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Integrated programmable circuits for microwave-photonics radio base stations

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for microwave-photonics radio base stations.**

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*A mis padres Carlos y María José,
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y a mis abuelos Miguel, Charo, Pedro y Amparo.*

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Abstract

Nowadays we are experiencing a true revolution in the fields of telecommunications and computing, driven by an exponential increase in the demand for resources from emerging applications. Technologies such as autonomous driving with LiDAR systems, the Internet of Things, and large language models require not only a high investment in terms of power consumption, but also extremely high operating speeds, due to the complexity and demands of their functionalities. In this context, photonics, and in particular programmable integrated photonics, is emerging as a key technology to address these challenges, by enabling higher processing speeds, reduced energy consumption, and lower manufacturing costs in the implementation of current and future communication systems.

This thesis aims to develop solutions based on programmable integrated photonics to address some of the main current demands in the field of communications, such as beamforming and signal processing. Throughout this work, several designs are presented and experimentally validated that not only significantly outperform previously reported results in the literature, but also show great potential for large-scale fabrication at low cost, making them promising candidates for improving the communication systems of the future.

Resumen

Actualmente estamos viviendo una auténtica revolución en el ámbito de las telecomunicaciones y la computación, impulsada por un crecimiento exponencial en la demanda de recursos por parte de nuevas aplicaciones emergentes. Tecnologías como la conducción autónoma con sistemas LiDAR, Internet of Things y los grandes modelos de lenguaje requieren no solo una elevada inversión en términos de potencia energética, sino también velocidades de operación extremadamente altas debido a la complejidad y exigencia de sus funcionalidades. En este contexto, la fotónica, y en particular la fotónica integrada programable, se perfila como una tecnología clave para afrontar estos retos, al permitir incrementar la velocidad de procesamiento, reducir el consumo energético y abaratar los costes de fabricación en la implementación de los sistemas de comunicación actuales y futuros.

Esta tesis tiene como objetivo desarrollar soluciones basadas en fotónica integrada programable para dar respuesta a algunas de las principales demandas actuales en el ámbito de las comunicaciones, como el conformado de haz y el procesado de señal. A lo largo de este trabajo se presentan y validan experimentalmente varios diseños que no solo superan significativamente los resultados previamente reportados en la literatura, sino que además muestran un alto potencial de ser fabricados a gran escala con costes reducidos, lo que los convierte en candidatos prometedores para mejorar los sistemas de comunicación del futuro.

Resum

Actualment estem vivint una autèntica revolució en l'àmbit de les telecomunicacions i la computació, impulsada per un creixement exponencial en la demanda de recursos per part de noves aplicacions emergents. Tecnologies com la conducció autònoma amb sistemes LiDAR, Internet of Things i els grans models de llenguatge requereixen no sols una elevada inversió en termes de potència energètica, sinó també velocitats d'operació extremadament altes, a causa de la complexitat i exigència de les seues funcionalitats. En aquest context, la fotònica, i en particular la fotònica integrada programable, es perfila com una tecnologia clau per a afrontar aquests reptes, al permetre incrementar la velocitat de processament, reduir el consum energètic i abaratir els costos de fabricació en la implementació dels sistemes de comunicació actuals i futurs.

Aquesta tesi té com a objectiu desenvolupar solucions basades en fotònica integrada programable, per a donar resposta a algunes de les principals demandes actuals en l'àmbit de les comunicacions, com són el conformament de feix i el processament de senyal. Al llarg d'aquest treball es presenten i es validen experimentalment diversos dissenys que, no sols superen significativament els resultats prèviament reportats en la literatura, sinó que a més mostren un alt potencial de ser fabricats a gran escala amb costos reduïts, cosa que els converteix en candidats prometedors per a millorar els sistemes de comunicació del futur.

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Chapter 1

Introduction

Human beings are intrinsically social animals with a fundamental need to communicate with others. This communication has driven our tremendous development as a species, evolving from rudimentary sound-based communication that could not be considered language, to written communication, and eventually to long-distance written communication through various messaging techniques. Today, we have achieved wireless communication over distances far beyond imagination, using light itself as the transport mechanism. We live in the Information Age, where large-scale information systems play a crucial and ever-expanding role in our daily lives. This role continues to grow with the emergence of new communication and computing paradigms such as the Internet of Things (IoT), autonomous vehicles, artificial intelligence, smart cities, and the ongoing development of 5G and 6G networks. These advancements raise questions about whether it will be possible to create mechanisms to support such immense demand [1], [2].

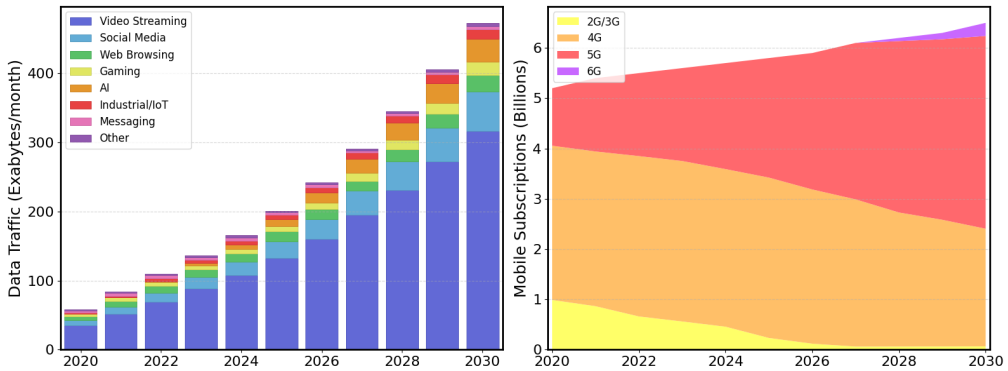


Figure 1.1: Annual growth in data traffic consumption and mobile user numbers, including projections for the coming years, with an estimated Compound Annual Growth Rate (CAGR) of approximately 18%. Data reproduced from CISCO and NOKIA. [1], [3].

One solution for enhancing the capacity of communication channels involves increasing the number of antennas used in both transmission and reception and, following this idea, technologies like Multiple-In Multiple-Out (MIMO) and microwave-phased array antennas (PAAs) have been enablers of the evolution of the past Long-Term Evolution

(LTE) [4], [5]. These technologies are now being adapted to include large-scale antenna arrays, with hundreds of elements, to optimize operations at mmWave carrier frequencies. Such architectures demand complex control and processing systems, along with significant power consumption—a trade-off for achieving high data rate transmissions. Traditionally, digital signal processors (DSPs) have been the backbone of wireless communications. However, the growing demands of advanced applications are pushing the speed limits of DSPs, which typically operate below a few gigahertz, constrained primarily by electronic sampling rates. Similar limitations are found in other conventional electronic components, such as beamforming networks and filters, whose bandwidths fall short of meeting current performance requirements. Therefore, developing novel technologies that enable high-speed, high-capacity, and low-latency performance while maximizing power efficiency is crucial. To achieve higher speeds and lower power consumption, optical analog processing engines can be employed before DSPs for both computation and signal processing—this is where microwave photonics plays a key role.

Microwave photonics (MWP) is a discipline that combines the principles of microwave engineering and photonics to address various challenges related to the generation, processing, transmission, and reception of microwave signals using optical techniques. In MWP, microwave signals are up-converted to optical signals and vice versa, enabling high-performance functionalities and advantages from optical systems that are not easily achievable with conventional microwave or classical electronic technologies alone [6]–[8]. Some of the main benefits include the wide bandwidth provided by optics, which allows for the transmission and processing of high-frequency microwave signals without significant loss or distortion. This is particularly advantageous for signal processing applications that require elevated data rates, such as microwave signal filtering, beamforming networks and distortion compensation [9]–[11]. Additionally, these benefits enable the generation of higher microwave frequencies, which are typically challenging to achieve with traditional electronic systems in terms of frequency stability, phase noise, and tunability. Another significant advantage is optics’ immunity to external electromagnetic interference, unlike traditional electronic transmission techniques. By mitigating signal degradation and noise, microwave photonics enhances system reliability and performance, making it ideal for deployment in electromagnetically intense environments such as aerospace radar systems and industrial settings. This has been highly influential, particularly when combined with the low losses, in the creation of Radio-over-Fiber Transmission Systems, where optical fiber links are used to distribute high-frequency microwave signals over long distances with minimal loss and distortion [12]. These systems are used in applications such as distributed antenna systems and wireless communication networks, where centralizing the signal processing chain and decoupling it from the antennas allows for a simplified antenna architecture. Finally, microwave photonics has emerged as a viable option for achieving fast-speed and high-resolution interrogation by translating optical wavelength shifts from the optical domain to microwave frequency shifts in the microwave domain [13], [14]. For all these fields of application, MWP serves as a high-speed analog alternative that significantly alleviates the workload of classical electronic devices. However, despite its numerous benefits and applications, discrete MWP systems suffer from certain drawbacks. Frequently, these systems are bulky and feature relatively expensive discrete fiber-optic components, leading to a lack of flexibility. Moreover, interconnections with fiber pigtailed can reduce system sturdiness, potentially diminishing the high power efficiency that MWP offer in some cases. Fortunately, the advancing maturity of

integrated photonic technology enables the fabrication of increasingly large and complex circuits directly on a chip. This represents a significant advancement for MWP systems, as many of the drawbacks mentioned earlier can now be effectively compensated [15], [16]. The immediate improvement lies in the size of the systems, allowing for deployment in portable and space-constrained environments. In addition, thanks to integration advancements, it is now feasible to have multiple functionalities that were once represented by separate devices onto the same chip. This, added to the reconfigurability of the integrated devices, leads to cost reduction, minimizes fiber coupling losses, and mitigates dissipation losses attributed to thermal control, as the same control source can be utilized for multiple functions [17], [18]. Advances on the technology have made possible the apparition of different fabrication platforms simultaneously. Each of these platforms has its own strengths and is chosen based on the specific requirements of the application, such as the need for active or passive components, integration density, cost, and performance characteristics [19], [20].

- Silicon-on-Insulator (SOI) is one of the most mature fabrication technologies, offering significant advantages for future development. Its compatibility with Complementary Metal-Oxide-Semiconductor (CMOS) manufacturing processes allows for the seamless integration of photonic and electronic components on the same chip. The high refractive index contrast between silicon and silicon dioxide enables strong light confinement in silicon waveguides, allowing for high integration density while maintaining moderate losses, around 1 dB/cm, making them suitable for a wide range of applications. However, the indirect bandgap of silicon presents a challenge for active elements. While doping processes enable the creation of photodetectors and modulators, integrating efficient light sources remains an issue.
- Indium Phosphide (InP) is well-studied technology, primarily due to its advantageous direct bandgap property. This characteristic makes InP the ideal choice for monolithic integration of both active and passive components, allowing for efficient light emission, amplification, and detection. Despite these benefits, InP is not well-suited for integration with CMOS technology, making the fabrication process expensive.
- Silicon Nitride (SiN) has garnered significant attention in recent years due to its exceptionally low optical losses, the lowest among all photonic platforms, making it uniquely valuable for specific applications. Like SOI, SiN is CMOS-compatible, facilitating integration with electronic circuits. Additionally, it offers a wide transparency range and high power handling capabilities. Nevertheless, despite its advantages, the refractive index contrast of SiN, while higher than that of InP, is still lower than that of silicon, resulting in less efficient light confinement compared to silicon photonics, potentially requiring larger device footprints .
- Other platforms: integrated MWP have been demonstrated using other materials. Gallium Arsenide (GaAs), known for its direct bandgap, is used for active circuits. Silica glass planar lightwave circuits have been demonstrated in wavelength division multiplexing (WDM) systems due to their excellent optical properties and ease of mass production. However, the low index contrast of this material limits the size of the circuits, requiring larger footprints. Finally, special mention goes to

Lithium Niobate (LiNbO₃). This material boasts unique and exceptional electro-optic, nonlinear-optic, and acousto-optic properties, making it ideal for applications such as high-speed modulators. It has historically faced significant challenges in material integration and processing but, despite these difficulties, LiNbO₃ is gaining renewed popularity due to recent advancements in thin-film integration, which promise to overcome previous limitations and expand its applications.

Overall, this dissertation highlights the fast-growing nature of this research field and its key role in developing future communication networks. It also presents the work done in this thesis, which aims to help bring cutting-edge technologies closer to real-world applications by enabling next-generation communication systems with flexible integrated optical circuits at the chip level.

1.1 Objectives

Given the numerous applications of integrated microwave photonics that have emerged in recent years, this thesis aims to push the boundaries of optical integrated circuits by developing innovative architectures and configurations. By combining the powerful processing capabilities of microwave photonics with the flexibility of programmable integrated circuits, we strive to enhance the performance and adaptability of modern communication standards. In particular, we focus on two key pillars of next-generation telecommunications: beam steering and massive MIMO signal processing. Our objective is to design high-speed, energy-efficient systems that can significantly improve the efficiency and scalability of future communication networks. Specifically, the main two objectives are:

- Develop an integrated photonic chip featuring a novel beamforming network architecture based on true-time delays, aiming to achieve the optimal balance between bandwidth, resolution, and ease of implementation. Secondary objectives include put into practice calibration techniques for delay lines without external monitoring, optimizing design space and electrical power, and carrying out complementary applications for optical systems based on true-time delays.
- Develop an integrated analog photonic MIMO processor with interferometric meshes as the core engine. This includes studying the latest integrated optical mesh technologies, formulating integration theories for their combination with high-speed wireless signal processing, and creating a scheme for implementation.

Regarding the methodology, the design process begins with semi-analytical modeling to establish the initial device concept. This concept is then validated through numerical simulations to assess its performance. Once the desired results are achieved, the system layout is designed and finalized for fabrication. All designs in this work were fabricated on a SOI platform by *Advanced Micro Foundry* in Singapore, chosen for its cost-effectiveness, high refractive index contrast, and suitability for complex, compact designs, while also allowing the fabrication of modulators and photodetectors through doping. Finally, the institute's laboratories and equipment are utilized for experimental measurements and validation of the work.

1.2 Structure

This thesis is submitted in fulfillment of the requirements for the degree of *Doctor of Philosophy* in Telecommunication Engineering under the doctorate program at the *Universitat Politècnica de València*. The research was conducted within the Photonics Research Labs group from the *Instituto de Telecomunicaciones y Aplicaciones Multimedia* (iTEAM), under the supervision of Prof. José Capmany Francoy.

The structure of this work is presented as a collection of papers published in peer-reviewed scientific journals. These articles showcase the author's original contributions and key findings of the thesis, along with the relevant state of the art. The two main objectives of this thesis will guide its structure, dividing it into two sections that address the key functionalities of a next-generation base station: a photonic beamforming network system and an analog photonic MIMO processor. This document is subsequently structured in the following chapters:

- Chapter 1: *Introduction* places this work within the context of photonics, emphasizing the utility of microwave photonics—and particularly integrated photonics—in enhancing the performance of future telecommunication systems. It also outlines the main goals of this dissertation and the methods used to achieve them.
- Chapter 2: *Antennas and Optical Beamforming Networks* covers the fundamentals of antenna arrays, from classical electronics to their implementation using microwave photonics. It reviews the latest advancements in integrated optics, highlighting the benefits, challenges, and areas for improvement.
- Chapter 3: *Integrated True-Time Delay Line Processor* includes our original work on the theoretical design and experimental demonstration of a true-time delay engine that achieves unprecedented levels of resolution and bandwidth when used as a beamforming network, and it can also be applied to other processing tasks such as filtering and signal shaping. The work is presented in Paper A, Paper B and Paper C that cover all the applications and the characterization technique.
- Chapter 4: *Analog Photonic MIMO Processor*. This chapter covers the fundamentals of integrated interferometric meshes and analog photonic vector/matrix multipliers—technologies primarily used in machine learning and neuromorphic computing, but also adaptable to telecommunications processing systems, as discussed in Paper D.
- Chapter 5: *Conclusions and Future work* outlines the main findings of this thesis and evaluates the extent to which the objectives have been met. Additionally, it suggests further steps for future work related to the topics discussed.

Chapter 2

Antennas and Optical Beamforming Networks

An antenna is a transducer that converts electromagnetic energy, in the form of electrical signals or electromagnetic waves, between a guided medium (such as a cable or a waveguide) and free space, and vice versa. Antennas can be fabricated in various shapes and from different materials, which can alter their aiming direction and efficiency. They can also be realized in different sizes, significantly influencing the frequency of radiation. Even though antennas can be designed to transmit and receive radio waves equally in all horizontal directions (omnidirectional antennas), the ideal objective is to transmit radio waves with the maximum possible energy in a specific direction. This requires antennas with high directivity and gain, enabling them to concentrate power effectively in the desired direction.

The directivity, D , of an antenna is a measure of how concentrated the radiated power is in a particular direction compared to an isotropic radiator, which is an idealized antenna that radiates power uniformly in all directions, forming a constant spherical radiation pattern centered at the antenna. It is defined as the ratio between the power density radiates in an specific direction (θ, φ) , and the power density radiated by the isotropic radiator, radiating the same amount of total radiated power, P_{rad} ,

$$D(\theta, \phi) = \frac{\frac{P_{\text{rad}}}{d\Omega}(\theta, \phi)}{\frac{P_{\text{rad}}}{4\pi}}. \quad (2.1)$$

This parameter is closely related to the antenna's gain, G , a more realistic measure of how effectively it converts input power, P_{in} , into radiated power in a specified direction. Gain combines the antenna's directivity with its radiation efficiency, η ,

$$G(\theta, \phi) = \frac{\frac{P_{\text{rad}}}{d\Omega}(\theta, \phi)}{\frac{P_{\text{in}}}{4\pi}} = \eta D(\theta, \phi). \quad (2.2)$$

High gain and directivity are highly desired attributes in antennas; however, designing and implementing such antennas is often challenging and highly specialized, especially for small antennas used in modern mobile telecommunication systems. Additionally, there is

a strong desire to be able to dynamically change the direction of maximum transmitted power, further complicating the design process.

2.1 Antenna Arrays

A method to enhance the gain of an antenna involves arranging multiple simple antennas. An antenna array consists of several elementary antenna elements in a specific spatial configuration and orientation, driven by a network known as a beamformer. This network provides a complex excitation to all the N -elements, $C_n = |C_n|e^{i\psi_n}$, with $n = 1, 2, \dots, N$. The total radiated field of the array depending on the vector electric field, $f(\theta, \phi)$, expressed in the far-field is:

$$\mathbf{E}(r, \theta, \phi) = f(\theta, \phi) \frac{e^{ik_0 r}}{4\pi r} \sum_{n=1}^N C_n e^{ik_0 \hat{\mathbf{r}} \cdot \mathbf{r}_n}, \quad (2.3)$$

where $k_0 = 2\pi/\lambda_0$ is the propagation constant and λ_0 is the wavelength in the medium. It depends on the direction from the observer (where the field is being evaluated) to the origin of the coordinate system in use, $\hat{\mathbf{r}}$, and the position of each element, $\mathbf{r}_n$, with respect to the origin [21], [22].

Directivity and gain are proportional to the radiated power per unit of solid angle, $P_{\text{rad}}/d\Omega$, which is proportional to $(4\pi r |\mathbf{E}|)^2$, thus we can write the directivity of the array proportional to the field in Eq.2.3:

$$\begin{aligned} D(\theta, \phi) &\propto |f(\theta, \phi)|^2 \left| \sum_{n=1}^N C_n e^{ik_0 \hat{\mathbf{r}} \cdot \mathbf{r}_n} \right|^2 \\ &\propto |f(\theta, \phi)|^2 |F(\theta, \phi)|^2, \end{aligned} \quad (2.4)$$

the directivity is proportional to the vector field pattern of the individual antennas and the array factor, $F(\theta, \phi)$, defined as the sum of the contribution from each antenna to the total field,

$$F(\theta, \phi) = \sum_{n=1}^N C_n e^{ik_0 \hat{\mathbf{r}} \cdot \mathbf{r}_n}. \quad (2.5)$$

This is a crucial result because it demonstrate that it is possible to alter the radiation characteristics of an antenna array simply by changing its spacial distribution ($\hat{\mathbf{r}} \cdot \mathbf{r}_n$) of the electrical feeding signal (C_n). Moreover, in most practical scenarios, the array factor primarily determines the characteristics of the array, enhancing the performance of antennas with low directivity, such as dipoles or patches. For this reason, the main focus in the design of multi-antenna systems is not on the quality of individual antennas, but on the behavior of the array as a whole.

2.2 Linear Array

The easiest distribution of an arrangement of N antennas is all of them one after other in the same line. This one-dimensional distribution is straightforward to analyze mathematically because, due to its cylindrical symmetry, it depends on only one angular parameter.

Since the origin of coordinates is located at the first antenna, it is more convenient to perform a translation in the summation so that the upper limit is $N - 1$ and the elements are numbered as $n = 0, 1, 2, \dots, N - 1$. Due to the symmetry of the array's position, the array function exhibits cylindrical geometry and can be expressed solely as a function of the angle between the observer and a perpendicular line to the x-axis.

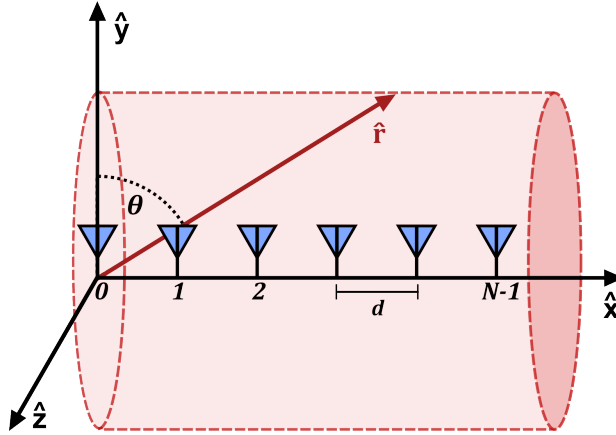


Figure 2.1: Schematic of a linear antenna array consisting of N equally spaced elements.

The position vector of the observer and the position of each element are as follows:

$$\hat{\mathbf{r}} = \sin \theta \hat{\mathbf{x}} + \cos \theta \hat{\mathbf{y}}, \quad (2.6)$$

$$\mathbf{r}_n = nd\hat{\mathbf{x}}, \quad (2.7)$$

and the array factor of a linear array can be written as follows:

$$F(\theta) = \sum_{n=0}^{N-1} |C_n| e^{i(k_0 nd \sin \theta + \psi_n)}. \quad (2.8)$$

Given a specific pointing angle, θ_0 , there is a combination of phases of the complex excitations, ψ_n , that maximizes the value of the array factor in that direction, thereby directing the antennas to aim there. The condition for this behavior on the n -th antenna is the following:

$$\psi_n = -k_0 nd \sin \theta_0, \quad (2.9)$$

with that the array factor becomes:

$$F(\theta) = \sum_{n=0}^{N-1} |C_n| e^{ik_0 nd(\sin \theta - \sin \theta_0)}. \quad (2.10)$$

Considering uniform amplitude excitation, $|C_n| = 1$, we can express the array factor as a geometrical series:

$$F(\theta) = e^{ik_0 \frac{N-1}{2} d(\sin \theta - \sin \theta_0)} \frac{\sin \left[\frac{k_0 Nd}{2} (\sin \theta - \sin \theta_0) \right]}{\sin \left[\frac{k_0 d}{2} (\sin \theta - \sin \theta_0) \right]}. \quad (2.11)$$

What truly interests us about the array factor function is its absolute value. To examine the characteristics of this function more effectively, we perform a variable change to better analyze its domain as follows:

$$|F(x)| = \left| \frac{\sin \left[\frac{N}{2}x \right]}{\sin \left[\frac{1}{2}x \right]} \right|. \quad (2.12)$$

It is easy to see that because of the nature of the sine function, Eq.(2.12) is a periodic function. It exhibits a maximum (main lobe) equal to N for all the values of $x = 2\pi m$ where $m = 0, \pm 1, \pm 2, \dots$ Figure 2.2 shows how increasing the number of antennas enhance the directivity of the array, making it more selective and less wide. The array also has secondary lobes on both sides of each main lobe. The first side lobe, which is the highest, has a height of approximately $2/(3\pi)$. This periodicity is crucial to consider during the design process. If not accounted for, certain configurations could result in two maxima of the array function, leading to two pointing directions in the antenna array, as shown in Fig.2.2.

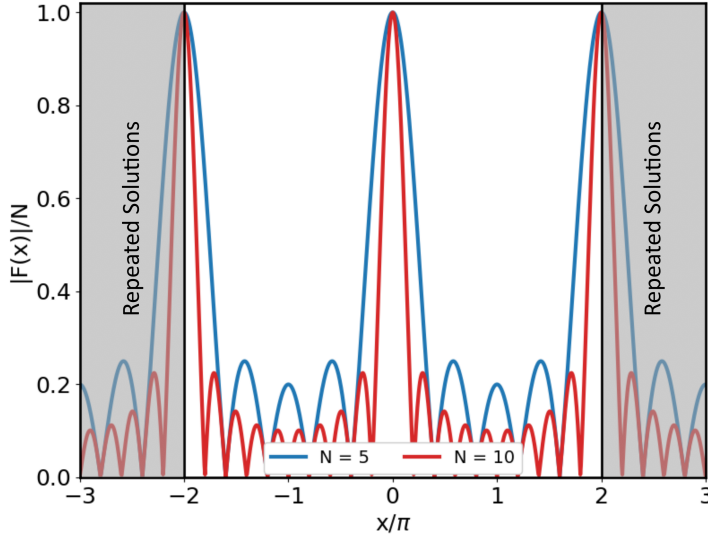


Figure 2.2: Normalized array factor as a function of x for 5 and 10 elements. Gray shading indicates the regions of the function's domain where multiple solutions appear, which are repetitions of the main solution.

To prevent this, we will restrict the domain to $x \in (-2\pi, 2\pi)$,

$$-2\pi < k_0 d (\sin \theta - \sin \theta_0) < 2\pi. \quad (2.13)$$

Given that the minimum and maximum values for the sines of the angles are ± 1 the possible values of the parenthesis range from -2 to 2 . With this in mind, we can split the previous equation into two different equations as a function of the minimum wavelength of radiation:

$$-2\pi < -\frac{4\pi}{\lambda_0} d \quad \text{and} \quad \frac{4\pi}{\lambda_0} d < 2\pi, \quad (2.14)$$

this way, we recover the constraint on the separation between antennas that ensures there is only one lobe of radiation in the array for all $\theta \in [-\pi, \pi]$,

$$d < \frac{\lambda_0}{2}. \quad (2.15)$$

This expression indicates that for a linear array of antennas, maintaining the separation between them below half a wavelength is sufficient to ensure the absence of grating lobes for any scanning angle.

2.2.1 Phase Shifters vs True-Time delays

From the previous study, we can conclude that properly exciting each antenna element is crucial for the array's performance. For a desired pointing angle, the necessary phases are calculated using Eq.(2.9). In a practical implementation, these phases correspond to real phase shifts applied to the transmitted microwave signal and the system responsible for feeding each antenna with the appropriate amplitude and phase shift is known as a beamforming network or beamformer. There are two methods for achieving this phase shift: constant-phase phase shifters and constant-delay phase shifters [23]. Each type results in a different frequency response for the beamformer, affecting its ability to process wideband signals. The wideband performance of an array is crucial in modern telecommunications due to the increasing data rates that require broader operating bandwidths. Additionally, wideband capability allows antennas to operate across multiple frequency bands, rather than being limited to a single central frequency, providing greater flexibility and efficiency in signal transmission.

Constant-Phase Phase Shifters

In the case of constant-phase phase shifters, these elements are designed with a central frequency as a reference. They operate by altering the propagation constant, resulting in the desired phase shift at this central radio frequency. If we consider a constant-phase phase shifter designed for a central frequency f_0 we can re-express Eq.(2.9) and Eq.(2.10) depending on the frequency of emission of the array f ,

$$\psi_n = -\frac{2\pi f_0}{c}nd \sin \theta_0, \quad (2.16)$$

$$F(\theta) = \sum_{n=0}^{N-1} |C_n| e^{i \frac{2\pi f}{c} nd (\sin \theta - \frac{f_0}{f} \sin \theta_0)}. \quad (2.17)$$

As previously discussed, the array factor reaches a maximum whenever the argument of the exponential is zero. For the case of $f = f_0$, the result is immediate and corresponds to the pointing angle for which the phase shifter was designed. However, for other radiation frequencies, the value differs:

$$\theta = \arcsin \left(\frac{f_0}{f} \sin \theta_0 \right). \quad (2.18)$$

For frequencies different from the design frequency, we observe a drift in the pointing angle, an effect known as *frequency squint* or *beam squint* [24]. Figure 2.3 depicts this

effect for various ratios between the design frequency and the actual frequency, as well as for different pointing angles, showing that as the pointing angle increases, the beam squint becomes more pronounced.

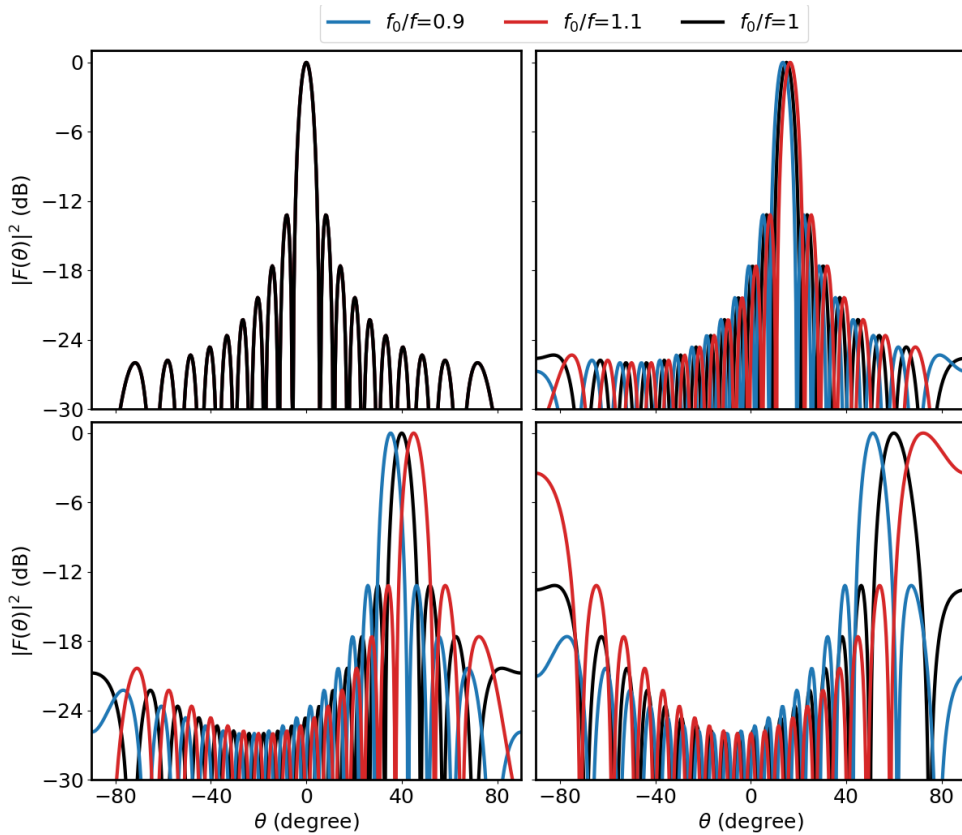


Figure 2.3: Exemplification of the beam squint effect for an array of 20 antennas for different nominal pointing directions, 0°, 15°, 40° and 60°.

Traditionally, electronic beamformers use either ferrite or ferroelectric ceramic phase shifters. These systems are significantly bandwidth-limited because the instantaneous phase shifts are characterized and designed for a single central frequency, resulting in the same phase shift being applied uniformly across all frequencies of the spectrum and causing adjacent frequencies to experience different phase shifts than intended[25], [26]. This has significant practical implications. Wideband signals generally cannot be processed using this type of phased array antennas, limiting their bandwidth to several hundreds of MHz in most commercial devices. In transmission, part of the signal is divided and sent to different pointing angles. In reception, only a portion of the signal's spectrum, corresponding to the pointing angle designed for the central frequency, is effectively processed. The rest of the signal experiences a passband filtering effect.

True-Time Delays

In this scenario, we aim to avoid artificial or instantaneous phase shifts on the signal, preferring instead for the phase shift to occur naturally through the propagation of the waves within a waveguide. The phase achieved by a delayed wave on its path through a medium of length L with a constant group index n_g , independent of frequency (i.e., without dispersion) is:

$$\psi = -\beta L = -\frac{2\pi f n_g}{c} L, \quad (2.19)$$

the phase shift is directly proportional to the frequency, ensuring the accurate rate of change in phase shift for frequencies other than the central one. Comparing this relation with Eq.(2.9) we obtain the condition of the different lengths L_n in the array:

$$L_n = \frac{nd \sin \theta_0}{n_g}. \quad (2.20)$$

After applying these lengths to the array factor summation, we achieve the ideal behavior, independent of frequency, when obtaining a zero in the exponential term,

$$\begin{aligned} F(\theta) &= \sum_{n=0}^{N-1} |C_n| e^{i \frac{2\pi f}{c} nd \sin \theta - i \beta L_n} \\ &= \sum_{n=0}^{N-1} |C_n| e^{i \frac{2\pi f}{c} nd (\sin \theta - \sin \theta_0)}. \end{aligned} \quad (2.21)$$

With true-time delay, the pointing angle remains consistent across all frequencies as long as the dispersion in the delay lengths is negligible, resulting in a broader operating bandwidth. However this solution presents a drawback: the length of a delay line cannot be changed once it has been fabricated. Most applications based on this technique switch between multiple delay line subsections with different lengths with digital control. Numerous designs of RF beamforming networks based on discrete delays have been proposed based on p-i-n diode, field-effect transistors, or microelectromechanical switches (MEMS) that can be used in these circuits to electronically switch the phase state [27]–[30]. Switching different lengths does not limit the operating bandwidth but results in a discretization of possible delays and, consequently, the number of achievable pointing angles.

When it comes to the practical implementation of antenna arrays, the options largely depend on the bandwidth required by the application. In some cases, exclusively using phase shifters is sufficient for narrow bandwidths and high angular resolution, as they are generally cheaper and simpler to implement than delay lines. On the other hand, when applications require wide bandwidths, the use of delay lines becomes imperative. Unfortunately, this often comes at the expense of angular resolution, typically resulting in only a limited number of achievable pointing angles [31]–[33]. Recently, new techniques have been developed that combine both phase shifts and true-time delays with complex processing techniques, achieving an intermediate point of bandwidths and resolution [34]–[36]. However, last application scenarios may require bandwidths reaching several tens of GHz while maintaining high resolution. This trend poses a significant challenge for the design of beamforming networks, as these requirements exceed current state-of-the-art

capabilities. Furthermore, there is an increasing demand for antenna arrays in satellite applications, where strict constraints on weight and bulk further complicate the design of phased array systems. For these reasons, researchers have begun exploring the potential of integrated photonic solutions and technology as a viable alternative for constructing beamforming networks. Integrated photonics offers several advantages over microwave transmission lines, including extremely large bandwidth, very low loss, reduced weight and bulk, reconfigurability, and immunity to interference.

2.3 Integrated Optical Beamforming Networks

Since the beginnings of microwave photonics, various methods have been explored to realize optical beamformers. One such technique involves using heterodyne detection to translate a phase shift on the optical carrier to the radio frequency signal. These techniques are highly complex, requiring single sideband modulation, filtering, and carrier reinsertion, and they suffer from beam squint due to the direct phase shift applied to the central frequency. Despite these challenges, they continue to be investigated for application in low bandwidth systems [37], [38].

Later on, alternatives based on true-time delay were explored to achieve broader bandwidths, trying to implement delays in the order of nanoseconds and picoseconds using discrete optical devices. Most of these techniques rely on the effect of various propagation velocity lines, based on highly dispersive fibers [39]–[41], fiber Bragg gratings [42], [43], or photonic crystal fiber [10] and even optical micro-combs [44], [45].

Although many of these demonstrations showed impressive results, they were never implemented in everyday communication systems. Despite improving on many aspects of classic RF alternatives, these architectures still relied on large devices such as wavelength multiplexers or fiber spools, making them bulky, expensive, and unstable, and in most cases, they did not offer real reconfigurability. Thanks to the constant development of integration technologies, complex integrated designs could be realized, opening the door to new approaches for implementing delays beyond propagation speed variation techniques.

2.3.1 Microring Resonators

One of the primary approaches for achieving true-time delay with integrated optics involves using microring resonators (MRR). Although these structures can take on an infinite number of shapes, the underlying principle remains the same: a resonant structure coupled to a waveguide, where the coupling strength, the losses and the optical length of the cavity determine the resonator's performance.

The length, L_r , is responsible of the frequency response of the ring, its periodicity and the frequencies affected by the resonance. The period is called free spectral range (FSR) and equals the inverse of the time spend by the light in completing one round trip, T_r ,

$$FSR = \frac{L_r n_g}{c} = \frac{1}{T_r}. \quad (2.22)$$

The coupling strength determines the ease of energy transmission between the waveguide and the cavity. Mathematically, this coupling strength can be represented by two

coefficients: k , the cross-coupling term and t , the self-coupling term. Here, k^2 and t^2 are power splitting ratios of the coupler, and satisfy $k^2 + t^2 = 1$. In an ideal scenario, without losses or reflections in the waveguide, and assuming continuous wave (CW) operation and matched fields, the ratio of the transmitted to the incident field in the bus waveguide can be expressed as:

$$\frac{E_t(t, \Omega)}{E_{in}} = \frac{t - \alpha e^{i\Omega}}{1 - t\alpha e^{i\Omega}}, \quad (2.23)$$

where Ω is the optical frequency normalized to the FSR and α is the single-pass amplitude transmission, which includes both propagation loss in the ring and loss in the coupler. The phase response of the ring can be extracted from this equation:

$$\Phi(t, \Omega) = \arctan\left(\frac{-\alpha \sin \Omega}{t - \alpha \cos \Omega}\right) - \arctan\left(\frac{-t\alpha \sin \Omega}{1 - t\alpha \cos \Omega}\right). \quad (2.24)$$

The group delay induced by the resonant structure can be calculated as the differentiation of the phase response with respect to frequency:

$$\frac{\tau(t, \Omega)}{T_r} = \left(\frac{\alpha^2 - t\alpha \cos \Omega}{t^2 + \alpha^2 - 2t\alpha \cos \Omega} \right) - \left(\frac{(t\alpha)^2 - t\alpha \cos \Omega}{1 + (t\alpha)^2 - 2t\alpha \cos \Omega} \right). \quad (2.25)$$

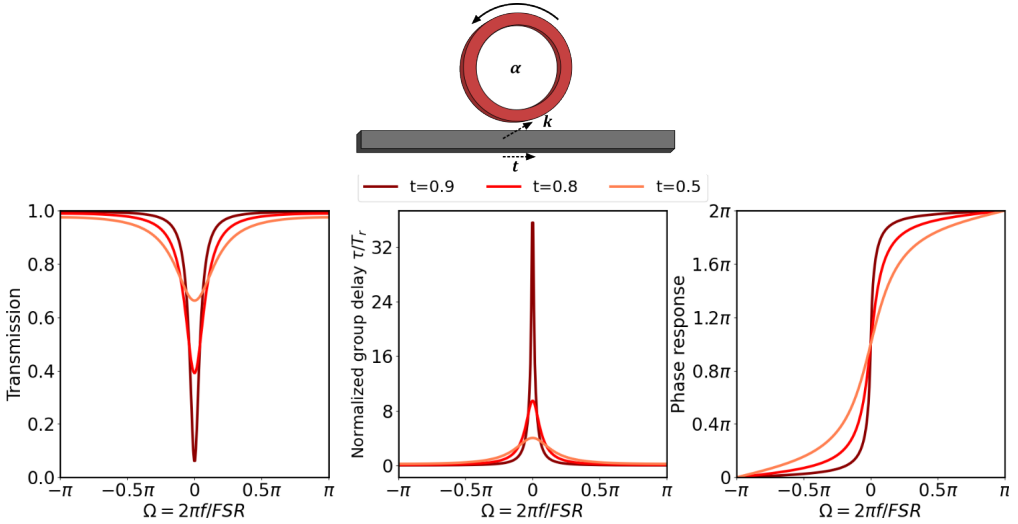


Figure 2.4: Transmission, group delay and phase response of a microring resonator for different coupling coefficients assuming $\alpha = 0.95$.

Figure 2.4 shows the response of a single MRR for different coupling coefficients. Both transmission and group delay can be continuously tuned depending on the coupling coefficient, achieving several times the value of a round trip at critical coupling, where the light is completely trapped in the resonator. It is worth noting how the ring exhibits a linear phase slope and an equivalently flat group delay, which are the desired characteristics for obtaining true-time delay. However, this behavior is restricted to frequencies extremely close to the resonance, thus it is very limited in bandwidth.

Thanks to its simple yet efficient design, it is possible to cascade several MRRs, independently adjust the central frequency and coupling of each one, and achieve a flat delay response over a wider bandwidth compared to using a single ring. This is equivalent to a phase response that maintains linearity over a larger frequency range. Figure 2.5 illustrates the response of three cascaded ring resonators. The central frequency and coupling strength of each ring have been adjusted to achieve a flat transmission and group delay, resulting in a delay exceeding 8 times that of a round trip. The linear region of the phase response has also been expanded, being three times larger than the bandwidth of a single resonator in this case.

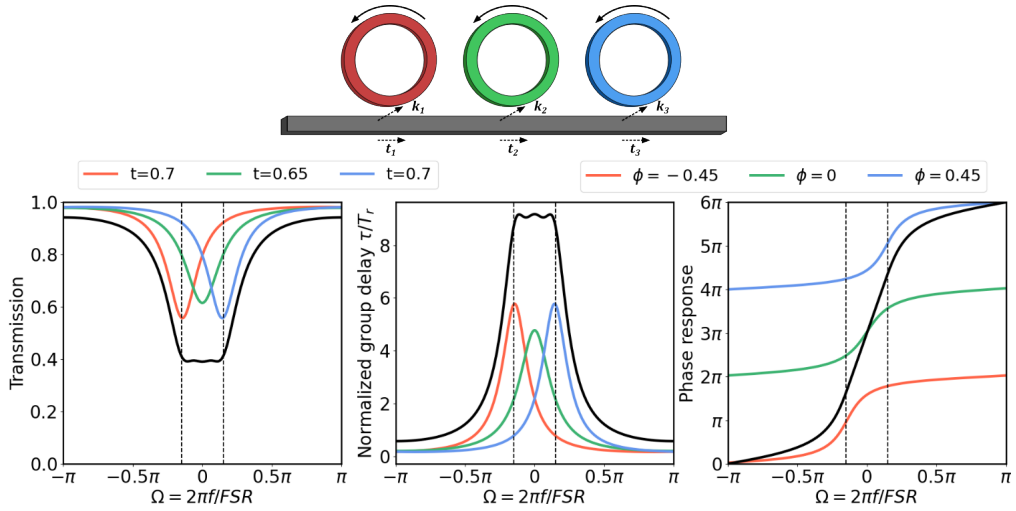


Figure 2.5: Transmission, group delay and phase response of 3 cascaded MRRs for different resonant frequencies and coupling coefficients assuming $\alpha = 0.95$.

Even though these architectures offer significant reconfigurability and continuous delays, they have some drawbacks. The first issue is their lack of flexibility for continuous tuning, owing to the trade-off between group delay and transmission. For instance, as shown in Figure 2.5, achieving the desired delay results in partial signal loss during transmission. The second problem is related with the real effective delay of these devices. The actual bandwidth for a flat delay response is determined by the number and length of the microring resonators; a larger number results in a wider bandwidth. However, working with a large number of rings is challenging because their resonances and couplings require precise positioning. This precision is highly sensitive to thermal crosstalk, a common issue in structures with many elements. This disadvantage is particularly dominant in schemes with direct detection and modulation. In these cases, the bandwidth of the MRRs should be at least twice the separation from the first frequency of the low sideband to the highest frequency of the upper sideband, expressed as $BW_\tau = 2f_{RF} + BW_{RF}$. With this type of detection, typical bandwidths do not exceed 10 GHz [46], [47].

Due to the difficulty of achieving such bandwidths, an alternative approach relies on using single sideband and carrier filtering with reinjection, similar to the procedure of the first beamformers implemented in the optical domain. In this case, the narrower bandwidth of the cascaded rings affects only one of the sidebands of the optical signal. This

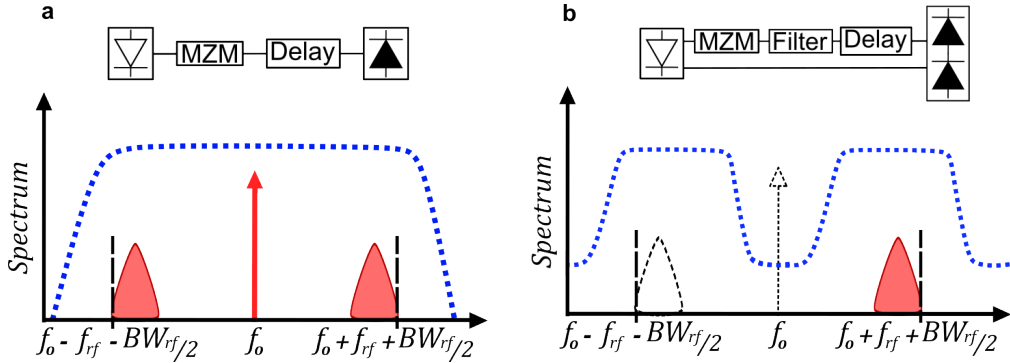


Figure 2.6: Schemes from the two main approaches for implementing true-time delays using MRR, depending on the group delay bandwidth. **a** Case where the bandwidth is enough compared to the bandwidth of the up-converted signal. The architecture is based on direct modulation and detection. **b** Case where the bandwidth of the MRRs is narrower. It is required single-sideband and carrier suppression along with coherent optical detection. MZM, Mach-Zehnder Modulator.

action applies the desired linear phase shift to the RF signal, and after reinjecting the carrier, the delay is recovered using coherent optical detection on a balanced photodetector. These complex systems pose significant challenges in implementations with more than a couple of antennas, primarily due to the intricacies involved in all stages, including filtering, reinjection, and photodetection. It is evident that simpler alternatives are necessary to achieve true-time delays with integrated optics, particularly when targeting high frequencies.

2.3.2 Delay Lines

Through the application of advanced manufacturing techniques, the concept of switchable delay lines, initially developed for classical antennas [27]–[30], was seamlessly adapted to optics.

The concept behind these architectures is both simple and robust, facilitating the translation from electronics to optics. Electronic switches based on MEMS have now been replaced by optical switches, typically using tunable Mach-Zehnder Interferometers (MZI). These are driven by either the thermo-optic effect, using embedded heaters, or the electro-optic effect, based on p-n junctions. Figure 2.7 presents habitual implementations of switched delay lines in electronic arrays and the general scheme translated to a photonic integrated circuit.

The advantage of integrated optics lies in its significantly higher group index compared to microstrip guides. This adaptation has resulted in the development of more compact and easier-to-manufacture devices, enabling the implementation of circuits with the same number of delay stages in significantly reduced space and with lower losses.

Despite the significant benefits of this architecture, such as enhanced bandwidth, thermal stability, and control, its main drawback in both electrical and optical implementations is the limited number of achievable delays or pointing angles. As we will explore in the following chapter, scaling the number of antennas exacerbates the issue, leading to large footprints for achieving few pointing angles, highlighting an intrinsic and

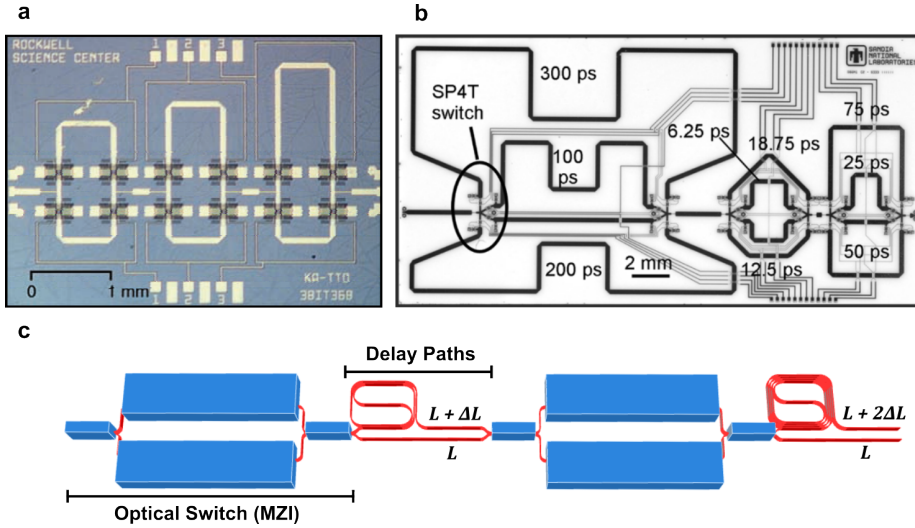


Figure 2.7: General schemes for true-time delays based on switched lines. Panels **a** and **b** show two implementations from standard electronic systems using MEMS, as described in [27] and [30]. Panel **c** shows the translation to photonic integrated circuits.

seemingly unbridgeable trade-off between bandwidth and resolution in optical beamforming networks based on MRRs or switched delays. This was one of the objectives of this thesis, and the following chapter provides a detailed explanation of a viable solution to this challenge.

Chapter 3

Integrated True-Time Delay Signal Processing

As discussed in the previous section, there are two main approaches to implementing true-time delays in integrated photonics, microring resonators and switched delay lines. This research focuses on switched delay lines because they are easier to implement, more robust against fabrication errors and thermal crosstalk, and offer higher bandwidth [48], [49]. This type of true-time delay implementation has been widely studied, with numerous experimental demonstrations across various fabrication platforms[50]–[52]. However, they all face the same challenge: achieving compact scaling of the number of delay combinations relative to the chip size or the number of active elements. This chapter presents original contributions to photonic true-time delay processors based on switched delay lines. It explores applications beyond traditional beamforming networks, leveraging the full capabilities of switched delay line architectures.

In Paper A, we present the theoretical design and experimental demonstration of a novel switched optical delay line beamformer architecture, which achieves unprecedented resolution compared to existing state-of-the-art designs. Paper B addresses the experimental characterization of switched delay lines using noninvasive methods and introduces a new procedure that reduces both power consumption and chip footprint for these structures. Finally, in Paper C, we introduce the design of a true-time delay signal processor that not only includes beamforming capabilities but also demonstrates excellent performance in signal filtering and generation.

3.1 Paper A

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Ultrabroadband high-resolution silicon RF-photonic
beamformer

Pablo Martínez-Carrasco, Tan Huy Ho, David Wessel and José Capmany

Ultrabroadband high-resolution silicon RF-photonic beamformer

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Microwave photonics aims to overcome the limitations of radiofrequency devices and systems by leveraging the unique properties of optics in terms of low loss and power consumption, broadband operation, immunity to interference and tunability. This enables versatile functions like beam steering, crucial in emerging applications such as the Internet of Things (IoT) and 5/6G networks. The main problem with current photonic beamforming architectures is that there is a tradeoff between resolution and bandwidth, which has not yet been solved. Here we propose and experimentally demonstrate a novel switched optical delay line beamformer architecture that is capable of achieving the desired maximum resolution (i.e. 2^M pointing angles for M -bit coding) and provides broadband operation simultaneously. The concept is demonstrated by means of a compact ($8 \times 3 \text{ mm}^2$) 8 (5-bit) delay line Silicon Photonic chip implementation capable of addressing 32 pointing angles and offering 20 GHz bandwidth operation.

3.1.1 Introduction

With the emergence of new communication paradigms like the Internet of Things (IoT) and 5/6G networks, the need for high-speed and high-capacity wireless communication has increased exponentially [1], [2]. In this scenario, microwave-phased array antennas (PAAs) have become an indispensable element of contemporary radar and wireless communication systems [3]. Electronic phased array antennas operate by manipulating the phases of signals transmitted from each antenna element in the array. By doing so, the array can focus its radiation in a specific direction, resulting in high directivity and reduced interference from other directions [4]. However, this use of phase shifts can lead to the beam squint effect, where changes in the microwave frequency cause a deviation in the pointing angle, limiting the operating bandwidth to less than a few hundred megahertz [5]–[7].

Microwave photonics technology has enabled the introduction of optical true-time delay line (OTTDL) technology into phased array PPAs [8] bringing key advantages in terms of low signal loss and immunity to electromagnetic interference [9] and more importantly, avoiding the appearance of the beam squint effect and extending the operating bandwidths of PAAs [10]. Numerous fiber-based architectures have been proposed for implementing OTTDLs, either based on Bragg gratings [11], [12], highly dispersive fibers [13], [14], or wavelength multiplexing [15], [16]. While these approaches have yielded impressive performance, they typically are bulky and require precise cleaving of their length to achieve accurate delays. On the other hand, on-chip integrated microwave photonics

opens new possibilities with a reduction in space weight and power consumption (SWAP) [17]–[19]. In this area, there have also been multiple OTTDL proposals based on different architectures and technologies which can be grouped into two main categories. PAA architectures based on single and cascaded microring resonators (MRRs), [20]–[22] provide continuous tuning of the time delays, resulting in high angular resolution, but face bandwidth limitations arising from the resonant behavior of their delays. Furthermore, they need complex stabilization circuits to avoid resonance drifting due to temperature fluctuations, which are greatly increased when the number of antennas in the array escalates. PAAs based on switched M -bit delay lines can offer significant delays with wide bandwidth and low sensitivity to temperature fluctuations [23]–[25] at the expense of discrete tuning of the time delays. There is thus a tradeoff between resolution and bandwidth which has led to a recent proposal [26] to substantially increase the resolution of switched delay line PAAs by means of a parallel configuration of N OTTDLs, where the basic delay increment is not constant in each OTTDL, instead it increases by one unit. This configuration is capable of achieving maximum resolution (i.e. 2^M pointing angles) but surprisingly suffers from beam squint for negative pointing angles due to the impossibility of providing negative incremental delays. Unfortunately, this prevents broadband operation. A similar architecture based on MRRs has been recently proposed [27], but suffers from the same operational problems, contrary to what is considered a feature of using true time delay.

Here we propose and experimentally demonstrate an architecture that is capable of achieving the desired resolution (i.e. 2^M pointing angles) providing broadband operation at the same time. This configuration is based on the inclusion of a delay equalization stage prior to the switched delay line array to avoid the need for negative incremental delays to implement negative pointing angles. We demonstrate the successful operation in a silicon photonics chip for a record spectral range spanning from 10 to 30 GHz (which is almost 3 orders of magnitude larger than the bandwidth reported for electronic beamformers) and provide the roadmap to enlarge this range to 50 GHz and beyond. Furthermore, we show via Monte Carlo simulation that this configuration is scalable and robust against design and fabrication deviations.

3.1.2 Results

Limitations of current switched delay line beamformers

Switched OTTDLs have been extensively studied and utilized in the literature due to their simple design and operation, which provides one of the most robust methods of obtaining optical delays. Figure 1(a) shows the simplest design of one of these delay lines, which in general comprises M delay stages or "bits". Each bit unit is implemented by means of a tunable switch and two optical paths with different lengths: the short path is constant for all delay stages, while the longer path increases its length with the number of bits as $2^{M-1}U$ where U represents the path imbalance of the first stage. The tunable switch is implemented by means of a balanced Mach-Zehnder Interferometer (MZI), composed of two 3-dB directional couplers and two parallel waveguides loaded with phase shifters as shown in the inset of Figure 1 (a). A traditional beamformer based on switched OTTDLs is implemented by combining in parallel N OTTDLs identical to the one shown in Fig. 1(b). Here, each of the identical OTTDLs feeds an independent radiating element. The

number of addressable pointing angles is given by (see "Methods"):

$$N_a = 2 \left\lfloor \frac{2^M - 1}{N - 1} \right\rfloor + 1. \quad (3.1)$$

The parenthesis on the previous equation denotes the integer part of a real number. It is immediate to observe that this figure (closely related to the beamformer resolution) scales quite poorly with the number of delay stages and active elements, due to its inverse dependence on the number of antennas. Ideally one would want the number of possible pointing angles to be independent of N and as close as possible to 2^M .

Zhu and co-workers [26] have recently proposed a novel architecture shown in Fig. 3.1(c), which in essence keeps the parallel OTTDL configuration but where the value of the basic switched delay τ is changed by one unit in each adjacent OTTDL (U for line 1, $2U$ for line 2, $3U$ for line 3 and so on). This configuration brings, in principle, the advantage of enabling 2^M pointing angles as desired providing broadband (i.e. squint-free) operation. In [26] the authors reported a 5-bit $N = 8$ microwave photonic beamformer on silicon a silicon photonics chip and demonstrated the capability of addressing 32 angles from -75.51° to 75.64° at 16 GHz (see Fig. 8 (c), and a reproduction here in the center upper part of Fig. 3.1(d)). Although it is claimed that this configuration is broadband (from 8 to 18 GHz) we found that beam-squint is produced in the negative pointing angles even in a narrow frequency region (15.8 GHz-16.1 GHz) around this frequency as shown in Fig 3.1(d). Moreover, as the operating frequency is decreased to lower values (i.e. 14, 12, and 10 GHz), the number of negative pointing angles is reduced as well, as shown in Fig 3.1(e). This striking result is against the common belief that OTTDL-based beamformers should be beam squint-free but can be explained by an analysis of the Array factor of the beamformer (see "Methods" and Supplementary Note 1). The conclusion is that the scheme proposed in [26] is not able to provide broadband operation.

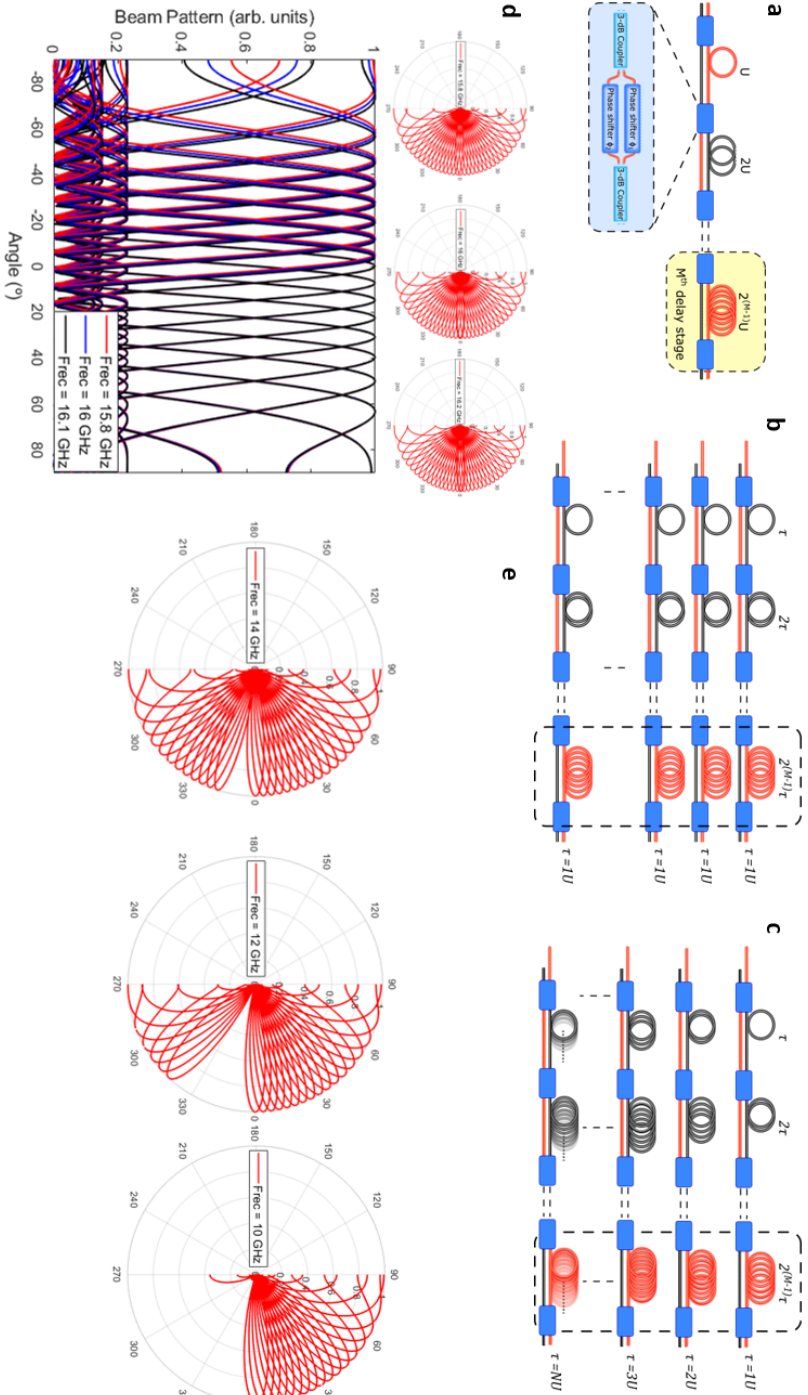


Figure 3.1: Current beamforming network architectures based on OTTDLs. **a** Scheme of switched OTTDL with M delay stages. **b** Classical beamformer based on N identical switched OTTDLs. **c** Beamformer composed of N OTTDLs with increasing positive delays [26]. The unit delay value increases one time unit U from one line to another (i.e., from $1U$ to NU). **d** Beam patterns generated by the architecture of ref. [26], for 15.8 GHz (red), 16GHz (blue), and 16.1 GHz (black). In polar coordinates, a superposition effect becomes evident near the broadside region. Upon closer inspection of half of the total angles, the beam squint effect appears for negative pointing angles. **e** Beam patterns in the architecture of ref. [26], for 14, 12, and 10 GHz. In addition to the beam squint effect observed for negative angles the same angles begin to disappear for lower frequencies

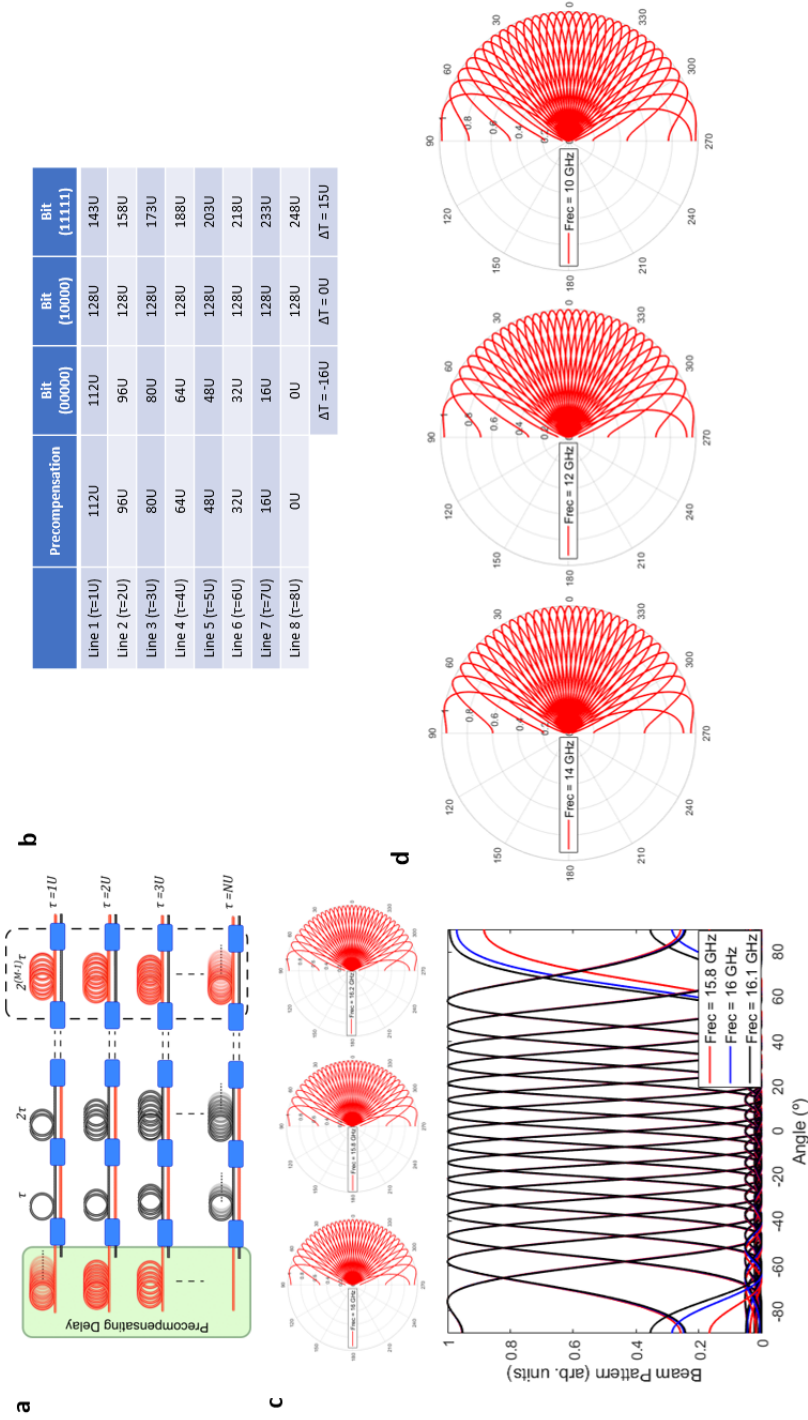


Figure 3.2: Novel Beamforming network architecture with equalization. **a** Layout of the architecture proposed in this work, emphasizing the pre-compensating/equalizing delay line stage component (highlighted in green). **b** Design table of the pre-compensation delay line stage for the unequalized 8 OTTDL beamformer and final overall output values for three different bit words, corresponding to the largest negative (00000), broadside (10000), and largest positive (11111) angle. **c** Beam patterns in the new architecture for 15.8 GHz (red), 16GHz (blue), and 16.1 GHz (black). Beam squint for the negative angles has disappeared. **d** Beam patterns in the new architecture for 14, 12, and 10 GHz. All the pointing angles appear regardless of the operating frequency.

Novel Beamformer architecture

The main reason behind the beam squint is that the configuration reported in [26] is not capable of providing negative incremental delays, which are essential to steer negative pointing angles independently of the frequency. These pointing angles, when achieved are due to phase shift and not true time delay conditions. A solution to overcome this limitation is to introduce a delay equalization stage prior to the switched delay line beamformer as shown in Fig. 3.2(a). This equalization stage must be designed so the bit word corresponding to the broadside radiation angle in the beamformer is rotated to the lowest negative radiation angle. In this way, both positive and negative pointing angles can be achieved using positive incremental time delays only and therefore beam squint-free operation is achieved.

The equalization delay design aims to transform the incremental delay range of the beamformer from $[0, (2^M - 1)U]$ to $[-2^{M-1}U, (2^{M-1} - 1)U]$. For line n the value of the equalization delay must be therefore:

$$\epsilon_n = (N - n)2^{M-1}U. \quad (3.2)$$

The incremental delays between antennas are then:

$$\Delta T = T_n + \epsilon_n - T_{n-1} - \epsilon_{n-1} = BU - 2^{M-1}U, \quad (3.3)$$

where T_n and ϵ_n represent the selected delay and equalization delay for the line n . The value of the selected delay depends on the chosen bit-word B (from $[0, (2^M - 1)]$) and the position of the line as $T_n = nBU$.

And the array factor transforms to:

$$|F(\theta)| = \frac{\sin \left[\pi N f \left(\frac{d}{c} \sin(\theta) - (B - 2^{M-1})U \right) \right]}{N \sin \left[\pi f \left(\frac{d}{c} \sin(\theta) - (B - 2^{M-1})U \right) \right]}, \quad (3.4)$$

where d is the space between radiating elements in the array, f is the RF frequency of the signal and c is the speed of light. From Eq.(3.4), we can compute the highest value for the basic unit delay that guarantees frequency-independent pointing angles. This is obtained by forcing the broadside condition ($B = 0$) in the unequalized beamformer and proceeding as shown in “Methods”:

$$\frac{d}{c} \sin(\theta_{min}) + 2M - 1U = 0. \quad (3.5)$$

From Eq.(3.5), the maximum value allowed for U given a negative minimum angle θ_{min} is $2^{1-M}(d/c) \sin \theta_{min}$ and it can be readily checked that for the maximum positive pointing angle ($B = 2^M - 1$):

$$\sin(\theta_{max}) = \left[1 - \frac{1}{2^{M-1}} \right] \sin(\theta_{min}) < 1 \longrightarrow \theta_{max} < \pi/2. \quad (3.6)$$

Hence, all pointing angles fulfill the beam squint-free condition. Figure 3.2b provides the data for the design of the equalization stage for the beamformer reported in [26] using Eq.(3.2) and the output delay results for each of the 8 OTTDLs in three relevant cases corresponding to the lowest negative pointing angle, broadside and the highest positive pointing angle (see Supplementary Note 2 for the complete table of values for

the 32 pointing angles). Figure 3.2c shows the beam patterns for the same frequency range frequency region (15.8–16.1 GHz) as in Fig. 3.1d. Note that the beam squint has disappeared. Furthermore, Fig. 3.2d shows that as the operating frequency is decreased to lower values (i.e., 14, 12, and 10 GHz), the number of negative pointing angles does not change and their directions remain unaltered. These results confirm the simultaneous achievement of the maximum number of pointing angles 2^M and broadband squint-free operation.

Beamformer chip design and fabrication

The beamforming network architecture has been demonstrated experimentally using a design with 8 OTTDL and 5 delay stages per line. For the demonstration, the equalization delay was added as an additional delay stage at the front end of each line. The design was made keeping in mind a PAA with a maximum operating RF frequency of $f_{max} = 30$ GHz and a spacing between antennas of 5mm, following the condition of single lobe of radiation $d \leq c = 2/f_{max}$. Based on the general design equation from Eq.(3.5), we calculated the value of basic unit delay setting a pointing angle range between -60° to 60° . For a value of $U = 0.92$ ps, we found the perfect balance between resolution and angle aperture while avoiding secondary lobes with too much intensity, resulting in an expected range from -61.87° to 55.7° .

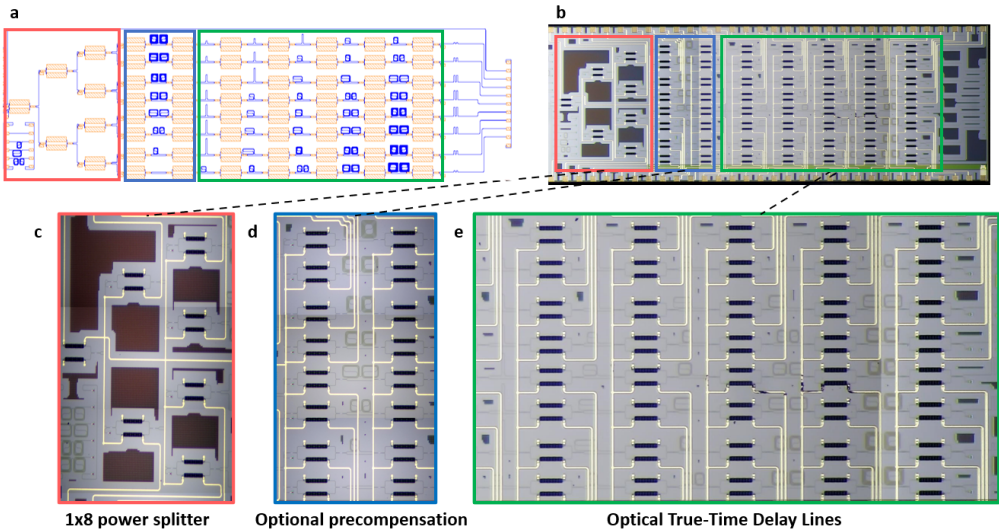


Figure 3.3: Layout and micrographs of the silicon photonics beamformer chip. **a** GDS layout with the optical elements from the 1×8 equalized beamformer. **b** Micrograph of the silicon photonics chip, including electrical connections ($8\text{mm} \times 3\text{mm}$). The three different parts are highlighted in color boxes. **c** Zoom of the 1×8 tunable splitter tree. **d** Zoom of the optional pre-compensating delay line unit as an additional bit. **e** Zoom of the eight 5-bit OTTDLs stage.

The photonic integrated circuit (PIC) was manufactured in SOI technology by Advanced Micro Foundry (see “Methods”). Typical manufacturing figures from this foundry are: waveguide loss 1.1 dB/cm, MMI insertion losses 0.15 dB and grating coupler insertion losses 3.5 dB. The total insertion losses measured for the longest line configuration on the

chip are 21.9 dB, taking into account 9 dB from the splitter tree and 7 dB from in/output grating couplers. All of these values for a wavelength of 1550 nm.

The phase shifter power consumption $1.35\text{mW}/\pi$, with a tuning speed around 3 kHz. Low-power phase shifters implemented using suspended waveguide and trenches are key in enabling a reduced power consumption for the 65 MZIs integrated in the PIC, Fig. 3.3a depicts the chip layout, while Fig. 3.3b showcases a micrograph of the fabricated PIC (footprint measuring $8\text{mm} \times 3\text{mm}$). The three main sections of the beamformer: (a) The 1×8 tunable splitter, (b) the delay equalization stage and the 5 bit 8 OTTDL units (offering delays ranging from 0.92 ps to 117.76 ps) are shown in more detail in Fig. 3.3c-e, respectively.

Experiments

The silicon photonic beamformer was tested and measured (see “Methods”) using the setup depicted in Fig. 3.4a. A first characterization was performed to measure the phase and group delay frequency responses from 100 MHz to 30 GHz of each separate OTTDL unit for each of the 32 states that can be programmed using 5 bits. Figure 3.4b shows the results for the group delay characterization of OTTDL #1, #4, and #8 (see Supplementary Note 3 for the complete phase and group delay characterization of the 8 OTTDL in the beamformer). Here, the incremental delay is $U = 0.92$ ps for OTTDL #1, $4U = 3.68$ ps for OTTDL #4 and $8U = 7.36$ ps for OTTDL #8 as expected. Note that the group delay response shows an almost flat frequency response within the 30 GHz spectral range. We then proceeded to measure the frequency response of the delays provided by each of the 8 OTTDL units for each one of the possible 5-bit coding words and from them the beam patterns corresponding to each of the possible pointing angles. Figure 3.4c shows the simulated and measured beam patterns (at operation frequencies of 10, 20, and 30 GHz) for three pointing angles of the beamformer, corresponding to the lowest negative pointing angle, broadside and the highest positive pointing angle, respectively. Insets depict for each case the delays of each of the OTTDLs. As it can be noted, the beam patterns are free from beam squint in a 20 GHz spectral range, which is over twice the highest value ever reported and furthermore, the number of available pointing angles is $2^5 = 32$, that is, the maximum possible for a 5-bit beamformer.

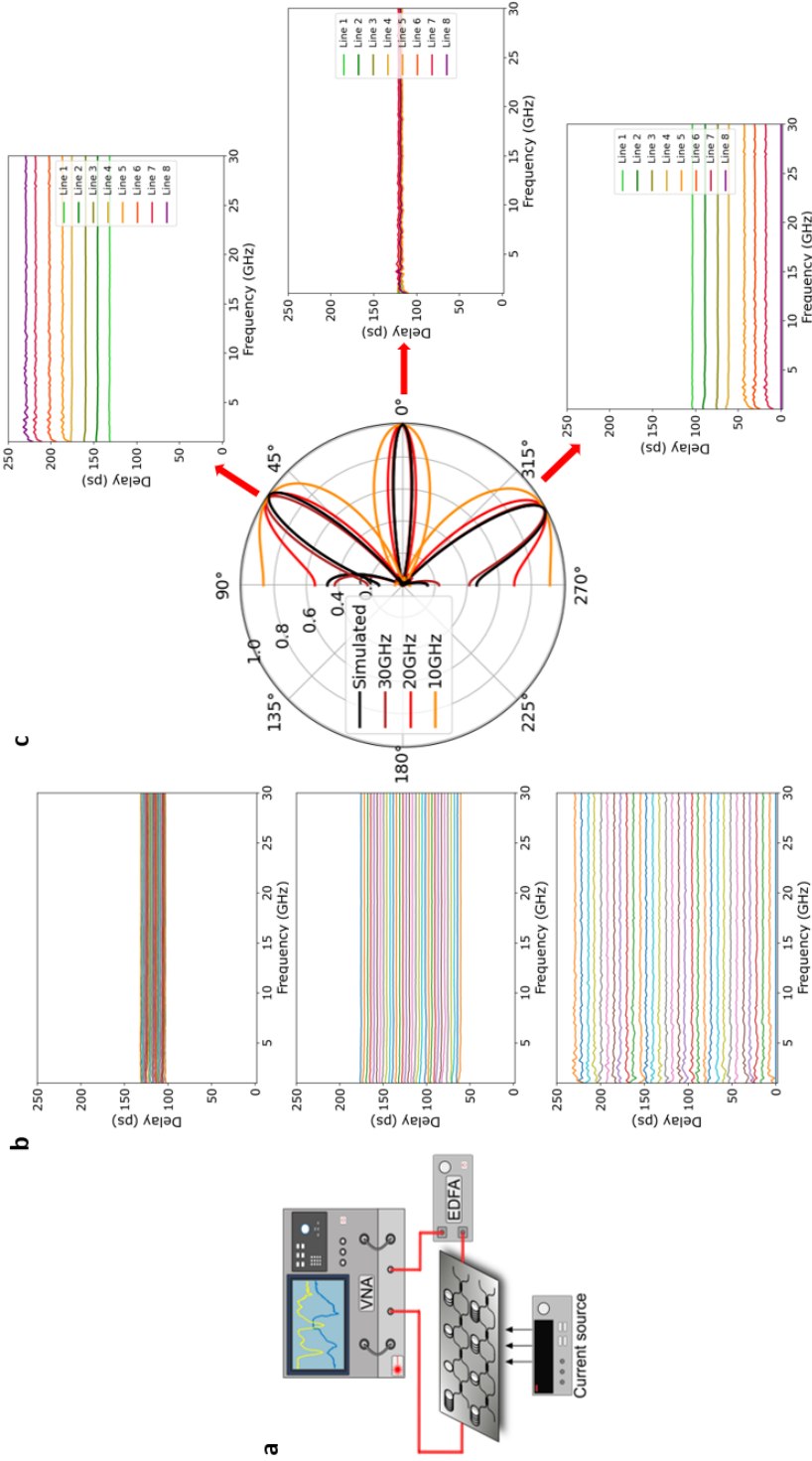


Figure 3.4: Experimental setup and results. **a** Experimental setup for the frequency response characterization of the Silicon photonics beamformer. **b** Group delay vs frequency measurement results for the first, fourth, and eighth lines. Each one shows the 32 delay curves corresponding to the whole set of 5-bit configurations. **c** Calculated beam patterns vs three different frequencies (10, 20, and 30 GHz) for the largest negative (00000), broadside (10000), and largest positive (11111). Insets correspond to the group delay frequency responses of each pointing angle configuration (for full results of the 32 possible pointing angles, see Supplementary Note 4).

3.1.3 Discussion

The proposed architecture provides a long-sought solution to the problem of achieving broadband operation and the highest possible number of pointing angles. It can be implemented using a compact layout, and there are several approaches to scale it in terms of OTTDL units and therefore of radiating antennas (i.e., from 1×8 to 1×16 or 1×32). A strategy to overcome the input power splitting loss in the beamformer is to use a Semiconductor Optical Amplifier (SOA) stage (preamplification configuration). InP SOA devices can routinely achieve gains in excess of 15 dB when incorporated into Silicon chips via micro-transfer printing. The number of bits per line can also be extended (i.e., from 5 to 7 or 8) leveraging the SOA gain but also by using novel low-loss building block designs in Silicon that can achieve propagation losses below 0.25 dB/cm with $2\text{-}\mu\text{m}$ waveguides [28], [29] and reducing MZI insertion losses to 0.22 dB [30].

Another benefit of this architecture is its low-power consumption. For instance, for the design reported here, the average power consumption is 79mW (see details in Supplementary Note 4). There are, however, several limitations that need to be addressed. The first one is related to the expected power imbalance due to the different compensating lengths used for the OTTDLs. The propagation losses of each line depend on the number n of the OTTDL and the pointing angle configuration chosen for the array:

$$Losses_n = e^{-\frac{(\alpha/20)}{n_g} \ln(10) [(N-n)2^{M-1} + nB]} = e^{-\gamma [(N-n)2^{M-1} + nB]}, \quad (3.7)$$

where α is the propagation loss coefficient in terms of optical power of the technology and n_g is the silicon waveguide group index. To facilitate the calculations, the constants have been grouped into a single constant, γ . This imbalance can act as a source of unwanted tapering, deviating the beam pattern from the desired configuration. To evaluate its impact, we can approximate the array factor by the field radiation pattern from a continuous apodized aperture of length N as a function of an angle p :

$$p = f \left[\frac{d}{c} \sin(\theta) + (B - 2^{M-1}) U \right], \quad (3.8)$$

which is given by:

$$\begin{aligned} |F(p)|^2 &= \left| \frac{1}{N} \int_{1/2}^{N+1/2} e^{-\gamma [(N-n)2^{M-1} + nB]} e^{-i2\pi p x} dx \right|^2 = \\ &= \frac{e^{-\gamma [B(N+1) + (N-1)2^{M-1}]} \left[4 \sin^2(N\pi p) - 2 + e^{-N\gamma(B-2^{M-1})} + e^{N\gamma(B-2^{M-1})} \right]}{N^2 \left[(\gamma(B-2^{M-1}))^2 + (2\pi p)^2 \right]}. \end{aligned} \quad (3.9)$$

Note that for an uniform (i.e $\gamma = 0$) field pattern, Eq.(3.9) gives the array factor of the uniform beamformer Eq.(3.4). The effect of imbalance loss can be evaluated though $\epsilon(p, \gamma) = (|F(p, \gamma)|^2 - |F(p, 0)|^2) / |F(p, 0)|^2$, which for the beamformer reported here renders negligible results for waveguide propagation loss coefficients as high as -3 dB/cm (even in this case it results in a negligible difference with respect to the uniform beam pattern in terms of shape, see Supplementary Note 5). In any case, full compensation

and even intentional apodization can be implemented by inserting p-n junction-based variable optical attenuators (VOAs) at the output of each line or by the use of a variable splitter tree.

Unavoidable fabrication errors and the evaluation of their potential impact on the beamformer sensitivity is also a main concern. By identifying potential issues early on, we can avoid costly and time consuming errors that would otherwise require significant resources to fix [31], [32]. The impact of fabrication errors can be estimated using statistical techniques and simulation. Of particular importance is the impact of common variations on the pointing angles, which arise from the asymmetric power splitting ratio on the 3-dB couplers, as well as from deviations in the phases of both arms of the MZI switches. Also relevant are the changes in the waveguide dimensions, which affect the group index in different sections of the structure [33], [34].

For the evaluation, the best option is to use a Monte Carlo analysis. The operation of each delay stage and MZI switch from each OTTDL in the beamformer can be modeled by three Gaussian random variables centered at their ideal values, having a standard deviation, σ_K , σ_ϕ and σ_{n_g} for random fluctuations around the mean of the coupling coefficient (K), phase term (ϕ) and group index, respectively. Manufacturing imperfections in parameters related to MZIs provided by different foundries and technologies are typically given by $\sigma_K = \sigma_\phi = 1\%$ [35], [36]. As far as the group index is concerned, this is highly sensitive to the specific technology used, the manufacturing technique, and any errors in waveguide dimensions. Fortunately, there are many available studies on these deviations for various integrated photonics technologies [37], providing information on manufacturing deviations from the usual group indices. For our experimental demonstration, the chip was fabricated at a standard 220 nm-thick Silicon on Insulator foundry. With these characteristics, knowing the type of guide and the manufacturing standards, we conclude that the expected group index is $n_g = 4.18$, with a variance of $\sigma_{n_g} = 1.2\%$. Figure 3.5a,b shows an example of the impact (over one realization) of manufacturing errors on the time delays for the eight different outputs of the beamforming network in the case of broadside radiation. For this case, all the delays provided by the beamformer lines should be equal. Figure 3.5a. Illustrates the impact of $\sigma_K = \sigma_\phi$, which translate into small amplitude random spike pulses. The impact of σ_{n_g} shown in Fig. 3.5b translates into slight random deviations from the mean delay. Experimental results on measured delay errors in our chip are shown in Fig. 3.5c. Despite variations of up to 3 ps in the worst cases, this error is well-tolerated when reconstructing the corresponding beampatterns, as shown in Fig. 3.5d.

To account for a more accurate description we ran a Monte Carlo process with 1000 realizations to evaluate the impact of errors on the beam patterns corresponding to the broadside and the highest positive pointing angles. The results are shown in Fig. 3.5e, f, respectively, and reveal that manufacturing errors impact performance mainly by reducing the Main Lobe to Secondary Ratio (MLSR) less than 1.5 dB, with negligible impact on the main lobe characteristics and their pointing angles, varying the beamwidth at 3 dB less than 1° . This indicates that the proposed beamforming architecture should provide excellent resilience to fabrication errors.

In conclusion, we have reported and experimentally demonstrated a novel tunable beamforming architecture based on silicon photonic integrated switched delay lines that is capable of providing simultaneous record broadband and squint-free operation and the maximum number of pointing angles 2^M , where M is the number of coding bits.

The key novel concept resides in inserting an equalizing stage to enable negative pointing angles using exclusively positive true time incremental delays. We have specifically demonstrated squint-free operation from 10 to 20 GHz of an 8-lines 5-bit OTTDL beamformer, capable of addressing 32 pointing angles ranging from -61.87° to 55.7° with a remarkably low power consumption (below 100mW). The features provided by this new architecture go remarkably beyond the current state of the art as shown in Table 3.1 and furthermore this configuration can be readily extended to both broader spectral regions (i.e., 10–50 GHz operation can be readily achieved by reducing U from 0.92 to 0.55 psec, which is possible with current Silicon photonics fabrication technologies) as well as to the higher number of radiating elements (from 8 to 16 or 32) and 2D beam steering by means of chiplet configurations. These flexible features make them a strong candidate to support massive MIMO architectures in future 6G systems.

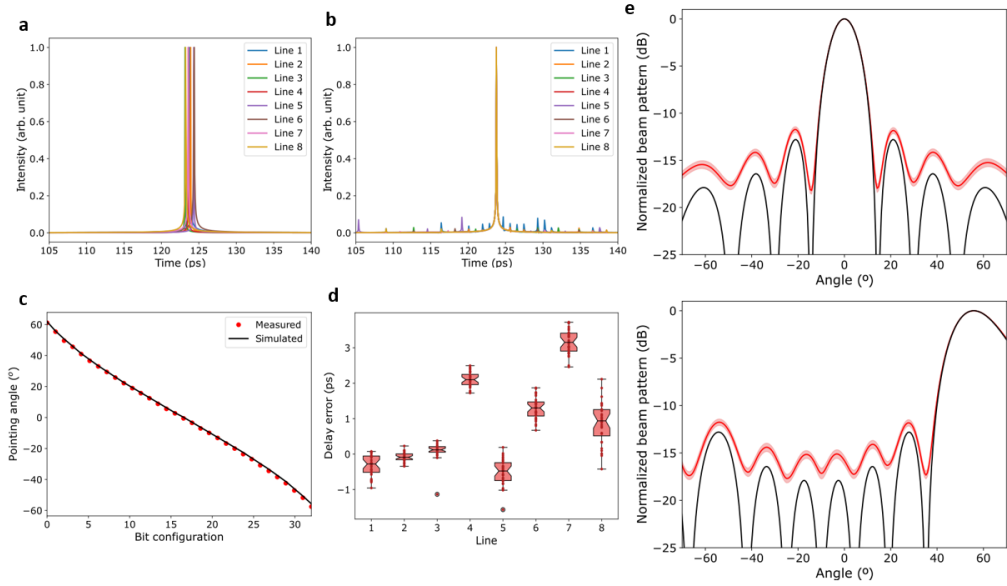


Figure 3.5: Experimental setup and results. **a** Experimental setup for the frequency response characterization of the Silicon photonics beamformer. **b** Group delay vs frequency measurement results for the first, fourth, and eighth lines. Each one shows the 32 delay curves corresponding to the whole set of 5-bit configurations. **c** Calculated beam patterns vs three different frequencies (10, 20, and 30 GHz) for the largest negative (00000), broadside (10000), and largest positive (11111). Insets correspond to the group delay frequency responses of each pointing angle configuration (for full results of the 32 possible pointing angles, see Supplementary Note 4).

Table 3.1: Performance comparison of various state-of-the-art integrated microwave photonic beamformers. (*)Only for positive beam steering angles.

Technical approach/reference	Structure	Platform	Space (mm^2)	Bandwidth (GHz)	Delay (ps)	Power (mW)	Ins. loss (dB)	Angle range ($^\circ$)	Angles	Lines
Switched Delay Lines[26]	MZI-TTODLs	SOI	11×3.9	10(*)	0-496	1450	35	0 to 75.64	16	8
Switched Delay Lines[38]	MZI-TTODLs	SiN	8×32	NS	0-22.5	NS	NS	-51 to 34	6	4
Switched Delay Lines[24]	MZI-TTODLs	SOI	7.4×1.8	NS	0-191.4	1251	27	-	-	1
Switched Delay Lines[28]	MZI-TTODLs	SOI	3.9×5.6	NS	0-210.6	NS	10	-60 to 60	9	16
Switched Delay Lines[29]	MZI-TTODLs	SOI	3.8×10.9	23.5	0-310.8	4560	11-7	NS	9	8
Continuous[39]	MRR	SiN	2×11	5	34-252	NS	11	-28 to 34	-	2
Continuous[40]	MRR	SOI	4.58×0.8	2	36-200	862	NS	-30 to 30	-	4
Continuous[41]	MRR	SiN	8×32	8.6	0-208	NS	NS	-22 to 22	-	4
Switched Delay Lines (this work)	MZI-TTODLs	SOI	8×3	>20	0-228.2	72	12.9	-55.7 to 61.87	32	8

3.1.4 Methods

Array factor of a true time delay 1D beamformer

The performance of a 1D beamformer from the pointing angle characteristic is given by its Array Factor:

$$F(\theta) = \sum_{n=1}^N \frac{1}{N} e^{-i2\pi f \left[\frac{d}{c} (n-1) \sin(\theta) - T_n \right]}, \quad (3.10)$$

where θ is the angle describing the pointing direction and T_n is the time delay for antenna element n . We can define ΔT as the time delay between successive antenna elements given in general by:

$$\Delta T = T_n - T_{n-1} = nBU - (n-1)BU = BU, \quad (3.11)$$

where U is the basic unit delay. For traditional beamformer architecture B ranges from $E[-(2^M - 1)/(N - 1)]$ to $E[(2^M - 1)/(N - 1)]$, where E represents the integer part. For the resolution-improved architecture proposed here B ranges from 0 to $2^M - 1$. By developing the summation and taking the absolute value we arrive at:

$$|F(\theta)| = \left[\frac{\sin \left(\pi f N \left[\frac{d}{c} \sin(\theta) - BU \right] \right)}{N \sin \left(\pi f \left[\frac{d}{c} \sin(\theta) - BU \right] \right)} \right]. \quad (3.12)$$

The squint-free pointing angles of the beamformer are determined by:

$$\frac{d}{c} \sin(\theta) - BU = 0. \quad (3.13)$$

Provided that $|cBU/d| \leq 1$. Note as well that if the condition is fulfilled and B is positive then squint-free angles are contained in the $[0, \pi/2]$ region. If the condition is not fulfilled, then the pointing angles are given by:

$$\frac{d}{c} \sin(\theta) - BU = -\frac{k}{Nf}, \quad (3.14)$$

with $k = 1, 2, 3, \dots$ which are frequency dependent and thus lead to beam squint.

Fabrication

The designed photonic integrated circuit was manufactured by Advanced Micro Foundry, according to a standard SOI process. The chip was fabricated from a SOI wafer with a 220 nm slab thickness, with 500 nm single-mode waveguides defined using deep ultraviolet lithography (193 nm). The waveguide sections with phase shifters are obtained by depositing a thin heater layer over the waveguide, made of 120 nm TiN, which will be powered by metal DC tracks of 2000-nm thickness deposited in the final manufacturing stages.

Phase and group delay measurements

The phase and group delay were measured using a phase-shift approach where RF-to-RF response is measured after going through the integrated circuit. In the laboratory, the chip was integrated into a test and measurement setup, where input and output signals were injected and extracted using optical fibers connected to the corresponding grating couplers (one input coupler and eight output couplers). The switched OTTDLs were programmed using thermo-optic phase shifters, which were controlled by electrical currents from an external source array. The input RF-modulated signal was generated by a Vector Network Analyzer Agilent N5245A capable of modulating signals up to 50 GHz. With a lightwave component analyzer module N4373C, we upconverted the signals to the optical domain using a 1550-nm internal source. Afterward, the output signal from each OTTDL was also downconverted using the same optical module from the VNA. The measured group delay response is obtained after normalizing the measurement by taking into account the external optical components employed in the experimental setup of Fig. 3.4a.

Data availability

All raw and processed data generated in this study are available from the corresponding author upon request.

Code availability

All codes are available from the corresponding author upon request.

Acknowledgements

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Author contributions

P.M. and J.C. carried out the original design and calculations for the beamformer, P.M. designed the silicon chip and carried out the measurements. P.M., J.C., and T.H. analyzed the results and J.C., T.H., and D.W. managed the project.

3.1.5 References

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3.1.6 Supplementary

Supplementary Note 1

Origin of beam-squint in the un-equalized true time delay beamformer architecture

In their work, Zhu et al. proposed an antenna spacing, d , of $\lambda/2$ for a frequency of 16 GHz, i.e. 9.4 mm, and a temporal resolution of 2 ps, which is equivalent to what we have referred to as the basic unit delay U . For the established value of d , it is easy to verify that $d/c \leq 32$ ps, which corresponds to the delay value for the seventeenth pointing angle. This is the reason why only 16 frequency-independent angles appear. By applying the equations developed in Methods, we can study when frequency-dependent solutions begin to appear in the Array Factor.

$$\frac{d}{c} \sin(\theta) - BU = -\frac{1}{Nf} \quad (3.15)$$

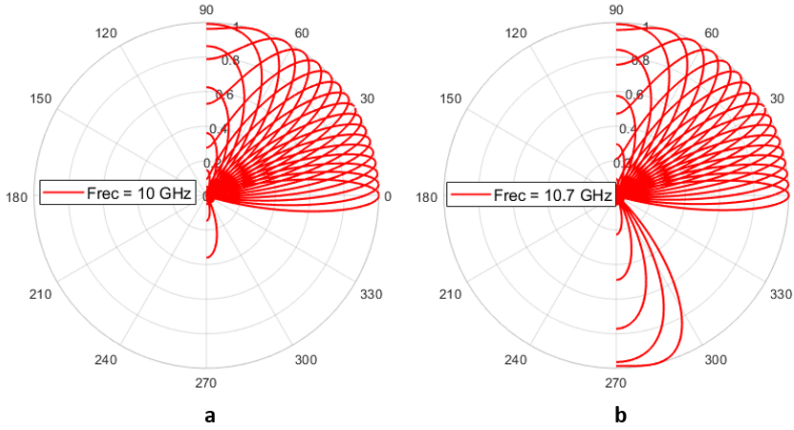


Figure 3.6: Beginning of the emergence of low-frequency negative pointing angles in the previous architecture. a Beam patterns for 10 GHz. **b** Beam patterns for 10.7 GHz. From this second frequency the lobe corresponding to the delay between 31U antennas starts to appear at -90° .

In this case, the first pointing angle observed in the region of negative angles for low frequencies corresponds to the maximum delay between antennas, which is $31U$.

$$f = \frac{1}{\frac{d}{c} \sin(\pi/2) + 31U} = 10.7 \text{ GHz} \quad (3.16)$$

Supplementary Fig.3.6 shows the beam patterns for lower frequencies and for this frequency, showing that negative solutions begin to appear

Below a frequency of 10.7 GHz the system is not capable of targeting any negative angle and it is at this frequency that these angles start to appear. In the same way, we can determine from which frequency we can recover the complete 32 pointing positions, even though half of them present beam-squint.

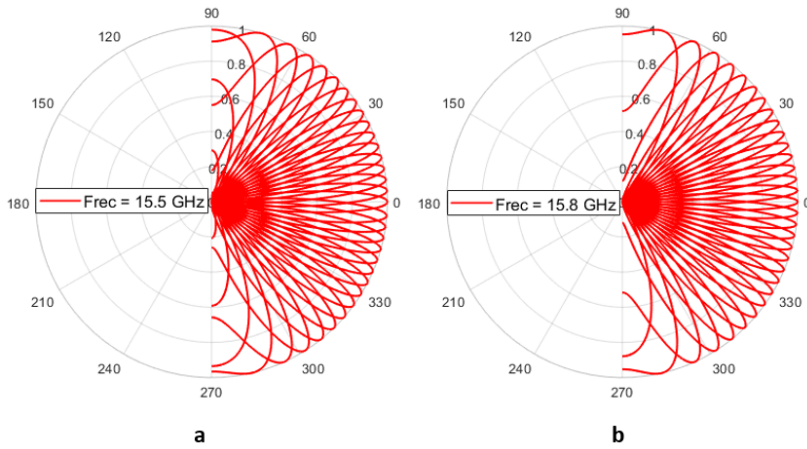


Figure 3.7: Beginning of appearance of all pointing angles for high frequencies in the previous architecture. a Beam patterns for 15 GHz, only 31 pointing angles are displayed. **b** Beam patterns for 15.8 GHz. We recover all the 32 pointing angles.

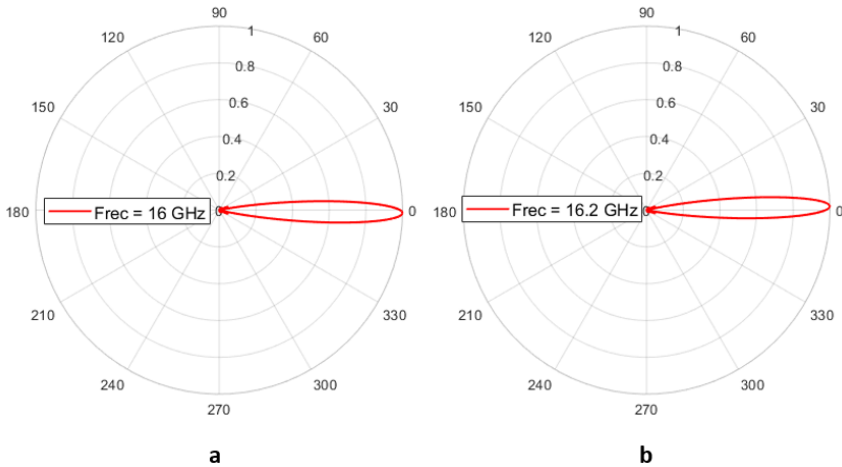


Figure 3.8: Overlapping of the pointing angle with other array configurations. a Beam pattern corresponding to 16 GHz. **b** Beam pattern corresponding to 16.2 GHz. In the second figure we can appreciate how the pointing angle is beyond 0° , thus overlapping with other array configurations.

$$f = \frac{1}{\frac{d}{c} \sin(\pi/2) + 16U} = 15.8 \text{GHz} \quad (3.17)$$

This is demonstrated in Supplementary Fig. 3.7, which shows how for frequencies below 15.8 GHz the 32 pointing angles do not appear, taking into account that half of them are frequency dependent.

At 15.8 GHz and above, 32 positive and negative pointing angles are displayed. Notice that the last pointing angle is really close to the broadband position and this distance becomes smaller with increasing frequency. It is also possible to determine the maximum frequency at which solutions with beam-squint begin to overlap with frequency-independent solutions in the region $[0, \pi/2]$, as shown in Supplementary Fig.3.8.

$$f = \frac{1}{\frac{d}{c} \sin(0) + 31U} = 16.1 \text{GHz} \quad (3.18)$$

This results in a limited set of frequencies, ranging from 15.8 GHz to 16.1 GHz, in which we find 32 different pointing angles that deceive us, promising to be frequency-independent but exhibiting beam-squint behavior.

Supplementary Note 2

Equalizer stage design tables and results

Supplementary Table 3.2 shows all bit configurations for the 8 OTTDLs, from $\Delta T = -16U$ to $\Delta T = 15U$. The first row corresponds to the value of the delay in each line and the second row presents the corresponding precompensation.

Table 3.2: All the configurations of the beamforming network together with the delay value in each line and the corresponding precompensation value.

	Line 1	Line 2	Line 3	Line 4	Line 5	Line 6	Line 7	Line 8
Line delay (τ)	1U	2U	3U	4U	5U	6U	7U	8U
Precompensation	112U	96U	80U	64U	48U	32U	16U	0U
Bit (00000)	112U	96U	80U	64U	48U	32U	16U	0U
Bit (00001)	113U	98U	83U	68U	53U	38U	23U	8U
Bit (00010)	114U	100U	86U	72U	58U	44U	30U	16U
Bit (00011)	115U	102U	89U	76U	63U	50U	37U	24U
Bit (00100)	116U	104U	92U	80U	68U	56U	44U	32U
Bit (00101)	117U	106U	95U	84U	73U	62U	51U	40U
Bit (00110)	118U	108U	98U	88U	78U	68U	58U	48U
Bit (00111)	119U	110U	101U	92U	83U	74U	65U	56U
Bit (01000)	120U	112U	104U	96U	88U	80U	72U	64U
Bit (01001)	121U	114U	107U	100U	93U	86U	79U	72U
Bit (01010)	122U	116U	110U	104U	98U	92U	86U	80U
Bit (01011)	123U	118U	113U	108U	103	98U	93U	88U
Bit (01100)	124U	120U	116U	112U	108U	104U	100U	96U
Bit (01101)	125U	122U	119U	116U	113U	110U	107U	104U
Bit (01110)	126U	124U	122U	120U	118U	116U	114U	112U
Bit (01111)	127U	126U	125U	124U	123U	122U	121U	120U
Bit (10000)	128U	128U	128U	128U	128U	128U	128U	128U
Bit (10001)	129U	130U	131U	132U	133U	134U	135U	136U
Bit (10010)	130U	132U	134U	136U	138U	140U	142U	144U
Bit (10011)	131U	134U	137U	140U	143U	146U	149U	152U
Bit (10100)	132U	136U	140U	144U	148U	152U	156U	160U
Bit (10101)	133U	138U	143U	148U	153U	158U	163U	168U
Bit (10110)	134U	140U	146U	152U	158U	164U	170U	176U
Bit (10111)	135U	142U	149U	156U	163U	170U	177U	184U
Bit (11000)	136U	144U	152U	160U	168U	176U	184U	192U
Bit (11001)	137U	146U	155U	164U	173U	182U	191U	200U
Bit (11010)	138U	148U	158U	168U	178U	188U	198U	208U
Bit (11011)	139U	150U	161U	172U	183U	194U	205U	216U
Bit (11100)	140U	152U	164U	176U	188U	200U	212U	224U
Bit (11101)	141U	154U	167U	180U	193U	206U	219U	232U
Bit (11110)	142U	156U	170U	184U	198U	212U	226U	240U
Bit (11111)	143U	158U	173U	188U	203U	218U	233U	248U

Supplementary Note 3

Phase delay and group measurements per OTTDL unit

The phase and time delay measurements were obtained as indicated in the Methods Section. Below, in Supplementary Fig. 3.9, there are the phase and group delay results for the 32 different bit combinations in each one of the eight lines.

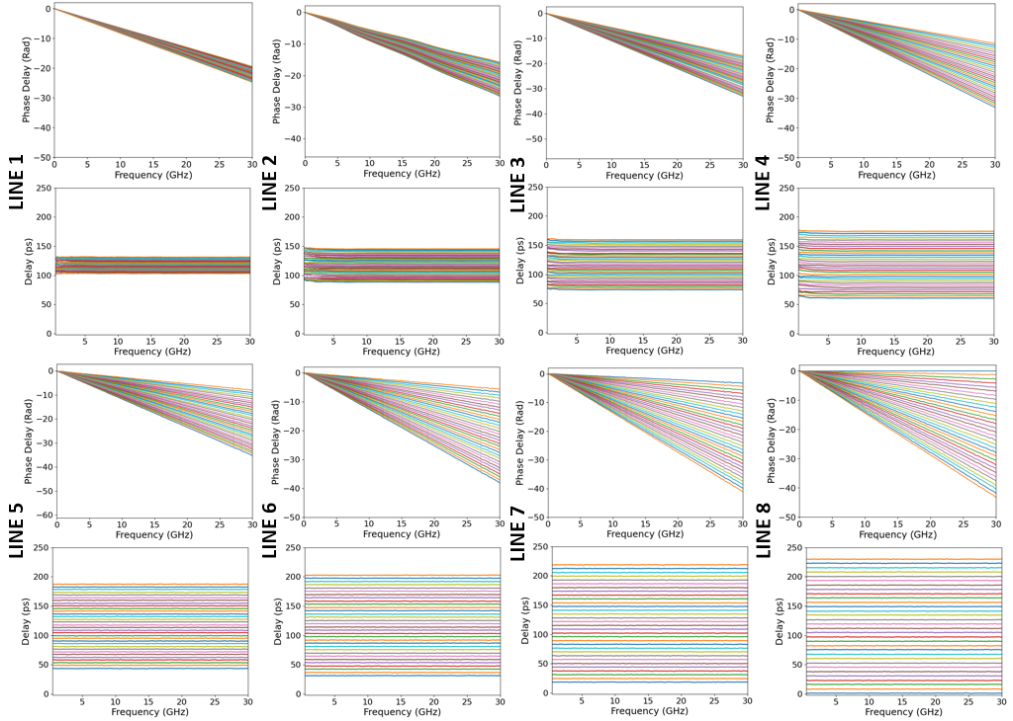


Figure 3.9: Phase and group delay all the lines and bit configurations.

In the first line, the time delays for all bit configurations are relatively close to each other, primarily because the delay per stage in this line is the smallest of the entire array. This, in turn, affects the phase delay, resulting in similar slopes for all the configurations. These measurements change with every line up to the eighth, where the delay is maximum, and we can see a big difference in time and slopes for every bit word. All measurements were taken using as reference the smallest optical path within the beamforming network, the eighth line for bitword 00000. The value of the precompensation delay in each line can be appreciated, from which the time delay starts to grow as the bit words increase. The delay value for each line in all the configurations of the beamforming network is represented in the following graphs, and it can be seen in a very visual way how the precompensation delay allows to obtain increasing delays in the two directions of the array.

Supplementary Fig. 3.10 shows the delays in each line for the first set of 5 bit word configurations (00000 to 01111) together with the corresponding beam pattern for each

configuration. Supplementary Fig. 3.11 shows identical information for the second set of 5 bit word configurations (10000 to 11111).

The system starts pointing to negative angles, with a decreasing delay from the first line to the eighth one. As the bit word increases, the relative delay between lines becomes smaller until reaching the bit word 10000, where all lines suffer the same delay, and we obtain the broadband configuration. After this, the temporal separation between lines grows again, but this time in the opposite direction of the array, allowing pointing at positive and negative angles.

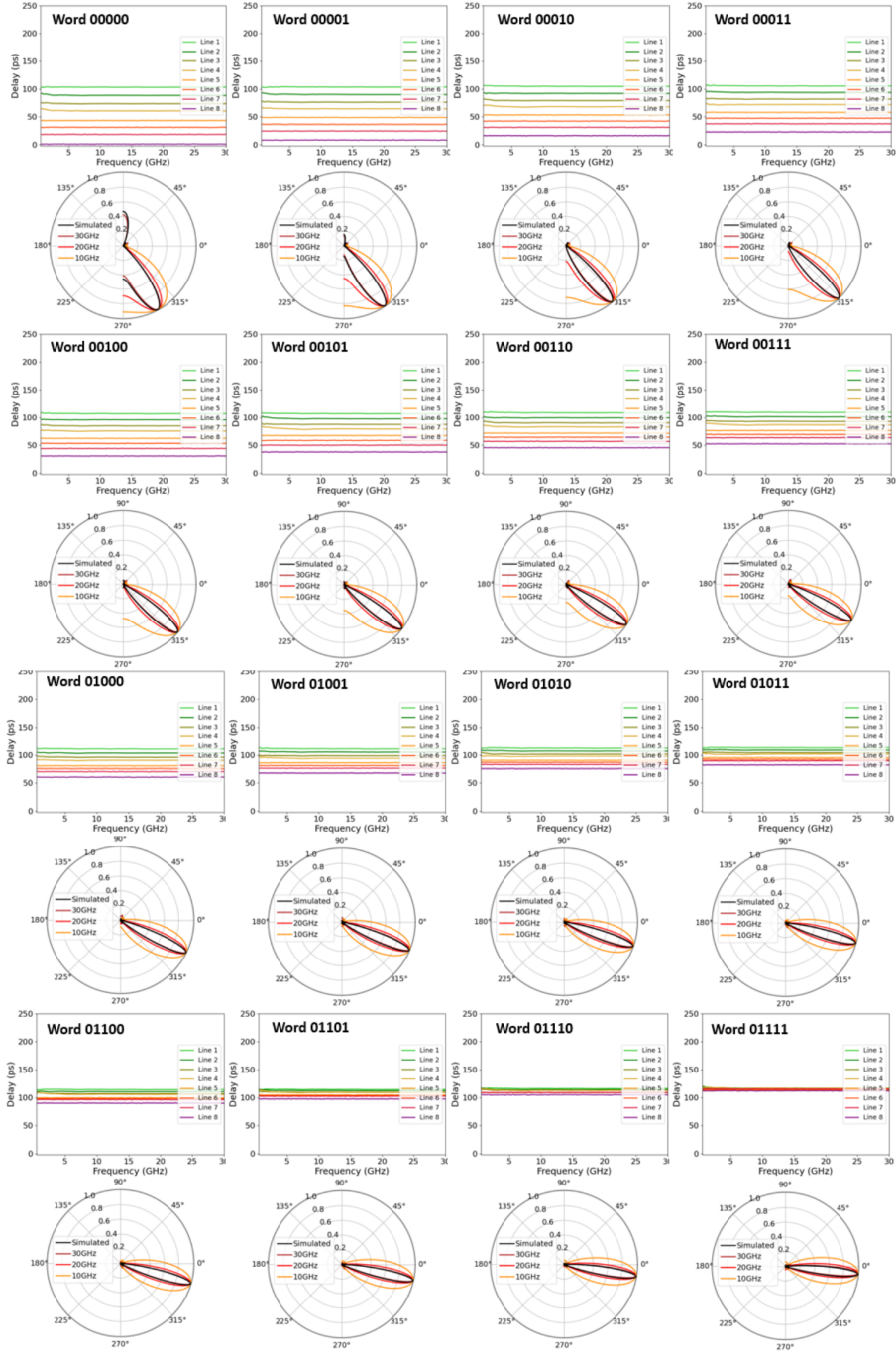


Figure 3.10: Delays and beam patterns for the first set of 16 different pointing angles (bit words 00000 to 01111).

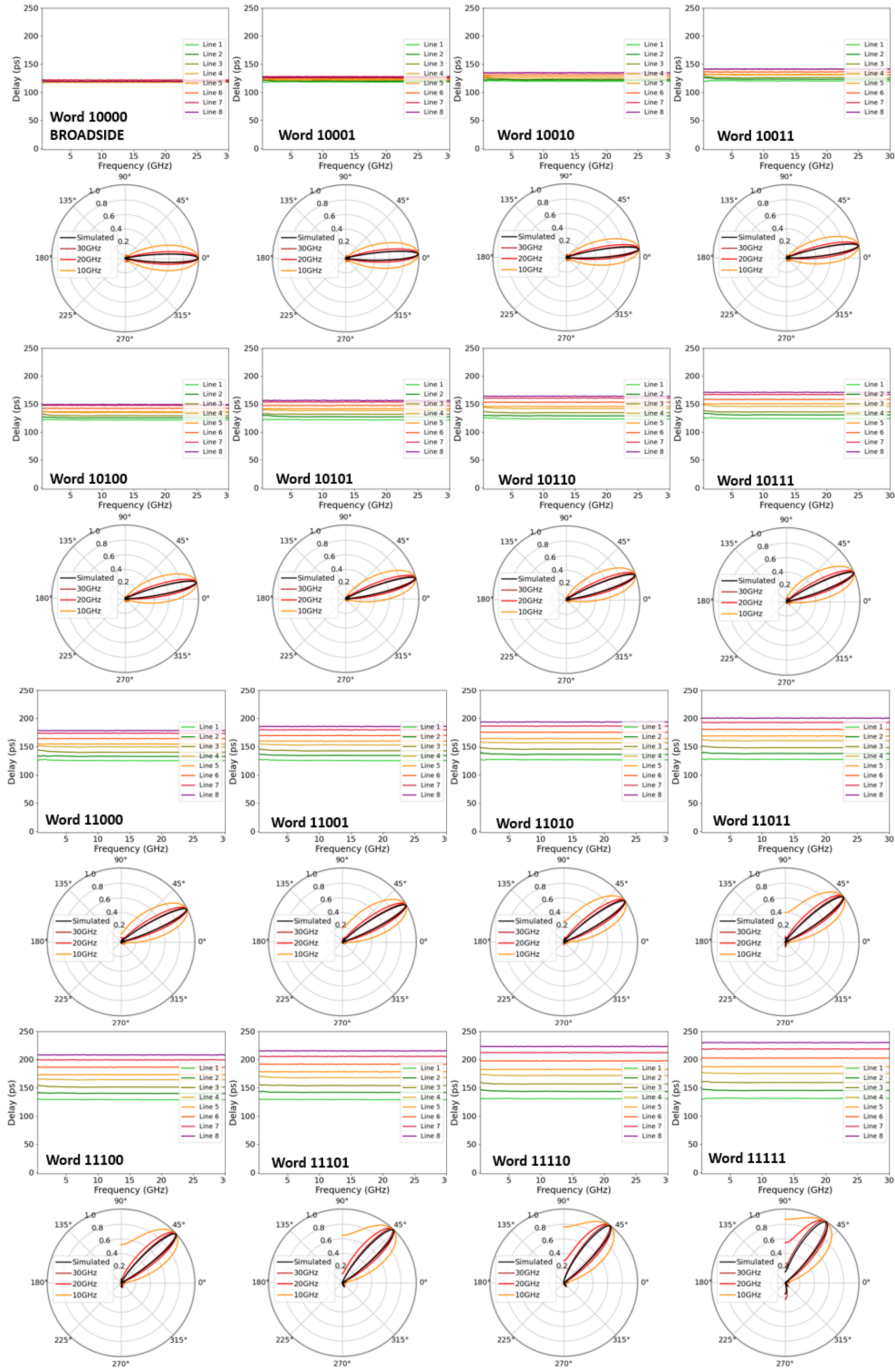


Figure 3.11: Delays and beam patterns for the second set of 16 different pointing angles (bit words 10000 to 11111).

Supplementary Note 4

Power Consumption

Table 3.3: Compilation of the power consumption of the MZIs corresponding to each OTTDL stage.

Switch	Cross (mW)	Bar (mW)	Switch	Cross (mW)	Bar (mW)
S_{10}	0	1.51	S_{20}	0	2.42
S_{11}	2.45	1.14	S_{21}	1.06	2.38
S_{12}	0.07	1.39	S_{22}	0.90	2.27
S_{13}	0.31	1.60	S_{23}	0.91	2.23
S_{14}	0.65	1.95	S_{24}	0.94	2.27
S_{15}	2.45	1.15	S_{25}	0.80	2.05
S_{16}	1.17	2.45	S_{26}	0.32	1.59
S_{30}	0	1.35	S_{40}	0	1.78
S_{31}	2.50	1.20	S_{41}	0.70	2.03
S_{32}	2.27	0.99	S_{42}	0.60	1.93
S_{33}	2.16	0.87	S_{43}	0.57	1.84
S_{34}	0	1.28	S_{44}	0.23	1.50
S_{35}	0.15	1.49	S_{45}	0.03	1.34
S_{36}	1.50	0	S_{46}	1.52	0
S_{50}	0	1.98	S_{60}	0	2.00
S_{51}	1.78	0.51	S_{61}	0.18	1.49
S_{52}	2.08	0.78	S_{62}	2.10	0.77
S_{53}	2.11	0.78	S_{63}	2.21	0.93
S_{54}	1.44	0.16	S_{64}	1.96	0.68
S_{55}	1.67	0.36	S_{65}	2.09	0.79
S_{56}	2.25	2.25	S_{66}	2.16	0.82
S_{70}	0	1.01	S_{80}	0	1.51
S_{71}	2.07	0.78	S_{81}	1.35	0
S_{72}	0.69	2.03	S_{82}	0.80	2.09
S_{73}	0.45	1.73	S_{83}	0.52	1.88
S_{74}	0.60	1.92	S_{84}	1.27	0
S_{75}	0.65	1.88	S_{85}	0.63	1.98
S_{76}	2.07	0.76	S_{86}	1.14	0.02

The power consumption for the beamforming network comes from the phase actuators based on thermo-optics form the switchable delay lines. Table 3.3 shows the power consumption for the bar and cross states of all the tunable couplers on the chip (including the optional precompensation stage). The consumptions for the cross and bar states are different for each switch due to differences in the manufacturing process but in general the power consumption for a π -phase shift is around 1.29 mW. The average consumption per line is approximately 9 mW giving a total consumption for the complete optical beamforming network of 72 mW. Once the correct operation of the architecture has been checked, power consumption can be reduced if the precompensation stage is fixed instead of having a specific delay stage for it.

Supplementary Note 5

Effect of loss imbalance in the Array Factor

Since the delay line stages in the beamformer have different waveguide lengths one may ask whether these can yield significant differences in propagation losses, which might potentially induce an undesired apodization effect in the beam pattern. Here we report the main simulation results of our architecture for different waveguide loss coefficients, ranging from 0 to 5 dB/cm (including our measured value of 1.1 dB/cm). Losses higher than 3 dB/cm are far beyond the worst-case scenario on any foundry nowadays, but we have nevertheless included them to prove that the apodization effect is negligible.

Supp. Figure 3.12 (a-d) illustrates the effect of various waveguide propagation loss values in each of the delay lines across all the 32 5-bit word configurations. Due to the precompensation stage, the most substantial dissimilarities for the extreme side angles, whereas for the broadside configuration, all lines exhibit uniform propagation losses. Therefore, in Supp. Fig. 3.12 (e-h) we plot the corresponding beam patterns for one of the two extreme side angles (-60°). Inevitably, as the propagation losses increase the directivity and the sidelobes' extinction ratio decrease. However, in terms of its shape, the pattern closely retains its ideal form compared to that obtained with a uniform apodization, even in the presence of losses.

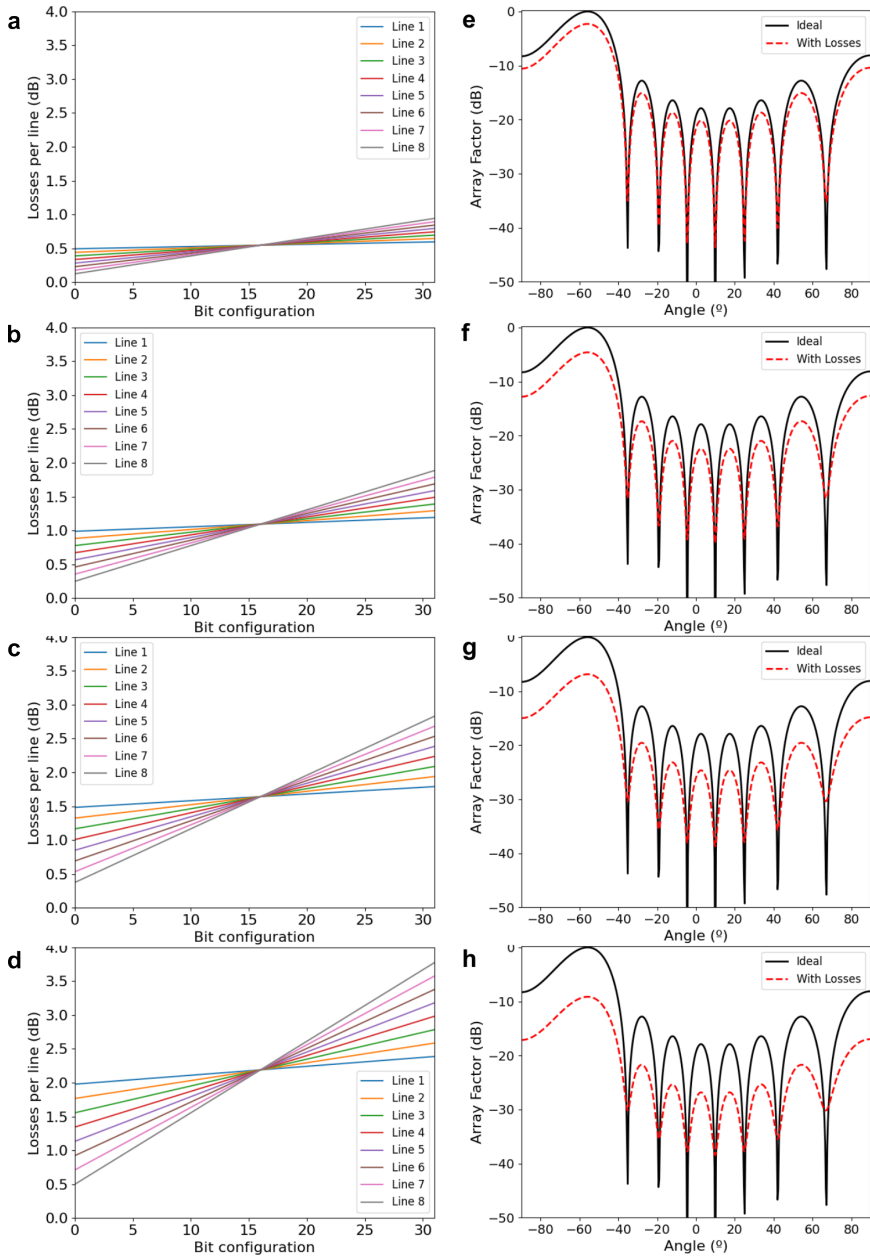


Figure 3.12: (a-d) Losses per line for 0.75, 1.5, 2.25 and 3 dB/cm as loss coefficient. (e-h) The corresponding side angle (bitword 00000).

In Supp. Fig. 3.13 we show the variation of the Main Lobe to Secondary Ratio (MLSR) with the loss coefficient for the worst-case extreme side pointing angle, proving that the value of tapering is negligible.

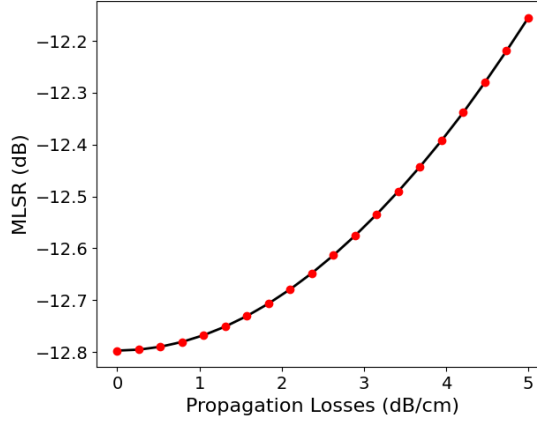


Figure 3.13: The value of the MLSR for different propagation losses. In order to observe an appreciable change, we have to allow for propagation losses much higher than those routinely attainable in the current state of the art.

Supplementary Note 6

Before measuring the beamforming network, we collected data to estimate the performance of the fundamental building blocks employed in the final design, in order to characterize their insertion losses and wavelength dependence. All the structures show a flat response in wavelength and the fittings were done for the principal wavelength of use, 1550 nm.

Propagation Losses measurements

We used the cutback technique to measure the propagation losses in straight waveguides of different lengths incorporated as test structures in the chip as shown in Supplementary Fig. 3.14

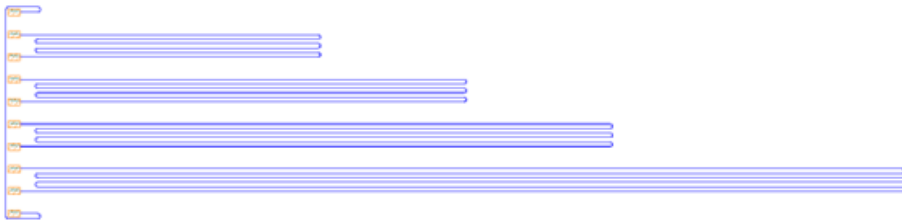


Figure 3.14: Structure used to measure the propagation losses of the waveguides. Input by grating coupler array.

The main results for the spectral dependence of losses in the 1520-1580 nm wavelength

range are shown in the left part of Supplementary Fig. 3.15, where each color corresponds to a waveguide of different lengths. From those, we can extrapolate for a given wavelength a linear regression curve as shown in the right part of Supplementary Fig. 3.15 for 1550 nm. At this wavelength the slope corresponds to a loss of 1.1 ± 0.3 dB/cm.

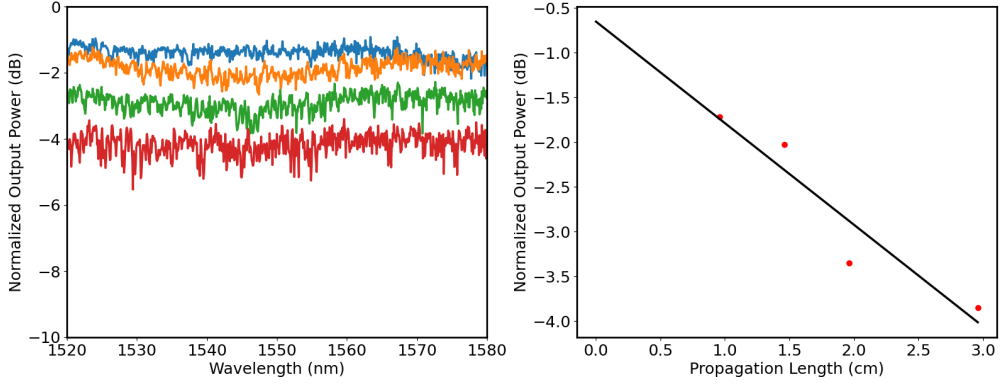


Figure 3.15: Wavelength spectrum and linear fitting for the different lengths. The responses correspond to each of the different lines, each one longer than the previous one. The propagation loss coefficient obtained is 1.1 ± 0.3 dB/cm.

MMI Losses measurements

Similarly, we used a set of test MMI configurations as shown in Supplementary Fig. 3.16 to allow us to measure the insertion losses of a cascade of MMIs ranging from 1 to 10.

The main results for the spectral dependence of losses in the 1520-1580 nm wavelength range are shown in the left part of Supplementary Fig. 3.17, where each color corresponds to a cascade of different numbers of MMI couplers. From there, we can extrapolate for a given wavelength a linear regression curve as shown in the right part of Supplementary Fig. 3.17 for 1550 nm. At this particular wavelength, the slope corresponds to an insertion loss of 0.15 ± 0.05 dB/MMI.

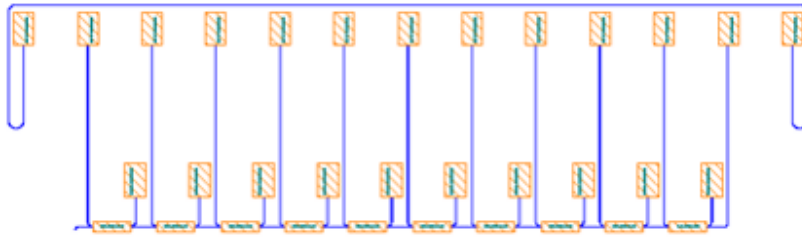


Figure 3.16: Structure used to measure the propagation losses of the MMIs. Input by grating coupler array.

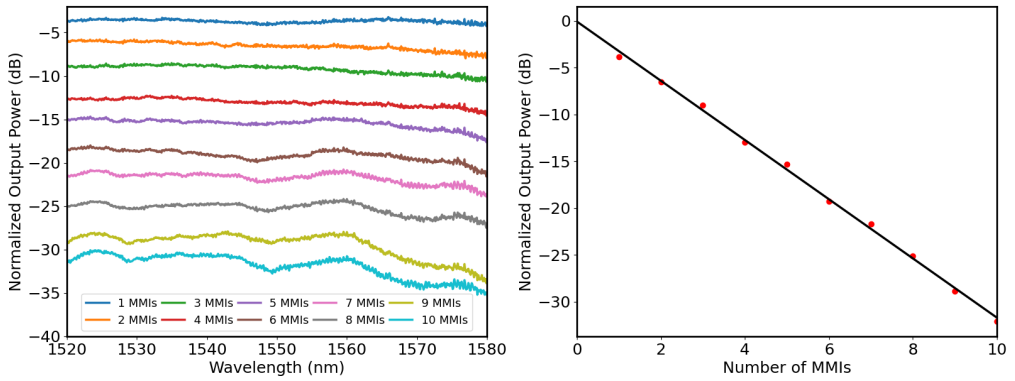


Figure 3.17: Wavelength spectrum and linear fitting for a different number of concatenated MMIs, from 1 to 10. The slope analysis provided us with the insertion loss value, obtained after accounting for the power loss resulting from the MMI's 50/50 splitting ratio. Our measurement indicates an insertion loss of 0.15 ± 0.05 dB/MMI

MZI measurements

Finally, we characterized the insertion losses for different switching MZI units employed to setup the signal routing in each delay line according to the programmed bit word. The main insertion losses versus the wavelength results are shown in Supplementary Fig. 3.18, where the different curves correspond to different currents applied to the thermo-optic actuators (0 to 0.7 mA). The extreme values correspond to bar and cross states while the intermediate correspond to partial coupling. As in the previous cases, the results at 1550 nm were employed to find the MZI insertion loss at that wavelength yielding a value of 0.45 dB.

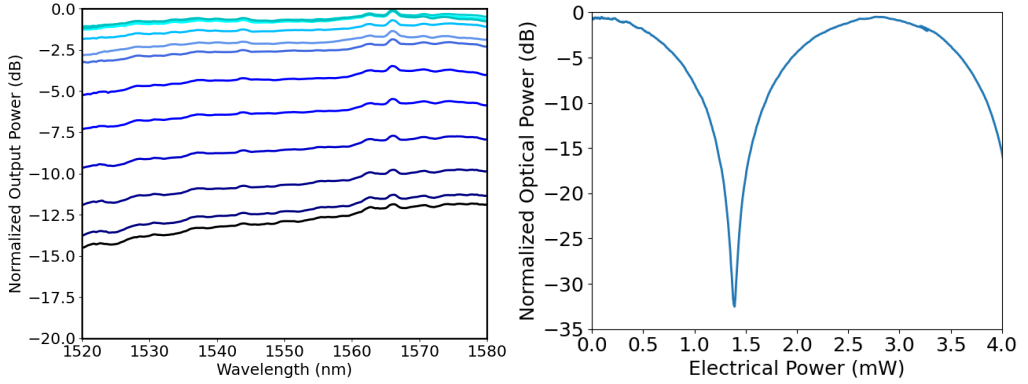


Figure 3.18: The colored lines show the wavelength response for different currents applied to the thermo-optic actuators. On the right side, the response of one of the MZI at 1550 nm.

3.2 Paper B

Journal of Lightwave Technology, (2024)

Nonintrusive characterization method for integrated
optical delay lines

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Nonintrusive characterization method for integrated optical delay lines

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In this work we present an automatic calibration technique for switched Optical True Time Delay Lines which allows the control of all of the switchers without relaying on intermediate attenuators nor external test ports, thus optimizing the number of active elements on chip and the footprint. This technique has been used for the characterization of a delay line fabricated on a silicon photonic chip. The accuracy and reliability of this technique were validated for applications beyond beamforming by utilizing the calibration data to synthesize variable optical interleavers.

3.2.1 Introduction and context

In recent years, microwave photonics has emerged as a technology with immense potential, destined to enhance the capacity and speed of modern telecommunications. This is especially critical in light of the escalating demands for high-speed and high-capacity communication driven by emerging applications like the Internet of Things (IoT) and future 5/6G networks [1], [2].

It accomplishes this by transcending the limitations of radiofrequency devices and systems, capitalizing on the unique optical attributes of low loss, minimal power consumption, broadband operation, immunity to electromagnetic interference, and extensive tunability [3]. This is achieved through the upconversion of radiofrequency signals into the optical domain, their processing, and subsequent downconversion, effectively capitalizing on the advantages of photonics. An area of substantial focus is optical true-time delay (OTTD), a technology enabling precise control over the propagation delay of optical signals. This capability finds application in various domains, including synchronization and buffering of optical signals [4], optical signal filtering and processing [5], and more recently, in reservoir computing [6].

However, among these applications, OTTD has proven to be an ideal solution for phase array antennas (PAA) and beamforming networks, greatly enhancing their performance [7], [8]. These notable improvements over traditional beamformers stem from the elimination of the beam squint effect associated with electronic phase shifters, where changes in the microwave frequency cause a deviation in the pointing angle, limiting the operating bandwidth to less than a few hundred megahertz [9]–[11]. This translates into higher bandwidth, increased speed, and reduced power consumption.

The ongoing progress in integrated photonics has made it possible to integrate on-chip OTTD systems with increasingly complex circuits. This integration not only encompasses all the previously mentioned advantages but also extends to devices with an

exceptionally compact footprint [12], [13]. Numerous schemes for implementing OTTD have been demonstrated, but among them, tunable OTTD based on microring resonators (MRRs) and optical switchable waveguides stand out for their simplicity and suitability for monolithic integration on silicon-on-insulator (SOI) chips.

PAA architectures relying on single and cascaded MRRs offer continuous time delay tuning, delivering high angular resolution [14]–[17]. However, they encounter bandwidth limitations due to the resonant behavior of their delays and exhibit high sensitivity to thermal crosstalk and manufacturing errors. Conversely, switched delay lines, while offering a more limited angular resolution, provide significant operational bandwidth and exhibit low sensitivity to thermal effects and manufacturing variations, which make them the central focus of this work. The uses of this type of integrated structures go beyond just PAA with high performance [18]–[21], including wavelength division multiplexing (WDM) filters [22], [23], gain and dispersion equalization filters [24].

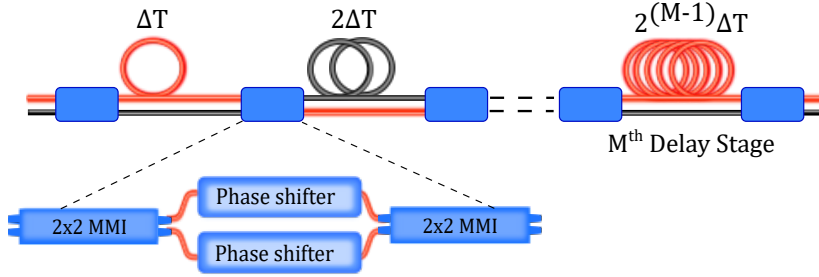


Figure 3.19: Scheme of a single M-bit switched OTTDL, where ΔT is the basic delay of the line.

Figure 3.19 depicts the conventional setup of a switched OTTDL. This configuration comprises M delay stages (bits) featuring two selectable paths. The lower path maintains a constant length for each bit, while the other one exponentially increases its length as the number of bits progresses. These delay lines are interconnected by tunable switches, consisting of 2×2 Mach-Zehnder Interferometers (MZIs). By switching between the long and short delay paths in each stage, it is possible to achieve discrete delays that are multiples of a fundamental delay ΔT , from 0 to $(2^M - 1)\Delta T$. This method of obtaining delays can be straightforwardly represented using binary notation, where a "0" denotes the selection of a short path in a delay stage, and a "1" represents the selection of a long path.

In this scheme, it is customary to incorporate variable optical attenuators (VOAs) [18], [19], [21] or test ports (such as photodetectors or grating couplers) [20] between the delay stages. These elements are strategically placed along all connection waveguides, serving two objectives. Firstly, they enable the calibration of the optical switches, ensuring accurate switching states. Secondly, they attenuate the residual signal in the case of having limited extinction ratio (ER) on the switches, thereby enhancing the quality of the transmitted signal. However, both configurations introduce additional complexity to the integrated circuit, accompanied by higher power consumption due to the VOAs and extra insertion losses caused by the test ports. Most notably, these additions result in an increase of the chip's footprint and more complex packaging.

Non-invasive characterization techniques have recently been studied in the literature, such as using intermediate phase shifters in the delay stages as monitoring elements that

avoid adding losses [25], or techniques that rely on optimization algorithms to reduce the standard deviation in the optical response of the delay lines [26].

In this paper, we extend the work presented in [27], and we present and validate a calibration method for switched delay lines that eliminates the necessity of supplementary structures. Our proposed approach allows for the development of novel architectures that are not only more compact but also exhibit reduced power consumption. Section 3.2.2 introduces the calibration method based on spectral Fourier transform. In Section 3.2.3, we show the details of the silicon photonic OTTDL used for the experimental demonstration of the proposed approach. Moving forward, Section 3.2.4 discusses the chip calibration measurements and demonstrates its application for a tunable optical interleaver based on a lattice filter, showcasing the method's efficacy. Finally, in Section 3.4.1, we provide a summary and conclusions of the study.

3.2.2 Calibration Method

While theoretically, the initial state of each MZI should be a perfect cross-state, this ideal state is often far from reality. Minor variations in the width and height of the waveguides in the two paths of the MZI result in discrepancies in the group index, leading to a phase difference between the arms. Typically, the default state of each MZI in the OTTDL is unpredictable and random due to these manufacturing imperfections and that is the reason of using VOAs or test ports. Nonetheless, we present an alternative approach that only utilizes Fourier transform analysis of the optical spectrum.

Equation (3.28) characterizes the transfer function of an individual tunable coupler, expressed as a function of the phase ϕ applied to one of its arms.

$$H_{TC}(\phi) = -ie^{i(\phi/2)} \begin{bmatrix} \sin(\phi/2) & \cos(\phi/2) \\ \cos(\phi/2) & -\sin(\phi/2) \end{bmatrix}. \quad (3.19)$$

The phase variation is uniform across both outputs and will be eliminated post-photodetection, rendering it inconsequential. Given this, we can simplify the preceding equation by expressing it in terms of the coupling coefficient κ .

$$H_{TC}(\kappa) = \begin{bmatrix} s(\kappa) & c(\kappa) \\ c(\kappa) & -s(\kappa) \end{bmatrix} = \begin{bmatrix} \sqrt{1-\kappa} & \sqrt{\kappa} \\ \sqrt{\kappa} & -\sqrt{1-\kappa} \end{bmatrix}. \quad (3.20)$$

When a signal is injected into one of the inputs of the delay line, the initial tunable switch divides the signal between the paths of the first delay stage and the subsequent tunable coupler combine them,

$$\begin{aligned} H_{DS} &= H_{TC}(\kappa_1) H_{Delay} H_{TC}(\kappa_0) \\ &= \begin{bmatrix} s_1 & c_1 \\ c_1 & -s_1 \end{bmatrix} \begin{bmatrix} e^{i\omega\Delta T} & 0 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} s_0 & c_0 \\ c_0 & -s_0 \end{bmatrix} \\ &= \begin{bmatrix} s_0 s_1 e^{i\omega\Delta T} + c_0 c_1 & c_0 s_1 e^{i\omega\Delta T} - s_0 c_1 \\ s_0 c_1 e^{i\omega\Delta T} - c_0 s_1 & c_0 c_1 e^{i\omega\Delta T} + s_0 s_1 \end{bmatrix}. \end{aligned} \quad (3.21)$$

The time difference between paths is related with the physical length difference, ΔL , and the group index, n_g , of the waveguide mode, $\Delta T = n_g \Delta L / c$. This analysis can

be extended for additional tunable couplers and delay stages and after photodetection, the contributions from all length imbalances mix in the photodiode. Hence, in the presence of optical leakages between delay stages resulting from inefficient switching, we will detect a distinctive interferometric pattern in the frequency response of the delay line. This pattern exhibits a combination of all the characteristic Free Spectral Range (FSRs) corresponding to all the length imbalances across each delay stage and the associated coupling coefficients.

$$\begin{aligned}
P_0(\omega) = & (s_0 s_1 s_2 s_3)^2 + (s_0 s_1 c_2 c_3)^2 + (c_0 c_1 s_2 s_3)^2 + (c_0 c_1 c_2 c_3)^2 + (s_0 c_1 s_2 c_3)^2 + (c_0 s_1 c_2 s_3)^2 + \\
& (c_0 s_1 s_2 c_3)^2 + 2s_0 c_0 [s_1 c_1 ((s_2 s_3)^2 + (c_2 c_3)^2 - (s_2 c_3)^2 - (c_2 s_3)^2) + c_1^2 s_2 c_2 (s_3^2 - c_3^2) - s_1^2 c_2^2 s_3 c_3] \cos(\omega \Delta T) + \\
& + 2s_1 c_1 c_2 [s_2 ((s_0 s_3)^2 + (c_0 c_3)^2 - (s_0 c_3)^2 - (c_0 s_3)^2) + c_2 s_3 c_3 (s_0^2 - c_0^2)] \cos(\omega 2\Delta T) + \\
& + 2s_0 c_0 c_2 [s_3 c_3 (2s_1 c_1 s_2 + c_1^2 c_2) + s_1^2 s_2 (c_3^2 - s_3^2)] \cos(\omega 3\Delta T) + \\
& + 2s_2 c_2 s_3 c_3 [(s_0 s_1)^2 + (c_0 c_1)^2 - (s_0 c_1)^2 - (c_0 s_1)^2] \cos(\omega 4\Delta T) + \\
& + 2s_0 c_0 c_1 s_2 s_3 c_3 [2s_1 c_2 - c_1 s_2] \cos(\omega 5\Delta T) + 2s_1 c_1 s_2^2 s_3 c_3 [s_0^2 - c_0^2] \cos(\omega 6\Delta T) + 2s_0 c_0 s_1^2 s_2^2 s_3 c_3 \cos(\omega 7\Delta T)
\end{aligned} \tag{3.22}$$

$$\begin{aligned}
FTT(t) = & \int_{-\infty}^{\infty} P_0(\omega) e^{-i2\pi\omega t} d\omega \propto \\
& \propto s_0 c_0 [s_1 c_1 ((s_2 s_3)^2 + (c_2 c_3)^2 - (s_2 c_3)^2 - (c_2 s_3)^2) + c_1^2 s_2 c_2 (s_3^2 - c_3^2) - s_1^2 c_2^2 s_3 c_3] \delta(t \pm \Delta T) + \\
& + s_1 c_1 c_2 [s_2 ((s_0 s_3)^2 + (c_0 c_3)^2 - (s_0 c_3)^2 - (c_0 s_3)^2) + c_2 s_3 c_3 (s_0^2 - c_0^2)] \delta(t \pm 2\Delta T) + \\
& + s_0 c_0 c_2 [s_3 c_3 (2s_1 c_1 s_2 + c_1^2 c_2) + s_1^2 s_2 (c_3^2 - s_3^2)] \delta(t \pm 3\Delta T) + \\
& + s_2 c_2 s_3 c_3 [(s_0 s_1)^2 + (c_0 c_1)^2 - (s_0 c_1)^2 - (c_0 s_1)^2] \delta(t \pm 4\Delta T) + \\
& + s_0 c_0 c_1 s_2 s_3 c_3 [2s_1 c_2 - c_1 s_2] \delta(t \pm 5\Delta T) + s_1 c_1 s_2^2 s_3 c_3 [s_0^2 - c_0^2] \delta(t \pm 6\Delta T) + s_0 c_0 s_1^2 s_2^2 s_3 c_3 \delta(t \pm 7\Delta T)
\end{aligned} \tag{3.23}$$

To illustrate the calibration procedure, consider a generic switched OTTDL with 3 delay stages with respective delay values of ΔT , $2\Delta T$ and $4\Delta T$. The transfer matrix of the delay line has been calculated similarly to the one expressed in Eq.(3.21), but with the inclusion of two additional delay stages alternated between tunable couplers. Equation (3.22) depicts the photodetected normalized optical power considering a bar-state transmission at the top port, $|H_{11}|^2$, the signal is injected through the top input, and the measurement is taken at the top output.

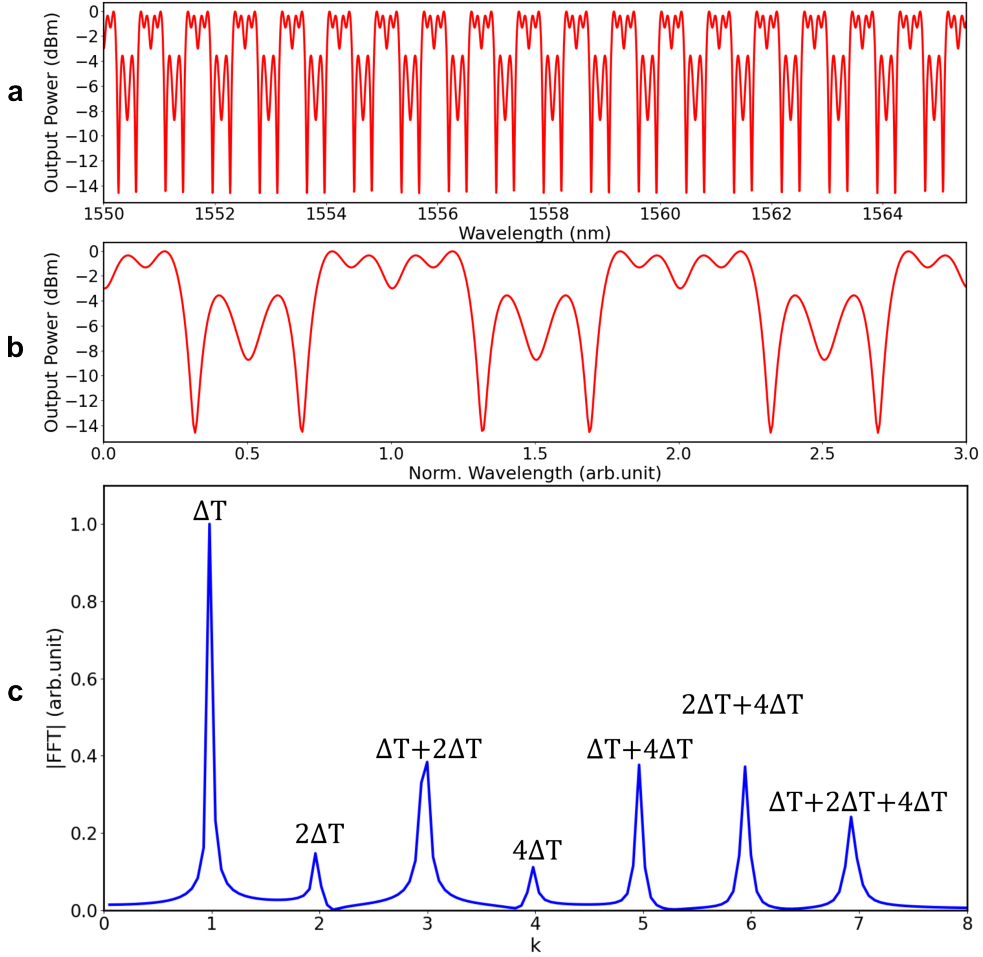


Figure 3.20: **a** Optical spectrum of a random state in the OTTDL example. **b** Zoom of a section of the spectrum normalized to the largest FSR, covering 3 repetitions. **c** Fourier transform of the spectrum through the OTTDL expressed as a function of all the possible time combinations.

The frequency-dependent nature of the response creates a periodic pattern. Only relying on the spectrum is insufficient for accurately characterizing the MZIs; it is necessary to take the Fourier transform. By doing so we move from frequency to time space, in other words, to a space in which the repeating patterns in the spectrum are quantified. One of the benefits of using the Fourier Transform is its independence from the absolute

optical power measured; only a relative measurement is needed. This is because the information provided by the FFT relies exclusively on the shape of the spectrum, not its absolute power. As shown in Equation (3.23), the Fourier transform exhibits peaks in time aligning with the inverse of the FSRs associated with the different delay interferences. Thus, each peak contains information about the 'intensity' in the number of repetitions in the spectrum corresponding to the different time combinations.

Figure 3.20 represents the optical response for a random configuration of this delay line. With the transformed spectrum we are able to discern the contributions of the different optical paths and to know which MZIs are in an intermediate state. Since all time differences in the delay stages are multiples of a fundamental basic delay, we can normalize the time to this value, $k = t/\Delta T$. That way the peaks will be located uniformly as shown in Fig.3.20(c). By systematically adjusting the currents from the first MZI to the latest one, we can effectively suppress the contributions on the spectrum. This process involves gradually diminishing the peaks of interference on the transformed spectrum, starting with those associated with the smallest optical delays.

Algorithm 1 Delay line calibration

- 1: **Stage 1:** Characterize the first M MZIs
 - 2: **for** $i = 1$ to M **do**
 - 3: Sweep current applied to MZI _{i}
 - 4: With $u \in \{n | n \in \mathbb{N}, 1 \leq n \leq 2^{(M-i)}, n \text{ odd}\}$
 - 5: Save first current that nulls the peaks $2^{(i-1)}u\Delta T$
 - 6: Save second current that nulls the peak $2^{(i-1)}u\Delta T$
 - 7: Apply first/second current to MZI _{i}
 - 8: **end for**
 - 9: **Stage 2:** Characterize the last MZI
 - 10: Sweep current applied to MZI _{$M+1$}
 - 11: Save current that maximizes the output optical power
 - 12: Save current that minimizes the output optical power
-

We commence the tuning process with the first MZI. Referring to Eq.(3.23), it becomes evident that the initial peak cancellation occurs only when this MZI is in either the bar state ($s_0 = 1$ and $c_0 = 0$) or the cross state ($s_0 = 0$ and $c_0 = 1$). By doing this, we not only suppress the first interference peak but also eliminate all odd contributions, as they all share a common factor of $s_0 c_0$.

Then, we repeat the same process with the second MZI, shutting down the peak corresponding to $2\Delta T$ when $s_1 = 0$ or $c_1 = 0$. At this point, the peak corresponding to $6\Delta T$ will also have vanished, leaving only the contribution of the third delay stage. We perform the same procedure once more for s_2 and c_2 , suppressing the peak at $4\Delta T$. Subsequently, we determine the bar and cross states of the final MZI by optimizing the optical power, adjusting s_3 and c_3 . Algorithm 1 shows a more detailed version of this process for a delay line with M delay stages, $(M + 1)$ MZIs.

Finally, when the spectrum has been completely flattened, it indicates that all the MZIs have reached either a perfect bar or cross state, this is $\kappa_n \in \{0, 1\}$. At this point, the remaining question is to determine which specific state each MZI is in. To accomplish this, we can utilize a Vector Network Analyzer (VNA) to measure the phase of the signal and extract the corresponding delay information.

3.2.3 Chip Structure

In order to evaluate the characterization method, an OTTDL was specifically designed for use as a reconfigurable lattice filter. The OTTDL configuration consists of 5 delay stages and 6 tunable couplers. The unit delay within the system has a value of $\Delta T = 9.5$ ps, resulting in a maximum delay of $16\Delta T = 152$ ps for the final stage.

The photonic integrated circuit was produced by Advanced Micro Foundry, using a conventional Silicon-on-Insulator (SoI) fabrication process. The chip was crafted from an SoI wafer with a 220 nm thick slab, and deep ultraviolet lithography (193 nm) was employed to define the 500 nm wide single-mode waveguides.

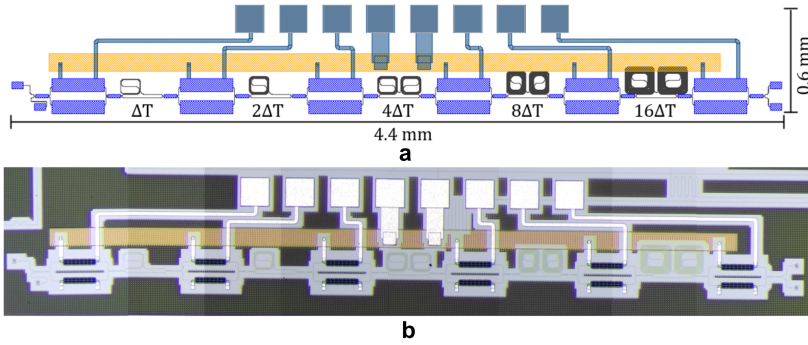


Figure 3.21: Layout (a) and micrograph (b) of the fabricated silicon photonics OTTDL (4.4 mm x 0.6 mm).

For the waveguide sections containing phase shifters, a thin heater layer consisting of 120 nm TiN was deposited over the waveguide, which is powered by metal DC tracks of 2000 nm thickness. These thermo-optic phase actuators are implemented using suspended waveguides and trenches, key elements that enable a reduced power consumption of only $1.35mW/\pi$. Figure 3.21 shows the layout and the micrograph of the fabricated silicon photonics OTTDL.

3.2.4 Experimental Results

Experimental Calibration

The calibration process was executed according to the technique described in Section 3.2.2. The chip was powered through a multicontact probe equipped with 8 electrical channels, and optical input and output were facilitated using a single fiber and grating couplers. For the wavelength sweep, we employed a tunable laser, the EXFO T100S-HP, in conjunction with a full-band component tester, the EXFO CT440 comb source. The wavelength span we utilized was 3 nm, with a resolution of 1 pm. Smaller delay lengths may necessitate larger spans to achieve the highest resolution in the reciprocal spectrum.

Figure 3.22 illustrates the characterization process, presenting both the optical spectrum and its corresponding Fourier transform in a unified view.

At first, the line arrived with the MZIs in a completely random state (first row of Fig.3.22) resulting in a spectrum with numerous interferences and contributions in time. We initiated the process by suppressing the first peak, which was all most negligible,

until we saw how all the odd terms disappeared (second row of Fig.3.22). As we systematically turned off contributions by configuring the MZIs in the bar or cross positions, the spectrum became smoother, and the contributions gradually vanished. This iterative process was repeated for the second and third MZIs, ultimately leaving us with only the contributions of the last two delay stages (third row of Fig.3.22). Unfortunately, due to limited ER, thermal crosstalk, and electrical precision, achieving a fully flattened spectrum is impossible, resulting in small, uncorrelated noise variations in both frequency and time domains. In our experimental demonstration, the best result we achieved exhibited a standard deviation in the optical spectrum of 0.4 dB compared to the ideal value.

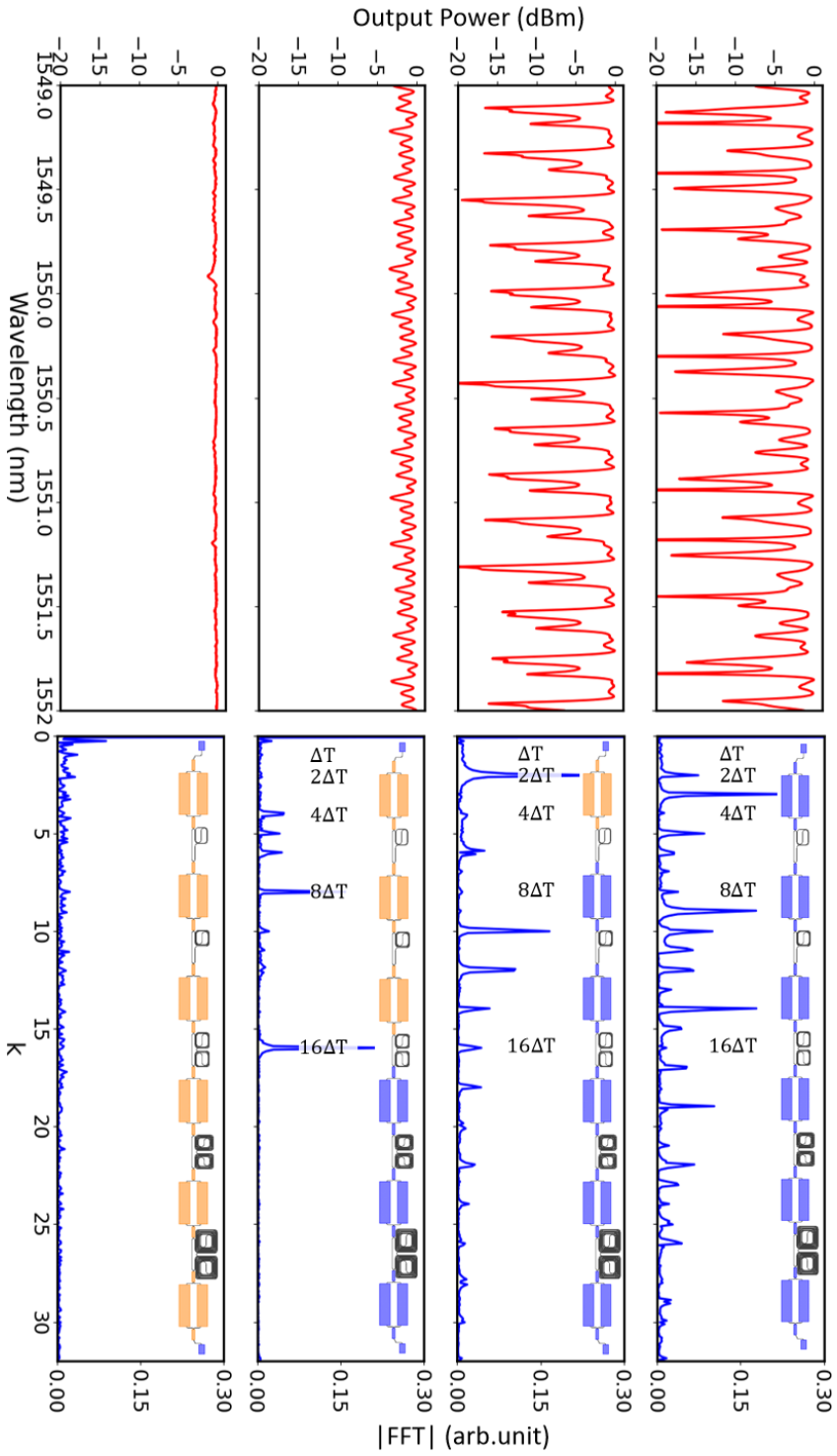


Figure 3.22: Experimental measurements of OTTDL calibration. Each row in the figure shows the spectrum, its Fourier transform, and the corresponding configuration of the MZIs. The MZIs that remain in their original random state are highlighted in blue, while the ones in the bar/cross state are highlighted in orange. As the entire delay line transitions to the bar/cross state the spectrum becomes completely flat, devoid of any peaks in its transform.

After achieving a this flattened optical spectrum, the next step involved measuring the phase and group delay for all the bar/cross configurations, thus 32 different positions. Utilizing a phase-shift approach, the RF-to-RF response was measured as the signal traveled through the integrated circuit, thereby providing information of the delay suffered by the signal for each line configuration. The input RF-modulated signal was generated by a VNA Agilent N5245A capable of modulating signals up to 50 GHz. With a lightwave component analyzer (LCA) module N4373C we upconverted the signals to the optical domain using a 1550 nm internal source. Afterwards, the output signal from the OTTDL was also downconverted using the same optical module from the VNA. The experimental setup is shown in Fig. 3.23(d).

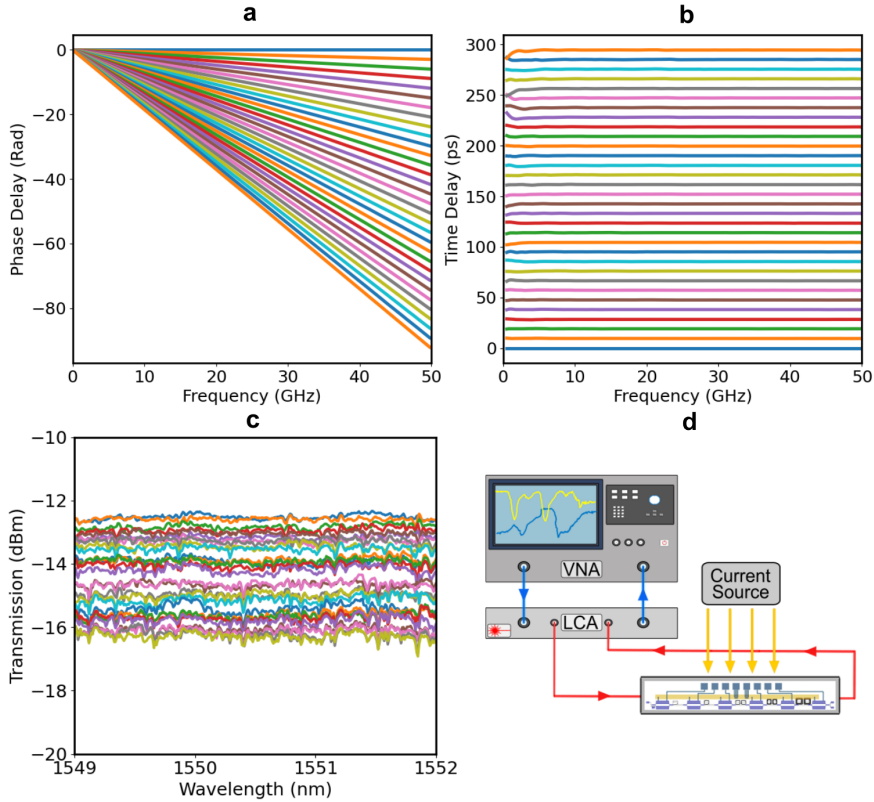


Figure 3.23: **a** Microwave phase and **b** group delay responses for all 32 delay states of the OTTDL. **c** Transmission spectra for all bit configurations. **d** Experimental setup.

Figure 3.23 shows the phase, time, and optical response of the line across its 32 delay states. We extract the applied delay across a broad spectrum of frequencies from the RF phase measurements. Once we have all the delays ordered in ascending order, it becomes straightforward to correlate the set of currents applied for each delay with the bar or cross position of the MZIs.

In Fig. 3.23(c), the measured optical transmission response is depicted, with the power difference between bit configurations attributed to variations in propagation and bend losses across each stage. The obtained delay values exhibit minimal deviation from

the theoretical delays, with an average discrepancy of only $\pm 0.16\text{ps}$, as shown in Fig. 4.18.

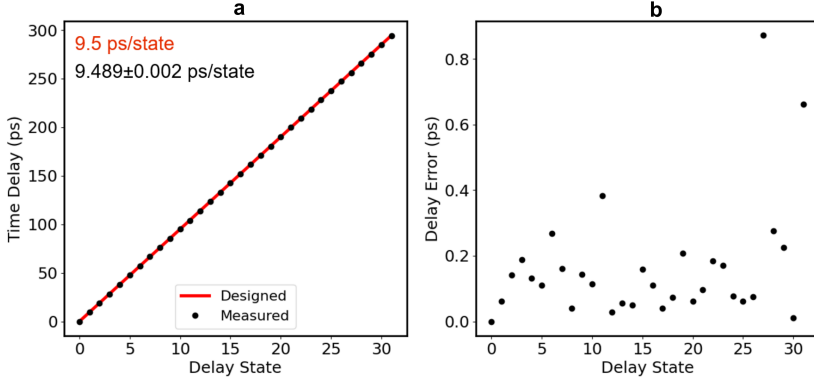


Figure 3.24: **a** Delay measurement results using the calibration method compared with the designed values. **b** The delay time errors of 32 delay states

Once we determine the precise locations of all the bar and cross states, we isolate each MZI one by one for independent characterization, as illustrated in Figure 3.25. We consistently measure the bar transmission of the OTTDL, ensuring all MZIs are set to the bar state except for the one being characterized, whose current is going to be swept. Figure 3.25(a) shows the characteristic curve of one of the MZIs of the line, demonstrating a large ER, an encouraging indication of the accurate outcome of the characterization method.

Utilizing these curves, we executed an adjustment of the coupling coefficient vs electrical current, $\kappa(I)$, a crucial step for the following synthesis of filters within the line. The fitting curve of the coupling coefficient can be obtained from Eq.(3.28) and Eq.(3.20). We compare both bar transmission terms for optical power, $|H_{11}|^2$, and fit them to the applied current, considering that the induced phase shift on the thermo-optic actuators follows a quadratic formula,

$$\kappa(I) = 1 - \sin^2 (\alpha_0 + \alpha_1 I + \alpha_2 I^2). \quad (3.24)$$

Application for optical filter synthesis

We have adapted the synthesis of lattice-based filters developed in Madsen's work [28] to the simplified design of our OTTDL. This design, specifically tailored for beamforming applications, omits phase shifters between the delay stages, thereby limiting flexibility in terms of filtering design.

The following equations construct the transfer matrix after the n^{th} delay stage, where $c_n = \sqrt{1 - \kappa_n}$, $s_n = \sqrt{\kappa_n}$ and $m = 2^{(n-1)}$,

$$\begin{bmatrix} X_n(z) \\ Y_n(z) \end{bmatrix} = \Phi_n(z) \begin{bmatrix} X_{n-1}(z) \\ Y_{n-1}(z) \end{bmatrix}, \quad (3.25)$$

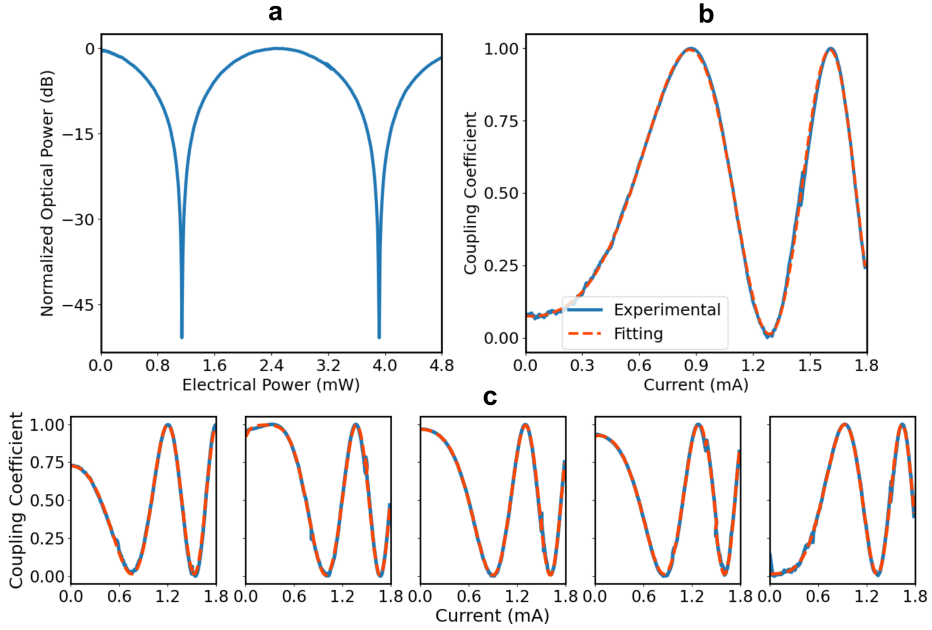


Figure 3.25: **a** MZI transfer function depending on the electrical power applied on the thermo-optic actuator. **b** Coupling coefficient fit depending on the given current. **c** Fittings of all the MZIs in the line.

$$\Phi_n(z) = \begin{bmatrix} c_n & -is_n \\ -is_n & c_n \end{bmatrix} \begin{bmatrix} z^{-m} & 0 \\ 0 & 1 \end{bmatrix}. \quad (3.26)$$

Iteratively applying the equations from a single delay stage to five, we derive the total transfer function in the Z-domain for our OTTDL,

$$\begin{bmatrix} X_{out}(z) \\ Y_{out}(z) \end{bmatrix} = \Phi_5 \Phi_4 \Phi_3 \Phi_2 \Phi_1 \Phi_0 \begin{bmatrix} X_{in}(z) \\ Y_{in}(z) \end{bmatrix}, \quad (3.27)$$

then, we utilize an optimization algorithm based on standard minimization of the filter response in comparison to an ideal mask.

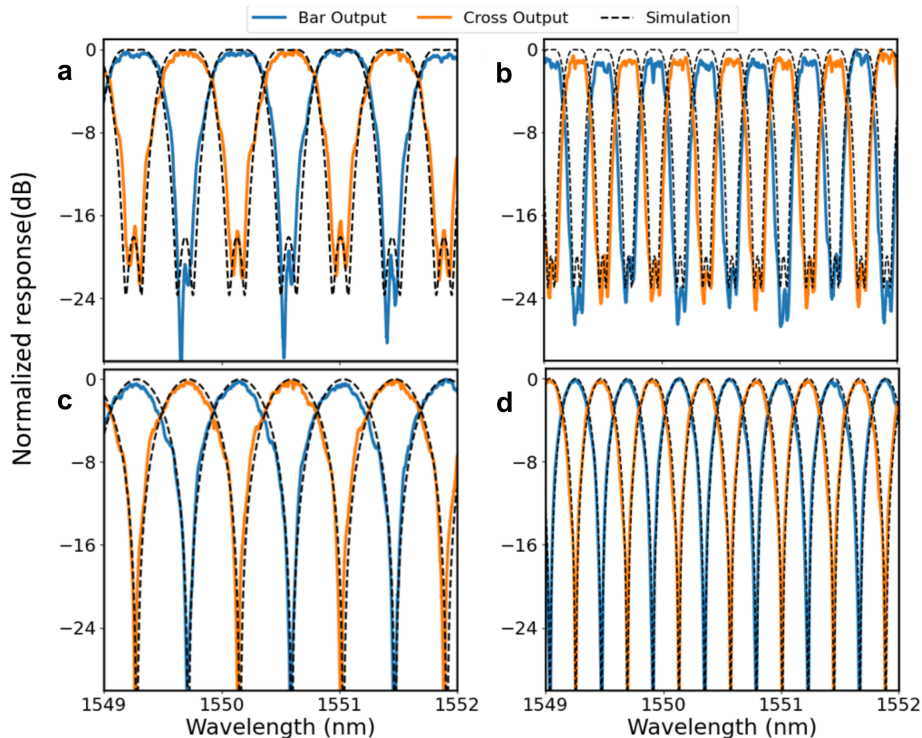


Figure 3.26: **a** Measured bandpass filter with 52 GHz of bandwidth. **b** Measured bandpass filter with 26 GHz of bandwidth. **c** Measured notch filter with 106 GHz of FSR. **d** Measured notch filter with 53 GHz of FSR.

Table 3.4: Filter Parameters

Filter	κ_0	κ_1	κ_2	κ_3	κ_4	κ_5
Bandpass (52 GHz)	0.5	0.6	0.04	0	0	0
Bandpass (26 GHz)	0	0.5	0.6	0.05	0.001	0
Notch (106 GHz)	0.5	0.5	0.002	0	0	0
Notch (53 GHz)	0	0.5	0.5	0.005	0	0

This fine-tunes the couplings of the MZIs, ensuring that the final response aligns with a desired filter mask, within the design parameters if feasible. The calculated coupling coefficients are then applied to the lattice filter making use of the previous characterization fittings. This enables us to synthesize interleavers that are based on notch and bandpass filters, shown in Fig.3.26. Table 3.4 presents the corresponding coupling coefficients for each filter. The results are very close to the theoretical values, demonstrating the accurate and precise performance of our proposed characterization method.

3.2.5 Conclusions

We have introduced a calibration method tailored for switched OTTDL, marking a significant advancement by eliminating the requirement for VOAs or dedicated test ports. This strategic elimination not only optimizes power consumption but also contributes to a more efficient utilization of device footprint. Through experimental validation, we have successfully demonstrated the efficacy of this calibration method on a silicon photonic chip. We successfully characterized an OTTDL and, leveraging the results obtained through this characterization, we extended the application to configure tunable optical interleavers. By introducing this approach, we pave the way for the development of more compact and simplified designs, addressing the critical need for miniaturization in next-generation communications technologies.

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3.3 Paper C

APL Photonics, (2025)

Integrated Microwave Photonics True-Time Delay Signal Processor

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Integrated Microwave Photonics True-Time Delay Signal Processor

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Silicon photonic integrated circuits offer significant improvements in processing bandwidth, power efficiency, and low latency, addressing the needs of future microwave communication systems. Several successful applications have been demonstrated in this field; however, the focus is now shifting toward integrating these applications into single programmable photonic circuits. This approach not only reduces fabrication costs but also makes photonics more accessible for everyday use. This paper presents a scalable silicon-based signal processor with advanced functionalities, including high-speed arbitrary waveform generation, tunable bandwidth filtering and ultrabroadband beamforming. These results highlight improvements in both scale and performance, representing a significant step forward in large-scale, high-performance, multifunctional photonic systems.

3.3.1 Introduction

The rapid growth of data volumes in communications, AI training, and cloud computing demands efficient data processing. This data is usually stored as digital electrical information and transmitted as wireless radio frequency (RF) signals. However, modern communication systems and computational electronics may struggle to handle the increasing workload, making it essential to explore alternative processing techniques. Microwave photonics (MWP) have emerged as promising solutions to these challenges. These systems harness the unique advantages of optics—such as flexibility, high bandwidth, and low power consumption—to process millimeter-wave signals directly in the optical domain [1], [2]. These techniques fit perfectly with the recent revolution in fabrication methods for integrated photonic circuits (PICs), which allow the creation of large, complex systems with exceptional precision [3]–[5]. This breakthrough has sparked significant interest across a wide range of research areas, including emerging computational paradigms such as quantum [6], [7] and neuromorphic computing [8], [9], advanced sensing and spectroscopy systems [10], [11], and the aforementioned optical signal processing technologies [12], [13].

Integrated microwave photonic circuits offer flexible, broadband analog solutions that bridge the limitations of digital signal processors (DSPs), currently constrained to a few gigahertz. These circuits are designed to meet the growing demands of modern telecommunications, driven by advancements in Internet of Things, 5G, 6G, and cutting-edge radar systems [14]–[16]. Many key RF functions have been successfully demonstrated using on-chip photonic signal processors, such as spectral filters [17], RF phase shifters [18], [19], integrators, differentiators [20], pulse shapers and microwave generation [21],

[22], and beamformers [23], all delivering exceptional performance. However, despite their exceptional performance, most of these circuits are still custom-designed for specific tasks. The future of this field lies in multifunctional integration, combining programmable functions onto a single chip to reduce fabrication costs and accelerate commercial implementation [24], [25].

In this work, we propose a scalable design for a photonic signal processor chip based on optical true-time delay (TTD). This device aims to address three major applications of MWP: signal generation, filtering and beamforming. Our goal is to achieve a simple, universal, and versatile design while maintaining high performance to enhance future telecommunication networks.

In Section 3.3.2, we present the proposed architecture, detailing its general implementation. This section also introduces the fabricated silicon chip used for the practical implementation and provides an explanation of the measurement setup and procedures. Moving forward, in Section 3.3.3 we introduce the first application of the architecture, an ultrafast RF arbitrary waveform generator. Then, in Section 3.3.4 we demonstrate the second application of the architecture: a programmable filter with variable bandwidth. This section showcases a wide range of filter implementations, all achieved using the same device. In Section 3.3.5, we present the results of the architecture’s final functionality, the beamforming network. In Section 3.4, we analyze the system’s scalability, and finally, in Section 3.4.1, we summarize the findings and present our conclusions.

3.3.2 Architecture Design

True-time delay is a widely used technique in microwave photonics for implementing beamforming networks that are free from beam squint. There are two main approaches to realizing TTD on integrated devices: microring resonators, which offer continuous delays but are sensitive to thermal fluctuations and fabrication errors, and switched delay lines, which are more robust and provide broader bandwidth, though they are limited to a discrete set of delay positions. For our design, we opted for the latter, aiming to maximize the number of achievable delay settings while ensuring high tolerance to deviations and imperfections. The architecture presented in this paper builds upon our recent work [26], which was originally designed as a beamforming network. In that study, we introduced a novel beamformer that minimized the number of delays while maximizing resolution, significantly enhancing resource efficiency compared to previous beamforming networks based on switched true time delay lines. However, despite these advancements, the architecture did not fully meet the increasing demands for multifunctional integration and flexibility. To overcome these limitations, the design proposed in this work retains the high-resolution, ultra-broadband performance of the original beamformer while incorporating additional functionalities—such as filtering and arbitrary waveform generation (AWG)—that were not achievable in the previous design. This is accomplished by incorporating a combined output at the back-end of the structure, formed by Multi-Mode Interferometers (MMIs).

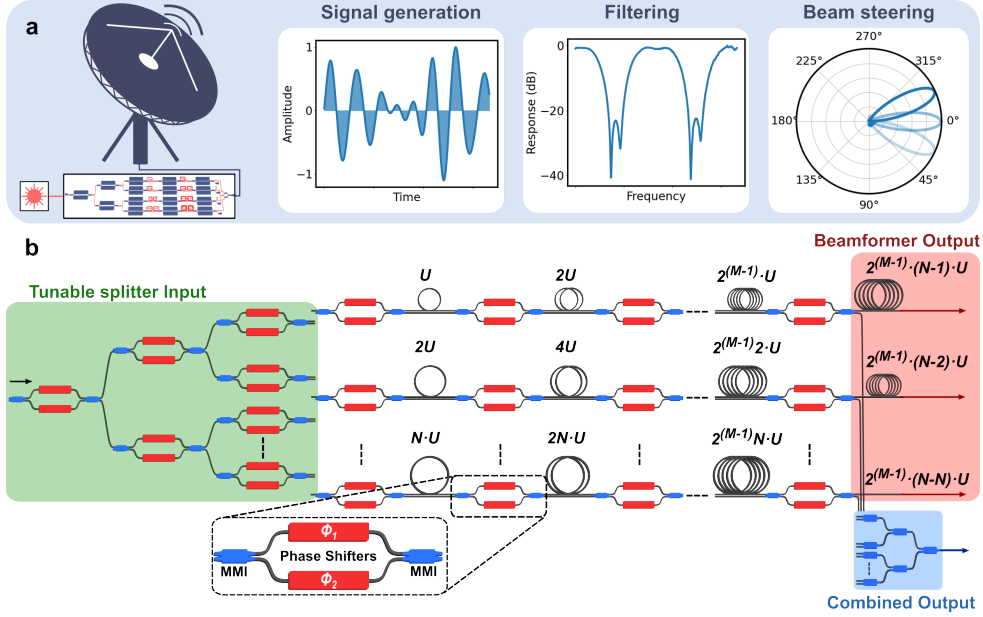


Figure 3.27: (a) Conceptual illustration of the proposed silicon photonic signal processor designed for multipurpose photonic processing applications. (b) General schematic of the proposed architecture featuring N lines and M delay stages. The tunable splitter tree at the input, highlighted in green, distributes light across the lines. MZIs guide the light through the delay stages, finishing at the backend, where they are used for switching between the two possible outputs: the beamformer output (highlighted in red) and the combined output (highlighted in blue).

The general scheme of the proposal is depicted in Fig.3.27. The main building block of this design is the tunable coupler, implemented here as Mach-Zehnder Interferometers (MZI) formed by two 2x2 MMIs with a phase shifter section between them. In this work, the phase shifter operates using the thermo-optic effect, however, for faster applications, the electro-optic effect could be considered, as demonstrated in recent advancements in programmable photonics, where lithium niobate is used as the fabrication platform, enabling high speeds while maintaining low losses [27], [28]. The general transfer function of a MZI can be expressed as follows:

$$H_{MZI}(\phi_1, \phi_2) = ie^{i(\phi_1 + \phi_2)/2} \begin{bmatrix} \sin\left(\frac{\phi_1 - \phi_2}{2}\right) & \cos\left(\frac{\phi_1 - \phi_2}{2}\right) \\ \cos\left(\frac{\phi_1 - \phi_2}{2}\right) & -\sin\left(\frac{\phi_1 - \phi_2}{2}\right) \end{bmatrix}, \quad (3.28)$$

here, i represents the imaginary unit, while ϕ_1 and ϕ_2 denote the phase shifts applied to each arm of the MZI. This matrix establishes the relationship between the amplitude and phase modifications of the electromagnetic fields at the interferometer's outputs and their corresponding values at the inputs. By adjusting the phase shifts in the segment between the MMIs, ϕ_1 and ϕ_2 , we can adjust the overall phase shift of the optical carrier at the output, which is highly beneficial for filter synthesis. More importantly, we can simultaneously vary the power splitting ratio between the arms of the MZI, a feature that enables the creation of diverse beam patterns or the selection of different windows

functions. Generally, when operating the core of the device, the tunable couplers are typically configured in either the bar or cross state. This configuration is well-suited for representing binary notation, with each state commonly encoded as a bit word.

At the front-end of the structure, there is a tunable splitter tree which depends on the number of total lines, N . This setup allows for customized power distribution to each line, rather than a uniform distribution across all lines, without introducing additional losses. Afterwards, connected to each output of the splitter tree, there is a delay line with M delay stages or bits, where the delay values increase exponentially in powers of 2 for each bit with respect to a fundamental delay, U . As a specific case of this architecture, the delay value also scales with the position of the line. For instance, the delay value for the second line is doubled compared to the first line, and for the N^{th} line, it is multiplied by N . This results in 2^M unique combinations of delays between lines at the output of the final MZI, with the total delay between antennas given by $\Delta T = BU$, where B is the selected bit word ranging from 0 to $(2^M - 1)$.

The final MZI manages the switch between the two distinct outputs of the architecture: the beamformer and the combined output. While the combined output simply transfers the original delays achieved in the device's core to the common combiner formed by MMIs, the waveguide of the beamformer output introduces an additional delay depending on the line position, given by $\epsilon_n = (N - n)2^{(M-1)}U$. This adjusts the range of delays between antennas while preserving all combinations, shifting the value of ΔT from $[0, (2^M - 1)U]$ to $[-2^{M-1}U, (2^{M-1} - 1)U]$. This adjustment allows the beamformer to aim without beam squint effects across both positive and negative angles, maintaining constant delays between antennas in both directions of the array while maximizing the number of delay stages to achieve the highest possible resolution.

Experimental demonstration

For the experimental demonstration, we included four delay lines, each with two delay stages (2 bits of resolution), fed by a tunable splitter tree, with outputs connected using a fixed combiner consisting of three 2x1 MMIs. The fundamental delay, U , was designed to be 10 picoseconds, slightly larger than the width of the pulse-shaped laser available in our laboratory, which was intended for measuring the AWG. This consideration is crucial to prevent the laser pulses from overlapping at the combiner. For this measurement, we used high-speed fiber-pulsed source that produced a train of Gaussian pulses with a width of 4.4 ps and a repetition frequency of 5 GHz. The repetition rate was chosen to be sufficiently low to ensure that there is no overlap between pulses at the chip's output.

We designed a photonic integrated circuit that was manufactured by Advanced Micro Foundry on a Silicon on Insulator platform. The layout and a micrograph of the chip are presented in Figure 4.12. The chip was produced using an SOI wafer with a 220 nm slab thickness, with 500 nm single-mode waveguides that were defined through deep ultraviolet lithography (193 nm). For the sections containing phase shifters, a thin heater layer of 120 nm TiN was deposited over the waveguide giving as a result MZIs with a measured power consumption of only $1.35mW/\pi$. The thermo-optic phase shifters were controlled electronically using Qontrol programmable power sources, managed via software.

The optical input and output of the chip were achieved through vertical coupling using grating couplers, with coupling losses of 3 dB. These losses were normalized during the measurements, isolating propagation and insertion losses as the primary loss mechanisms

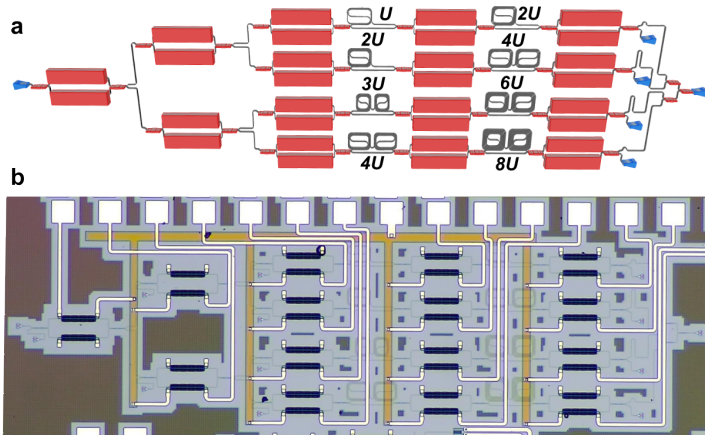


Figure 3.28: (a) Layout of the experimental demonstration chip. (b) micrograph of the fabricated device on a SOI platform (4mm x 1.2mm).

within the chip. The propagation losses were approximately 1 dB/cm, while the insertion losses for each MZI were around 0.4 dB.

3.3.3 AWG application

Ultrafast optical and radio-frequency waveforms with bandwidths reaching tens to hundreds of gigahertz have potential applications in numerous fields, including high-speed optical communications, radar, microwave imaging, and instrumentation test measurements [29]. However, generating high-frequency (e.g., 20–80 GHz) and ultrabroad-bandwidth arbitrary waveforms through electronic means is challenging due to the limitations of digital-to-analog converter technology and significant timing jitter [30], [31]. Because of that, AWG has become a focal area in microwave photonics due to the ability of optical techniques to produce high-frequency and large-bandwidth signals. The most common methods for generating arbitrary waveforms in photonics rely on bulky frequency-to-time mapping techniques. This process involves adjusting a broadband source using a spectral filter and then passing the broadband signal through a dispersive element. In the dispersive element, different wavelengths are delayed relative to each other, resulting in their arrival at a photodetector at different times, thus creating the desired waveform [21], [32].

Here, we demonstrate a compact device capable of generating arbitrary waveforms using a temporal pulse shaping system synthesized in the time domain. This approach eliminates the need for external dispersive elements while keeping control over both the shape and frequency of the generated waveform. This is achieved by generating delayed and scaled replicas of short input optical pulses in the optical domain, with each replica experiencing a different time delay. Through precise programming, we ensure that the four output pulses are evenly spaced in time with delays of U , $2U$, or $3U$, depending on the bit configuration shown in Table 3.5. Finally, the delayed pulses from all lines are combined at the common output, generating various RF wavepackets. The splitter tree enables the application of different weights, facilitating the creation of complex

waveforms.

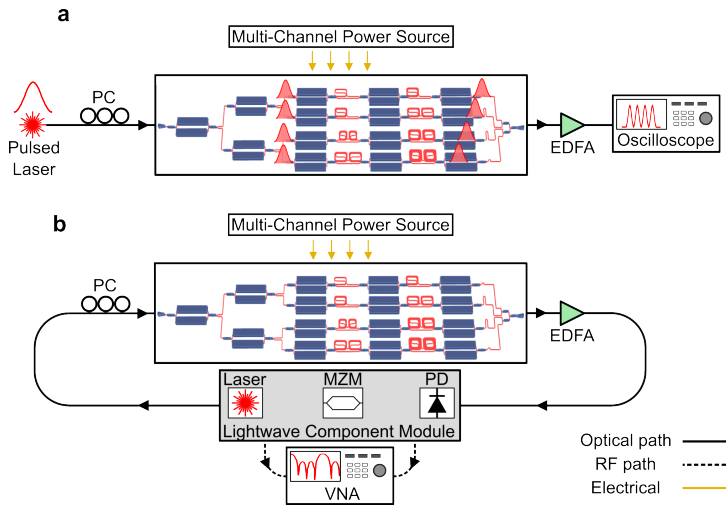


Figure 3.29: (a) Schematic diagram of the experimental setup used for AWG measurements. (b) Schematic diagram of the experimental setup used for delay characterization and filter measurements. PC, polarization controller; MZM, Mach-Zehnder modulator; EDFA, erbium-doper fiber amplifier; PD, photodetector.

	Bit (00)	Bit (01)	Bit (10)	Bit (11)	ϵ
Line 1	0	U	2U	3U	+6U
Line 2	0	2U	4U	6U	+4U
Line 3	0	3U	6U	9U	+2U
Line 4	0	4U	8U	12U	0

Table 3.5: Delay values applied to each line for all bit configurations, including the additional fixed delay used in the beamforming application.

Figure 3.29 depicts the two measurement schemes used in this work. Figure 4a provides details of the AWG measurement setup. For the 4.4 ps pulse generation, we used a high-speed fiber laser (Calmar Optcom PSL-10-6T), while the time response was measured using the optical sampling module of our Digital Serial Analyzer Sampling Oscilloscope (DSA8200), which supports measurements up to 100 GHz. Time delay and transmission measurements for filter characterization in the following sections were performed using a Vector Network Analyzer (VNA) Agilent N4373C, paired with a Lightwave component module for measurements up to 50 GHz, as shown in Figure 3.29(b). This module integrates an optical modulator and photodetector, enabling direct electrical-to-optical up-conversion within a single device.

First, we characterized the processor by precisely controlling the optical coupling of all the MZIs in the design with respect to the applied electrical power. After completing the MZI characterization, we proceeded with delay measurements using the VNA to verify that all delays were accurately implemented. Figure 3.30(a) presents the results obtained from measuring the RF transmission, phase and time responses. The time

and delay measurements were highly satisfactory, aligning perfectly with the theoretical predictions.

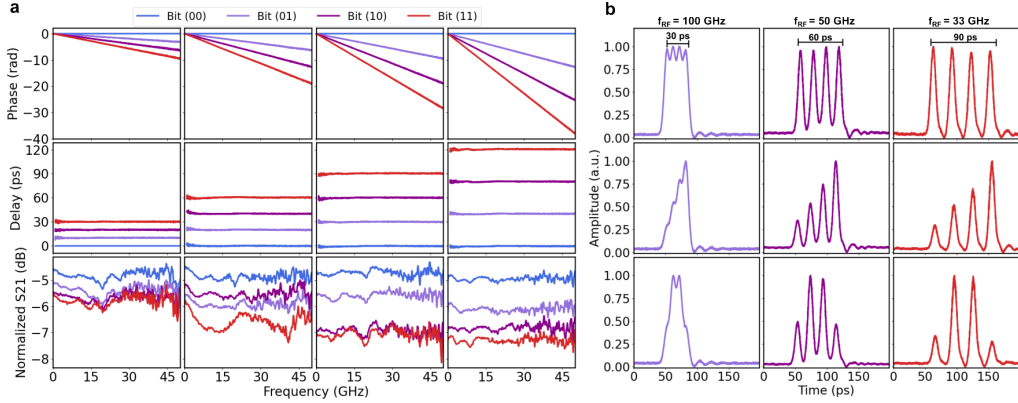


Figure 3.30: (a) Results from the RF characterization of the delay lines. Each column represents measurements from the first to the last delay line in the architecture, while the colored lines indicate the different bit configurations of all the lines. (b) Measured waveforms for different time separations between peaks and shapes, including rectangular, sawtooth, and Gaussian profiles.

The RF transmission spectra exhibit some noise, which may be attributed to the finite extinction ratio of the MZIs, thermal crosstalk, and electrical variations caused by the use of multi-contact probes. However, these variations are relatively minor, remaining within 1 dB, and are expected to improve significantly in fully packaged devices. On the other hand, the losses progressively increase from one line to the next due to propagation losses, which are inherent and unavoidable. This effect, however, was accounted for and mitigated using the tunable splitter tree, which compensates for the higher losses by distributing more optical power to the affected lines.

In our proof-of-concept design, the total number of delay combinations is four, as summarized in Table 3.5. The first bit word is not relevant for this application, as the lack of delay between pulse replicas causes them to arrive simultaneously at the combiner, avoiding any waveform generation.

Figure 3.30(b) displays the measured optical waveform on our chip before photodetection. The separation between the peaks is inversely related to the resulting radio frequency signal generated on a high-speed photodetector. Using our fabricated device, we can produce optical replicas with three distinct separations, corresponding to generated RF center frequencies of 33 GHz, 50 GHz, and 100 GHz, depending on the selected bit word: (11), (10), and (01), respectively.

3.3.4 Filter application

Microwave filters are crucial components in RF links and are among the applications that benefit most from photonic integration, as photonics enhances these systems through RF-to-optical frequency upconversion and flexible optical-domain filtering. Recent advancements in integrated optics present significant opportunities for innovative filter designs, including both infinite impulse response (IIR) and finite impulse response (FIR) filters,

each offering distinct characteristics that make them well-suited for specific performance requirements and applications [33].

In this work, we demonstrate two types of FIR filters achievable with our proposed architecture. The first type, multi-tapped delay filters, also known as transverse filters, are inspired by the digital filter concept commonly used in RF systems and represent the most traditional approach for implementing MWP filters. In a transverse filter, the input signal undergoes discrete sampling, followed by delaying and weighting, before being summed. All these steps can be implemented using optics, creating a transfer function described by:

$$H(\omega) = \sum_{n=0}^{N-1} a_n e^{-i\alpha_n} e^{-in\omega\Delta T}, \quad (3.29)$$

where a_n is the weight of the sampled signal at the n -th tap, α_n is the carrier phase shift of the n -th sample, and ΔT is the intertap signal time delay. The frequency selectivity of the filter is primarily governed by the number of effective taps, the inter-tap delay sets the frequency periodicity with the free spectral range (FSR) given by $1/\Delta T$, and the phase coordination establishes the spectral positions of both the filter's passbands and stopbands [17].

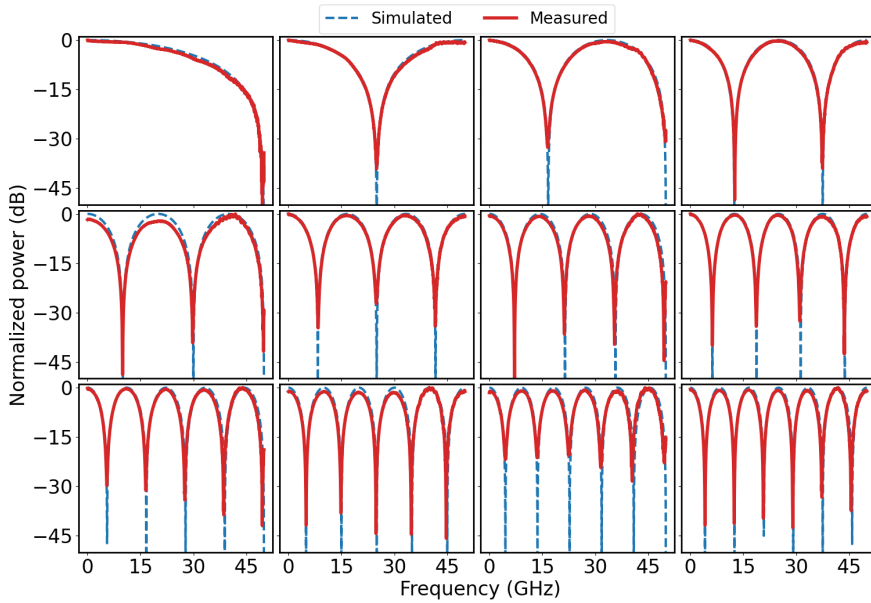


Figure 3.31: Comparison between simulated and measured results for all 12 two-tap filter combinations, arranged in order of increasing ΔT values, from the smallest to the largest.

Thanks to the flexible splitter tree at the system's front end, we can select between different feeding modes, simultaneously driving two, three, or all four lines, enabling a wide variety of filter shapes and bandwidth combinations. When using only two lines, up to 12 distinct intertap delays can be configured, ranging from U to $12U$. This flexibility arises because the only constraint is the maximum achievable delay between the two taps,

a far less restrictive condition than in larger filters with three or four taps. In those cases, maintaining a constant delay between repetitions becomes crucial, significantly limiting the number of possible combinations. For example, to achieve an intertap delay of $12U$ with two taps, we simply split the signal between a zero-delay line and the fourth line, which provides the maximum delay of $4U + 8U$.

Consequently, for two taps, the free spectral range varies resulting in a tunable notch filter with an FSR ranging from 100 GHz when $\Delta T = U$ to 8.3 GHz when $\Delta T = 12U$. Figure 3.31 shows all the different filter combinations of two taps.

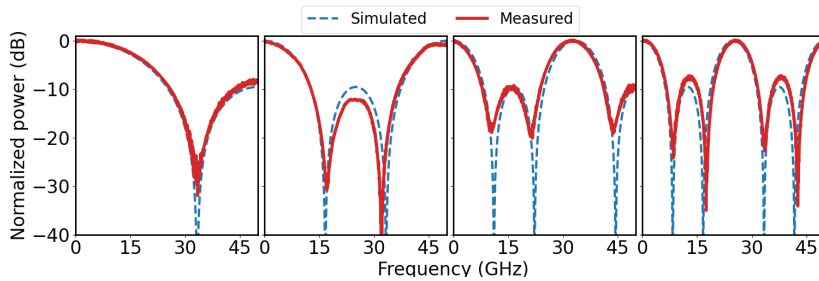


Figure 3.32: Comparison of simulated and measured filter responses with three lines in use.

Similarly, by feeding three lines, we obtain an additional tap for the filter. In this configuration, there are only four possible delay combinations, ranging from $\Delta T = U$ to $\Delta T = 4U$. As a result, the filter's response transitions from a notch filter to a typical transverse filter, characterized by variable bandwidth. Specifically, we measured 3 dB bandwidths of 32.1 GHz, 14.8 GHz, 8.9 GHz, and 8.45 GHz, corresponding to FSRs of 100 GHz, 50 GHz, 33.3 GHz, and 25 GHz, respectively. The results from the measurements are shown in Fig. 3.32.

Finally, all lines can be used simultaneously, resulting in three possible intertap delays while maintaining uniform spacing, as shown in Table 3.5. This results in narrower filter bandwidths while preserving the same FSR. Figure 3.33 depicts the results for the four-tap measured filters.

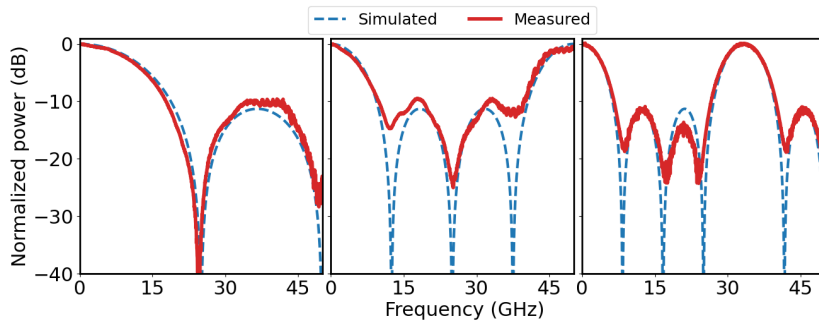


Figure 3.33: Measured filter responses with all lines fed simultaneously, showing bandwidths of 21.5 GHz, 10.3 GHz, and 7.8 GHz.

Table 3.6 presents a comparison between simulation and measurement for key multi-tap filter parameters. Specifically, we analyze the average error and deviation for the notch positions, 3-dB bandwidth, and Main Lobe to Secondary Lobe Ratio (MLSR) across all filters, categorized by their tap count. Notch deviations remain below 1 GHz, indicating a strong similarity between the measured and ideal filter responses, although higher-order filters exhibit more discrepancies. The 3-dB bandwidth relative error increases at lower tap due to inaccuracies in synthesizing narrower filters, but in all cases, the error remains below 10%. Furthermore, the deviation of less than 2 dB in the MLSR of the higher-order filters reinforces the notion of solid results.

	Notch deviation	3-dB Bandwidth relative error	MLSR deviation
Two-taps Filter	0.16 GHz	8.4 %	-
Three-taps Filter	0.6 GHz	7.8 %	1.5 dB
Four-taps Filter	0.65 GHz	6.1 %	1.4 dB

Table 3.6: Comparison of the average deviation in multi-tap filter parameters between simulations and experimental measurements.

Additionally, the tunable splitter tree at the input of the architecture allows for flexible distribution of optical power across all the lines. This capability enables the implementation of window functions, a common tool in signal processing, within the synthesized filters. By applying them we can enhance the MLSR of the filter, as illustrated in Fig.3.34. However, this improvement comes with a slight trade-off, as it increases the bandwidth of the low-pass band.

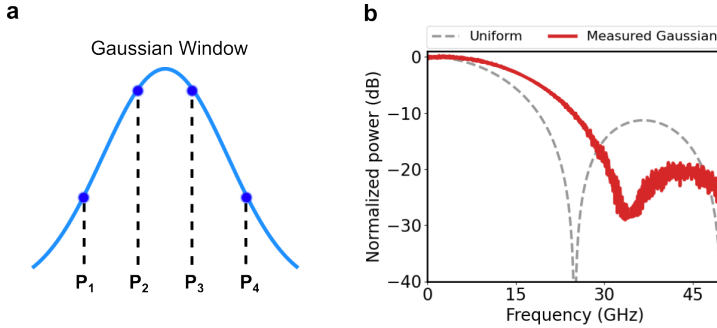


Figure 3.34: (a) Illustration of a Gaussian window applied to a four-tap filter, where the optical power, P , distributed across each line follows the profile of a Gaussian function. (b) Measurement results for the filter with an intertap delay $\Delta T = U$, comparing the performance of a Gaussian window with uniform power distribution.

The second type of filters that can be implemented with this approach are two-stage lattice filters. In this configuration, the light is routed through a single line, and by adjusting the coupling strength of each MZI (rather than strictly using the bar or cross states as in the previous case), it is possible to achieve different band-pass filter responses. The general transfer function for a filter implemented on the N^{th} line, with M delay stages

(corresponding to $M + 1$ MZIs), is given by:

$$H(\omega) = H_{MZI_{M+1}} \prod_{m=1}^M \begin{bmatrix} e^{-i\omega 2^{(m-1)}NU} & 0 \\ 0 & 1 \end{bmatrix} H_{MZI_m}. \quad (3.30)$$

It is straightforward to observe that for identical coupling configurations of the MZIs across different lines, the only distinction in their transfer functions lies in their FSR, associated with the value of the smallest delay in the line, given by $FSR = 1/(NU)$. The common phase can be ignored because it will be absorbed by the photodetector, simplifying the analysis. The transfer function of the MZIs from Eq.(3.28) can be simplified to that of a tunable coupler using the relations: $\sin(\Delta\phi_m/2) = \sqrt{1 - \kappa_m}$ and $\cos(\Delta\phi_m/2) = \sqrt{\kappa_m}$, depending only on the coupling coefficient κ .

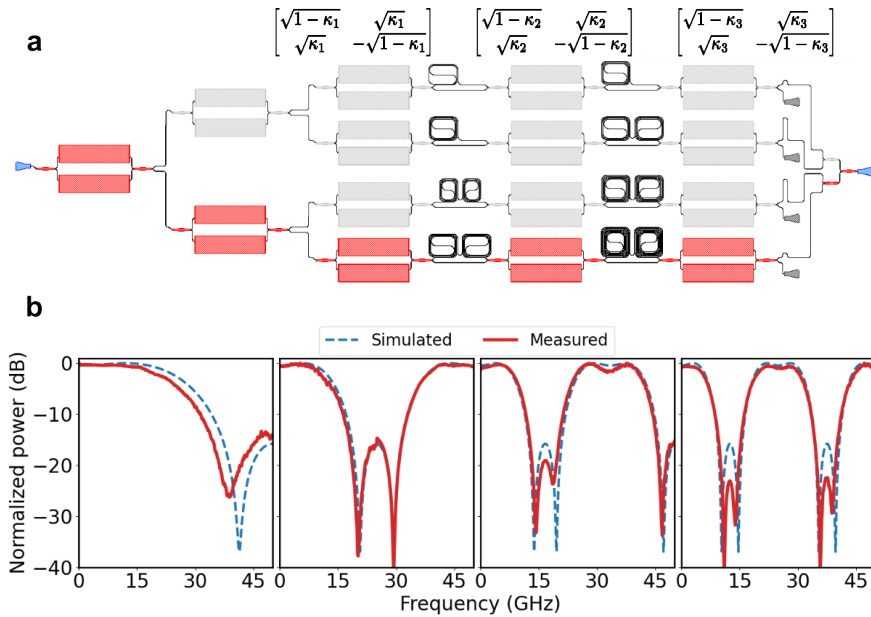


Figure 3.35: (a) Schematic illustrating the method for driving the fabricated chip to synthesize a lattice filter using the fourth line. (b) Results from implementing the same filter across all lines using a consistent set of coupling coefficients: $\kappa_1 = 0.8$, $\kappa_2 = 0.7$ and $\kappa_3 = 0.5$.

This concept is illustrated in the upper part of Figure 3.35, which also demonstrates the implementation of one of these lattice filters in the fourth line of our device. Additionally, the figure presents measurement results using the same set of coupling coefficients applied across all the lines, demonstrating that while the filter shape remains consistent, the FSR varies.

3.3.5 Beamforming application

The final demonstrated application is the operation of the device as an optical beamforming network. This functionality was extensively covered in our previous work [26]; therefore, additional compensation delays were omitted in this design to optimize space

in the proof-of-concept device. Instead, we prioritized demonstrating the other applications, as the added waveguide length used for beam reorientation does not introduce measurement challenges or affect the results. Thus, for beam reconstruction we assume the presence of additional delays, as described in Section 3.3.2.

To evaluate the performance of the integrated beamformer, we calculate the far-field pattern of the emitted beams using the delay measurements from each channel and the array factor equation for uniform feeding, which is expressed as follows:

$$F(\omega, \theta) = \sum_{n=0}^{N-1} \frac{1}{N} e^{-i\omega \left(\frac{d}{c} n \sin(\theta) - T_n \right)}, \quad (3.31)$$

where T_n represents the total delay at the output of the n^{th} line and d the spacing between adjacent antenna elements. In this case, d was chosen to be 10 mm, ensuring that the resulting four steering angles range from -35° to 18° , while maintaining frequency independence for a maximum emission frequency of 16.5 GHz.

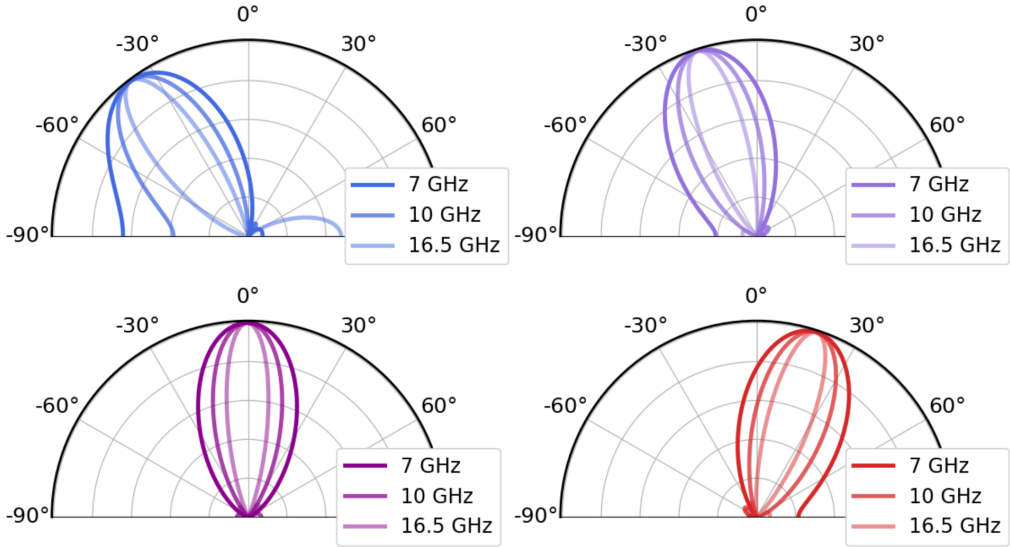


Figure 3.36: Simulated radiation patterns for all the bit configurations based on the measured delay values at emission frequencies of 7 GHz, 10 GHz, and 16.5 GHz.

Figure 3.36 shows the beampatterns for a range of frequencies from 7 GHz to 16.5 GHz. Each one of the four subgraphs represents a different delay combination, with delays between lines ranging from $-1U$ to $2U$, using the first line as the time reference. These values are determined by the bit configuration shown in Table 3.5, along with the additional delay applied to each line in the beamforming setup. The color scheme matches the delays displayed in Figure 3.30(a), to which we have added the additional delay contribution. The blue beam patterns correspond to the maximum delay of $2U$, resulting in the largest pointing angle. Next, the violet lines represent a delay of $1U$, followed by the purple lines, which depict the broadside pointing angles when all lines share the same time delay. Finally, the red curves correspond to a time delay of $-1U$.

Since the array consists of only four antennas, operating at lower frequencies is not recommended, as the beam width increases, reducing the array's directivity. On the other hand, operating at frequencies higher than 16.5 GHz is also not recommended, as it may lead to the emergence of undesired radiation lobes at the largest pointing angles.

3.4 Scalability

A key advantage of this architecture is its scalability; the number of possible filter configurations, beam steering angles, and synthesized RF frequencies with the AWG depends solely on the number of delay stages in the lines and grows exponentially as 2^M . This rapid increase is highly beneficial, as it allows a significant expansion in possible configurations across all applications with minimal additional resources in design and space for extra delay stages. However, scaling this system comes with challenges, with losses being the most limiting factor. These losses depend on the number of lines, the number of delay stages, and the length of the basic delay unit. To address this, we conducted a scalability study considering the maximum losses observed in the system, specifically, the losses experienced by light traveling through the N^{th} line while undergoing the maximum possible delay.

One of the main contributors to the total system losses is the power splitting loss, which is unavoidable in any architecture where the input signal power is divided among N outputs,

$$\textit{Splitting Losses}(dB) = 10 \log_{10}(1/N). \quad (3.32)$$

Another source of loss comes from MZI imperfections, primarily due to inefficiencies in the MMIs, which can cause reflections or inaccurate splitting ratios. However, these losses can be minimized to below 0.1 dB per MMI, resulting in a total MZI loss of approximately 0.2 dB [34]. An alternative approach is to implement MZIs using directional couplers, a solution that is gaining popularity in high-density integration due to its significantly lower losses compared to MMI-based implementations, reaching around 0.08 dB per MZI [35]. In our study, we will consider an intermediate value of 0.3 dB per MZI, between the loss values demonstrated in this work and the ideal low-loss value. The total losses from the MZIs depend on the loss factor of a single MZI and the number of delay stages, M , which contribute as $M + 1$ since the light must pass through this many MZIs to traverse all delay stages. Additionally, the losses scale with the number of lines, N , as at least $N - 1$ MZIs are required in the input splitter tree to ensure full flexibility,

$$\textit{MZI Losses}(dB) = \textit{MZI}_{\text{Loss}}(M + N). \quad (3.33)$$

Lastly, propagation losses contribute to the total system losses, depending on the propagation loss factor α and the maximum distance traveled by the light. This corresponds to the largest bit word in the last line, undergoing the greatest optical path length, which can be expressed as a function of the basic unit delay U , the speed of light c , and the group index of the fabrication platform n_g ,

$$\textit{Prop. Losses}(dB) = \alpha N 2^M \frac{Uc}{n_g}. \quad (3.34)$$

For the study we assumed a design with a basic unit delay of 5 ps, which represents an intermediate point between the 10 ps used in this proof of concept and the sub-picosecond delays typically used in some beamformer designs. This choice aims to strike a balance between different applications of the architecture: those requiring shorter delays, such as high-frequency beamforming networks, and those operating with longer delays, such as the AWG, depending on the characteristics of the Gaussian pulse used. Regarding the fabrication platform, despite the advantages of SOI due to its CMOS compatibility, it still exhibits high propagation losses, which can be significant in large-scale designs. Therefore, we have considered propagation losses from alternative fabrication platforms, such as Si_3N_4 , where losses below 0.001 dB/cm have been reported [36], significantly lower than the 1 dB/cm typically found in SOI. However, extensive research efforts are currently focused on drastically reducing propagation losses in SOI delay lines, showing promising results that could make SOI a highly relevant option in the near future [37]. For this simulation, we have assumed an intermediate value of 0.03 dB/cm, based on the TriPleX fabrication platform, which is a well-established and mature technology [38]. Figure 3.37 presents the total losses results from the study.

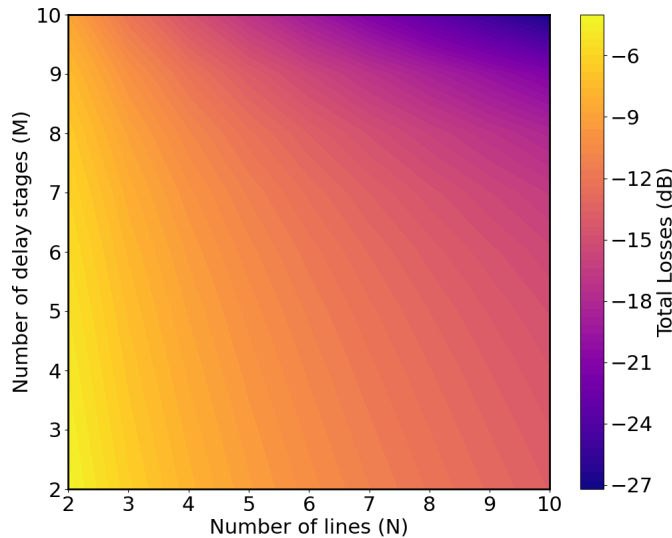


Figure 3.37: Maximum system losses calculated in the scalability study for varying numbers of lines and delay stages.

The results show that losses increase rapidly as delay stages are added, due to the exponential growth of path lengths, especially as the number of lines in the system increases, leading to significant propagation losses. However, with 8 delay stages, the losses are around -15 dB, a very reasonable value for microwave links and comparable to similar works in the literature. These losses can be mitigated through amplification, while maintaining a total of 256 delay combinations. This substantial number of combinations highlights that, despite some inherent losses, the benefits of the system outweigh the drawbacks. It is also important to note that this study does not incorporate the latest state-of-the-art parameters, which could further enhance these results and potentially enable a higher level of on-chip integration in the future.

3.4.1 Conclusion

In this paper, we have demonstrated a multifunctional and programmable architecture that operates as an arbitrary waveform generator, a tunable filter with adjustable bandwidth, and a broadband RF beamforming network. These functionalities were successfully demonstrated on a fabricated silicon chip, achieving RF signal generation with various waveforms up to 100 GHz, a reconfigurable filter capable of operating as either a notch or band-pass filter with a wide range of FSRs and bandwidths, and broadband optical beamforming from 7 GHz to 16.5 GHz with a scanning range from -35° to 18° .

The results obtained are very satisfactory given the size of the test sample; however, the design is easily scalable by incorporating additional delay stages or lines, significantly expanding the range of achievable configurations across all functionalities while maintaining a small footprint. This opens the door to potential implementation in future base stations or wireless communication systems, contributing to the advancement of next-generation communications.

Acknowledgements

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3.4.2 References

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Chapter 4

Analog Photonic MIMO processor

4.1 Multiple-Input Multiple-Output Theory

As discussed in previous chapters, phased array antennas are a major advancement in optimizing energy efficiency for telecommunications by focusing radiated power in a desired direction. However, this is not their only advantage. The use of multiple transmitters and receivers is closely linked to Multiple-Input Multiple-Output technology, a key component of 5G that plays a critical role in addressing the essential requirements of future wireless standards [53]. MIMO technology takes advantage of multiple antennas at both the transmitter and receiver to enable the simultaneous transmission of multiple data streams over the same channel through spatial division multiplexing. This significantly boosts the data capacity of wireless networks while offering two key advantages: enhanced spectral efficiency without the need for additional spectrum, and improved energy efficiency through array gain, reducing radiated power while supporting multiple users simultaneously [54], [55].

The success of this technology arises from the complex and demanding signal processing involved in both transmission and reception, along with a comprehensive analysis of the transmission channel. While coding and signal processing are crucial for the successful implementation of a MIMO system, the communication channel represents a major component that determines overall system performance [56], [57]. A considerable volume of work has been performed to characterize communication channels for general wireless applications and right now multi-user radio channel modeling is a discipline in itself. Developing a generalized channel model that accurately predicts path loss and spatio-temporal properties across varying environmental conditions and propagation scenarios is particularly challenging. Therefore, in this work, we will focus on the fundamentals, aiming to provide an introduction and foundation to better understand the results obtained using integrated photonics.

Consider a general scenario with a $r \times t$ MIMO system where t is the number of transmitting antennas and r is the number of receiving antennas. In this configuration, each receiving antenna collects signal contributions from all transmitting antennas, allowing us to model each transmission-reception link as a time-variant single-input single-output (SISO) wireless communication channel. These channels are typically characterized by multipath propagation, where signals travel multiple paths due to reflections and scattering from physical objects in the environment. As a result, the receiving antenna detects

different replicas of the original signal at various time intervals, each subject to delay and attenuation [58]. A more advanced step in studying multi-antenna wireless communication channels is to analyze a single-input multiple-output (SIMO) scenario, where one transmitter communicates with multiple receptors.

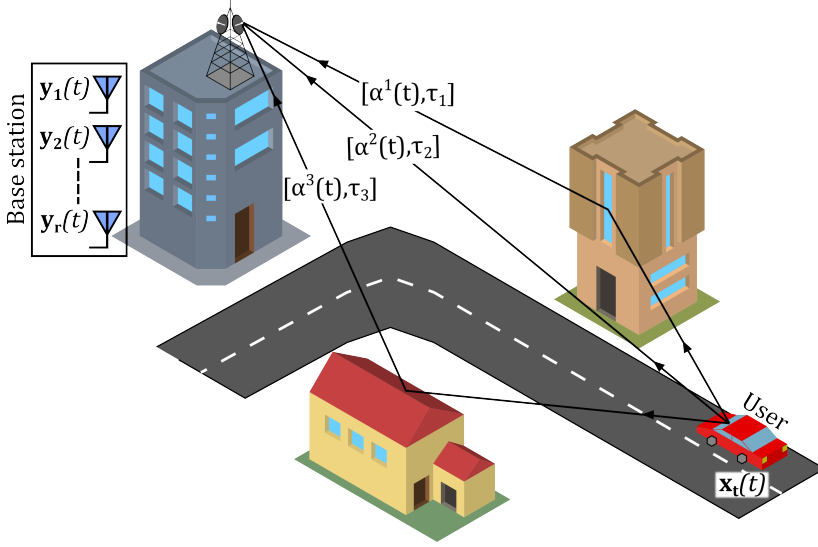


Figure 4.1: SIMO channel scheme

In this case, the channel behavior can be modeled as a vector that captures the impulse response at each receiving antenna. Extending this concept to MIMO systems is straightforward and involves scaling the model's dimensions to account for additional transmitters. Since each receiving antenna picks up signals from all transmitters, the individual links between antennas can be collectively represented by a complex MIMO channel matrix. Each column of this matrix corresponds to the vector that would represent an independent SIMO communication channel.

$$H(t, \tau) = \begin{bmatrix} h_{11}(t, \tau) & h_{12}(t, \tau) & h_{13}(t, \tau) & \cdots & h_{1t}(t, \tau) \\ h_{21}(t, \tau) & h_{22}(t, \tau) & h_{23}(t, \tau) & \cdots & h_{2t}(t, \tau) \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ h_{r1}(t, \tau) & h_{r2}(t, \tau) & h_{r3}(t, \tau) & \cdots & h_{rt}(t, \tau) \end{bmatrix}, \quad (4.1)$$

where $h_{rt}(t, \tau)$ is the time-varying impulse response of the channel between the t^{th} transmit antenna and the r^{th} receive antenna. This response can be calculated as a function of the time-varying gain coefficient of the link for each impulse corresponding to the different paths, $\alpha_{rt}^{(n)}(t)$, for a total of N impulses

$$h_{rt}(t, \tau) = \sum_{n=1}^N \alpha_{rt}^{(n)}(t) \delta(\tau - \tau_n). \quad (4.2)$$

Let $y_r(t)$ denote the received signal at the r^{th} antenna element at a certain instant t . Then, the received signals at the receptor are denoted as $\mathbf{y}(t) = [y_1(t), y_2(t), \dots, y_r(t)]^T$. The same way, the transmitted signals can be denoted as $\mathbf{x}(t) = [x_1(t), x_2(t), \dots, x_t(t)]^T$. Thus, the relationship between the transmitted and received signals can be expressed as:

$$\mathbf{y}(t) = \int H(t, \tau) \mathbf{x}(\tau) d\tau + \mathbf{n}(t), \quad (4.3)$$

where $\mathbf{n}(t)$ denotes the noise and interference that has passed through the receiver and has been sampled at the matched-filter output, typically modeled as vector of additive white Gaussian noise.

The values of the matrix components in Eq.(4.1) are influenced by several physical factors in the system, with antenna defects, geometry, and the correlation between transmitting and receiving antennas—reflecting their mutual influence—being the most significant. Correlation, along with the array's orientation, plays a crucial role in communication performance, and have a huge impact on the channel behaviour. High correlation can lead to self-interference, while multipath propagation can cause the channel to become highly frequency-selective. The ideal scenario involves a direct line of sight between transmitters and receivers with no correlation between antennas, leading to a MIMO channel with significantly reduced multipath propagation, resulting in a flatter frequency response, that does not change in time. Correlation can be reduced by using orthogonal (cross) polarization of the antennas on an array or by ensuring sufficient physical separation between them, though the latter is difficult to achieve in compact devices. As a result, the channel matrix will exhibit a more or less diagonal structure, but it will never be the identity matrix.

Let us now consider a time-invariant channel, which occurs when there is no movement of the transmitters, receivers, or the primary multipath contributors. Under these conditions, we have $H(t, \tau) = H(t - \tau)$. For instance, if the channel has a flat frequency, meaning there will be one single tap denoted by $H\delta(\tau)$, then Eq.(4.3) reduces to Eq.(4.4) [59]

$$\mathbf{y}(t) = H\mathbf{x}(t) + \mathbf{n}(t). \quad (4.4)$$

While this assumption may appear simplistic, it reflects the actual approach used in MIMO processing. In practical applications, the bandwidth of the transmitted signal is adjusted to match the flatness of the channel, specific and time slots are designated during which the channel remains stable to ensure effective MIMO transmission; this is referred to as the channel's coherence time. This coherence time varies depending on the communication scenario and the channel necessitates repetitive estimation at a frequency that is inversely proportional to the coherence time.

One of the most common approaches to channel estimation is pilot-based estimation. In this method, predefined pilot signal sequences, chosen according to a periodic pattern, are transmitted prior to the actual data. The base station receives these pilot signals, which have been altered by the channel, and applies one of several estimation techniques to estimate the channel. The most widely used estimators include Maximum Likelihood (ML), Least Squares (LS), and Minimum Mean Squared Error (MMSE). Other

approaches rely prior knowledge of the physical transmission scenario and the statistical properties of the received symbols, which have been distorted by the channel, without the need for pilot signals. These are known as blind channel estimation methods [60]. But the process does not end here. After characterizing the channel and determining the relationship coefficients between antennas, we still need to reverse the alterations experienced by the transmitted signal in Eq.(4.4) in order to recover the original information, $\mathbf{x}(t)$. This is achieved by implementing a linear detector based on the known channel matrix. By multiplying it with the matrix inverse, either before transmission or after reception, the original signal can be recovered with only the noise contribution, eliminating interference caused by multiple users.

Traditionally, all signal processing based on matrix multiplication was handled by DSPs in base stations, but electronic processors encounter inherent limitations in bandwidth and efficiency due to their digital nature. Additionally, matrix multiplication is a highly power-intensive and time-consuming operation, with complexity that scales rapidly, posing significant challenges for large-scale implementations. That is why the next step toward large-scale Massive MIMO systems should focus on achieving fast, real-time implementations with minimal power consumption [61].

4.2 Photonic Analog Computing

The quest for faster forms of computation brings us to a key element in the development of our photonic MIMO processor: the analog implementation of matrix operations using optics. This technology has been developed recently within the framework of a new computational paradigm that has emerged in recent years, driving the significant advances we are now witnessing in artificial intelligence, neuromorphic computing.

Inspired by biological neural networks, this approach traces its origins back to 1943, when Warren McCulloch and Walter Pitts introduced the concept of neural networks in their work, *A Logical Calculus of Ideas Immanent in Nervous Activity* [62]. Following the pioneering work of McCulloch and Pitts, many related studies emerged, but it was not until 1958 that the foundations of modern machine learning were firmly established by Frank Rosenblatt. In his influential paper, *The Perceptron: A Probabilistic Model for Information Storage and Organization in the Brain* [63], Rosenblatt introduced the concept of a learning algorithm for a single-layer neural network, capable of learning linearly separable functions. The perceptron can be combined with others to form more complex, multi-layer neural networks with significantly higher computational capacity. A single perceptron (or artificial neuron) in a multi-layer structure can be mathematically expressed as $y_i = \phi(\sum w_{i,k}x_k + b_i)$, where y_i is output of neuron i , ϕ is a nonlinear activation function, x_k represents the input signals from the previous layer, $w_{i,k}$ denotes the weight connecting neuron i with input x_k and b_i is the bias term.

Early neural networks were implemented in very rudimentary ways compared to today's technology, primarily due to limitations in hardware and computational resources. Initially, the first neural networks were simple, single-layer architectures with a few artificial neurons, often built as electromechanical devices. As digital computers developed, these models transitioned to software, though they remained basic due to computational constraints. This shift from hardware to software, however, laid the foundation for more complex systems. It wasn't until the 1980s, with the re-discovery and popularization of

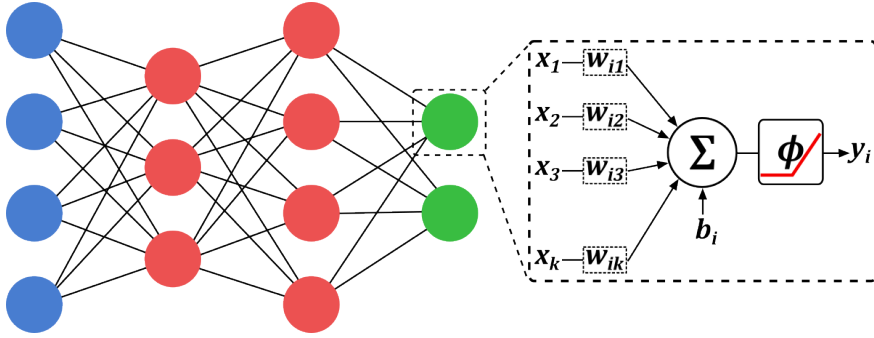


Figure 4.2: Illustration of a multi-layer neural network consisting of several neurons: the input layer shown in blue, multiple hidden layers in red, and the output layer in green.

the backpropagation algorithm, that multi-layer neural networks capable of solving non-linearly separable problems became feasible, marking a crucial advancement in neural network training. Modern neural models consist of many layers, leading to significant computational demands and high power consumption, particularly in matrix-vector multiplication operations that occur in each node of the network. To address these challenges, researchers are exploring alternatives such as implementing these networks in analog circuits, which, though more susceptible to noise, offer the potential for greater speed and energy efficiency.

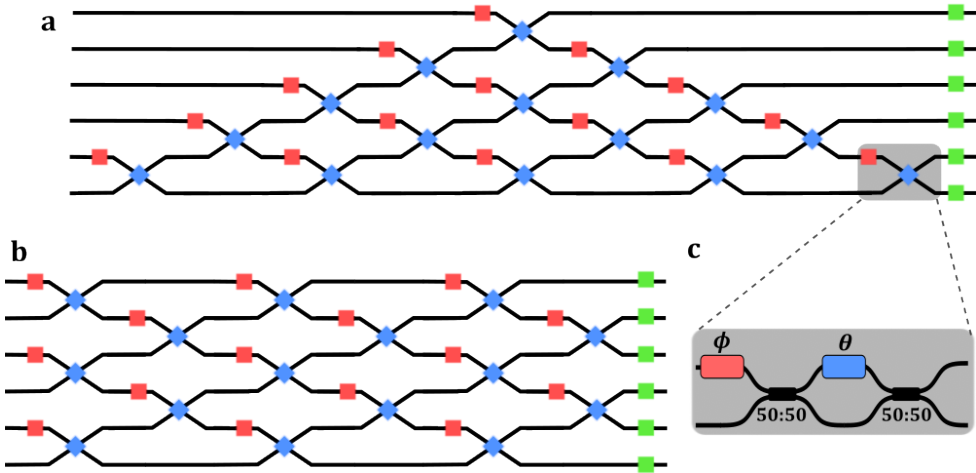


Figure 4.3: A universal N -mode multiport interferometer (illustrated here for $N = 6$) can be implemented using a mesh of $N(N - 1)/2$ beam splitters. **a** Matrix implementation based on Reck (triangular) decomposition [64]. **b** Matrix implementation based on Clements (rectangular) decomposition [65]. **c** Variable beam splitters can be implemented using a Mach-Zehnder interferometer, which consists of two 50:50 directional couplers, with a phase shifter placed at one of the input ports.

From a photonic perspective, various architectures have been proposed to perform matrix operations at extremely high speeds with minimal power consumption. Initially, these designs relied on discrete components to implement recurrent neural networks through

reservoir computing [66], [67]. With the advent of integrated photonics, new architectures have been proposed, some based on large multimode interferometers [68]–[70], and others utilizing wavelength multiplexing with microring resonators [71], [72]. However, the most successful approach for performing optical matrix-vector multiplication is based on the work of Reck [64] and later Clements [65], through their implementation of universal multiport interferometers. The implementation of this type of architecture is illustrated in Figure 4.3. Both configurations rely on the fact that any general unitary transformation matrix can be realized through a combination of beam splitters and phase shifters, as shown in Figure 4.3c. A final column of phase shifters, represented as green rectangles in the figure, is included to complete the transformation in higher order matrices.

This breakthrough paved the way for the rapidly growing field of integrated photonic neural networks [73]–[76], driving advancements in matrix-based architectures tailored for these applications. Moreover, it gave a boost to quantum computing by enabling the implementation of quantum circuits (primarily unitary matrices) on photonic circuit fabrication platforms [77], [78].

However, the immense potential of optical matrix multiplication extends beyond neuromorphic computing. It also applies to areas like optical switching, image compression through optical Fourier transforms, and more recently, signal processing applications. In Paper D, we lay the groundwork for developing an integrated analog photonic MIMO processor, building on classical MIMO theory and recent advances in optical matrix multiplication.

4.3 Paper D

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Integrated microwave photonics coherent processor for
Massive-MIMO systems in Wireless Communications

Pablo Martínez-Carrasco Romero, José Roberto Rausell-Campo, Diego Pérez-Galacho,
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Integrated microwave photonics coherent processor for Massive-MIMO systems in Wireless Communications

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Massive-MIMO systems can achieve high capacities and data rates by increasing the number of operational antennas in the base station. As more antenna elements are introduced, the complexity of the signal processing operations increases accordingly and current electronic processors struggle to reach those requirements. To overcome these hurdles, we propose a novel theory to process the RF signals in the optical domain. We experimentally demonstrate the capabilities of theory using a non-integrated Dual Parallel Modulator as a 2×2 optical matrix multiplier, recovering satisfactorily previously mixed RF signals modulated following the standards in 5G communications. Finally, we propose an integrated photonic architecture based on a core of Mach Zehnder Interferometers to physically process the information. Together with the optical core, we proposed the integration of RF modulators, microwave photonic filters and high-speed photodetectors. We present simulations on the expected performance of the circuit and analyze the impact of fabrication errors.

4.3.1 Introduction

It is undeniable that large scale wireless telecommunications are playing a great role in our day to day. Together with the current deployment of the 5G network, we will face an increasing demand for high capacity, high data rates and low latencies coming from the introduction of emerging Internet of Things (IoT) services, autonomous vehicles, artificial intelligence, smart cities and the development of the future 6G [1], [2]. An improvement in the capabilities of current available technologies and the development of novel systems will be essential to reach the required specifications [3]. One of these technologies is the Multiple-In Multiple-Out (MIMO) that has been one of the enablers of the evolution of the past LTE and is being adapted to bring in large-scale antenna arrays with hundreds or thousands of antennas in order to optimize the operation with mmWave carrier frequencies [4], [5].

Massive-MIMO systems allow the transmission of several signals over the same radio channel augmenting the capabilities of the system and achieving high spectral efficiencies and low latencies. To fully harness these benefits, high dimensional processing operations are required [6]. This requires not only a great computational capacity but also large power consumption [7], [8]. Current approaches rely on Digital Signal Processors (DSP), but electronic processors present some intrinsic limits in terms of bandwidth and efficiency [9], which, together with the end of the Moore's era [10] makes necessary the exploration

of alternative ways of computation. A possible solution may be found in photonics. Photonic systems are immune to electromagnetic interference and have some advantages over electronic devices such as high bandwidths, high parallelism and low latency. The processing of RF signals in the optical domain have been deeply studied and promising results can be found in the literature. High-speed modulators, RF filters, delay lines and beamforming networks have been experimentally demonstrated [11]. Moreover, general purpose microwave photonic processors that can be reconfigured via software have also been proposed [12], [13].

For the case of MIMO processing the main purpose is to unscramble the received RF signals that have been mixed due to interference during their free space propagation. There have been multiple attempts to solve this problem aided by photonic systems. Interference cancellation, where the signal of interest is separated from added noise have been experimentally demonstrated using tunable delays [14], [15], semiconductor optical amplifiers [16], [17], dispersive elements [18], microring resonators (MRRs) [19] or the nonlinear response of optical modulators [20]. Furthermore, a more general scheme based on the broadcast-and-weight (B&W) protocol using MRRs to perform the matrix multiplications in optical domain have been theoretically studied and great improvements in terms of bandwidth and energy consumption over DSP systems have been calculated [21].

In the following work we present a coherent architecture based on meshes of integrated Mach Zehnder Interferometers (MZI) on a Silicon-on-Insulator platform instead of microring resonators to perform matrix multiplications in parallel. One of the advantages of interferometric approaches over the B&W protocol is that complex transformations can be directly implemented without any pre- or post- processing. MZI meshes can perform any unitary transformation $|y\rangle = U|x\rangle$ and can be extended to any general linear transformation via Single Value Decomposition (SVD) [22]. This type of linear architectures has been previously implemented for deep learning [23], quantum computing [24] or Ising machines [25]. Mode scrambling due to crosstalk in multimode fibers has been also tackled with these systems [26]–[28]. We will extend the photonic matrix multiplier core by adding high-speed RF modulators, photonic filters and high-speed photodetectors to adapt the system to the requirements of massive-MIMO systems.

This work brings as a novelty the combination of previous investigated ideas, that is, photonic matrix multiplication, up-down conversion of RF signals and photonic assisted RF signal processing, in the context of wireless MIMO communications. More concretely, two key concepts have been added to the standard optical matrix multipliers. First, the introduction of microwave photonic filters for the suppression of the optical carrier and one of the RF sidebands. Second, the reinjection of the optical carrier for the recovery of the amplitude and phase with direct photodetection. To the best of our knowledge, there is no work that demonstrates both theoretically and experimentally the benefits of photonics-based RF MIMO and analyse the limitations and necessary upgrades of the integrated components to achieve a full competitive position.

The remaining of this article continues as follows. In Section 4.3.2, we introduce the basics of the RF MIMO. In Section 4.3.2, we develop the theory of up-converting the RF signal into the optical domain and unscrambling the corresponding signals with a coherent analog processor and direct photodetection. We study the need for side-band and carrier filtering to maintain the phase information of the RF signals. In Section 4.3.3, we carry out an experimental demonstration of the presented theory in a non-

integrated optical setup. Finally, in Section 4.3.4 we show the proposed architecture and explained the different modules, along with theoretical simulations of the performance of the architecture.

4.3.2 MIMO RF Theory

Consider a massive-MIMO uplink system with t single-antenna transmitters and a Base Station (BS) equipped with r receiving antennas, where the channels between the users and the BS are modelled as narrowband time-invariant wireless channels with Rayleigh fading. The behaviour of the channel is described by the deterministic matrix [29], [30] $H \in \mathbb{C}^{r \times t}$.

The theoretical development has been made using Dirac's notation which simplifies expressions and calculations when working with vectors and matrices. If we take a sample of a discrete time instant we can express the received signal at the BS as follows:

$$|Y\rangle = H|X\rangle + |n\rangle, \quad (4.5)$$

where $|X\rangle \in \mathbb{C}^{t \times 1}$ denotes the transmitted signal, with any modulation, $|Y\rangle \in \mathbb{C}^{r \times 1}$ denotes the received signal and $|n\rangle \in \mathbb{C}^{r \times 1}$ represents the white symmetric Gaussian noise, $|n\rangle \sim N(0, \sigma^2)$. After the signals transmitted by the users are received, the objective is to obtain the best estimation of the signal transmitted by each user, $|\hat{X}\rangle$, by solving the following optimization problem:

$$|\hat{X}\rangle = \underset{X}{\operatorname{argmin}} ||Y\rangle - H|X\rangle|. \quad (4.6)$$

Different detection schemes have been proposed in the literature e.g. linear methods as Zero Forcing (ZF) or Minimum Mean-Square Error (MMSE). These methods follow the same idea, multiply by a matrix A that cancels the interaction with the channel to the greatest extent

$$|\hat{X}\rangle = A|Y\rangle = AH|X\rangle + A|n\rangle. \quad (4.7)$$

Traditionally, MIMO models in wireless communications have been performed in the digital domain since there is no way to linearize the received signal in the analog domain it becomes impossible to perform the required matrix multiplications. Photonics brings us the possibility of performing this matrix multiplication analogically in the optical domain, taking advantage of the fact that an interaction between electromagnetic fields occurring in RF can be undone optically.

Photonic Signal Processing

We start from the RF matrix signal received at the BS, $|Y\rangle$. The original signal $|X\rangle$ was modeled as an analytical signal, which has undergone changes in the channel in the form of attenuations, combinations and phase shifts, so the received signal is also analytical. Using optical amplitude modulators the RF signal is up-converted into the optical domain. For the development of the theory we assume a perfectly linear double-sideband modulation of the optical field. Although this approach may seem unrealistic, the field of linearization and filtering of modulated signals has been extensively studied and has very advanced linearization techniques if needed [31]. At first stages of study this

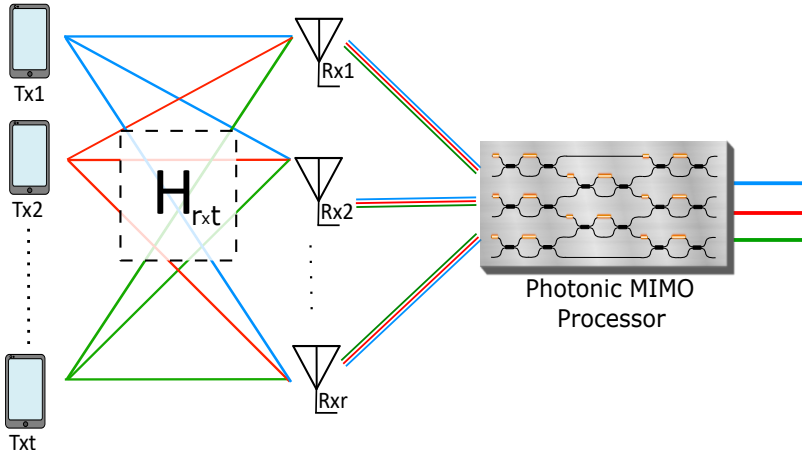


Figure 4.4: Schematic of the complete system. Several Rf signals are sent through a physical channel. In the process of propagation, the signals are mixed thus each antenna on the reception receives a linear combination of the original signals. The output of these antennas is up-converted to the optical domain on the photonic MIMO processor (more details in Fig.4.12), where the changes of the channel on the Rf signals are undone. Finally, the original wireless signals are recovered.

approximation gives satisfactorily accurate results (see Section 4.3.3). Optical modulation can be expressed mathematically as follows:

$$|S\rangle = e^{i\omega_0 t} \left[\begin{bmatrix} 1 \\ 1 \\ \vdots \\ 1 \end{bmatrix}_{r \times 1} + m \Re[|Y\rangle] \right] = e^{i\omega_0 t} [| \alpha \rangle + m \Re[|Y\rangle]], \quad (4.8)$$

where the exponential term represents the optical laser multiplying a vector formed by ones and the vector formed by the received analytical signal together with the modulation index m . The unit vector is just a mathematical tool used for the representation of the optical carrier in an specific time instant. We will denote it as $|\alpha\rangle \in \mathbb{N}^{r,t \times 1}$ and its dimension depends on the dimension of the vector to which it is added. The symbol $\Re$ denotes the real part of the element.

At this point we must operate on the modulated signals, applying the appropriate combinations and phase shifts in a matrix manner to undo the changes produced by the channel. However, if a modulated signal passes through an optical phase shifter, it acquires a phase shift which will be lost on the photodiode due to the square-law of the photodetection process, assuming an envelope detector. Because of that, in order to see changes on the RF signal by applying changes on the optical signal we must filter out the optical carrier and one of its side-bands, allowing for just single-side band (SSB) with carrier suppression propagation through the optical computational tool. After this operation the optical carrier must be re-inserted with the objective of generate a photocurrent with a phase shift in the microwave domain equal to the phase shift in the optical domain [32]. In order to maintain the phase coherence of the complete signal both filtering and carrier re-injection must be done on-chip. A possible option could

be splitting the optical carrier on-chip before modulation, so the re-injection preserves the phase of the signal. Mathematically, filtering carrier and side-band is equivalent to eliminating $|\alpha\rangle$ and recovering $|Y\rangle$ as an analytical signal with half the intensity, since one of the side-bands has been lost

$$|S_F\rangle = e^{i\omega_0 t} \left[\frac{m|Y\rangle}{2} \right]. \quad (4.9)$$

After the filtering stage we can operate with the mathematical tool that performs matrix multiplication

$$|\hat{S}_F\rangle = e^{i\omega_0 t} \left[\frac{m}{2} A|Y\rangle \right] = e^{i\omega_0 t} \left[\frac{m}{2} |\hat{X}\rangle \right]. \quad (4.10)$$

On the resulting side-band there is the information of the estimated transmitted signal. We can recover these RF signals after re-injecting the optical carrier and performing direct photodetection.

$$|\hat{S}\rangle = e^{i\omega_0 t} \left[\begin{bmatrix} 1 \\ 1 \\ \vdots \\ 1 \end{bmatrix}_{t \times 1} + \frac{m}{2} |\hat{X}\rangle \right] = e^{i\omega_0 t} \left[|\alpha\rangle + \frac{m}{2} |\hat{X}\rangle \right] \quad (4.11)$$

Envelope photodetection can not be expressed as a matrix, it must be calculated element by element. As a general result, the photodetected current of the signal in an instant t at the output k would be:

$$\begin{aligned} I_k = |\hat{S}_k|^2 &= e^{i\omega_0 t} \left[\alpha_k + \frac{m}{2} \hat{X}_k \right] e^{-i\omega_0 t} \left[\alpha_k + \frac{m}{2} \hat{X}_k^* \right] = \\ &= 1 + \frac{m^2}{4} |\hat{X}_k|^2 + m \Re \left[\hat{X}_k \right], \end{aligned} \quad (4.12)$$

discarding the direct current term and filtering the high order RF frequencies, that result from the signal-to-signal beat noise, we can recover the transmitted information from $\Re \left[\hat{X}_k \right]$.

Alternative cases: Filtering just one SB

Carrier filtering and reinsertion is performed to avoid unwanted rotations of the signals at the different outputs of the matrix. Usually, in classical wireless communications there is the presence of Phase Locked Loops (PLLs), responsible for coordinating the arrival phase of the incoming signal and correctly orienting the received constellation. On the ideal case of having a PLL on the output of each photodetector it would not be necessary filtering and reinserting the carrier, it would only be necessary to use SSB modulation before photodetection, resulting on an easier design in the photonic processor

$$|S_F\rangle = e^{i\omega_0 t} [|\alpha\rangle + m|Y\rangle]. \quad (4.13)$$

By filtering only one of the sidebands we recover the received analytical signal with half the intensity and still keeping the contribution from the optical carrier. Then, just

as in the previous case, we operate the mathematical tool to estimate the original sent signal

$$|\hat{S}\rangle = e^{i\omega_0 t} \left[A|\alpha\rangle + \frac{m}{2} A|Y\rangle \right] = e^{i\omega_0 t} \left[A|\alpha\rangle + \frac{m}{2} |\hat{X}\rangle \right], \quad (4.14)$$

$$\hat{S}_k = e^{i\omega_0 t} \left[\sum_{j=1}^r A_{kj} \alpha_j + \frac{m}{2} \hat{X}_k \right], \quad (4.15)$$

$$I_k = \left| \sum_{j=1}^r A_{kj} \right|^2 + \frac{m^2}{4} |\hat{X}_k|^2 + m \Re \left[\sum_{j=1}^r A_{kj} \hat{X}_k \right]. \quad (4.16)$$

In this case the multiplication $A|\alpha\rangle$ alters the value of the DC term after photodetection, as well as the amplitude and phase of the photodetected signals. This effect would worsen the information of the received signals, mainly due to the presence of noise.

4.3.3 Proof of Concept: Experimental demonstration

As a proof of concept, we experimentally demonstrate the theoretical proposal in Section 4.3.2. We implemented known RF channels in order to undo its changes in optical domain. Two hybrid couplers, 90° and 180° , have been used to simulate the channels. Although these are very low-dimensional and simple matrices, they serve to demonstrate the possibility of operating in the optical domain to undo the changes of RF signals.

In both cases we implement 2×2 matrices, which combine the signal produced by two different signals, with the possibility of using different frequencies, bandwidths and modulation schemes. Assuming a channel without noise and known by both the receiver and the transmitter, the implementation of the estimation matrix is very simple, especially considering that these are square matrices. Therefore, for the particular case of our experimental set-up $A = H^{-1}$.

The RF signals were generated using two different Vector Signal Generators (VSGs), E8267C PSG Agilent and SMW200A Rhode & Schwarz. Both signals are mixed using the hybrids and each one of the outputs is up-converted to the optical domain. For this up-conversion we used a Dual-Parallel Single-Drive Mach-Zehnder modulator (DPMZM) where each one of the IQ modulators is driven simultaneously and independently with the RF signals from the outputs of the hybrids. Notice that this is not the common use for a DPMZM, we are feeding each arm with different RF signals. The idea is to use the DPMZM as two independent Single-Drive Mach-Zehnder modulators biased in quadrature and implement at the same time the desired matrix multiplication by tuning the bias voltage. After this operation, the resulting optical signal gets transformed back to the RF domain using a photodetector. Finally, we use a Vector Signal Analyzer (VSA), SMW200A Rhode & Schwarz, to study the resulting signal.

Since we just can study the outputs separately and our VSA has a PLL implemented we only need to perform the single side band filtering described in Section 4.3.2. An scheme of the experimental set up is shown in Fig.4.5.

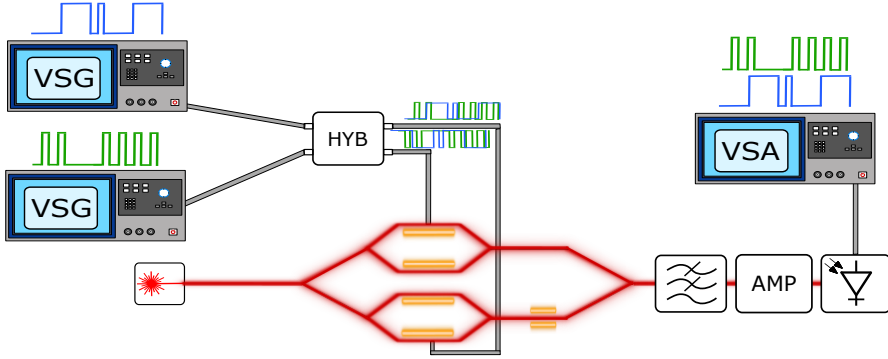


Figure 4.5: Experiment scheme. Using two Vector Signal Generators we generate the independent RF signals which will be combined using the different hybrids. Both mixed signals are up-converted on the Dual-Parallel and with the phase shifter from one of its arms we can undo the combination caused by the hybrids.

Hybrid 90°

The Hybrid coupler 90° can be expressed as a matrix in terms of electromagnetic fields as follows:

$$H = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & i \\ i & 1 \end{bmatrix} \quad (4.17)$$

In order to implement the inverse matrix with the DPMZM we set one arm with a phase shift of $\pm\pi/2$ with respect to the other and we sum both signals.

$$A|\alpha\rangle = H^{-1}|\alpha\rangle = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & -i \\ -i & 1 \end{bmatrix} \begin{bmatrix} 1 \\ 1 \end{bmatrix} = \frac{e^{-i\frac{\pi}{4}}}{\sqrt{2}} \begin{bmatrix} 1 \\ 1 \end{bmatrix} \quad (4.18)$$

$$\sum_{j=1}^2 A_{kj} \hat{X}_k = \frac{1}{\sqrt{2}} (1 - i) \hat{X}_k = \frac{e^{-i\frac{\pi}{4}}}{\sqrt{2}} \hat{X}_k \quad (4.19)$$

As shown in Eq.(4.18) and Eq.(4.19), after the photodetector both signals presents the same DC term and undergo the same phase shift resulting in a rotation of the received constellation, which is easily correctable by our Vector Signal Detector with the PPL. To check that our scheme worked, we started by sending two RF tones at different frequencies looking for separate them at the output of the modulator. We found that by adding the predicted phase shift between arms we could recover both RF tones independently achieving a suppression of 30 dB of one of the tones with respect to the other as shown in Fig.4.8. This suppression corresponds to the extinction ratio of the DPMZM.

After this first test we tried a more complex experiment related directly with MIMO technology. We modulate two different RF signals with the same frequency, 15 GHz,

and bandwidth, 20 MHz. Seeking to discern between both signals, we modulate them following 16QPSK and 8PSK modulation, the evolution of both constellations is shown in Fig.4.6.

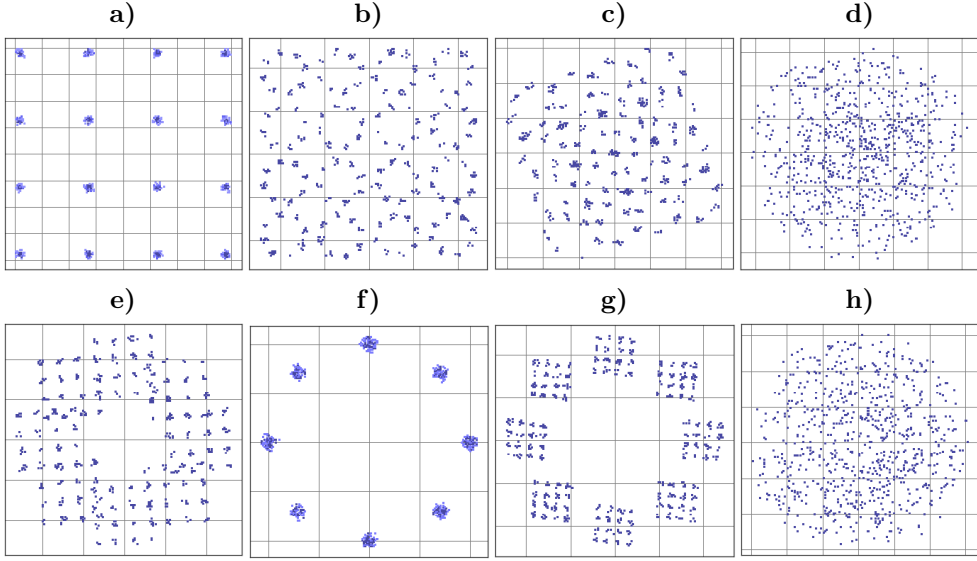


Figure 4.6: Constellation evolution using 90° hybrid. We can see how the signal changes depending on the phase applied by the modulator[a) to h)], going from a), a perfect 16QPSK in $-\pi/2$, to f) an 8PSK in $\pi/2$. Both constellations disappear completely when we are in the phases that combine both signs equally, 0 and π in d) and j).

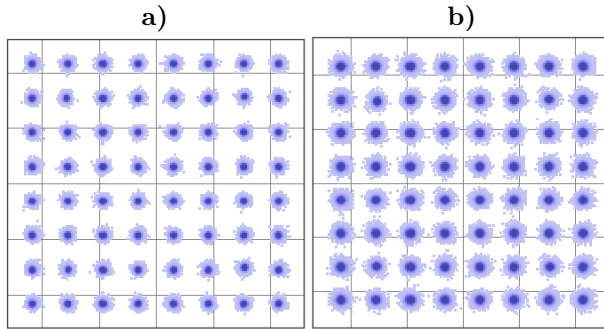


Figure 4.7: Constellation for 5G signals at 50 MHz a) and 100 MHz b) for 180° hybrid.

As seen in the spectrum of Fig.4.8, when the phase shift added by the dual-parallel modulator is $\pm\pi/2$ we can separate perfectly the original modulated signals. Following the same procedure, we recover the constellation of the two signals. The evolution of the received signals as a function of the applied phase is depicted in Fig. 4.6 where the noisy mix of both signals can be seen when we apply a phase different from $\pm\pi/2$. The same measurements can be performed on 5G signals, in this case for 50 and 100 MHz bandwidths. The constellations of the signals after their separation is shown in Fig. 4.7. This reaffirms the possibility of using optical processing to drive MIMO processing in the next generations of wireless communications.

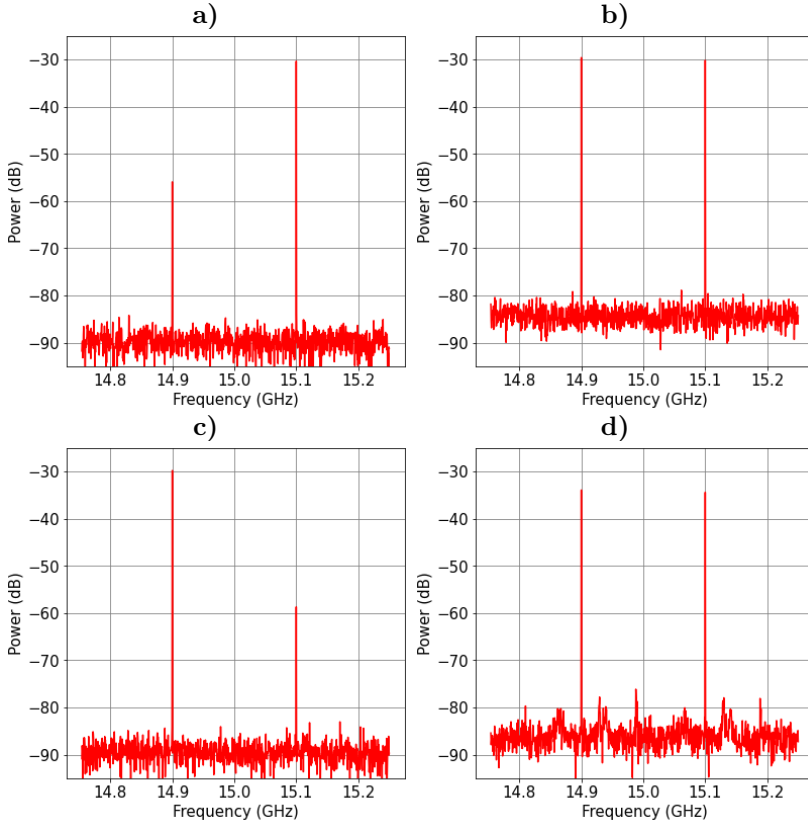


Figure 4.8: Spectrum 90° for a phase shift between the arms of the DPMZM of a) $-\pi/2$, b) 0, c) $\pi/2$ and d) π degrees. We can recover both tones with a power difference of 30 dB.

Hybrid 180°

Similar to the 90° hybrid, with the 180° hybrid we can separate the original signals using the optical modulator, just adding or subtracting the optical signals. In other words,

adding a relative phase shift between arms of 0 or π degrees.

$$H = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 1 \\ -1 & 1 \end{bmatrix} \quad (4.20)$$

In this specific case, when multiplying by the estimated matrix one of the output signals disappears because the carrier is canceled when the modulated signals are added in counter phase, despite of the imperfections on the hybrid and the actuation of the amplifier we are not able to recover this signal. As a result of this we changed the input ports of the RF signals to the modulator, seeing that way only the sum output of the estimation matrix.

$$A|\alpha\rangle = H^{-1}|\alpha\rangle = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & -1 \\ 1 & 1 \end{bmatrix} \begin{bmatrix} 1 \\ 1 \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 \\ 2 \end{bmatrix} \quad (4.21)$$

$$\sum_{j=1}^2 A_{1j} \hat{X}_1 = \frac{1}{\sqrt{2}} (1 - 1) \hat{X}_1 = 0 \quad (4.22)$$

$$\sum_{j=1}^2 A_{2j} \hat{X}_2 = \frac{1}{\sqrt{2}} (1 + 1) \hat{X}_2 = \frac{2}{\sqrt{2}} \hat{X}_2 \quad (4.23)$$

The suppression of one of the tones is shown in Fig.4.9 where we again achieve a power difference of 30 dB.

We then recovered the constellations of two RF signals at 15 GHz of central frequency modulated using 16QPSK and 8PSK modulation. The results are shown in Fig.4.10 with the independent signals shown in a) and c) and the received constellation when signals are not separated is presented in b).

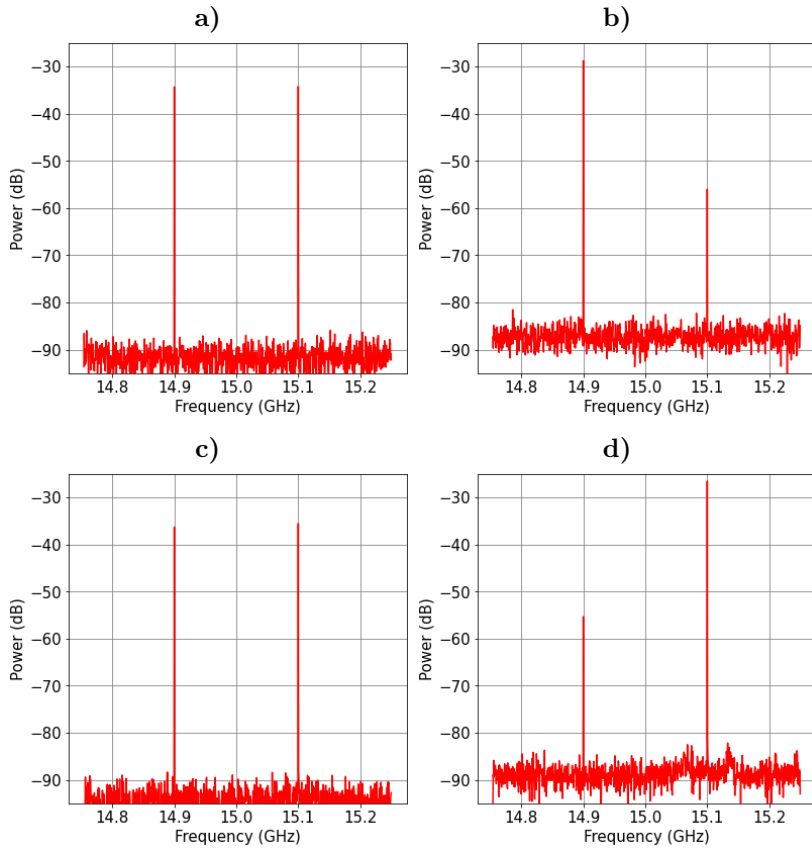


Figure 4.9: Spectrum 180° for a phase shift between the arms of the DPMZM of a) $-\pi/2$, b) 0, c) $\pi/2$ and d) 0 degrees after changing the input ports. We can recover both tones separately at the sum output of the estimation matrix.

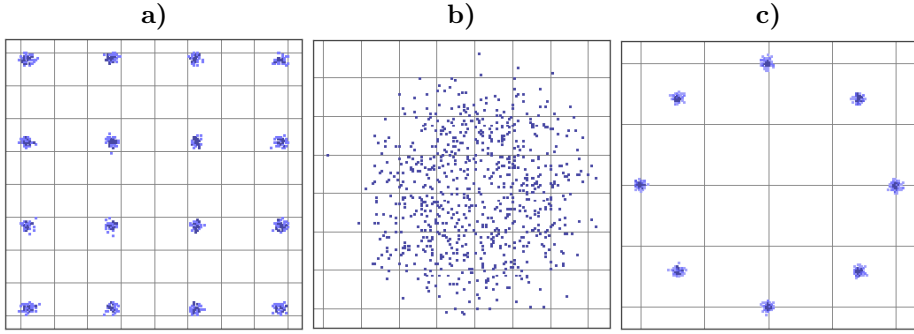


Figure 4.10: Constellation evolution using 180° hybrid, only for the sum with an phase difference of 0 degrees, a) and c), do we recover the original constellations, for intermediate offsets, b), we observe an interference between the two of them.

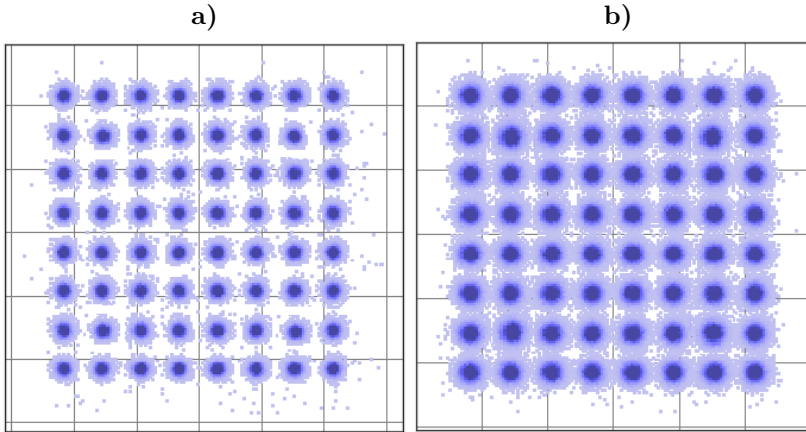


Figure 4.11: Constellation for 5G signals at 50 MHz a) and 100 MHz b) for 180° hybrid. The result for the transmission of 5G signals is similar to those presented with the 90° hybrid, however we can observe a slight deterioration of the constellations. This is due to the different transmission characteristics of the hybrids used.

4.3.4 Analog Photonic MIMO: Proposed Architecture

One of the most computational expensive processes in massive-MIMO corresponds to the multiplication of large matrices. Standard matrix multiplication algorithms require $\mathcal{O}(n^3)$ arithmetic operations. For large-scale computation, as it is the case in massive-MIMO systems, there exist more efficient algorithms that reduce the complexity to $\mathcal{O}(n^{2.81})$ (Strassen [33]) or $\mathcal{O}(n^{2.373})$ (Le Gall [34]) [35]. In the following section, we introduce a novel analog system with a coherent photonic processor for the unscrambling of RF signals in the optical domain. The proposed architecture is shown in Fig. 4.12 and is based on a Silicon-On-Insulator (SOI) platform. SOI has become the predominant platform for optical links and future optical processors thanks to its compatibility with

the available CMOS technology [36].

The proposed system consists of an up-conversion stage where the RF signal is converted into the optical domain. The resulting signal is filtered and only one of the SB is left. Then we have the optical matrix multiplier which can implement any complex-valued matrix $B \in \mathbb{C}^{N \times M}$ via SVD decomposition, where N is the number of inputs and M is the number of outputs. The final stage corresponds to the direct photodetection of the optical signal and the amplification of the detected RF signals. In order to preserve the amplitude and the phase changes of the RF signal, carrier reinsertion is necessary before the photodetection, see Section 4.3.2.

In a previous work, Salmani, M. et al [21] presented an optical processor for massive-MIMO systems that is based on B&W. This type of systems rely on the use of MRRs and wavelength division multiplexing (WDM). It presents some advantages over our presented architecture, specially in terms of footprint, but on the other hand, are inherently unable to implement complex matrix multiplications introducing extra pre- and post-processing steps. An extensive review of the capabilities of both architectures in terms of energy efficiency, optical losses and scalability as already been studied [37].

Up-conversion and filtering

The first step of the analog photonic architecture is to up-convert the incoming RF signal into the optical domain. To do so, a single wavelength light source is necessary. We proposed the use of a 1550 nm off-circuit laser to be compatible with current optical communication schemes. Emitted light is guided using single mode optical fiber (SMF) and fiber polarization controllers. External light sources present temperature stability and high light emission efficiency but are affected by the usually high coupling losses with the silicon microwave integrated circuit. Fiber-to-chip interface can be done via in-plane or out-of-plane couplers. In-plane couplers offer high coupling efficiencies, achieving theoretical losses of 0.6 dB with negligible polarization dependence but optical-quality facets need to be realized on integrated circuit. Meanwhile, vertical coupling offers easier lithographic fabrication [38]. As a future alternative, on-chip light sources promise to bring higher integration density and lower energy consumption but the technology is not mature enough [39].

Once the light is coupled into the photonic microwave integrated circuit, an splitter tree consisting of cascaded 3-dB MMI spatially multiplexes the optical signal into $N + 1$ independent paths, where N corresponds to the number of inputs in the optical matrix multiplier. An extra path will be needed for the reinsertion of the optical carrier before photodetection.

Next, the RF signals coming from the N antenna receivers are modulated into the optical domain using high-speed silicon modulators. High-performance Mach Zehnder Modulators (MZM) based on the plasma dispersion effect are commercially available from current silicon foundries, achieving bandwidths up to 40 GHz [40], [41]. Hybrid integration can open the path to higher modulation bandwidths on the order of 100s GHz [42], [43]. At the end of this stage, the optical carrier and one of the SB are filtered. Microwave filters integrated in silicon photonics have been widely studied and experimentally demonstrated [44].

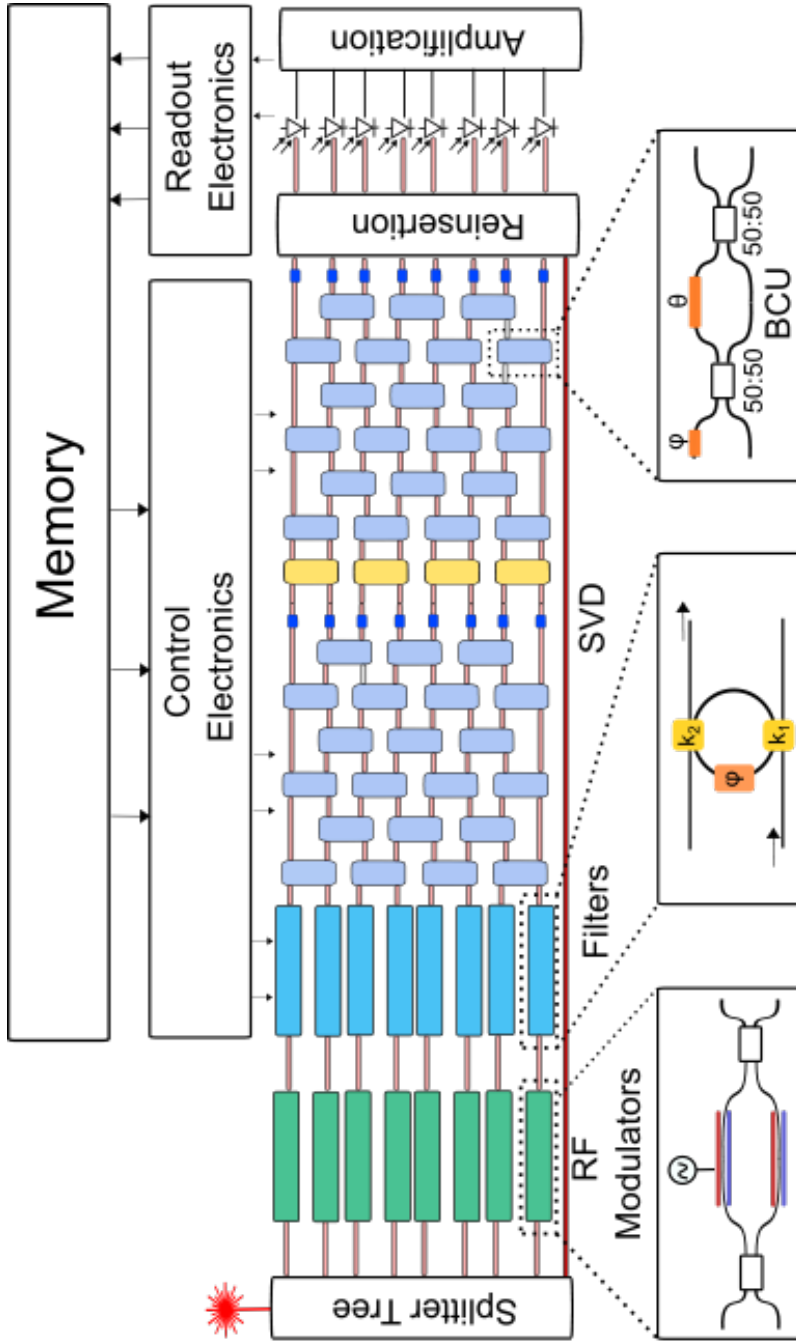


Figure 4.12: General scheme of the proposed photonic MIMO processor. The input optical laser is spatially multiplexed using a splitter tree made up of 1x2 MMI. The optical signals are then modulated with the RF signals coming from the antenna receivers, and one of the side-bands is filtered along with the optical carrier. One of the optical signals (lower path) is not modulated and serves for the carrier re-injection. The filtered signals are unscrambled using the optical matrix multiplier. A general matrix is implemented using the SVD decomposition. At the output of the matrix multiplication the optical carrier is reinserted and the resulting signals are photodetected and amplified. The complete system is controlled by a set of electronic devices.

Optical matrix multiplier

The optical matrix multiplier consists of a mesh of optical interferometers. The basic computational unit (BCU) of the mesh is a 2×2 balanced Mach Zehnder Interferometer made up of two 3-dB MMI couplers, a phase actuator (θ) in one of the internal arms and another phase actuator (ϕ) in one of the external arms [45]. An example of this BCU is shown in Fig. 4.12 and can implement the following unitary transformation:

$$U(2) = \begin{pmatrix} e^{i\phi} \sin(\frac{\theta}{2}) & e^{i\phi} \cos(\frac{\theta}{2}) \\ \cos(\frac{\theta}{2}) & \sin(\frac{\theta}{2}) \end{pmatrix} \quad (4.24)$$

The individual BCUs can be combined to create higher dimensional interferometers. The triangular arrangement proposed by Reck, et.al [46] and the rectangular arrangement proposed by [47] are the most common approaches. These architectures allow the implementation of universal unitary $U(N)$ transformations over an N -dimensional Hilbert space as $|y\rangle = U|x\rangle$. In our proposed architecture, the unitary matrices are implemented using a rectangular mesh of interferometers which provide a more compact and loss balanced scheme than the triangular architecture. The phase shift values of each of the BCU that bring the desire $N \times N$ unitary transformation can be calculated using the decomposition algorithm [47]. After this process, the transformation U is divided in a product of the unitary transformations of each individual BCU as

$$U(N) = D \left(\prod_{(n,m) \in S} T_{n,m}(\phi, \theta) \right) \quad (4.25)$$

where $T_{n,m}(\phi, \theta)$ is the transformation of a single BCU between modes n, m in an N -dimensional space and D is a diagonal matrix where the nonzero elements have unit modulus. The number of necessary BCUs is $\frac{N(N-1)}{2}$ in this arrangement. The total number of phase actuators scales as $\mathcal{O}(N^2)$, two per BCU plus a final row of N actuators that correspond to the diagonal matrix. Phase actuators can be design to harness the thermo-optic effect [48], a micro-electromechanical effect (MEMS) [49] or the Pockels effect, that aims to reduce the switching time under the microsecond scale [50].

Real channel matrices in massive-MIMO are, in general, not unitary and our analog photonic MIMO processor must implement any arbitrary complex valued matrix. It can be shown that any real or complex matrix $B^{N \times M}$ can be factorized using the SVD decomposition method as $B = U\Sigma V$, where U and V are $N \times N$ and $M \times M$ unitary matrices respectively, and Σ is a non-squared $N \times M$ diagonal matrix. The combination of these method with the mesh of interferometers has been previously discussed in the literature [22]. The diagonal matrix can be straightforwardly implemented using single MZI as optical attenuators, see the orange column of BCU at Fig.4.12. Although the SVD decomposition presents a computational complexity of the same order as the matrix multiplication it should be noted that the SVD step must be done only once for each coherence time of the channel. During this time a large number of multiplications are implemented and no advantage of the photonic processor over electronic systems is lost. Alternatively, the BCUs can also be tuned to implement the desire matrix without the SVD step using self-configuration techniques [51]. Along with the photonic mesh interferometer there is a non-interacting optical path in the lower part of the optical

Table 4.1: Parameters used for the analysis of the figures-of-merit of the system

Data Rate	DR	10 GS/s
Laser wall-plug efficiency	η_{WPE}	10%
Input optical power	P_{in}	30 mW
Thermo-optic phase shifter π -shift power consumption	$P_{\pi-TO}$	20 mW [57]
Power consumption of Mach Zehnder Modulator	P_{MZM}	480 mW [58]
Energy consumption of photodetection and amplification	E_{PD-TIA}	2.3 pJ/b [59]
Memory interface power consumption	P_{mem}	5.77 mW [37]

matrix processor. Here, the optical carrier is guided for its reinsertion before photodetection. Mathematically our $N \times M$ optical matrix becomes an $(N + 1) \times (M + 1)$ matrix where the extra row and column are full of zeros except on the element $(N + 1, M + 1)$ which is equal to 1. This expansion of the integrated photonic matrix was experimentally demonstrated for complex-valued operations by Zhang, H. et al at [52] although their photodetection scheme differs from our proposal. The calibration of the system can be performed with external photodetectors as presented by [53].

Reinsertion and photodetection

Once the signals have been multiplied and, in principle, unscrambled, the outputs of the matrix have to be photodetected and the corresponding RF signals detected. Our architecture will use high-speed envelope photodetectors. High-speed photodetectors compatible with the silicon photonics foundries technology have been reported and validated [40], [54], [55] As explained in Section 4.3.2, when no balanced photodetection is used, the reinsertion of the optical carrier is necessary. For this reinsertion we propose to divide the optical carrier that is at the $M + 1$ output of the optical matrix into M paths using an optical splitter tree made up of 1×2 MMI couplers. Each of these paths is combined with each of the M remaining outputs of the matrix using 2×1 MMI couplers. The resulting signal is photodetected and the received RF signals preserves the changes suffered during the matrix multiplication in both amplitude and phase. After the photodetection, a high-speed CMOS transimpedance amplifier (TIA) [56] is located out of the microwave integrated circuit to enhance quality of the signal.

Electronics

The photonic analog processor works together with a set of electronic systems. On the one hand, it is necessary to have the proper control electronics to configure the weight parameters of the optical matrix multiplier and manage the working regimes of the filters, modulators and photodetectors. On the other hand we need a readout circuitry to process the RF output signals. The correct design of a Printed Circuit Board (PCB) perfectly adapted for high frequencies and without interferences is essential. Achieving precise timing synchronization between ADCs and DACs in high-speed data processing systems is a critical challenge. This requires accurate clock distribution, delay compensation techniques, calibration procedures, and careful design of the system layout and signal routing to minimize signal reflections and crosstalk [60]. Finally, an electronic processor

Table 4.2: Computational speed for different matrix sizes

N	8	16	32	64	128
TMAC/s	2.56	10.24	40.96	163.84	655.36

takes control of said circuits and is responsible of loading and saving the information in the external random access memory.

System analysis

In this section we analyze some figures-of-merit of the analog photonic processor. We focus our study in the inference stage where the RF signals are unmixed using matrix multiplications. We consider square $N \times N$ matrices. Channel estimation is still realized by electronic processors. Future work will be needed to compare the capabilities of photonic processors during the estimation stage with current DSP systems. The parameters used for the estimates are presented in Table 4.1.

Computational Speed

When matrix weights are kept constant for a wide range of input vectors the computational speed is limited by the data preparation and readout speed. For a complex-valued matrix multiplication, a conventional electronic processor will take $4N^2$ multiply-and-accumulate (MAC) operations. The total number of MAC/s achievable by the computational unit is:

$$MAC/s = 4N^2 DR \quad (4.26)$$

where DR stands for data rate. We assume a conservative DR of 10 GS/s that is easily obtained with the available technology. In Table 4.2 we show the computational speed for different matrix sizes.

To satisfy the previous claim, photonic integration technology should solve the resolution challenge associated with the signal-to-noise ratio of the system. In particular, under the conditions expressed in Table 4.1, the insertion loss of the BCUs should be reduced below 0.1 dB for larger scale systems beyond 32x32 multiplications (See Section 4.3.4), in order to reach an effective number of bits of 4. Other options are the use of avalanche photodetectors or the augmentation of the power handling capabilities at PIC level followed by a higher optical power value.

Power consumption and efficiency

The electrical power consumption of the entire system can be divided into the contributions of the different subsystems, the data preparation (laser and modulation), mesh configuration (phase shifters), readout (photodetection and amplification), and communication with the memory.

$$P = P_{\text{laser}} + NP_{\text{mod}} + N^2P_{PS} + NP_{PD-TIA} + P_{\text{mem}} \quad (4.27)$$

Using the parameters of Table 4.1 we calculate the power consumption and energy efficiency for different matrix sizes. The results are presented in Table 4.3.

Table 4.3: Power consumption and energy efficiency for different matrix sizes

N	8	16	32	64	128
Power (W)	5.6	13.47	36.88	114.42	392.37
E/MAC (pJ/MAC)	2.19	1.32	0.90	0.70	0.60
TMACs / W	0.46	0.76	1.11	1.43	1.67

The photonic architecture can achieve energy efficiencies one order of magnitude higher than current GPU, that work in the order of 100s GFLOPs per Watt [61].

Scalability

The size of the matrix that can be implemented in the analog processor is limited by the optical losses that degrade the signal when it goes through the modulators, heaters and splitters. The optical power at the PD can be described as:

$$P_{PD} = P_{\text{laser}} - IL_{\text{SplitterTree}} - IL_{\text{mod}} - 2NIL_{\text{MZI}}, \quad (4.28)$$

where IL represent the insertion losses of the devices. The $IL_{\text{SplitterTree}}$ are $10\log(N) + \log_2(N)EL$, where EL are the excess losses of the individual splitter. If we want a conservative 10 dB SNR at the photodetectors we need an optical power of -21.6 dBm [62]. Assuming splitter excess losses of 0.1 dB, modulator losses of 4 dB and MZI insertion losses of 0.4 dB and an input optical power of 15 dBm the maximum size of the matrix core is 23×23 . The size of the core can be improved if optical amplifiers are placed in the diagonal matrix between the two unitaries. If the matrix to be multiplied has a higher dimension than the available at the optical core the matrix must be decomposed into lower dimension blocks and sequential algorithms must be applied. Note that the 10 dB SNR is directly associated to a resolution of less than 2 effective bits. To meet larger scale demands (transformations of 32 or 64 antennas) and resolutions of 10 bits the system would require a loss per BCU below 0.06 dB and 0.02 dB, respectively. Current commercial systems have already achieve a size of 64x64 with a computing capacity of 8 TOPS [63]. In addition, the system could be pumped with more optical power at the laser front, but power handling capabilities of silicon photonics is typically limited to a maximum of 17 dBm due to inherent system nonlinearities. A platform combining silicon nitride and silicon is ideal to cope with the high power, high-performance systems and large-scalability demands.

Simulation

In this part, we simulate the performance of the proposed architecture in a real MIMO problem using the theory developed in Section 4.3.2. We generate random channel matrices and assume that they are known by both the transmitter and the receiver. The RF signals are modulated using a QPSK modulation at 25 GHz carrier frequency. The signals are optically modulated and filtered assuming ideal behaviour. We chose a single-wavelength optical source at 1550 nm.

The optical processor is simulated following the MZI-based architecture explained in Section 4.3.4. We study the accuracy of the system in terms of Bit Error Rate (BER) as a

function of the MZI manufacturing errors. We have compared the error in implementing a desired matrix in the optical processor when the MZIs have imperfections in the MMI coupling coefficients (K) and phase terms (ϕ and θ) using Monte Carlo analysis. The operation of each MZI from the matrix can be modeled with Gaussian random variables centered at their ideal values, having a standard deviation σ_K , σ_ϕ and σ_θ [64]. Following results from the literature, we found reasonable assuming a deviation of 1% both for coupling coefficients and phase terms.

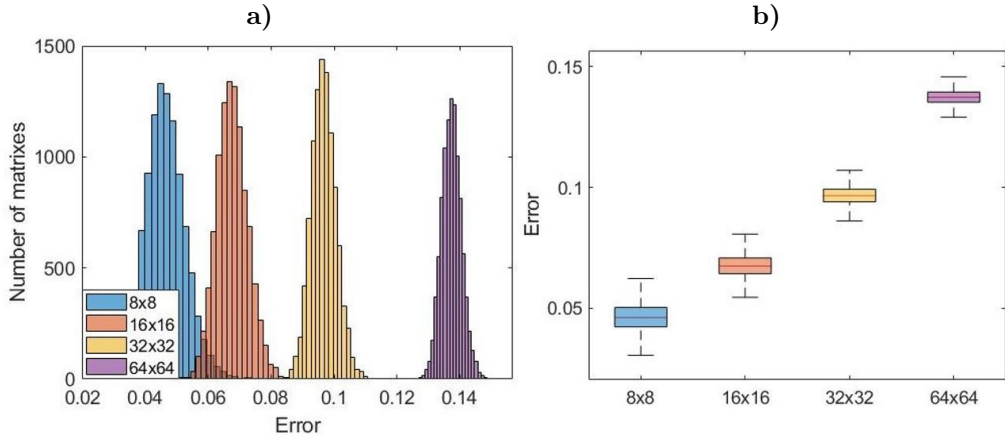


Figure 4.13: Simulation of unitary matrices with manufacturing errors for different dimensions, a) and b) its statistical errors with respect to the desired matrix

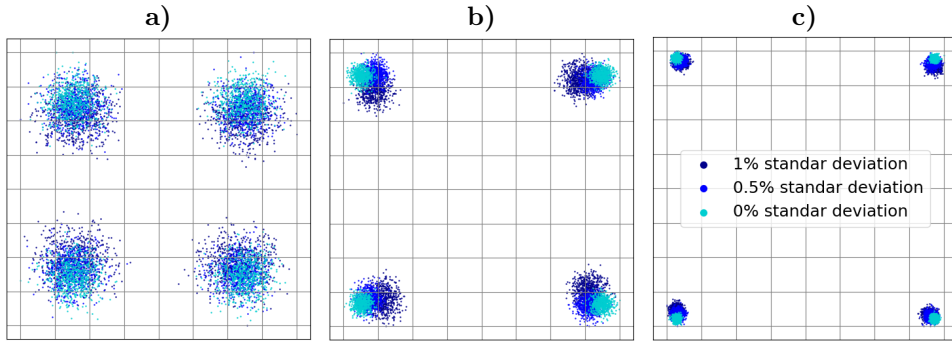


Figure 4.14: Recovered constellations on an 8x8 dimension system for a) 0dB, b) 10dB and c) 20dB of SNR. As the noise value is reduced, it is easier to notice the contributions to the error in the constellations caused by the deviations in the parameters from the optical matrices.

The impact of these deviations on the generated matrices for the generated matrices are shown in Fig.4.13. For the case of an 8x8 matrix we study the impact on the recovered constellations for different values of SNR. Results are presented in Fig.4.14.

The error is calculated as the squared norm of the difference between the desired matrix and the one obtained with the MZIs. The results of the error simulation are presented in Fig.4.15 and show that scaling the dimensions of the matrices increases the

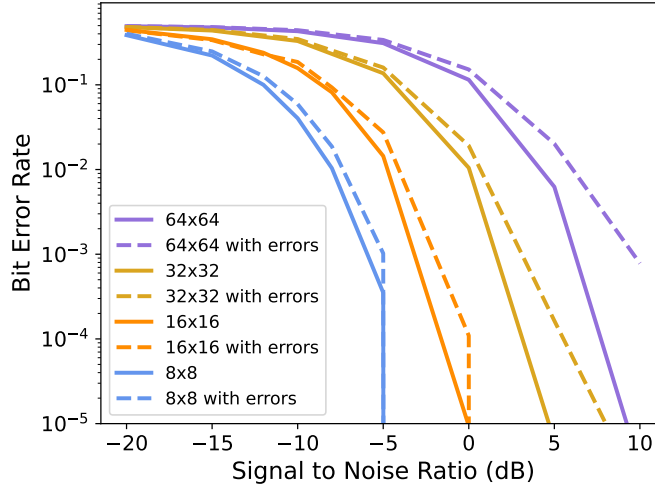


Figure 4.15: Effect of errors for real channel matrices, with dimensions from 8x8 to 64x64 for different SNR.

difference with respect to the desired matrix, reaching a difference of 14% for 64×64 unitary matrices. Regarding the errors caused by the matrix engine, recent works have proposed innovative approaches to reduce the impact of fabrication errors [65] and adapt the engine to be aware of the intrinsic noise of high-speed systems [66].

Photonic core is formed by two unitary matrices due to the SVD. We have studied how this would affect to the system performance for different Signal to Noise Ratio of the RF signals. The larger the dimension of the matrix, the higher sensitivity to inter-channel interference, multipath propagation and noise and also the greater the effect of the accumulated errors in the two units that compose the photonic core. The larger the dimension of the matrix, the greater the effect of the accumulated errors in the two units that compose the photonic core.

4.3.5 Conclusion

The increase of computational requirements in wireless communication systems has bring the need to explore alternatives ways of computation, faster and with less power consumption. In this paper, a theory to perform the unscrambling of RF signals by up-converting them into the optical domain has been developed, thus speeding up the processing of massive-MIMO signals and optimizing power consumption. We carry out an experimental proof of concept of this theory, where a dual-parallel Mach Zehnder Modulator is used for implementing 2×2 unitary transformations in the optical domain, recovering RF signals at 15 GHz with 20 MHz of bandwidth using 16QPSK and 8PSK modulation respectively and then with 5G signals with 50 and 100 MHz of bandwidth.

Finally, we introduce a novel photonic computing architecture based on a core of Mach Zehnder Interferometers in a feed-forward rectangular configuration that implements complex matrix multiplications in parallel. This architecture consists of an optical core, high-speed RF modulators, microwave photonic filters, high-speed photodetectors and the control electronics are introduced. To evaluate the capabilities of the photonic

processor we performed theoretical simulations and studied the figures-of-merit, seeing that the processing of RF signals in the optical domain takes advantage of the high-bandwidth, parallelism and low latencies of the photonic systems.

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4.3.8 References

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4.4 Experimental Demonstration

The final stage of this research involved the fabrication of an integrated photonic processor to experimentally validate the proposed theory. This work was carried out as part of a collaborative project between the Universitat Politècnica de València (UPV) and the Huawei Technological Center, under the framework of the UPV-Huawei Wireless Joint Innovation Center Agreement (Project YBN2020095120). Once the theoretical framework was established, we began designing a system based on the proposed analog photonic MIMO architecture. As a proof of concept, we developed an innovative and highly challenging layout designed to function as a 4×4 processing engine, shown in Fig. 4.16

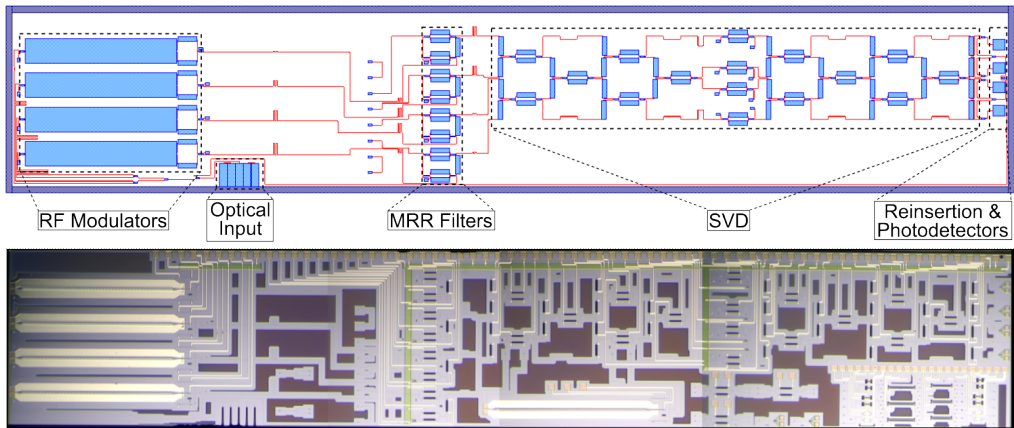


Figure 4.16: Simplified layout (excluding electrical tracks and pads) alongside the complete micrograph of the fabricated SOI chip (16 mm $\times$ 3 mm).

The system has a single optical input located at the lower edge of the chip, accessed through edge couplers. Unlike conventional designs, it does not include an optical output, as the processed RF signals are directly extracted using photodetectors at the system's back end. Upon entering through the edge coupler, the optical signal is split into two paths: one portion is directed to the final coherent detection system, while the other is sent to four high-speed modulators. These modulators play a crucial role in upconverting the RF signals, enabling further processing in the optical domain. Following modulation, a bank of add-drop microring resonators selectively filters out the carrier and one of the sidebands from the modulated signal. The remaining sideband is then fed into the matrix multiplication engine, which is designed based on a Singular Value Decomposition approach. This engine consists of two unitary matrices implemented using Clements' architecture, with an additional column of MZIs positioned between them to encode the diagonal values of the SVD. Ultimately, after carrier reinsertion through the four photodetectors, the recovered RF signal is downconverted back to the electrical domain.

4.4.1 Measurements and results

After fabrication, all the discrete elements that made up the layout were independently characterized using vertical coupling through grating couplers and single-mode fibers

for optical input, along with electrical probes for control signal driving. Figure 4.17 shows a close-up view of the main building blocks of the design, which are based on the modulators, phase shifters, MMIs, and photodetectors provided in the Process Design Kit from *Advanced Micro Foundry*. The measurement results for the discrete elements are shown in Figure 4.17e-f, where it can be seen that both the modulators and the photodetectors are capable of reaching high speeds, approaching 35 GHz and 47 GHz, respectively.

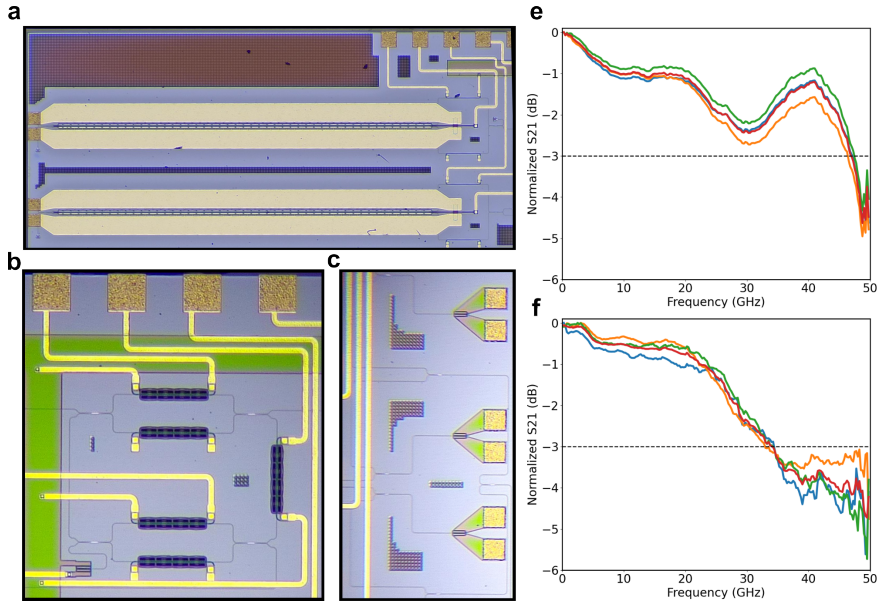


Figure 4.17: Detailed images captured from the fabricated photonic chip. **a** Subset of the four integrated intensity modulators, showing two units. **b** One of the ring resonator filters composed of two MMIs and an additional phase shifter. **c** Three integrated photodetectors and the carrier reinjection mechanism at the back end of the device. **e** Measured transmission spectrum of the integrated photodetector. **f** Spectral transmission response of the modulator.

Once all possible discrete measurements were completed, a custom PCB, shown in Fig.4.18a, was designed to simultaneously control all elements of the chip and enable the final proof-of-concept test using mixed RF signals. Unfortunately, despite all the efforts made, it was not possible to complete the full experiment. During the packaging process, the waveguide responsible for carrier reinjection was damaged, significantly limiting the device's functionality. An electrostatic discharge occurred during the measurement process, causing damage to both the photodetectors and phase shifters, as shown in Figure 4.18b-d, and ultimately rendering the chip non-functional.

Despite this setback, the theoretical work has had significant impact, with several research groups now allocating resources to further develop this line of investigation. On our side, we are planning a second chip design with the goal of successfully completing the work that was started.

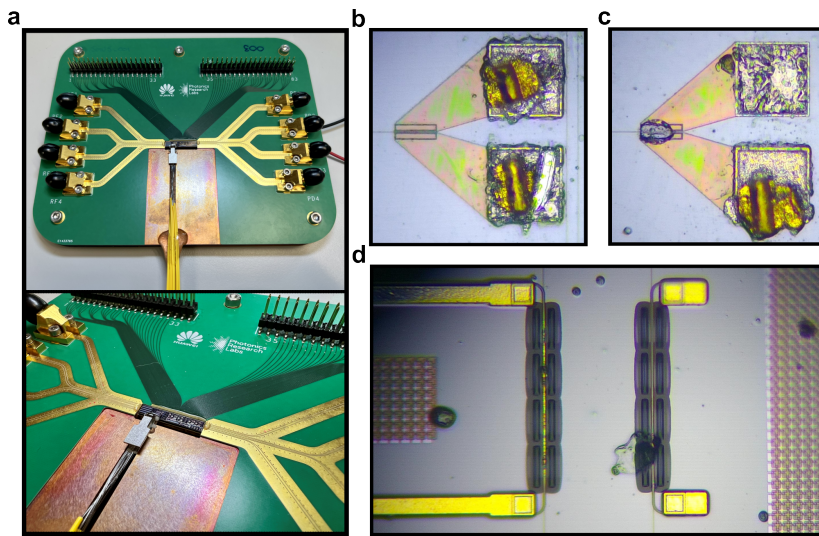


Figure 4.18: **a** Printed control board fabricated to drive the photonic processor. Electrical connections were established via wire bonding, and optical input was provided through a fiber array. Detailed images showing damaged elements of the chip. **b** Micrograph of a functioning photodetector, compared with **c** a photodetector damaged by electrostatic discharge. **d** Micrograph comparing two phase shifters from the same MZI, where the left one is damaged and shows a broken waveguide.

Chapter 5

Conclusions and Future work

In this final chapter, we summarize the main results achieved throughout the thesis. The main findings are discussed in depth, with particular attention to how they address and support the initial objectives set out in the introduction. Finally, we suggest possible directions for future work and reflect on the importance and impact of the research.

5.1 Conclusions

The focus of this thesis was to explore the latest advancements in integrated programmable photonics, with the goal of pushing the boundaries of current knowledge in this rapidly evolving technological field. The ultimate objective was to apply these insights to the development of future telecommunications base stations based on microwave photonics, taking full advantage of the benefits that this technology offers.

To tackle the main goal, the work was split into two parts, based on the physical structure of a modern base station. The first part looked at the outer section of the communication link within a base station, the beam steering system and the antennas. This was the first area we explored, aiming to build a microwave photonics-based beamforming system that could handle wide steering angles and high bandwidth, while also keeping resource use efficient. The second part focused on the inner, less visible part of a base station: the signal processing unit. In this case, the goal was to develop a programmable analog optical MIMO processor capable of handling the signals received by the antennas, leveraging the reconfigurability and high speed offered by integrated optics.

Following this structure, the document is organized as follows. Chapter 1 provides the background and outlines the benefits of microwave photonics and chip integration. The first part of the thesis centers on beam steering systems, beginning with Chapter 2, which establishes the theoretical foundation for antenna arrays. This chapter explores both conventional implementations and recent advances in photonic technologies, particularly those based on true-time delay. Building on this theoretical background, Chapter 3 presents the research results in the form of published work, highlighting not only practical applications in beam steering but also the signal processing techniques developed throughout the investigation. Finally, Chapter 4 explains the second major research topic, beginning with the fundamentals of MIMO communications and signal processing. Then we examine how these techniques can be adapted for photonic integration, with a focus on photonic matrix multipliers that were originally developed in the context of

neural networks as a promising solution.

The following points summarize the key conclusions drawn from this work:

- Paper A presents the most relevant and impactful results of the thesis, without diminishing the value of the other contributions. It showcases both theoretical and experimental validation of a beamforming network based on switched true-time delays, capable of simultaneously achieving unprecedented resolution and bandwidth for an integrated photonic system, while also optimizing the number of components and minimizing the required chip area.
- In Paper B, we present a novel calibration technique developed during the characterization of the beamforming network. This method enables precise tuning of the delay lines without the need for external monitoring components, thereby avoiding additional losses and preserving the compact footprint of the device.
- Paper C concludes the first research topic of this work and presents an integrated photonic chip that evolves from the design introduced in Paper A. This chip not only performs beamforming but also integrates additional signal processing functionalities such as filtering and high-frequency signal generation. Uniquely, this design combines the previously discussed benefits with the advantages of programmability, reconfigurability, and multifunctionality, which are essential features for the successful integration of photonic technology in future markets.
- Paper D explores the application of programmable photonics as an analog matrix multiplier for MIMO signal processing. This groundbreaking theoretical work lays the foundation for the design of a comprehensive photonic MIMO analog processor, aimed at significantly enhancing processing speed and reducing power consumption in these tasks.

5.2 Future work

Given the rapid growth and rising demands of modern telecommunications networks, this thesis has focused on exploring improvements in system speed and power efficiency through the use of integrated photonics. The work has introduced several solutions that offer clear advantages over traditional electronic approaches, with strong potential for integration into commercial systems in the near future. However, many aspects remain open for further investigation to refine the current results and to unlock new directions for research.

One of the main unresolved aspects of this thesis is the experimental demonstration of the integrated MIMO processor. Due to a combination of setbacks including the ambitious nature of the project and limited time for design refinement, measurements could not be completed. Nevertheless, the work has been well received by the research community and has sparked growing interest in this promising field. We are currently pursuing a new design using thin-film lithium niobate, a platform that has been gaining significant popularity recently, with the aim of finally achieving the long awaited practical integrated demonstration.

Future work also includes exploring the possibility of using the programmable integrated chip based on true time delays for implementing a silicon photonic neural network for chromatic dispersion compensation. This would represent an additional application using the same chip, inspired by recent research advances in machine learning techniques [79]–[82].

Finally, one of the key areas for future improvement involves a broader challenge faced by integrated photonics, which is scalability. Although many strategies to enhance scalability lie beyond the scope of this work and depend on advances in fabrication technologies such as reducing propagation losses or developing platforms that support higher device density in smaller areas, there are still meaningful opportunities to study scalability at the system level using the current manufacturing capabilities. We are currently exploring methods to scale the beamforming system by extending it from a one-dimensional antenna array to a fully two-dimensional configuration, as part of a collaborative project with Huawei.

Author's merits

JOURNALS:

- [Paper A] **Pablo Martínez-Carrasco Romero**, Tan Huy Ho, David Wessel and José Capmany. "Ultrabroadband high-resolution silicon RF-photonic beamformer". *Nature Communications*, 15(1), 1433 (2024)
- [Paper B] **Pablo Martínez-Carrasco Romero**, Tan Huy Ho and José Capmany. "Nonintrusive Characterization Method for Integrated Optical Delay Lines". *Journal of Lightwave Technology* 42, 4844-4850 (2024)
- [Paper C] **Pablo Martínez-Carrasco Romero**, Tan Huy Ho and José Capmany. "Integrated microwave photonics true-time delay signal processor". *APL Photonics*, 10(5) (2025)
- [Paper D] **Pablo Martínez-Carrasco Romero**, José Roberto Rausell-Campo, Diego Pérez-Galacho, X. Qing, T. Wang and Daniel Pérez-López. "Integrated microwave photonics coherent processor for massive-MIMO systems in wireless communications". *IEEE Journal of Selected Topics in Quantum Electronics*, 29(6), 1-12 (2023).

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- [Conf. 1] **Pablo Martínez-Carrasco Romero**, Daniel Pérez-López, Diego Pérez-Galacho, Tan Huy Ho, David Wessel and José Capmany. "High-precision broadband silicon photonics beam-former". Oral, *IEEE International Topical Meeting on Microwave Photonics (MWP)*, (pp. 1-4), Online (2022)
- [Conf. 2] José Roberto Rausell-Campo, **Pablo Martínez-Carrasco Romero**, X. Qing, T. Wang, and Daniel Pérez-López. "Integrated microwave photonics coherent architecture for Massive-MIMO". Oral, *IEEE Photonics Society Summer Topicals Meeting Series (SUM)*, (pp. 1-2), Sicily (2023)
- [Conf. 3] **Pablo Martínez-Carrasco Romero**, Tan Huy Ho and José Capmany. "Self-characterization method for integrated optical delay lines". Oral, *IEEE International Topical Meeting on Microwave Photonics (MWP)*, (pp. 1-4), Nanjing (2023)
- [Conf. 4] **Pablo Martínez-Carrasco Romero**, Tan Huy Ho and José Capmany. "Integrated photonics true-time delay signal processor". Oral, *Conference on Lasers and Electro-Optics/Pacific Rim. Optica Publishing Group*, (p. We2H2), Seoul (2024)
- [Conf. 5] **Pablo Martínez-Carrasco Romero**, Tan Huy Ho and José Capmany. "Programmable photonics processor based on true-time delay". Oral, *Conference on Lasers and Electro-Optics. Optica Publishing Group*, Los Angeles (2025)

PATENTS:

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OTHER CONTRIBUTIONS:

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List of Acronyms

RF	Radiofrequency
IoT	Internet of Things
MIMO	Multiple-In Multiple-Out
PAAs	Phased Array Antennas
LTE	Long-Term Evolution
DSPs	Digital Signal Processors
CAGR	Compound Annual Growth Rate
MWP	Microwave Photonics
SOI	Silicon-on-Insulator
CMOS	Complementary Metal-Oxide-Semiconductor
InP	Indium Phosphide
SiN	Silicon Nitride
GaAs	Gallium Arsenide
WDM	Wavelength Division Multiplexing
LiNbO₃	Lithium Niobate
MEMS	Microelectromechanical Switches
MRR	Microring Resonators
FSR	Free Spectral Range
CW	Continuous Wave
MZI	Mach-Zehnder Interferometers
SISO	Single-Input Single-Output
SIMO	Single-Input Multiple-Output
ML	Maximum Likelihood
LS	Least Squares
MMSE	Minimum Mean Squared Error

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