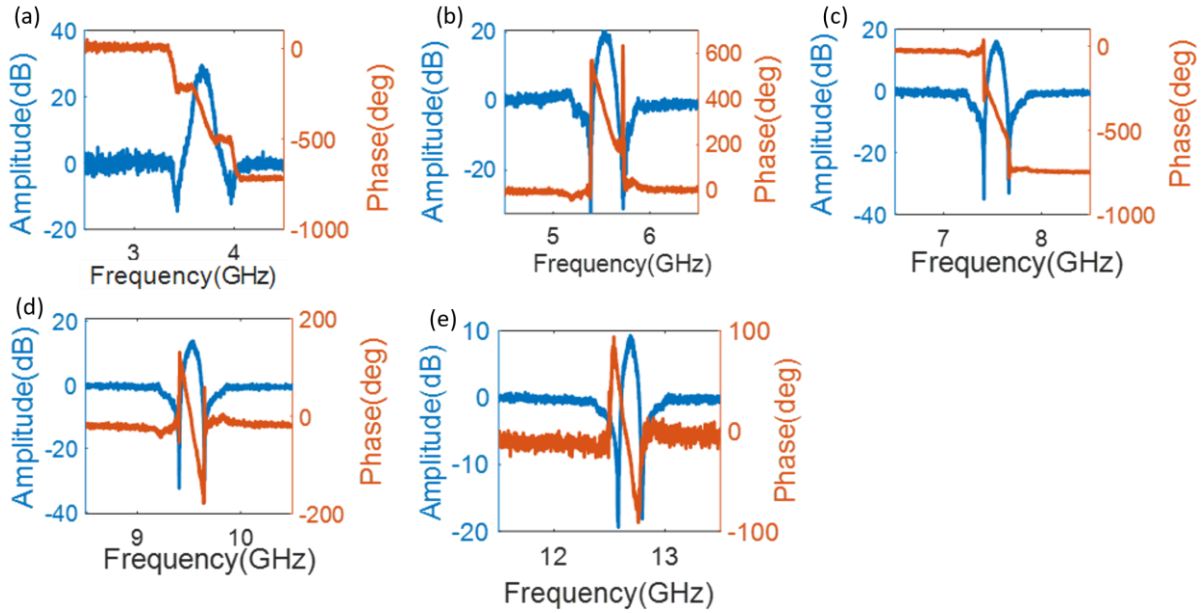


Fig. 4.20: EIT filter normalized amplitude and phase response variation over different frequency ranges from 3 GHz to 12 GHz.

EIA analogue filters with a high rejection ratio have been achieved using two SBS Stoke's responses. The two pump lasers, pump1 and pump2, have their frequencies tuned to have SBS Stokes response on each of the LSB and USB of the PM signal. The SBS gain responses are applied at $\omega_c + \Omega_{RF} + \Omega_B$ on USB and $\omega_c - \Omega_{RF} + \Omega_B + \Delta\omega$ on LSB. The RF frequency is swept using a VNA from 9 GHz to 10.5 GHz. Detection of this PM-modulated signal with the SBS gain resonances results in an EIA filter with a bandwidth tunable by varying $\Delta\omega$ of the pump2 laser.

The additional frequency shift in the pump2 laser has been varied from -300 MHz to -700 MHz. The EIA filter with a 3 dB bandwidth from 300 MHz to 650 MHz has been generated, as shown in Fig. 4.21(a) (amplitude response) and Fig. 4.21(b) (phase response). With an increase in the frequency shift, the depth of the EIA also increases as the phase response gets closer to 180° resulting in better cancellation of the two sidebands. Depth of EIA varies from 28 dB to 63 dB. The edges of the EIA response are sharper compared to the EIA using two loss responses, due to the SBS gain response having a narrower bandwidth than the SBS loss response.

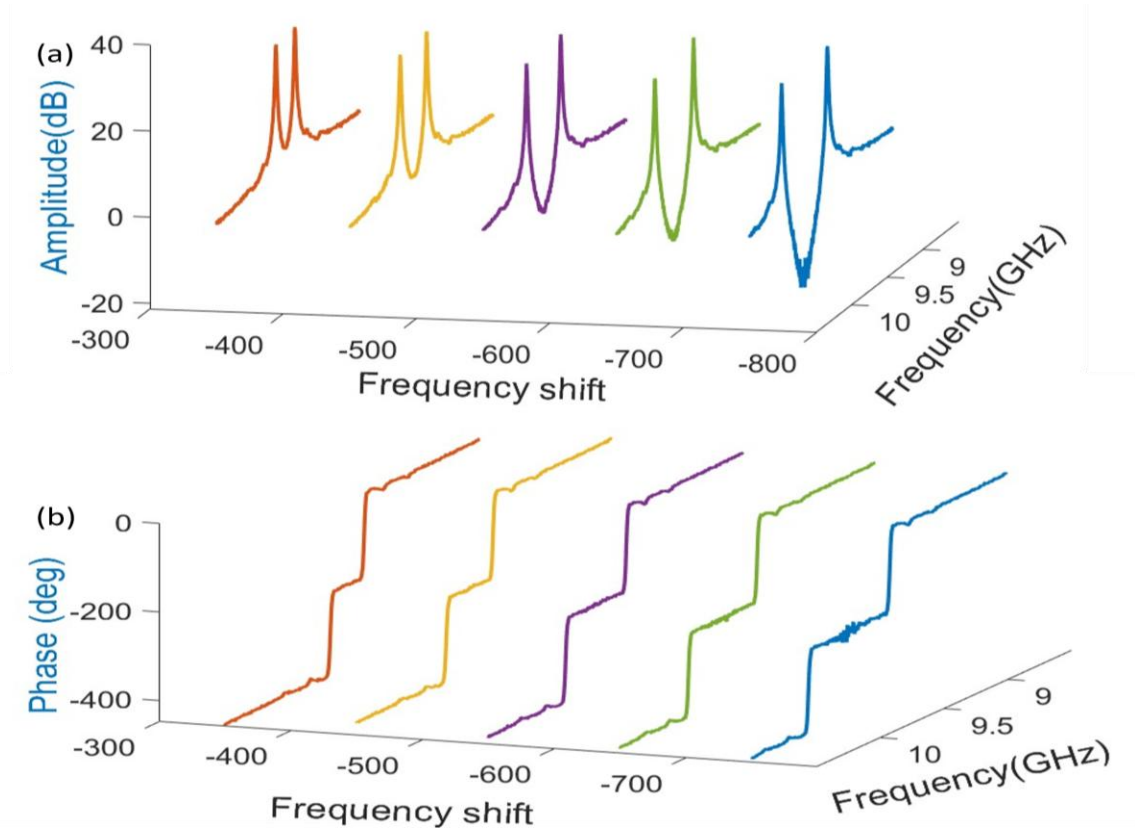


Fig. 4.21: (a) EIA filter bandwidth variation using two SBS stoke responses with frequency shift variation from -300 MHz to -700 MHz in pump2 laser. (b) phase response variation of EIA filter over frequency shift

The absorption bandwidth of an EIA analogue with two SBS gain responses can also be varied by varying the pump power level. The EDFA has been used to vary the pump power levels from 14 dBm to 7 dBm. The absorption spectra of EIA increase from 280 MHz to 340 MHz over a pump

power variation from 14 dBm to 7 dBm. The rise in bandwidth with a decrease in pump power signifies the high power efficiency of the filters for high-speed communication systems. The improvement in bandwidth with a decrease in pump power is due to a decrease in the slope of the SBS gain over the decrease in pump power, which increases the interference region with a close to π phase difference.

The rejection ratio of EIA filters with two SBS gain responses also increases with a decrease in pump power, making the system highly power efficient. The depth of the EIA filter increases from 28 dB to 47 dB. The increase in depth is due to enhanced cancellation of sidebands at low pump power. This EIA analogue filter's frequency tunability is achieved by tuning the RF frequency input of phase modulator. EIA filter frequency variation from 5 GHz to 15 GHz is shown in Fig. 4.22(a) to Fig. 4.22(e).

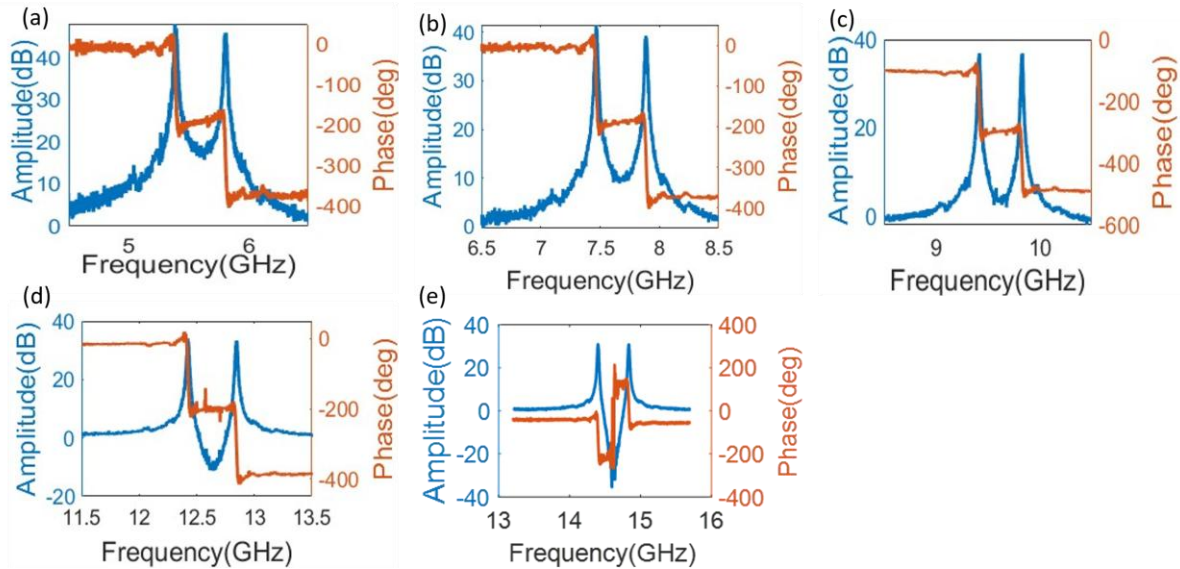


Fig. 4.22: EIA filter frequency variation from 5GHz to 15GHz using two Stokes response in a phase modulated analog photonic link

Spectra hole burning analogue filter responses have an absorption spectrum between two transmission spectra with different transmission gains. The different transmission gains are achieved by varying the pump power levels, resulting in SBS loss responses on LSB and USB with different rejection ratios. To demonstrate the SH burning spectral response, the probe laser ω_c is set at 194.1212 THz frequency, and the two pump lasers are set to have SBS loss response at $\omega_c - \Omega_{RF} - \Omega_B$ and at $\omega_c - \Omega_{RF} - \Omega_B + \Delta\omega$ for USB and LSB, respectively. The power of pump1 is set at -0.6 dBm and pump2 at 4 dBm to create a difference in the two pump power levels, as the pump power determines the amplitude of the RF signal detected at the PD. The EDFA gain on both pump lasers is set at 12dB. The amplitude and phase response of the SH burning response are shown in Fig. 4.23(a).

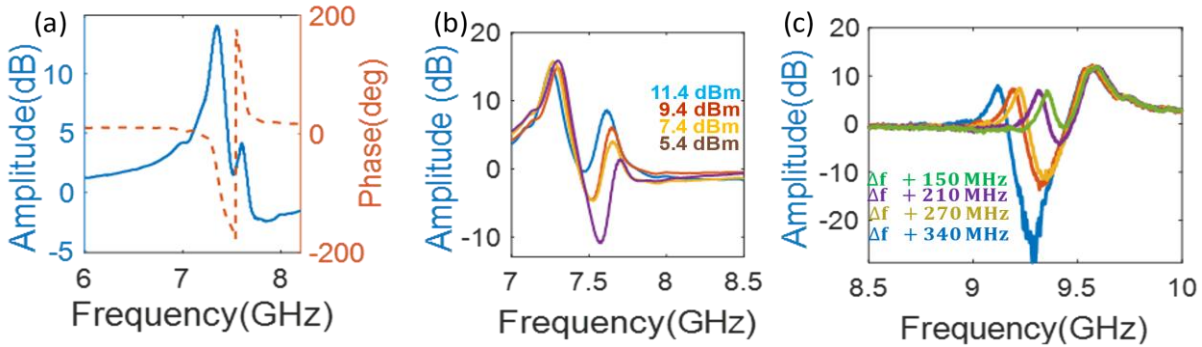


Fig. 4.23: (a) Spectra hole frequency response (b) variation of SH burning depth from 15 dB to 25 dB over pump2 power variation from 11.4 dBm to 5.4 dBm (c) SH burning hole depth variation over frequency shift in pump2 laser from 150 MHz to 340 MHz.

The transmission bandwidth of the SH filter can be changed by tuning the frequency shift of pump1 or pump2. The SH burning levels can be controlled by adjusting the pump power levels of either pump1 or pump 2. The burning hole depth over pump2 power variation, from 11.4 dBm to 5.4 dBm, is shown in Fig. 4.23(b). To demonstrate the variation in SH burning hole depth over different frequency shifts from 150 MHz to 340 MHz, and depth variation from 11dB to 36dB, as shown in Fig. 4.23(c).

The proposed filter system has demonstrated highly flexible EIA, EIT, and SH analogue filter responses in the microwave frequency band using stimulated Brillouin scattering. The tunability in bandwidth of the filter is achieved by tuning the frequency shift of a CW laser, making the system highly power-efficient with the capability of real-time tunability. The variation in rejection ratio and bandwidth is also demonstrated over a range of pump powers. Moreover, the spectra hole analogue filter response, achieved by varying the pump power level, is also demonstrated with tunability in bandwidth, rejection ratio, and frequency of operation. The flexibility in bandwidth, rejection ratio, and frequency of operation demonstrates the scalability of photonics for radar applications, enabling highly reconfigurable multiband operation. This demonstrates microwave photonics as a viable solution for generating filters for RF and microwave frequency applications, surpassing electronic-based methods.

4.7 Conclusion and Future Outlook

In this chapter, a novel technique for generating highly flexible electromagnetically induced absorption analogue and electronically induced transmission analogue filters in the microwave frequency band using a phase-modulated analog photonic link with two SBS anti-Stokes responses has been demonstrated. The tunability of absorption spectra for EIA and transmission spectral window for EIT is achieved by tuning the pump laser wavelength. EIA filters with a bandwidth of over 1 GHz and a rejection ratio of more than 50dB have been achieved. Spectra hole burning analogue frequency response with tunability in hole depth up to 36dB is also demonstrated using two SBS anti-Stokes responses. Moreover, an EIA analogue filter with an absorption depth of over 60dB, featuring two SBS Stokes responses, along with bandwidth tunability of greater than 1 GHz, is also demonstrated. The absorption spectra have a flat phase response, resulting in no delay. The switching from EIA to EIT frequency response is achieved just by tuning a pump laser, making

the filters highly reconfigurable. The proposed technique utilizes RF phase interferometry to achieve high bandwidth tunability, resulting in a highly power-efficient and reliable system. The filter frequency of operation has been tested up to 15 GHz, limited by the phase modulator bandwidth. The rejection ratio of the EIA and transmission gain of the EIT filter decrease over a rejection or transmission bandwidth greater than 1 GHz due to insufficient phase shift from the SBS around the center frequency. This can be further enhanced using higher pump power levels.

The proposed EIA, EIT, and SH analogue filters are highly useful for highly reconfigurable radars with multiband operation and configurable waveform bandwidth for detecting both long-range and small-size targets. The flexibility of EIA, EIT, and SH analogue filters can be utilized for applications with high control over the transmission/absorption frequency band to mitigate interference and jamming. This technique can also be used to generate band-pass to bandstop switching filters with a filter configuration variable over the pump laser wavelength for highly sensitive applications. The current progress in integrated photonic systems for generating on-chip lasers and modulators can be leveraged to achieve on-chip EIA/EIT and SH frequency responses for a highly reconfigurable on-chip photonic radar.

The radar system with high reconfigurability to operate from a low frequency band to a very high frequency band for the detection of long-range astronomical observations and high-speed low RCS flying aircraft. These radars require a remote antenna system to separate the complex signal processing systems and antenna systems using a low loss cable. The optical fibre with loss around $\sim 0.2\text{dB/Km}$ enables remote antenna system, however, requires a system to convert an optical signal to an RF signal for transmission. The phase modulator does not require DC bias, which makes it suitable for remote antenna locations, it requires a phase-to-intensity converter to get detected at the photodetector. The dispersive device, such as MZI, is used after the phase modulator

to get an RF signal at PD, however phase modulated link demodulated with MZI suffers from periodic nulls due to chromatic dispersion in a dispersive device, decreasing the operational bandwidth of the link. The null compensation in a phase modulated link demodulated with MZI is discussed in the next chapter.

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Chapter 5

Frequency-Response Tailoring of a Phase-Modulated Link using Stimulated Brillouin Scattering

The analog photonic link, utilizing optical phase modulators, is widely employed for high-speed communication links, photonic radar, and radio-over-fiber networks, facilitating signal transportation over optical transmission networks [1] [2]. The phase-modulated link does not require a DC bias and exhibits higher linearity compared to intensity-modulated links, thereby improving signal quality. However, the phase-modulated analog photonic link (APL) has a π phase difference between the two sidebands, resulting in a null frequency response with direct detection. The APL Link with a phase modulator requires a dispersive device, such as a Mach-Zehnder interferometer (MZI), with a path length difference between its two arms to convert the phase information into intensity, which is then detected at a photodetector. The chromatic dispersion due to path length difference in the MZI results in a dispersion-induced periodic transmission null frequency response, thereby decreasing the operational bandwidth of the APL. High operational bandwidth is essential for applications such as radars, where high-bandwidth chirps are transmitted to achieve high range resolution, and data communication links, where a high data rate is necessary to transport extensive amounts of data. SBS has been used for several microwave photonic

functionalities, including tunable filters, phase shifters, and delay lines.

In this chapter, we compensate for the chromatic dispersion-induced null frequency band of a phase-modulated APL demodulated by an MZI, utilizing an SBS gain response at the receiver to linearize the APL response at a fixed null frequency. The enhancement of the operational bandwidth of an APL link with PM+MZI by SBS gain response has been experimentally verified up to 23 GHz. Moreover, the system's capability to have multiple null frequency band compensation is also demonstrated. The null frequency compensation has also been verified over different MZI lengths to compensate for null frequency bands with different bandwidths.

The remainder of this chapter is organized as the introduction, and challenges are addressed in Section 5.1, followed by principles and theory in Sections 5.2 and 5.3. Section 5.4 explains the simulations, followed by the experimental setup in Section 5.5, with discussions & results in Section 5.6, and the conclusion and future scope in Section 5.7.

This chapter presents results from the journal and conference contributions listed below.

- M. Jha, R. Dhawan, R. Parihar and A. Choudhary, "Frequency Response Tailoring of a Phase Modulated Link With Stimulated Brillouin Scattering," in *IEEE Photonics Technology Letters*, vol. 34, no. 19, pp. 1054-1057, 1 Oct.1, 2022
- M. Jha, R. Parihar, R. Dhawan, and A. Choudhary, "Stimulated Brillouin Scattering based Null compensation in a Phase Modulated Link," in *Frontiers in Optics + Laser Science 2022 (FIO, LS)*.

This chapter includes results from a patent filed given below.

- Amol Choudhary, R. Dhawan, Reena Parihar, Mukund Jha, "Tunable Microwave Photonic Bandpass Filter using the Mach Zehnder Interferometer and Stimulated Brillouin Scattering," Indian Patent No. 202211010792, Feb2022.

5.1 Introduction

Microwave photonics is widely used in transporting high-speed data over long distances with high reliability [3]. The optical modulators convert the microwave signal into an optical frequency, which is sent through low-loss fiber and converted back to a microwave frequency at the receiver. The intensity modulator requires DC bias, which changes the linearity of the link in extreme environmental conditions. The optical phase modulator does not require any DC bias, significantly improving signal fidelity and reducing link complexity. However, the phase modulator has π out-of-phase sidebands, resulting in complete cancellation with direct detection at the photodetector. Various methods of phase modulator (PM) based analog photonic link (APL) demodulation have been demonstrated using dispersion-inducing devices [4], fiber Bragg gratings [5], ring resonators [6], dispersive fiber optic links [7], and stimulated Brillouin scattering (SBS) [8]. The dispersive device, such as MZI [9], has a path length difference between its two arms, which generates a frequency-dependent phase shift. The detection of a phase-modulated link with MZI results in periodic frequency nulls when demodulated at a photodetector. MZI uses the phase difference between its two branches to convert phase information to intensity information, resulting in an interference pattern with a periodic null response at the photodetector. The periodic null response results in a decrease in the operational bandwidth of the link and, hence, the overall data link capacity. The large operating bandwidth of the link is essential for radar systems, which require a wide bandwidth linear frequency-modulated signal to be transmitted in order to detect small, slow-moving targets.

Stimulated Brillouin scattering (SBS) [10] is widely used in microwave photonic signal processing due to its inherent characteristic of narrow bandwidth gain and loss resonance. The SBS gain and phase response can be easily tuned over a wide frequency range by counterpropagating the phase-

modulated carrier and a tunable pump laser. The narrow frequency response of SBS allows even small frequency variation, making the system highly flexible. While SBS has been used for several microwave photonic functionalities, including filtering [11]-[16], phase shifting [17], and delay lines [18], its potential has not yet been realized for link gain response engineering to increase the operational bandwidth of an PM+MZI link by selectively inserting gain responses on the nulls to tailor the frequency response of the APL.

The bandwidth of the communication link is vital for high-speed data links. The periodic null response in phase-modulated links with MZI decreases the available bandwidth, thereby limiting the link capacity, which is crucial for applications such as photonic radar. The periodic null response of the phase-modulated APL link can, in principle, be compensated by tailoring the frequency response of the APL link by selectively inserting SBS gain responses on the nulls. This method can also be utilized to selectively amplify the desired signal in the presence of a jammer. SBS gain based frequency response engineering enables recovering data from high noise by improving the signal-to-noise ratio of desired frequency signal.

This chapter discusses a method to enhance the operational bandwidth of an analog photonic link utilizing a phase modulator with a Mach-Zehnder interferometer. The stimulated Brillouin scattering gain response is used at the receiver to compensate for the nulls of the APL link, thereby increasing the operational bandwidth. Only the addition of a circulator and a tunable pump laser at the receiver can achieve this null compensation, leveraging the transmission fiber as the SBS gain medium without altering the transmission network. The center frequency of the null to be compensated is tuned by changing the pump laser frequency. The multiple null frequency band compensation is achieved by using multiple pump lasers to have gain responses at the desired null bands. This technique is useful for multiband radar operation, where the frequency of operation

needs to be tuned for long range or short range target detection. The SBS gain response can be utilized to compensate for the desired null frequency band or reject a frequency band by using the SBS loss response. The enhancement of the operational bandwidth of an APL link with PM+MZI by SBS has been experimentally verified up to 23 GHz with 16dB of null compensation. Furthermore, this technique can also be beneficial for rejecting image frequencies at the harmonics of the frequency of interest by selectively amplifying the fundamental frequency.

5.2 Principle of Operation

The proposed enhancement of operational bandwidth in a phase-modulated analog photonic link demodulated by a Mach-Zehnder interferometer (PM+MZI) is realized by imparting a stimulated Brillouin scattering (SBS) Stokes (gain) response at a selected null frequency band. The schematic of the PM+MZI link with null compensation by SBS is shown in Fig. 5.1. The optical carrier signal generated by a laser diode (LD) at a frequency ω_0 is injected into a phase modulator optical input. A swept RF signal is coupled to the RF port of the phase modulator using a vector network analyzer. The phase modulator generates 180° out-of-phase sidebands at its output. The phase-modulated signal is propagated through a transmission fiber. The received PM signal is injected into the Mach-Zehnder interferometer consisting of two lossless 50/50 couplers and a path difference of length L between the two arms. The path length difference results in a delay between the PM signals of the two arms of the MZI, resulting in an additional change in the phase shift of the PM signals. The additional phase shift enables the PM sidebands to come into phase, which is essential for the phase-to-intensity conversion required for detection at the photodetector. However, the phase imparted by the delay in the two arms of the MZI varies with frequency, resulting in a continuous change in phase shift between the two arms, which generates a frequency response with periodic nulls at PD. To compensate for these null frequency bands, a tunable pump

laser is counter-propagated in fiber using a circulator to generate an SBS Stokes response at a frequency shifted by the Brillouin shift from the tunable laser frequency. The tunable laser frequency is tuned to have an SBS gain response at the center frequency of the null frequency band, thereby enhancing the operational bandwidth of the APL link.

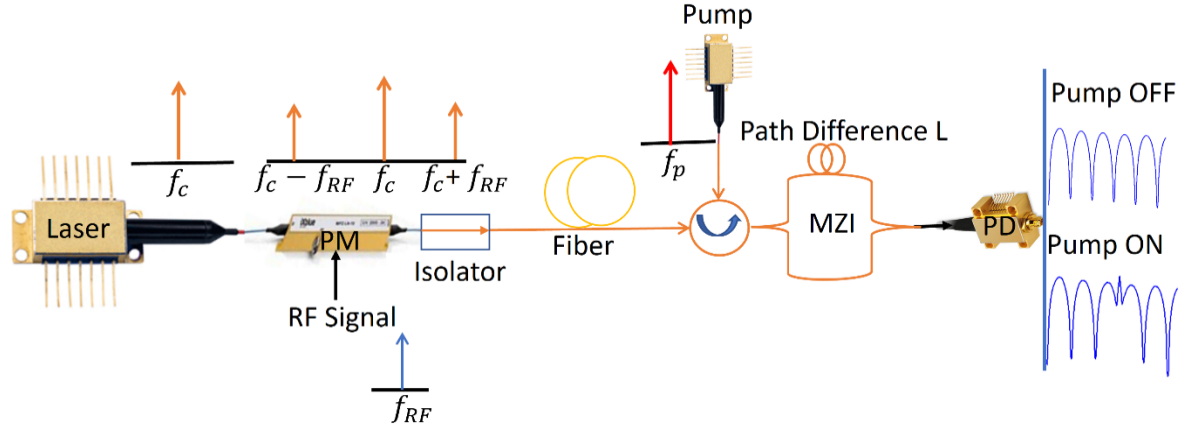


Fig. 5.1: The schematic of the proposed architecture of null compensation based on PM with SBS+MZI link. PM: phase modulator, RF: radiofrequency, MZI: Mach-Zehnder interferometer, PD: photodetector.

The Mach-Zehnder interferometer with different path length differences between its two arms generates different phase shift variations over frequency, resulting in different free spectral ranges (FSR) of the null frequency band. The bandwidth of the null frequency band also changes over different path length differences in MZI. The SBS has a narrow linear phase response, with the slope of the phase response varying over the pump power, enabling compensation of a narrow null frequency band. The multiple null compensations could also be achieved by using multiple pump lasers tuned to fall at different frequency nulls, significantly improving the operational bandwidth of the APL link.

5.3 Theoretical Analysis

The delay between the two arms of MZI with a path length difference L is represented as follows,

$$t = n_{eff} \frac{L}{c} \quad (5.1)$$

where the delay t is determined from the path length difference L and speed of light c , and n_{eff} is the effective refractive index of the fiber.

The phase modulator coupled with the probe frequency f_c , and a swept RF frequency is applied through the VNA to generate 180° out-of-phase sidebands.

The SBS gain response is precisely positioned on the nulls of the MZI for null compensation. The pump laser frequency is tuned as shown in equation (5.2) for the SBS gain response at the desired null frequency present at the frequency, f_{null} away from the probe frequency:

$$f_{pump} = f_c + f_{sbs} + f_{null} \quad (5.2)$$

where, f_{pump} , is the pump frequency, f_c , is the probe frequency, f_{sbs} , is the Brillouin shift generated by the optical fiber and f_{RF} is the center frequency of one of the nulls of the PM+MZI link, which needs to be compensated to double the operating bandwidth. When the pump power is applied, an SBS gain and phase response are generated at f_{sbs} away from the pump frequency due to Brillouin scattering in the optical fiber. The RF gain at the center frequency depends on the pump power, as the SBS gain can be increased or decreased by adjusting the pump power. The numerical expression of SBS gain and the phase response is given by equations (5.3) and (5.4) [19]

$$G_{sbs} = \frac{\gamma P_p}{1 + 4 \left(\frac{\Delta\omega}{\Gamma_B} \right)^2} L_{eff} \quad (5.3)$$

$$\phi_{sbs} = \frac{\gamma P_p \left(\frac{\Delta\omega}{\Gamma_B} \right)}{1 + 4 \left(\frac{\Delta\omega}{\Gamma_B} \right)^2} L_{eff} c \quad (5.4)$$

where $\gamma = (g_B/A_{eff})$ is the nonlinear SBS parameter defined as the ratio of the SBS gain coefficient g_B and effective area A_{eff} . $\Delta\omega$ is the detuning of the frequency from the peak SBS gain and Γ_B , is the Brillouin linewidth.

The transfer function of the SBS gain response H_{SBS} is represented as,

$$H_{SBS} = \exp(G_{sbs} + j\phi_{sbs}) \quad (5.5)$$

The Electric field at the output of phase modulator coupled with probe laser of frequency ω_c and RF frequency Ω with amplitude V_{RF} is represented as,

$$E_0(t) = A \exp(j(\omega_c t + \phi(t))) \quad (5.6)$$

ω_c is defined as the optical carrier frequency of the laser, and $\phi(t)$ is defined as the phase shift added by the phase modulator and is specified as

$$\phi(t) = \phi \sin(\omega t) \quad (5.7)$$

ϕ is specified as $\phi = \pi \frac{V_{RF}}{V_\pi}$, where V_π is the voltage required by the phase modulator for

introducing a 180° phase shift to the signal.

The optical electric field at the output of the MZI is expressed as,

$$E_1(t) = B (E_0(t - \tau) - E_0(t)) \quad (5.8)$$

where B includes the losses in the link. The beating of the electric field $E_1(t)$, at the photodetector with a responsivity $\mathfrak{R}$, results in the generation of RF signals with null response determined by MZI delay τ .

Applying $E_0(t)$ from (6) in (8) and solving the optical electric field at the output of MZI is represented as,

$$E_1(t) = B\{\exp(j(\omega_c(t - \tau) + \phi(t - \tau))) - \exp(j(\omega_c(t) + \phi(t)))\} \quad (5.9)$$

The MZI output is coupled to a photodetector for optical to electrical conversion, and the current detected at the photodetector is expressed as,

$$I(t) = E_1(t) \cdot E_1(t)^* \quad (5.10)$$

Applying $E_1(t)$ from (5.9) in (5.10),

$$I(t) = B\{\exp(-j(\omega_c(t - \tau) + \phi(t - \tau))) - \exp(j(\omega_c(t) + \phi(t)))\} \cdot$$

$$B\{\exp(-j(\omega_c(t - \tau) + \phi(t - \tau))) - \exp(-j(\omega_c(t) + \phi(t)))\}$$

$$I(t) = [2B^2 - B^2\{\exp(-j(\omega_c\tau + (\phi(t) - \phi(t - \tau)))) + \exp(j(\omega_c\tau + (\phi(t) - \phi(t - \tau))))\}] \quad (5.11)$$

Considering the real part and applying $I_{dcq} = \frac{\mathfrak{R}B^2P_0}{2}$ and $\Delta\phi = \phi(t) - \phi(t - \tau)$, where $\mathfrak{R}$ is the responsivity of the photodiode, solving (11), the current is represented as,

$$I(t) = [I_{dcq} - I_{dcq} \cos(\omega_c\tau + \Delta\phi)]$$

$$I(t) = I_{dcq} - I_{dcq} \cos(\omega_c \tau) \cos(\Delta \phi) + \sin(\omega_c \tau) \sin(\Delta \phi)$$

$$I(t) = I_{dcq} - I_{dcq} \cos(\omega_c \tau) \cos(\phi \sin(\Omega t) - \phi \sin(\Omega(t-\tau))) - \sin(\omega_c \tau) \sin(\phi \sin(\Omega t) - \phi \sin(\Omega(t-\tau))) \quad (5.12)$$

$$I(t) = I_{dcq} - I_{dcq} \left\{ \cos(\omega_c \tau) \cos\left(2\phi \left(\cos\left(\Omega t - \frac{\Omega \tau}{2}\right) \sin\left(\frac{\Omega \tau}{2}\right)\right)\right) - \sin(\omega_c \tau) \sin\left(2\phi \left(\cos\left(\Omega t - \frac{\Omega \tau}{2}\right) \sin\left(\frac{\Omega \tau}{2}\right)\right)\right) \right\} \quad (5.13)$$

Applying the Jacobi Anger series on (13), $I(t)$ is represented as,

$$\begin{aligned} I(t) = I_{dcq} - I_{dcq} & \left\{ \cos(\omega_c \tau) J_0\left(2\phi \sin\left(\frac{\Omega \tau}{2}\right)\right) \right. \\ & + 2 \sum_{n=1}^{\infty} (-1)^n \cos(\omega_c \tau) J_{2n}\left(2\phi \sin\left(\frac{\Omega \tau}{2}\right)\right) \cos\left((2n)\left(\Omega t - \frac{\Omega \tau}{2}\right)\right) \\ & \left. - \sin(\omega_c \tau) 2 \sum_{m=1}^{\infty} (-1)^m J_{2m+1}\left(2\phi \sin\left(\frac{\Omega \tau}{2}\right)\right) \cos\left((2m+1)\left(\Omega t - \frac{\Omega \tau}{2}\right)\right) \right\} \end{aligned}$$

Considering only the fundamental terms, $I(t)$ is represented as,

$$I(t) = I_{dcq} - I_{dcq} \left\{ \sin(\omega_c \tau) 2J_1\left(2\phi \sin\left(\frac{\Omega \tau}{2}\right)\right) \cos\left(\left(\Omega t - \frac{\Omega \tau}{2}\right)\right) \right\} \quad (5.14)$$

Considering, $2\phi \sin\left(\frac{\Omega \tau}{2}\right) < 1$, the $J_1\left(2\phi \sin\left(\frac{\Omega \tau}{2}\right)\right)$ can be approximated as,

$$J_1\left(2\phi \sin\left(\frac{\Omega \tau}{2}\right)\right) \approx \phi \sin\left(\frac{\Omega \tau}{2}\right)$$

Putting the value of $\phi = \pi \frac{V_{RF}}{V_\pi}$ and considering the responsivity as $|H_{pd}|$ and the SBS gain transfer function as H_{SBS} , the final current detected at photodetectors is represented as,

$$I(t) = I_{dcq} - I_{dcq} \left\{ \sin(\omega_c \tau) 2\pi \frac{V_{RF}}{V_\pi} \sin\left(\frac{\Omega \tau}{2}\right) |H_{pd}| H_{SBS} \right\} \quad (5.15)$$

The small signal power gain of the phase-modulated link demodulated with MZI and SBS gain is represented as,

$$g = \left(\frac{I(t)^2 R_o}{\frac{V_{RF}^2}{R_i}} \right)$$

Putting the value of $I(t)$, the small signal gain can be represented as,

$$g = \left(\frac{\sin(\omega_c \tau) 2\pi \sin\left(\frac{\Omega \tau}{2}\right) |H_{pd}| H_{SBS}}{V_\pi} \right)^2 R_o R_i \quad (5.16)$$

where, R_i is the input resistance and R_o is the output resistance H_{pd} is the frequency response of the photodiode and $I_{dc,q}$, is the current at quadrature. Hence, the link gain of the PM+MZI link is tailored to match the SBS gain response, which improves the operational bandwidth of the link.

The mathematical model presented above is simulated in MATLAB to verify null frequency band compensation in a phase-modulated link demodulated with an MZI. The frequency of the pump laser is tuned to compensate for the desired null frequency band. Furthermore, multiple null compensation is simulated by using multiple pump lasers to compensate for the null frequency band at different frequencies.

5.4 Simulation

Simulations are carried out using MATLAB to verify the feasibility of the proposed concept. In the simulation, an APL is modeled with a laser source, phase modulator, MZI, a pump laser source, and a photodetector. Here, the MZI with a path length difference of 22 cm is considered. The RF input of the PM and the pump laser frequency have been changed to set the SBS Stokes response on different nulls of the MZI-demodulated PM signal at the PD.

5.4.1 Null compensation

To verify the concept of null compensation or broadening of the operational bandwidth in a phase-modulated link demodulated with an MZI, the SBS Stokes response has been applied to selected nulls to compensate for the transmission dips. Fig. 5.2(a) shows the phase-modulated link demodulated with an MZI having a 22cm path length difference. The FSR of the periodic null frequency band is 925 MHz. Fig. 5.2(b) demonstrates the first null frequency band

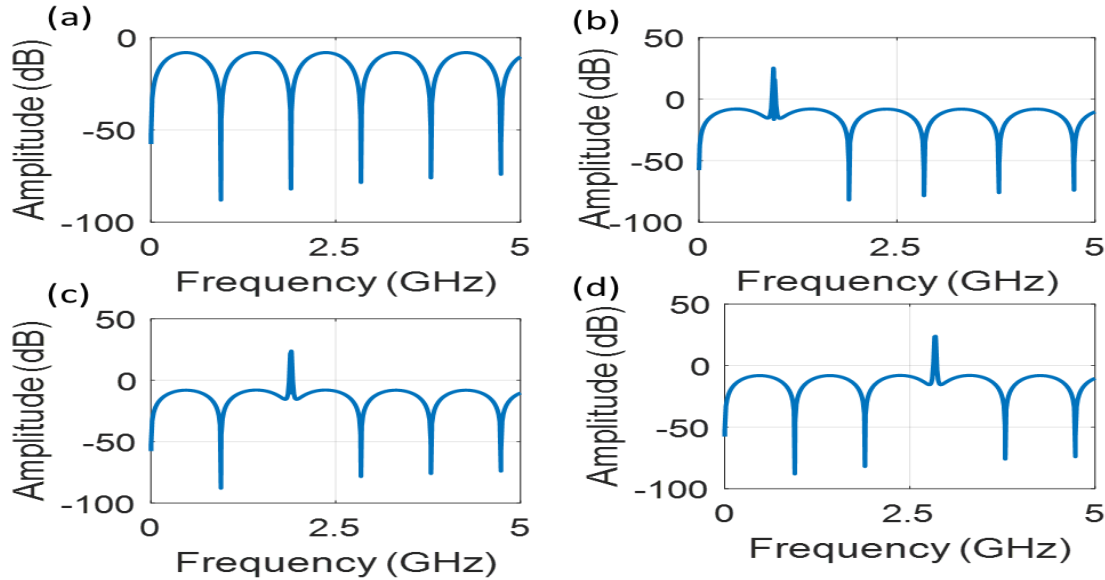


Fig. 5.2: (a) Phase modulated link demodulated with MZI with 22cm path length difference between two arms, (b) SBS gain response applied on first null frequency band at 925 GHz (c) SBS gain response applied on second null frequency band at 1.85GHz, (d) SBS gain response applied at third null frequency band at 2.75GHz

compensated with SBS gain response. Fig. 5.2(c) and Fig. 5.2(d) demonstrate the second and third nulls, respectively, compensated by tuning the laser frequency to achieve a gain response at 1.85 GHz and 2.75 GHz.

To achieve null responses with different null frequency bands, the MZI's path length difference has been varied, which changes the FSR of the nulls. Simulation of the PM with MZI and SBS link with different path length differences in MZI of $\sim 26\text{mm}$, $\sim 22\text{cm}$, and $\sim 46\text{cm}$ is shown in Fig. 5.3(a). The delay due to path length difference in MZI are 124.8ps(26mm), 1.05ns (46cm), 2.2ns(22cm). A null compensation of 19 dB has been achieved, and it is observed that the compensation technique improves the operational bandwidth of the link by a factor of two.

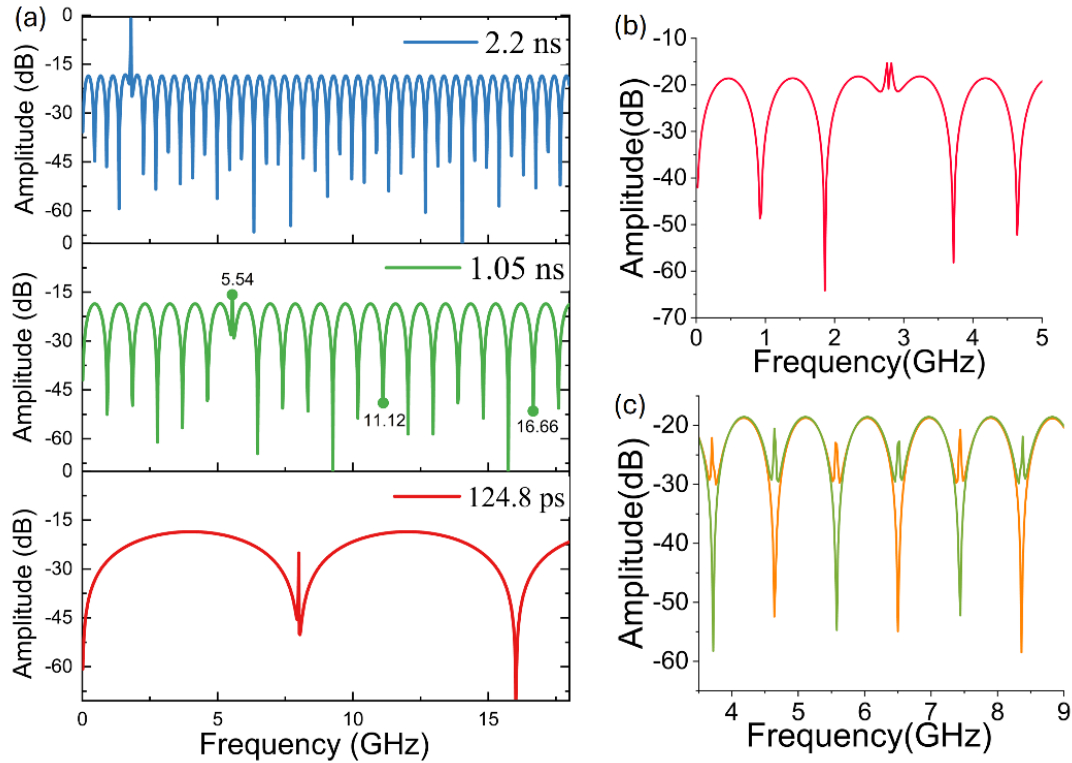


Fig. 5.3: Simulation of MZI delay tuning for generation of MZI nulls response at different FSRs of 420 MHz, 925 MHz, and 8.12 GHz for delays of 2.2 ns, 1.05 ns, 124.8 ps in the MZI. (b) Broadband null compensation by using a broadened pump. (c) Multiple-null compensation by using sets of three pump signals at 14.5 GHz, 16.35 GHz, 18.2 GHz, and 15.425 GHz, 17.275 GHz, 19.125 GHz to have filter response at 3.7 GHz, 5.54 GHz, 7.4 GHz in green and at 4.62 GHz, 6.46 GHz, 8.34 GHz specified in orange colors, respectively.

Furthermore, by using a broadened SBS pump, broadband null compensation can be achieved with high flatness, significantly improving the bandwidth of the APL link, as shown in Fig. 5.3(b). The multiple null compensations have been achieved by using multiple pumps to generate SBS gain response at different desired null frequency bands. By introducing three SBS pumps separated by the FSR of the link and positioning them on the nulls, three nulls can be compensated, as seen in Fig. 5.3(c), which can be extended using more SBS pumps.

5.4.2 Image Frequency Rejection

This technique of null compensation using stimulated Brillouin scattering gain response is also of interest for image rejection at the harmonics of the fundamental frequency. The desired RF frequency to be transmitted is centered at one of the nulls where the SBS gain response is positioned. SBS gain response enables the transmission of the fundamental frequency signal, but its image frequency is placed at other nulls of the PM+MZI link, thereby suppressing the image harmonic frequency. For example, in Fig. 5.3(b), for a delay of 1.05 ns, only the signal at 5.54 GHz is detected, whereas all the image frequencies at 11.1 GHz and 16.62 GHz are rejected. Image frequency rejection is beneficial for radar, where a high signal-to-noise ratio is necessary to detect small RCS targets.

5.5 Experimental Setup

An experimental setup to demonstrate the periodic nulls in a phase-modulated link demodulated with MZI by imparting SBS gain response is shown in Fig. 5.4. The setup consists of a phase modulator driven by a vector network analyzer (VNA), which modulates a CW laser carrier signal. A polarization controller (PC1) is used to optimize the PM transmission, and the modulated signal is coupled into a 3.2-km fiber through an isolator. A CW signal from a tunable laser (SBS pump)

is sent through a polarization controller (PC2), which controls the relative polarization of the pump and signal paths and is used to optimize the SBS gain. The output from PC2 is fed to the 1st port of a circulator, which is made to counter-propagate the carrier signal through port 2. The output from the 3rd port of the circulator (signal) is coupled to an MZI with a tunable path-length difference. The output from the MZI is fed to a photodiode connected to a VNA.

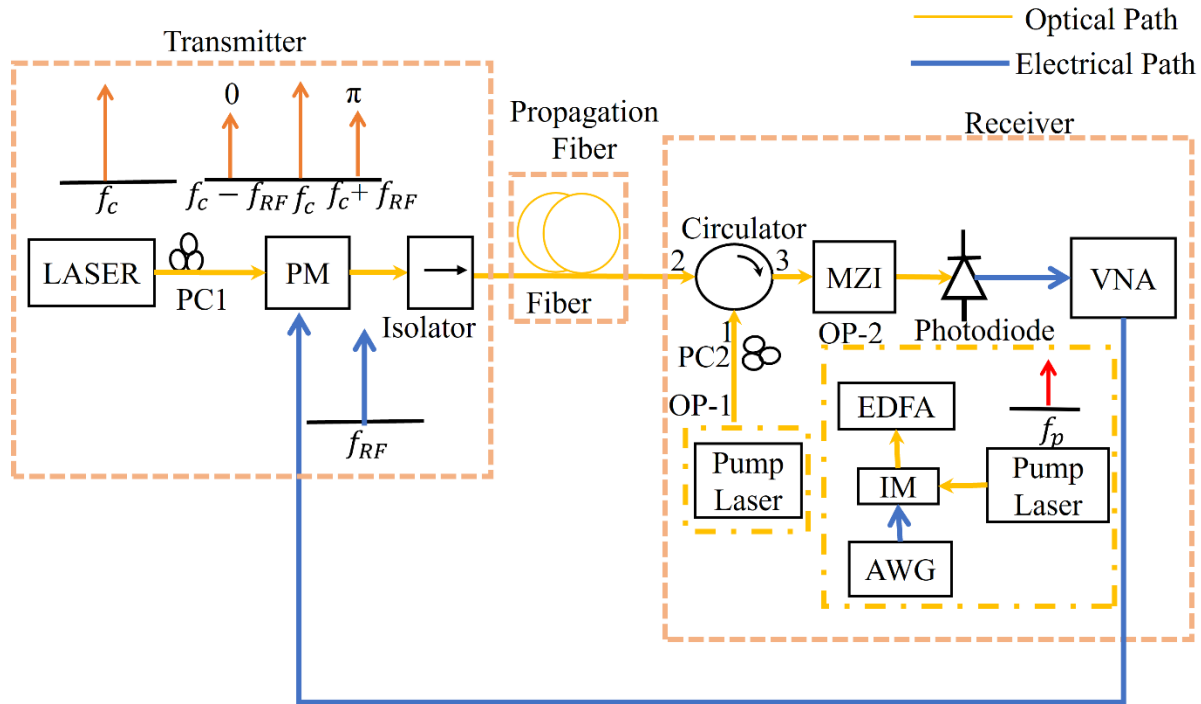


Fig. 5.4: Experimental setup of the proposed APL with MZI null compensation; OP-1(pump laser) and OP-2(multiple pump laser generation); PM, phase modulator; PC, polarization controller; MZI, Mach Zehnder interferometer; VNA, vector network analyzer; IM: intensity modulator; AWG: arbitrary waveform generator.

.which also measures the frequency responses.

To demonstrate multiple null compensation, an intensity modulator has been used to generate two sidebands along with the pump laser signal. The two intensity-modulated sidebands and the pump laser signal act as three pumps to generate the SBS gain response at three different frequencies, tuned by an RF input to the intensity modulator, as shown in Fig. 5.4 (OP-2). The pump power

level is controlled by the EDFA to control the SBS gain response.

5.6 Discussions & Results

To demonstrate the wideband APL using PM+MZI, the SBS pump laser (OP-1 in Fig. 5.4) frequency has been varied to position the SBS gain response at a null frequency band. A probe laser signal with a power of 18.5 dBm is supplied to a phase modulator driven by the VNA. A tunable pump laser of 12dBm power is coupled at the receiver to impart an SBS gain response at a Brillouin frequency shift of 10.8 GHz from the pump laser frequency. The pump laser is tuned to have an SBS response at different null frequency bands, thereby increasing the operation bandwidth of the link. It is observed from Fig. 5.5 part I-V that null compensation of up to 16 dB

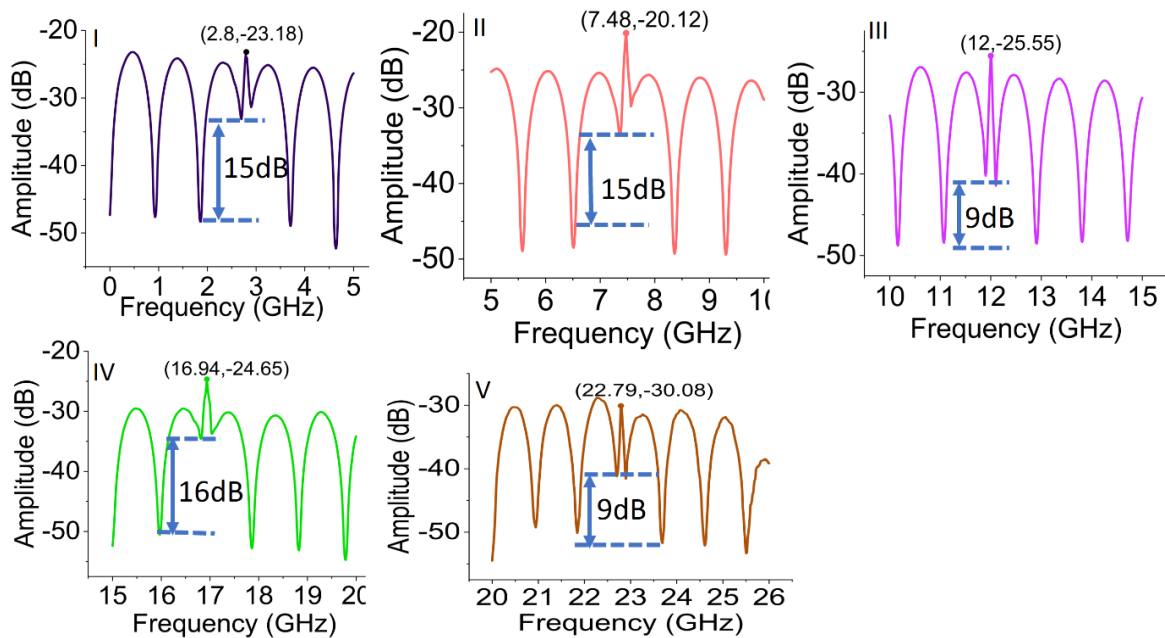


Fig. 5.5: The doubling of the operational bandwidth in the PM+MZI link is obtained using a vector network analyzer. Null compensation of up to 16 dB has been achieved using an MZI delay of 1.05ns (~22 cm path length difference) as shown in part I-V.

in amplitude and up to a frequency of 23 GHz has been achieved. The range of null frequency compensation can be increased by using high-frequency modulators and photodetectors. The

operational bandwidth of a PM+MZI link is doubled by compensating for one null frequency band. The change in path length difference in MZI changes the delay, which changes the FSR of the null frequency band. The path length difference in MZI has been varied to demonstrate the compensation of the periodic null frequency band with different FSRs. The MZI used has a path length difference of 26 mm (delay of 124.8ps), 22 cm (delay of 1.05 ns), and 46 cm (delay of 2.2ns), showing an FSR of 8.12 GHz, 925 MHz, and 420 MHz, respectively, as shown in Fig. 5.6 part (I-III). The pump laser is tuned to position the gain response at the center of the null frequency band at 8.19 GHz, 2.8 GHz, and 1.29 GHz for path differences of 26mm, 22cm, and 46cm, respectively. Null compensation of up to 16dB is achieved. To demonstrate the variation in link gain of a PM+MZI link with SBS, the pump laser power is varied to achieve different amplitudes of SBS gain response at the nulls of the PM+MZI link. The frequency response of the null frequency band compensation over pump power variation from 6 dBm to 12 dBm is shown in Fig. 5.7 (a). With the increase in pump power, the null compensation level also increases, allowing variable null depths to be configured flexibly through pump power control. For practical applications, optical phase-locked loops at the receiver can be used to frequency-lock the pump lasers, thereby preventing the relative drift of the pump laser with respect to the signal laser.

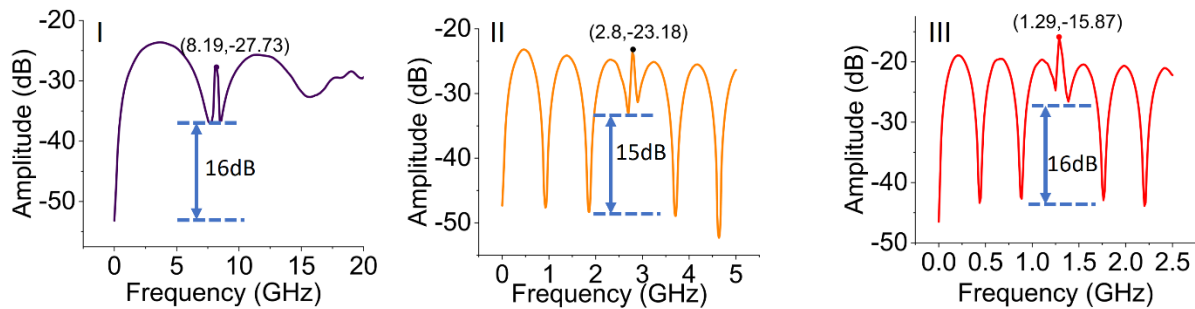


Fig. 5.6: Null compensation using SBS with MZI delay of 124.8ps, 1.05ns, and 2.2ns, as shown in Part I-III.

An important advantage of using a phase modulator in an APL is that it provides a link gain

advantage as compared to an IMDD link [11]. In our experiment, the link gain has been further improved using a balanced detector compared to a single photodiode. Here, the frequency response of a PM+MZI link with SBS on the first null, using a balanced photodetector, is shown in Fig. 5.7(b) in red, with a 6.5 dB improvement compared to the single photodiode (light blue). The link gain is currently limited to ~ -20 dB and can be improved through various link optimization techniques, including increased optical power and the introduction of RF amplifiers at the transmitter [20].

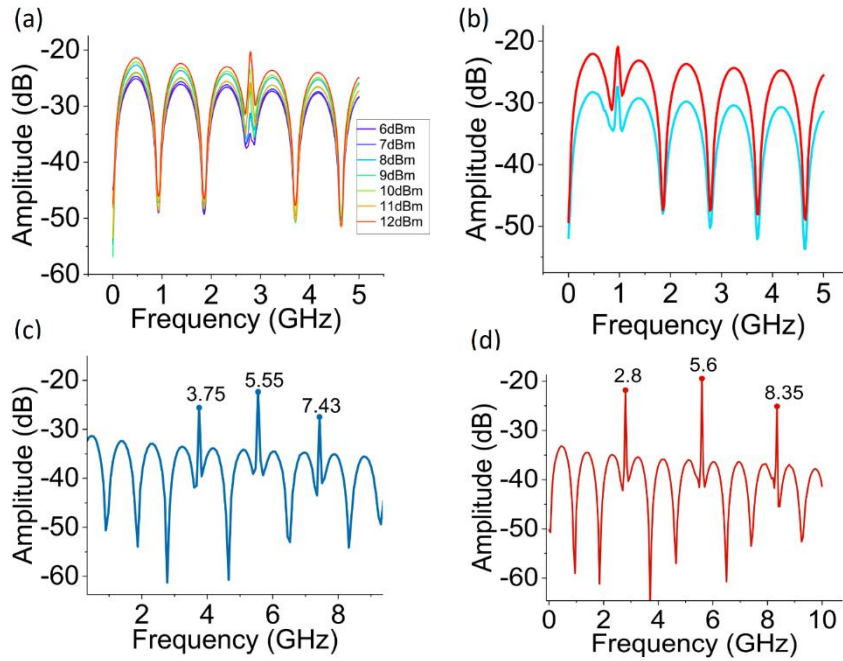


Fig. 5.7: Variation of the null compensation level by changing the pump power from 6 dBm to 12 dBm with an MZI with a ~ 22 cm path-length difference. (b) Improvement of the APL link response by using a balanced photodiode (red) when compared to a single photodiode (blue). Multiple-null compensation using three pump lasers with (c) FSR of 1.85 GHz, and (d) FSR of 2.75 GHz.

Furthermore, to demonstrate multiple null compensations, three pump frequencies were generated by modulating the pump laser using an intensity modulator, resulting in the addition of SBS Stokes responses at three adjacent nulls of the PM+MZI link. An AWG provides a tunable RF signal to the intensity modulator, generating two sidebands with the same amplitude as the carrier. The two

sidebands and carrier are used as pumps to impart three SBS gain responses on the three nulls of the link (shown as OP-2 in Fig.5.4). Two different configurations are tested by modulating an intensity modulator with an RF tone and using the three resultant optical tones as SBS pumps. By using two different RF tones, different sets of pump frequencies separated by 1.85 GHz and 2.75 GHz are generated. These resulted in null compensation, as shown in Fig. 5.7(c) at frequencies of 3.75 GHz, 5.55 GHz, and 7.43 GHz, and in Fig. 5.7 (d) at frequencies of 2.8 GHz, 5.6 GHz, and 8.35 GHz, respectively. The number of nulls that can be compensated for can be increased by using more than three pumps important for high-bandwidth electrical waveform generation [21].

The image rejection capability has been demonstrated by tuning the MZI length and SBS gain in a manner that eliminates undesired harmonics within the rejection bandwidth of the MZI. The image frequency rejection is further improved by having SBS gain response within the desired frequency bandwidth, resulting in a significant improvement in the rejection of undesired harmonics with a high rejection ratio. Here, MZI of length 22cm is used, and SBS gain is placed at 1.375GHz. In Fig.5.8 (part I), the rejection ratio between the desired frequency at 1.37 GHz and the image frequency at 2.75 GHz is 19.5dB. After enabling SBS gain response at 1.375 GHz, the rejection ratio increases to 34.5dB, as shown in Fig. 5.8 (Part II). This technique is useful in jamming conditions to amplify the desired frequency band with narrow SBS gain response while rejecting the jamming frequencies, resulting in the recovery of signal in a jammed environment.

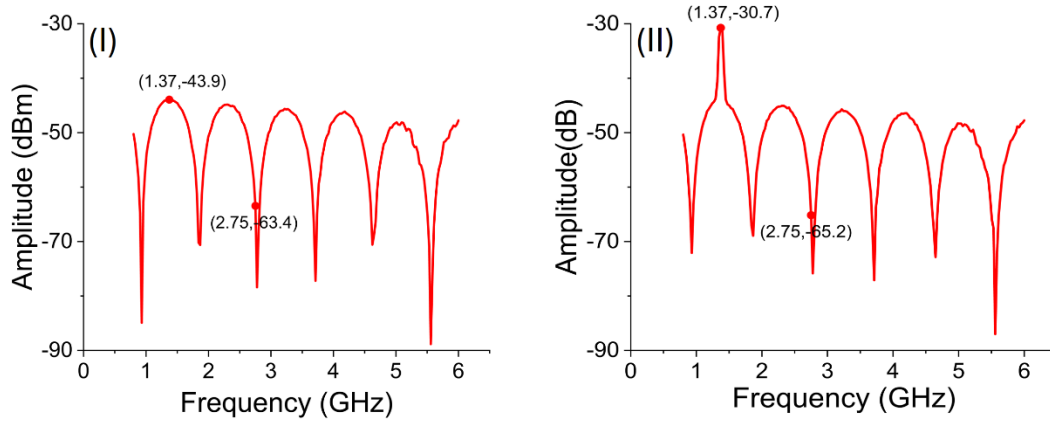


Fig. 5.8: Image frequency rejection, (I) phase modulated signal demodulated with MZI of path length difference of 22cm with desired frequency around 1.375GHz and image frequency at 2.75GHz with 19.5dB of rejection ratio, (II) SBS gain response applied at desired frequency band at 1.375GHz to have image frequency rejection ratio of 34.5dB.

The proposed system has demonstrated chromatic dispersion in a PM+MZI link due to a path length difference in the MZI, utilizing SBS gain response. A null frequency band compensation of up to 16dB has been achieved up to a frequency of 23 GHz. The multiple null frequency band compensation has been demonstrated by generating three pump lasers using an intensity modulator. The level of null compensation has been tuned by tuning the pump power level.

5.7 Conclusion and Future Work

In this work, a method of increasing the operational bandwidth of a phase-modulated link demodulated by an MZI is demonstrated using stimulated Brillouin scattering. This technique is useful for compensating for the dispersion-induced transmission nulls. Adding a tunable pump laser and a circulator at the receiver helps increase the bandwidth, while leveraging the transmission fiber as the gain medium, thus keeping the system's complexity very low. The null compensation level of up to 16dB has been achieved. Null compensation responses up to 23 GHz have been tested by tuning the pump laser frequency. Furthermore, multiple null compensation responses have also been demonstrated by using an intensity modulator to compensate for three

null frequency bands. The same setup has been utilized to demonstrate image frequency rejection by selectively amplifying the desired frequency band using the SBS gain response, thereby improving the image frequency rejection ratio.

The proposed technique for compensating the chromatic dispersion-induced periodic nulls enables a phase-modulated link for high-speed data transport by utilizing the larger bandwidth. The phase modulator does not require a DC bias, which improves the system's linearity in extreme environmental conditions. This enables the use of the PM+MZI link for multiband radar applications, where the SBS gain response can be tuned to enable transmission in specific frequency bands. This technique is highly useful for antenna remoting application with optical signal from signal processing system is transported to antenna system at geographically different location over low loss optical fiber which converted back to RF using PM+MZI without requiring any power. The increase in operational bandwidth can also be leveraged in radio-over-fiber networks and next-generation telecommunication systems.

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Chapter 6

Conclusion and Future work

6.1 Summary

The research presented in this thesis addresses the growing application of microwave photonics (MWP) in radar technology. The work explores MWP-enabled FMCW radar enhancement with stimulated Brillouin scattering based signal processing. The final chapter in this thesis provides a summary of the thesis and an outlook on novel prospects for future research on photonic radars.

MWP technology offers significant opportunities to achieve exceptional performance characteristics in radar systems. MWP enables high-resolution radar with high reconfigurability in frequency and bandwidth. FMCW radars have applications across diverse domains, including self-driving cars, aerospace, and satellites. Microwave photonics for FMCW radar, exploring simultaneous range and velocity estimation, is explored. Radars are required to operate in multiple frequency bands with different bandwidths to detect targets of different sizes and velocities. This requires a real-time tunable filter in the microwave frequency band to change the filter's frequency response, bandwidth, or to switch between transmission and rejection filters. A SBS based highly flexible EIT analogue and EIA analogue filters are also explored in this thesis. Chapter 2 establishes the theoretical background for FMCW radar and its mathematical modelling. It highlights the advantages of using microwave photonics for FMCW radar applications and explores signal processing techniques, such as filtering using SBS, along with the fundamentals of

MWP and SBS.

Chapter 3 demonstrates photonically generated highly tunable dual chirp waveform generation using a tunable reference laser. This technique requires a low-bandwidth modulator to generate dual-chirp waveforms at high frequencies and significantly decreases the system's cost. This technique can be used for a multipurpose radar, where the frequency of operation is tuned to detect different shapes and sizes of targets. This chapter also demonstrates a novel multiband complementary dual chirp waveform for simultaneous range and velocity estimation with higher accuracy. The experimental validation of the waveform has been demonstrated by simultaneously estimating the range, velocity, and direction of motion of a slow-moving toy car target in a laboratory environment. This waveform is highly useful for high-accuracy range estimation, along with high-resolution velocity estimation, for detecting both high-speed moving targets and small, slow-moving targets, such as drones.

Chapter 4 The multipurpose radar required operation in different frequency bands with waveforms of varying bandwidth, depending on the target's shape and size. The generation of a waveform with tunability in frequency and bandwidth necessitates a real-time microwave frequency band filter with tunability in center frequency and transmission or rejection bandwidth. The generation of highly tunable electromagnetically induced transparency (EIT) analogue filters, electromagnetically induced absorption (EIA) filters, and spectra hole (SH) analogue filter responses is demonstrated using two SBS anti-Stokes responses. The real-time switching between EIT analogue and EIA analogue filter response is achieved by frequency shifting of a pump laser. The variation in bandwidth, frequency of operation, and rejection ratio of the filters is achieved by tuning the pump lasers frequency and power. These highly tunable filters are useful for multiband radar operation, enabling the detection of both long-range, high-speed targets and small, slow-

moving targets. Bandwidth tunability is necessary to resolve very closely spaced targets, achieving high range resolution, which also requires filters to have real-time bandwidth tunability. The reconfigurable filters are also required for real-time control over transmission and reception frequencies, high-speed optical communication links, and radio-over-fibre systems.

Chapter 5 focuses on improving the operational bandwidth of the phase-modulated microwave photonic link, demodulated by a Mach-Zehnder interferometer. As the high-speed optical communication link and microwave photonic radar primarily use intensity modulators as photodetectors, they can detect only intensity fluctuations. However, the intensity modulators require one or more DC biases, which necessitate a DC voltage for operation and cause fluctuations in linearity due to changes in voltage under different environmental conditions. The operation of radar in remote locations, especially with antenna remoting and the availability of DC supply, is limited. The phase modulator does not require a DC bias, resulting in better linearity in challenging environmental conditions. However, the phase-modulated link has a π phase difference between the sidebands, resulting in no detection at the photodetector. MZI is used before PD for phase-to-intensity conversion; however, the additional path length difference in MZI leads to chromatic dispersion, resulting in a periodic null response that reduces the operational bandwidth necessary for high-speed communication links and radars. In this work, a periodic null at a specific frequency of operation is compensated by applying a SBS gain response at the centre of the null frequency band, thereby improving the operational bandwidth of a phase-modulated link demodulated with an MZI. The multiple null compensation is also achieved by using multiple pumps. Experimental validation is performed to compensate for the periodic nulls due to chromatic dispersion in MZI by utilizing the SBS gain response. With the availability of higher operational bandwidth and the capability to compensate for the desired frequency band, the phase-modulated link can be highly

useful for multiband radar operation, where the SBS gain response can be tuned to a desired null frequency band for transmission in a specific band and rejecting other bands with a periodic null band.

6.2 Future works

This section presents viewpoints on future research and challenges for next-generation photonic radars. These points aim to identify potential future research and opportunities.

6.2.1 On-chip photonic radar with waveform generation, transmitter, receiver, and signal processing, like filtering, in a single integrated photonic chip

With the advancement of integrated photonics, lasers [3] [4], modulators [5], and photodiodes [6] [7] are integrated on a single hybrid integrated photonic chip [8] [9]. This enables the integration of tunable radar waveform generation, transmitter, receiver, and signal processing, such as filtering, on a photonic chip, making the FMCW radar highly compact and efficient. The presented generation of a dual chirp waveform with tunable frequency by adjusting the laser frequency [10] can be easily integrated onto a photonic chip. The waveform generation can be further explored to generate multiband LFM, triangular chirp, or cross-chirp waveforms on a photonic chip. The presented electromagnetically induced transmission (EIT) and electromagnetically induced absorption (EIA) analogue filters [11], which offer tunability in frequency, bandwidth, and rejection ratio by adjusting the pump laser frequency, can also be integrated into a photonic chip as real-time tunable filters. Moreover, the advantages of photonic integrated circuits can be leveraged to enhance the performance, miniaturization, and capabilities of FMCW radar systems, thereby creating new opportunities for high-tech sensing and imaging applications.

6.2.2 Large bandwidth waveform generation for high resolution radar imaging

The bandwidth of linear frequency-modulated signals is critical for achieving high time-bandwidth

products [12], which in turn provide high range and Doppler resolution. The range resolution is inversely proportional to the bandwidth of the transmitted waveform, enabling the detection of multiple targets that are very close to each other [13] [14]. The temporal period of the waveform is crucial for frequency resolution when detecting slow-moving targets. The higher time-bandwidth product waveform can be used for better target discrimination, enhanced range resolution, and interference rejection, as well as an ISAR image of the target [15]. The large time-bandwidth waveform can be generated by using a small bandwidth LFM waveform and multiple RF tones to multiply the LFM waveform over time to generate a large time-bandwidth product LFM waveform, as shown in Fig. 6.1.

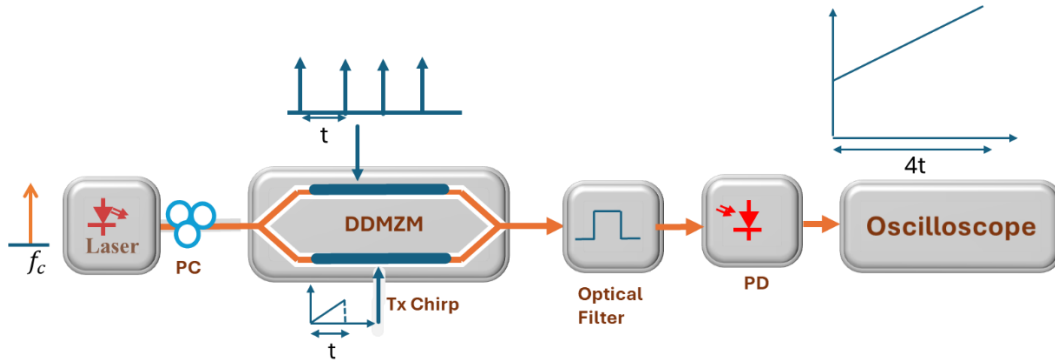


Fig. 6.1: Schematic to generate large time-bandwidth product waveform using small bandwidth LFM and multiple frequency tones. DDMZM: dual drive Mach Zehnder modulator, PD: photodiode, PC: polarization coupler

6.2.3 Pulsed photonic radar for long range multiple target detection

The frequency-modulated continuous-wave microwave photonic radar is used for detecting range and velocity over a small range. The detection of long-range targets requires an interleaved transmission and reception period. The pulse photonic radar [16] is used for detecting low RCS targets over a long range. We propose a photonic-based pulsed radar architecture for detecting lower RCS targets, along with long-range target detection, by utilizing longer receive periods compared to transmission times, as shown in Fig. 6.2. In the FMCW radar architecture, an LFM

signal is transmitted, and the received echo signal beats with the transmitted reference LFM signal to generate beat frequencies [17] - [20]. Here, the receiving time is the same as the LFM temporal width to obtain the beat signal. To improve the detection range, a large-bandwidth LFM signal is generated, and the transmission time is controlled using an RF switch to transmit a fraction of the generated large temporal-width LFM signal. The echo signal received from large-distance targets beats with a large temporal LFM signal for range and velocity estimation. This can be used to detect multiple targets.

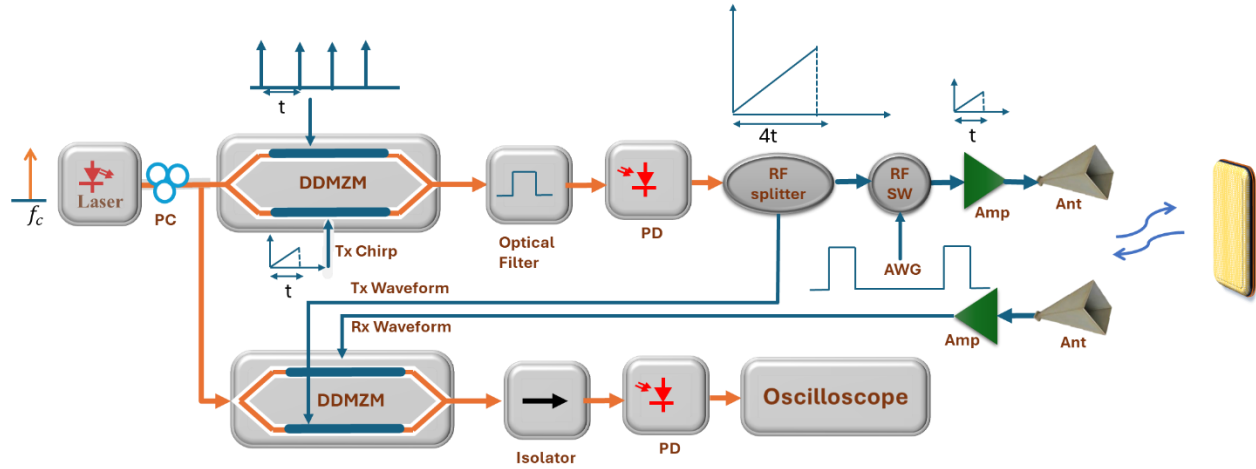


Fig. 6.2: Pulsed photonic radar architecture for long range multiple target detection. DDMZM: dual drive Mach Zehnder modulator; PC: polarization coupler; PD: photodiode; AWG: arbitrary waveform generator; Ant: antenna

6.3 Conclusion

The realization of a highly reconfigurable and reliable microwave photonic radar system on a photonic integration platform requires waveform generation, transmitter, receiver, and signal processing on a single photonic chip. The recent advancements in hybrid integrated photonic circuits, which incorporate lasers, modulators, and photodetectors, enable the realization of a single PIC-based radar. The implementation of signal processing elements, such as filters, delay lines, and phase shifters, can lead to a photonic radar system with significant improvements over conventional electronic-based radars. This thesis presents a method to generate radar waveforms,

such as dual chirp, with high tunability in frequency, requiring a very low bandwidth modulator. The frequency tuning is achieved using a CW laser, making it suitable for implementation on a PIC. A novel multiband complementary waveform is also demonstrated for simultaneous range and velocity estimation. A novel, highly flexible EIA and EIT analogue filter response with frequency bandwidth and rejection ratio tuning using a CW laser is also demonstrated and can be implemented on a PIC for real-time filtering in multiband radar applications. This thesis also presents a novel technique to improve the operational bandwidth of a phase modulator demodulated with an MZI, enabling its application in antenna remoting with no DC power availability. The analytical expression and experimental validation have been performed for waveform generation, range, and velocity detection of toy cars, along with optical signal processing, including filtering and chromatic dispersion compensation, in a phase-modulated link. The future work presented in this chapter aims to address the bottleneck in large-time-bandwidth waveform generation and the realization of pulse photonic radar for various applications, ranging from self-driving cars, radars, high-speed optical communication, and radio-over-fiber. Ultimately, by working together, academia, industry, and research institutions can drive the future of photonic FMCW and pulse radar technology, thereby supporting its wide-scale adoption.

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List of Publications

Contributions to patents contributing to the thesis:

[P1] Amol Choudhary, Rajveer Dhawan, Reena Parihar, **Mukund Jha**, “Tunable Microwave Photonic Bandpass Filter using the Mach Zehnder Interferometer and Stimulated Brillouin Scattering,” Indian Patent No. 202211010792, Feb 2022.

[P2] Amol Choudhary, **Mukund Jha**, R. Dhawan, Anu Sharma, “System And Method For Velocity Detection of a Radar Target,” Indian Patent No. 202411039577, May 2024 (filed).

Contributions to peer-reviewed journals contributing to the thesis:

[J1] **Mukund Jha**, Rajveer Dhawan, Reena Parihar, and Amol Choudhary, “Frequency Response Tailoring of a Phase Modulated Link with Stimulated Brillouin Scattering,” in Journal of Photonics Technology Letters, PTL-39358-2022.

[J2] **Mukund Jha**, Rajveer Dhawan, Reena Parihar, and Amol Choudhary, “Highly Flexible RF Filter Responses Using Stimulated Brillouin Scattering With Bandwidth up to 1 GHz,” J. Lightwave Technol. 41, 6675-6683 (2023)

[J3] **Mukund Jha**, Lakshay Kumar, Rajveer Dhawan, Atul Khanna, Anu Sharma, and Amol Choudhary, “Multiband complementary dual-chirp waveform for the simultaneous range and velocity detection,” Appl. Opt. 64, 6746-6756 (2025)

Contributions to other peer-reviewed journals:

[J4] Rajveer Dhawan, Debashis Parida, Reena Parihar, **Mukund Jha**, and Amol Choudhary, “Multiband LFM waveform generation and band-selection using stimulated Brillouin scattering,” Appl. Opt. 62, 6737-6745 (2023)

[J5] R. Dhawan, A. Sharma, **M. Jha** and A. Choudhary, “Simultaneous Range and Velocity Detection With Asymmetric Dual-Chirped Photonic Radar,” J. Lightwave Technol., 43, 1222-1230 (2024)

Contributions to peer-reviewed conferences contributing to the thesis:

- [C1] **M. Jha**, A. Khanna, L. Kumar, R. Dhawan, A. Choudhary, “Frequency Agile Dual-Chirp Microwave Waveform Generation with MZM,” *Conference on Lasers and Electro-Optics/Europe (CLEO/Europe 2025)*
- [C2] **M. Jha**, R. Parihar, R. Dhawan, and A. Choudhary, “Stimulated Brillouin Scattering based Null compensation in a Phase Modulated Link,” in *Frontiers in Optics + Laser Science 2022 (FIO, LS)*, paper JTU5A.46.
- [C3] **M. Jha**, R. Parihar, R. Dhawan, and A. Choudhary, “Switchable Analogue of Electromagnetic Induced Absorption and Transmission-like Filter Responses using Stimulated Brillouin Scattering,” in *Conference on Lasers and Electro-Optics/Europe (CLEO/Europe 2023)*
- [C4] **M. Jha**, R. Dhawan, A. Sharma and A. Choudhary, “Photonically-generated Multiband Complementary Dual Chirp Waveforms for Range and Velocity Detection,” in *Conference on Lasers and Electro-Optics (CLEO PR 2024)*.
- [C5] **M. Jha**, A. Khanna, L. Kumar, R. Dhawan and A. Choudhary, “Frequency Agile Dual-Chirp Microwave Waveform Generation with MZM,” in *Conference on Lasers and Electro-Optics Europe (CLEO/Europe-EQEC 2025)*.

Contributions to other peer-reviewed conferences:

- [C6] L. Kumar, A. Khanna, **M. Jha**, R. Dhawan and A. Choudhary, “MZM-Based FMCW Radar Receiver for Multi-Target Detection,” in *Conference on Lasers and Electro-Optics Europe (CLEO/Europe-EQEC 2025)*.
- [C7] A. Sharma, R. Dhawan, **M. Jha** and A. Choudhary, “Distance and Velocity Measurement Using a Single Asymmetric Dual-Chirp with FMCW RADARs,” in *Conference on Lasers and Electro-Optics Pacific Rim (CLEO-PR 2024)*.

Mukund Jha

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CAREER OBJECTIVE

To contribute my skills and knowledge to a design, development, and test team, working on cutting-edge technology in Microwave photonics, Photonic Radar, radio-over-fiber, Navigation, fiber optics, and protocol software development for mission-critical systems.

EDUCATION

2020-2026	Doctor of Philosophy (PhD): Electrical Engineering, Microwave photonics, Indian institute of Technology Delhi, Hauz Khas, New Delhi, India - Cumulative Grade – 8.125/10 - Thesis Title: Microwave Photonics for Ranging and Filtering Applications
2006-2010	Bachelor of Technology (B.Tech): Electronics and Communication Engineering, Maharaja Agrasen institute of Technology, New Delhi, India

Technical Skills

Design & Simulation Tools	MATLAB, OptSim, Optisystem, Ansys Lumerical
Programming Languages	C, C++, JAVA, VHDL, VERILOG
Test Equipment	Lasers, Optical modulators, EDFA, PD's, OTDR, optical spectrum analyser, Splicing, RF amplifiers, arbitrary waveform generators, Vector Network analyser (VNA), oscilloscope, and electrical spectrum analyser
Testing	Environmental Testing – JSS 55555, EMI/ EMC, MIL-STD 461 E/F

PROFESSIONAL BACKGROUND

Presently working at **Dynamatic Technologies limited**, New Delhi, as optical design architect, my primary job function includes the design and development of hardware and software products for defense and aerospace Applications.

WORK EXPERIENCE

[1.] Project: Development of Unmanned air vehicle for the Indian Airforce

- Design and development of sensors like EO gimbal, IR gimbal, and dual EO/IR gimbal for different classes of UAV.

- Testing of gimbal functionalities and performing the gain tuning.
- Development of real-time video stabilization to produce stabilized video on GCS.
- Development of interface software for the autopilot to integrate various sensors.

[2.] Project: Surface Surveillance Radar for Indian Navy

- Lead the Design, Development, Testing & Validation of a Multi-Function Console with military standards JSS 55555 and MIL-STD 461F for Surface Surveillance Radar.
- Successful Development and Production of special test equipment for all Radar sub-modules automated testing.
- Design and Development of Analog Signal Distribution board and high voltage amplifier board PCBs for interfacing ship sensors with Radar.
- Development of Microwave boards for Radar
- Testing of Scanter 6002 Surface Surveillance Radar with Radar Stabilization platform and Sensor's simulators and Successful completion of FAT of 15 Radars.

[3.] Project: Tactical Access Switch Development for Indian Army

- Developed an Elementary Management system to manage Tactical Access Switches located at geographically different locations.
- Development of Interactive GUI with SNMPv1, SNMP-v2, and SNMP-v3 Protocol supports sending requests and receiving traps from all internal components like switches, Routers, and VOIP systems.

[4.] Project: DVB-T S band Video Transceiver Design & Development for UAV Application

- Designed 2.1 GHz S-band Transmitter and Receiver for UAV
- Selection and Validation of components with respective evaluation boards like Mixer, LNA, Power Amplifier etc., used in the design.
- System Level Design and Simulation of Transmit and Receive chain in System Vue
- Developed Schematic and Layout for both Transmit and Receive Chain.

[5.] Project: Microwave Planning System Software Development

- Implemented Microwave Planning models Longley Rice and ITU models for Point to Point and Point to Multipoint link calculations.

➤ Designed GIS to load all types of DTED files and use these files for Link design calculations and displaying the Results on GIS interactively.

- Successful Testing of Microwave planning software by SQAG.

PUBLICATIONS

- **Mukund Jha**, Rajveer Dhawan, Reena Parihar, and Amol Choudhary, "Frequency Response Tailoring of a Phase Modulated Link with Stimulated Brillouin Scattering," in Journal of Photonics Technology Letters, PTL-39358-2022.
- **Mukund Jha**, Rajveer Dhawan, Reena Parihar, and Amol Choudhary, "Highly Flexible RF Filter Responses Using Stimulated Brillouin Scattering With Bandwidth up to 1 GHz," J. Lightwave Technol. 41, 6675-6683 (2023)

- **Mukund Jha**, Lakshay Kumar, Rajveer Dhawan, Atul Khanna, Anu Sharma, and Amol Choudhary, "Multiband complementary dual-chirp waveform for the simultaneous range and velocity detection," *Appl. Opt.* 64, 6746-6756 (2025)
- Rajveer Dhawan, Debashis Parida, Reena Parihar, **Mukund Jha**, and Amol Choudhary, "Multiband LFM waveform generation and band-selection using stimulated Brillouin scattering," *Appl. Opt.* 62, 6737-6745 (2023)
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- **M. Jha**, A. Khanna, L. Kumar, R. Dhawan, A. Choudhary, "Frequency Agile Dual-Chirp Microwave Waveform Generation with MZM," *Conference on Lasers and Electro-Optics/Europe (CLEO/Europe 2025)*
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- **M. Jha**, R. Dhawan, A. Sharma and A. Choudhary, "Photonicallly-generated Multiband Complementary Dual Chirp Waveforms for Range and Velocity Detection," in *Conference on Lasers and Electro-Optics (CLEO PR 2024)*.
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- A. Sharma, R. Dhawan, **M. Jha** and A. Choudhary, "Distance and Velocity Measurement Using a Single Asymmetric Dual-Chirp with FMCW RADARs," in *Conference on Lasers and Electro-Optics Pacific Rim (CLEO-PR 2024)*.

PATENTS

- Amol Choudhary, Rajveer Dhawan, Reena Parihar, **Mukund Jha**, "Tunable Microwave Photonic Bandpass Filter using the Mach Zehnder Interferometer and Stimulated Brillouin Scattering," Indian Patent No. 202211010792, Feb 2022.
- Amol Choudhary, **Mukund Jha**, R. Dhawan, Anu Sharma, "System And Method For Velocity Detection of a Radar Target," Indian Patent No. 202411039577, May 2024 (filed).