

Planar Periodic Leaky-Wave Antennas for Radar Applications

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Amar Al-Bassam, M. Sc.

aus Bagdad, Irak

Berichter: Universitätsprofessor Dr.-Ing. Dirk Heberling
 Universitätsprofessor Christophe Caloz, Eng., Ph.D., KU Leuven

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Summary

The ever-growing demand for connected and autonomous systems requires new concepts of antennas. With the coming next generation of mobile system standard (6G), the utilization of the terahertz spectrum (> 300 GHz) is envisioned. At these frequencies, the increased material and interconnection losses of the integrated chips demand high-gain antennas with large frequency bandwidth. Periodic leaky-wave antennas (P-LWAs) feature both high gain and large bandwidth. In addition, they are capable of steering the main beam by changing the frequency without requiring complex feeding networks or phase shifters. However, they suffer from the so-called broadside problem, where the gain degrades when the main beam points to the broadside direction. The thesis aims to develop new concepts of efficient planar P-LWAs for radar applications, starting from the theory through the design and ending with radar demonstration.

A description of the broadside problem is first presented. Also, the solution for the broadside problem is analyzed utilizing the symmetries of the unit cell of the periodic structure. The analysis is based on generic network models that establish analytical formulas for the propagation constant and radiation efficiency. Additionally, an electromagnetic perspective is provided by deriving the coupled-mode equations for asymmetric unit cells. It is shown that both the network model and the coupled-mode theory approaches can be on an equal footing. Specifically, both approaches can accurately determine the conditions to eliminate the broadside problem. Furthermore, asymmetric P-LWAs operating at 78 GHz are demonstrated with an equalized beam scanning through the broadside.

Furthermore, three concepts are investigated for polarization-diverse P-LWAs by exploiting the symmetry of the unit cell. The first concept enables orthogonal circular polarization with longitudinal asymmetry by exciting the structure at both ends. The second concept leverages the first design and extends it to achieve quad polarization, two linear and two circular polarizations, by combining two longitudinally-asymmetric P-LWAs. The third concept combines longitudinally and transversally asymmetric unit cells and achieves orthogonal linear polarizations, seen from both ports of the P-LWA. The first two and the third concepts are validated experimentally with designs operating at 78 GHz and 24 GHz, respectively.

Moreover, frequency-scanning concepts are introduced for 1D and 2D P-LWAs. First, 1D P-LWAs with an enhanced frequency-scanning rate are presented based on unit cell loading and the control of the dispersion diagram. Second, a 2D P-LWA concept is described, which combines a series-feeding network and an array of 1D P-LWAs, each with a unique dispersion curve, allowing for the control of scanning planes in the 2D space.

Finally, an FMCW-radar demonstration is presented, which combines the frequency-scanning and polarization diversity features achieved by one of the developed P-LWAs. These features allow for enhanced target detection enabled by the angular resolution using frequency-scanning and the target classification using radar polarimetry.

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Introduction

The next generation mobile system standard (6G) envisions many new potential applications besides the increased data throughput, thanks to the large available frequency bandwidth and low latency [1]. Some actively discussed applications are connected robotics, autonomous systems, extended reality, automation and manufacturing, and many others [2]. These applications rely on crucial technologies for localization, sensing, and imaging [3, 4]. Furthermore, 6G promises an increased spectral efficiency by allowing spectrum sharing between systems, such as joint communication and radar sensing (JCRS) [5]. A *radar*, which stands for *radio detection and ranging*, is a pivotal component in realizing these technologies.

Radar systems gained massive interest in the last two decades due to the advent of RF-CMOS chip manufacturing, which enabled their vast use in various civilian applications at the mm-wave frequencies, such as automotive radars [6] and personnel security scanners [7]. The current state-of-the-art radars use frequency-modulated continuous-wave (FMCW) waveforms, allowing for both velocity and range measurements. Furthermore, commercially available FMCW radar chips feature multiple-input multiple-output (MIMO) antennas, which enable angle measurement achieved by *digital beamforming* in addition to the velocity and range [8]. Additionally, MIMO radars have the advantages of high spatial resolution, enhanced signal-to-noise (SNR) ratio, and immunity to interference [9]. Conversely, they have the disadvantages of high cost and the need for high computation power, and therefore, high energy consumption. These problems scale up when these chips are synchronously cascaded to achieve very high angular resolution. Some examples of cascaded MIMO radars are used in driver assistance systems in automotive [10, 11], 3D non-invasive imaging [12], and concealed weapon detection at airports [13]. However, these examples operate in the frequency range of 70 – 110 GHz, where the RF-CMOS technology reaches its limit due to the increased phase noise and complex synchronization between the chips.

The *terahertz band* (THz) (0.3 – 3 THz) is widely discussed in 6G due to the ultra-wide frequency bandwidth available. However, at these frequencies, the free-space path loss increases dramatically with increasing range. In addition, the high material and chip interconnection losses limit the MIMO concept's implementation. Hence, for THz localization and imaging applications [14], a mechanically-steered pencil beam antenna is preferred, which allows the use of a single transmitter and receiver antenna [15]. The drawbacks of such an approach are the increased weight and slow measurement time, which are both not applicable to the mobile

application envisioned for 6G.

A promising approach of using frequency-scanning antennas, *leaky-wave antennas* (LWAs), allows for reducing output power requirements and the number of antennas. Moreover, LWAs are electrically large and, therefore, have high gain. Also, the frequency-scanning feature adds an additional dimension that can be exploited to increase the spectral efficiency of a radar system and hence, increase its angular resolution. This work deals with the design of efficient LWAs and the development of new concepts for radar applications.

1.1 Periodic Leaky-Wave Antennas (P-LWAs)

LWAs date back to 1940, when Hansen first patented the design of a traveling structure with a long slit [16]. They are defined in the IEEE standard for definitions of terms for antennas as [17]:

“An antenna that couples power in small increments per unit length, either continuously or discretely, from a travelling-wave structure to free space.”

Furthermore, their *uniform* class features the scanning of its main beam by changing the frequency from the *backfire* and close to the broadside without the need for phase shifters or complex feeding networks. In contrast, the *periodic* class can scan the full space from the *backfire* to the *endfire* and through the *broadside* [18].

From their first introduction, periodic LWAs (P-LWAs) suffer from gain degradation when the main beam scans through the broadside direction. This gain degradation is caused by an internal mismatch between the antenna’s *unit cells* [19]. Hence, this problem is called the *broadside problem*, which limits the useability of P-LWAs in an application where broadside radiation is required. In the last two decades, a solution to the broadside problem inspired by 1D metamaterial-based transmission lines has been proposed by equalizing the frequencies of the right-hand and left-hand transmission line branches [20]. However, this solution solves only part of the problem, which is called the *frequency balancing* of the *two* radiation mechanisms of a P-LWA. In [21], an extension to the frequency balancing is proposed to eliminate the broadside problem at broadside based on impedance matching between the unit cell. Recently, a network model-based analysis of the broadside problem has been established in P-LWAs with longitudinally and transversally symmetric unit cells [22].¹ The analysis shows that in the broadside direction, a P-LWA can achieve a maximum of 50 % radiation efficiency if one of the radiation mechanisms of P-LWAs has 0 % radiation efficiency. Later, it was shown that the broadside problem may be systematically eliminated if the unit cell’s symmetry is broken [23–26]. The broadside problem and its solution will be discussed in Chapter 3.

¹In Chapter 3, the longitudinal axis corresponds to the antenna or the propagation axis.

1.2 Frequency-Scanning Radars

In the 1960s, just two decades after their invention, the first use of P-LWAs in pulsed radar systems was reported [27, 28], where short pulses are transmitted at different frequencies allowing the detection of targets in different directions. The aim was to use P-LWAs for airport surveillance radars to detect the aircraft's position without requiring an expensive rotating mechanical parabolic antenna. The main issue was the pulse dispersion when such a system is used in the combination with pulse-compression technique, which correlates the received and transmitted pulses.

Three decades later, the first implementations of P-LWAs in FMCW radar systems were reported [29, 30]. Recently the topic has been trending again due to the low-cost availability of off-the-shelf radar components [31–34]. P-LWA implementations are also reported for imaging applications in the THz band [35, 36]. Mainly, all reported P-LWAs for imaging applications are one dimensional (1D) and capable of scanning the main beam with frequency in one direction, e.g., elevation, where the scanning in the other direction, e.g., azimuth, is achieved by mechanically moving the antenna front end.

The drawback of using P-LWAs for radar applications is the reduction of the range resolution due to the fractional bandwidth used for each beam. This problem will be discussed later in Chapter 6 in more detail, followed by a demonstration with an FMCW radar.

1.3 Research Contribution

The thesis aims to develop new concepts for P-LWAs, starting from the theory through the design and ending with radar demonstration. This work mainly focuses on P-LWAs, with a minor emphasis on radar applications. Hence, the contributions to P-LWAs can be divided into three main parts, each treated in a separate chapter, as follows

1. Electromagnetic modeling of P-LWAs: the coupled-mode theory is utilized to describe the broadside problem and its solution based on the asymmetry of the unit cell. Furthermore, the resolution of the broadside problem is extended to non-TEM structures, i.e., waveguides operating in the TE₁₀ mode, which to the author's knowledge, is not reported elsewhere.
2. Polarization-diverse concepts: based on the asymmetry approach, three concepts are presented to achieve orthogonal linear and/or circular polarization in P-LWAs.
3. Frequency-scanning concepts: 1D and 2D P-LWA structures are presented. First, 1D P-LWAs with an enhanced frequency-scanning rate are presented. Second, a 2D P-LWA concept is described, which can achieve scanning in two directions and is theoretically capable of scanning the full hemisphere above the antenna.

All concepts are supported by analytical analysis, full-wave simulations and measurements of planar P-LWAs. Additionally, two broadside frequencies are used for the designs, 24 GHz and 78 GHz.

Moreover, a contribution to the radar application is also presented. A P-LWA radi-

ating quad polarization, two orthogonal circular and linear, is used with an FMCW radar operating at 78 GHz. Furthermore, the concept of polarization decomposition, usually used for synthetic aperture radar (SAR) applications, is also implemented with the addition of angular discrimination by frequency scanning.

1.4 Thesis Organization

The thesis consists of seven chapters, including the current one. Each chapter is dedicated to a specific topic for reasons of self-consistency and better readability. The theory of LWAs is described in Chapter 2 with an emphasis on P-LWAs. Also, the radiation properties of P-LWAs are provided, such as radiation pattern, frequency scanning, and scanning range and rate. Examples of uniform and periodic LWAs are also presented.

Chapter 3 discusses the broadside problem in P-LWA and shows that its solution is based on the asymmetry of the unit cell. The asymmetry approach is first described utilizing network models and later analyzed by the couple-mode theory. The two equalization conditions, *frequency-balancing* and *quality-factor-balancing*, are supported by both the network model and coupled-mode theory approaches. Two periodic structures operating at 78 GHz are investigated. Specifically, a slotted P-LWA and a series-feeding network (SFN) based on substrate-integrated waveguide operating with the TE₁₀ mode are shown.

Chapter 4 deals with the polarization of P-LWAs. First, a series-fed patch (SFP) P-LWA with longitudinal asymmetry is presented. This P-LWA is capable of radiating orthogonal circular-polarization with the opposite handedness, which is achieved by exciting the P-LWA at its ports sequentially. Second, a P-LWA consisting of two parallel SFPs with longitudinal asymmetry in a parallel configuration is shown. This P-LWA can radiate quad polarization depending on the excitation of the four ports. Orthogonal circular or linear polarizations are achieved by exciting either ports of the two parallel SFPs or by exciting these ports simultaneously with even/odd configurations, respectively. Third, the concept of attaining orthogonal linear polarization is presented. The idea is based on combining longitudinal and transverse asymmetry, resulting in double asymmetry.

Chapter 5 presents frequency-scanning concepts in 1D and 2D P-LWAs. First, frequency-scanning rate enhancement methods are introduced with two implementations: 1) a stub-loaded series-fed coupled patch (SFCP) P-LWA operating at 24 GHz and 2) a dispersive SFN terminated with Vivaldi radiating elements operating at 78 GHz. Second, the 2D P-LWA concept is introduced, capable of scanning both azimuth and elevation directions with a pencil beam. The scanning planes are analyzed by the array factor theory, and later design of a 2D P-LWA based on the SFN and 1D array of the slotted substrate integrated waveguide P-LWA operating at 78 GHz is presented.

The FMCW-radar demonstration is presented in Chapter 6, operating in the frequency range 76–81 GHz. The quad-polarized P-LWA presented in Chapter 4 shows a combined polarimetric and frequency-scanning radar operation.

Chapter 7 concludes the thesis and provides an outlook for future works based on

the presented results.

Citations of articles by the author are highlighted in bold print, e.g., **[26]**.

Theory of Leaky-Wave Antennas

This chapter provides a brief introduction to the classification, theory and radiation properties of *leaky-wave antennas* (LWAs). First, the principle of leaky-wave radiation will be provided and then followed by the classification of LWAs with emphasis on the periodic class due to its importance to this work. Next, the theory of periodic structures is presented to understand the radiation mechanism, in particular of *periodic leaky-wave antennas* (P-LWAs). The following sections lay the foundations for the forthcoming chapters.

2.1 Principle of Leaky-Wave Radiation

A *leaky wave* (LW) is a wave traveling through a guided structure and progressively *leaks* its energy out to the free space as it propagates. A generic 1D LWA with a wave excited at $y = 0$ and traveling along the y -axis is depicted in Figure 2.1. The electromagnetic field of the wave in free space at any point can be expressed as [18]

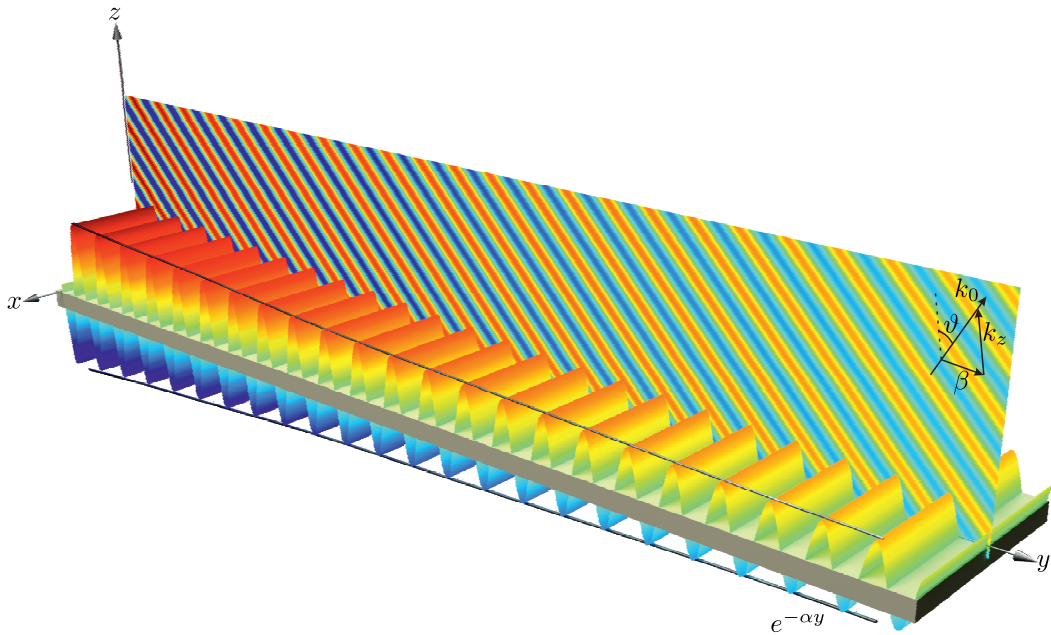


Figure 2.1: Radiation principle of leaky-wave antennas. A wave excited at $y = 0$ traveling with a phase constant, $\beta \leq k_0$, decays exponentially along the y -axis due to the loss factor, α , and produces a radiating wave directed in the yz -plane.

$$\psi(y, z) = \psi_0 e^{-(\gamma y + j k_z z)}, \quad (2.1)$$

where ψ_0 and k_z are the amplitude of the field and the wavenumber of the propagating wave perpendicular to the structure, respectively. The complex propagation constant of the traveling wave is given by¹

$$\gamma = \alpha + j\beta, \quad (2.2)$$

where β and α are the *phase* and *attenuation or leakage constants* of the guided wave, respectively. The wavenumber k_z (perpendicular to γ) relates the propagation constant γ and the wavenumber of free space k_0 by

$$k_z = \sqrt{k_0^2 - \gamma^2}, \quad (2.3)$$

with $k_0 = \omega/c$, ω and c are the radial frequency of the wave and the speed of light, respectively. In many practical implementations of LWAs where the conductor and dielectric losses in addition to the radiation leakage are small, the assumption $\beta \gg \alpha$ can be made. Therefore, Equation 2.3 reduces to

$$k_z = \sqrt{k_0^2 - \beta^2}. \quad (2.4)$$

The LW radiation depends on the solution of Equation 2.4, which can be either real or imaginary. If $\beta > k_0$ then k_z is imaginary and therefore, Equation 2.1 describes a decaying or *evanescent* field in the z -direction, i.e., $e^{-k_z z}$. In this case, the wave is a purely *guided wave*. If $\beta \leq k_0$, then k_z is real and therefore, Equation 2.1 describes a propagating wave, i.e., *radiating wave*. Additionally, the phase constant, β , of the guided wave depends on frequency. Hence, the radiated wave is directed in the yz -plane, as shown in Figure 2.1. More details on the radiation properties of LWAs will be provided in Section 2.4.

2.2 Classification of Leaky-Wave Antennas

LWAs are divided into two classes: *uniform* and *periodic*.² In this work, all LWAs considered are periodic but for the sake of consistency and completeness, a brief introduction to uniform LWAs is followed.

2.2.1 Uniform LWAs

Uniform LWAs represent a class of structures that are invariant with respect to their longitudinal axis, i.e., propagation axis (y). The design of W. W. Hansen published in his patent in 1940 [16] is the first known LWA. His design was based on perturbing the surface currents on the sidewall of a hollow *rectangular waveguide* (RWG) with a uniform longitudinal slot cut, or *slit*, operating in the fundamental mode TE_{10} , as

¹The imaginary part of a complex number in this work is always denoted by the imaginary unit j , where $j = \sqrt{-1}$, as per electrical engineering convention.

²This classification is valid for both 1D and 2D LWAs.

shown in Figure 2.2(a). Additionally, Hansen did not provide an analytical solution for his invention. By the late 1950s, Oliner and Goldstone provided a complete and rigorous analytical solution for this type of LWAs [37], and later they extended their work to *circular waveguide* LWAs [38].

This class of LWAs suffers from *two* major problems. First, disturbing the surface currents causes an abrupt change in the mode of the guided wave, which results in a high attenuation constant, α . This problem leads to a very small aperture even for physically long structures and hence, wide beamwidth, i.e., low directivity. Second, the phase constant of the dominant mode is always positive, i.e., $\beta > 0$, which leads to radiation only in the forward direction.³ More importantly, radiation in the broadside direction is not possible, as this would require $\beta = 0$, which is only possible if $\omega = 0$. To overcome the high leakage problem, Figure 2.2(b) shows a tapered slot in the broad wall of the RWG to achieve direction-dependent leakage factor $\alpha(y)$ [39]. Although the slot in this case is not uniform, the taper follows a continuous variation, which in the limiting case can be considered as uniform. The problem of the radiation direction, specifically the broadside radiation, can be solved by exciting the structure at the center with an appropriate feeding mechanism [18].⁴

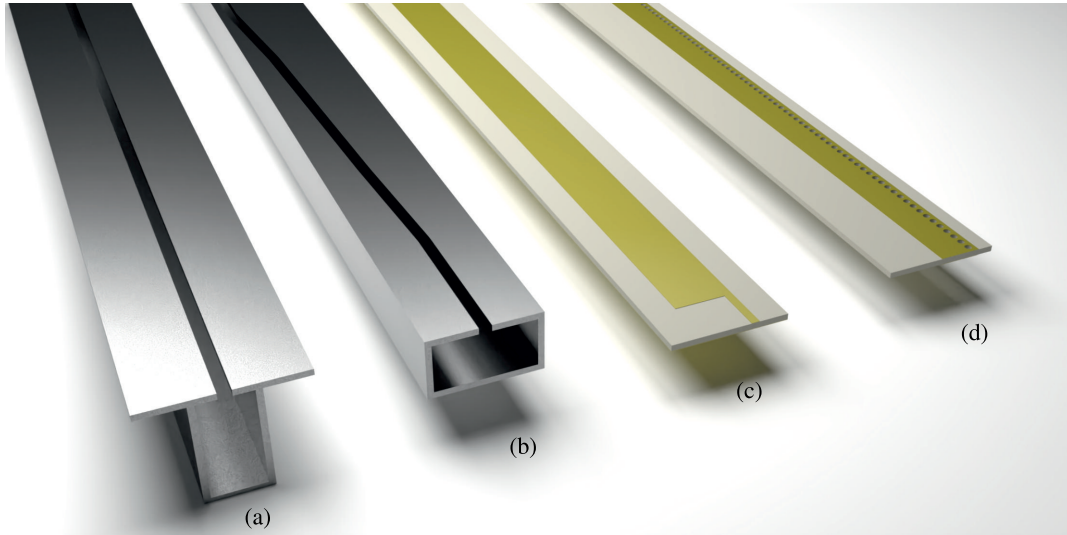


Figure 2.2: Examples of uniform leaky-wave antennas. (a) Slitted RWG, (b) tapered slitted RWG, (c) uniform microstrip line, and (d) half-mode SIW LWAs.

Additionally, different implementations of uniform LWAs have been reported. For example, Figure 2.2(c) shows a wide microstrip line structure that was invented by W. Menzel [40], operating in the non-TEM high-order mode EH_1 . Later, Oliner analyzed this type of LWAs in [41]. Finally, a *substrate-integrated waveguide* (SIW) operating in the half-mode TE_{10} is shown in Figure 2.2(d) [42], which is based on its classical RWG counterpart [43].

³The relationship between the phase constant, β , and the direction of radiation will be discussed in Section 2.4.2.

⁴Exciting an LWA at the center would generate two traveling waves propagating in opposite directions, which results in a single-beam but eventually a split-beam radiation as the frequency changes.

2.2.2 Periodic LWAs (P-LWAs)

Periodic LWAs represent a class of structures that are periodically modulated with respect to their transversal axis. The periodic modulation can be achieved by introducing periodic perturbations or discontinuities along the structure's axis (y). The first known implementation of a P-LWA was the *holey waveguide* reported by Hines and Upson [44], which replaced the long uniform slot [Figure 2.2(a)] with tightly spaced holes in the narrow wall of a dielectric-filled RWG. The spacing between the holes is called the *period* (p). Due to the nature of the periodicity, a P-LWA supports an infinite number of *space harmonics*, which allow the high-order space harmonics of non-radiative fundamental β_0 to radiate from the *backward* to the *forward* directions. The theory of periodic structures will be explained in greater detail in Section 2.3.

Figure 2.3 shows different examples of P-LWAs. Figure 2.3(a) depicts the holey waveguide P-LWA [44]. Figure 2.3(b) shows an RWG with longitudinal slots on its broad wall [45, 46]. The reason for alternating the slots is to closely space them by taking advantage of the phase jumps of the surface current phase due to symmetry [47]. Planar implementations of P-LWAs with either patch [48, 49] or stub [50] discontinuities are shown in Figure 2.3(c) and Figure 2.3(d), respectively. Planar implementation of the slotted RWG shown in Figure 2.3(a) based on SIW can be seen in Figure 2.3(e) [51].

Finally, P-LWAs based on 1D metamaterial transmission lines were investigated in the early 2000s [20, 52]. Figure 2.3(f) and Figure 2.3(g) show two different implementations of *composite right/left-handed* (CRLH) P-LWA based transmission lines,⁵ metal-insulator-metal (MIM) [54] and interdigital capacitor (IDC) [55], respectively.

In this work, only the periodic class of LWAs will be considered due to the aforementioned frequency-scanning capability that potentially covering the full upper hemisphere above the antenna.

2.3 Theory of Periodic Structures

Periodic structures are divided into *shielded* and *non-shielded* or *open* classes, where the former guides the electromagnetic wave along the structure and the latter provides radiation. Both classes can be characterized by a theory developed by the French mathematician G. Floquet, hence “Floquet’s theorem” [56],⁶ which deals with the solution to periodic linear differential equations. A myriad of literature deals extensively with Floquet’s theorem in the context of 1D periodic structures [19, 58–61, and references therein].

⁵This type of P-LWAs may be considered as *quasi uniform* due to the sub-wavelength period, i.e., $p \ll \lambda_g$, where λ_g is the wavelength of the guided wave [19, 53]. In this thesis, this type will be treated as periodic since it exhibits the same properties as all P-LWAs shown in Figures 2.3(a-e)

⁶Although G. Floquet first described the solution for periodic linear differential equations, it is F. Bloch who applied it to a physical structure, specifically in a 2D crystal lattice [57]. Therefore, some refer to this theorem as “Floquet-Bloch theorem” [20].

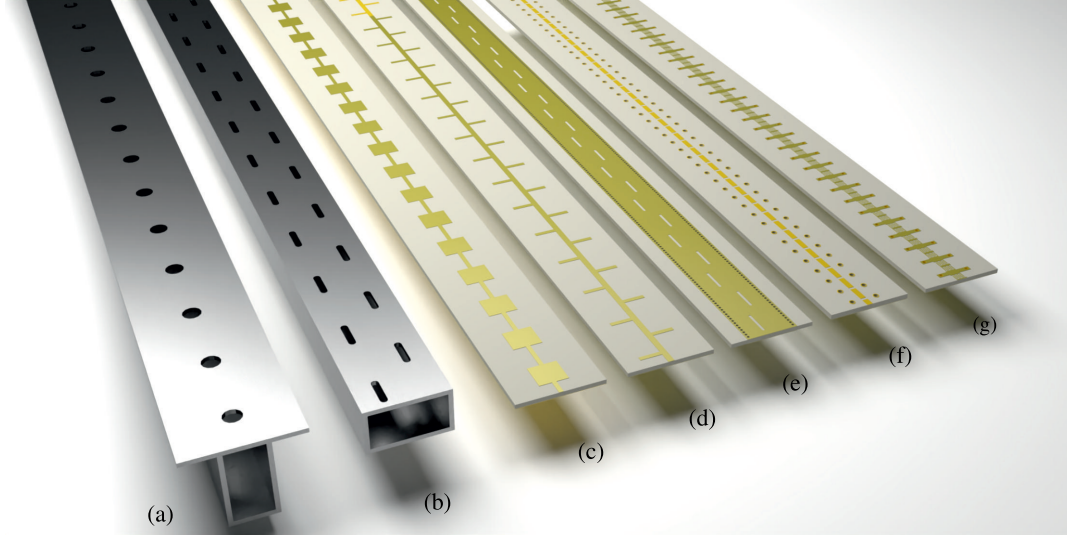


Figure 2.3: Examples of periodic leaky-wave antennas. (a) Slotted and (b) holey rectangular waveguides, (c) series-fed patch, (d) comb-line, (e) slotted substrate-integrated waveguide, and (f-g) interdigital and metal-insulator-metal antennas.

2.3.1 Floquet Theorem

Examples of P-LWAs are shown in Figure 2.3 with an arbitrary geometries with infinitely cascaded, or repeated, unit cells. The field of the propagating electromagnetic wave, i.e., a *Bloch wave*, at an arbitrary location along the structure's axis (y), differs from the field at an arbitrary integer multiple of the period, p , only by the complex propagation constant γ . Floquet's theorem describe mathematically the relationship of the time-harmonic⁷ electric field $\mathbf{E}$ ⁸ at positions separated by p as

$$\mathbf{E}(x, y + mp, z) = \mathbf{E}(x, y, z) e^{-\gamma mp}, \quad (2.5)$$

where $m \in \mathbb{Z}$ represents together with the period p a translational shift along the y -direction. As mentioned in Section 2.1, the complex propagation constant γ here can be expressed by rewriting Equation 2.2 as

$$\gamma = \alpha + j\beta = \alpha + j\beta_0(\omega), \quad (2.6)$$

where β_0 is the phase constant of the *fundamental* propagating mode of the guided wave. Furthermore, the attenuation constant, α , is equal for all space harmonics and is null for a guided and lossless structure. For a better representation of the periodic nature of the field, Equation 2.5 may be rewritten in terms of a periodic vector function $\mathbf{P}$ and the complex propagation constant, as

$$\mathbf{P}(x, y, z) = \mathbf{E}(x, y, z) e^{+\gamma y}. \quad (2.7)$$

In order for Equation 2.7 to obey the *periodic boundary condition* (PBC), $\mathbf{P}$ must satisfy

$$\mathbf{P}(x, y, z) = \mathbf{P}(x, y + mp, z). \quad (2.8)$$

⁷Throughout the thesis, a time-dependence $e^{j\omega t}$ is always assumed. Furthermore, the quantity $e^{j\omega t}$ together with $e^{-j\beta y}$ describe a propagating wave in the forward (+ y) direction.

⁸The derivation for the magnetic field $\mathbf{H}$ is analogous.

2.3.2 Space Harmonics

The electric field of a periodic structure can be represented by an infinite number of *space harmonics*. Since $\mathbf{P}$ is a periodic function, it can be represented by a Fourier series. The Fourier series expansion of $\mathbf{P}$ in Equation 2.7 is

$$\mathbf{P}(x, y, z) = \sum_{n=-\infty}^{\infty} E_n(x, z) e^{-jn\frac{2\pi}{p}y}, \quad (2.9)$$

where $n \in \mathbb{Z}$ is the number of the space harmonic and $E_n(x, z)$ is the cross-sectional electric field of the n th space harmonic, which represents the complex coefficients of the Fourier series. Inserting Equation 2.9 into Equation 2.7 results in an expression for the electric field

$$\mathbf{E}(x, y, z) = \sum_{n=-\infty}^{\infty} E_n(x, z) e^{-(\gamma + jn\frac{2\pi}{p})y}. \quad (2.10)$$

The propagation term in Equation 2.10 can be defined for the n th space harmonic as

$$\gamma_n = \gamma + jn\frac{2\pi}{p} = \alpha + j\left(\beta_0 + n\frac{2\pi}{p}\right). \quad (2.11)$$

Finally, the expression for the electric field in a periodic structure with a period p is obtained as

$$\mathbf{E}(x, y, z) = \sum_{n=-\infty}^{\infty} E_n(x, z) e^{-\gamma_n y}. \quad (2.12)$$

The last expression states that the electric field, or the Bloch wave in general, in a periodic structure is the expansion or *superposition* of an infinite number of space harmonics and can be solely described by the sum of cross-sectional Fourier coefficients E_n and $\hat{y}$ -dependent propagation constants γ_n . Furthermore, these space harmonics do not satisfy the PBC individually but they have to be all present simultaneously. It is worth noting that the space harmonics propagate with different phase constants, β_n , and experience the same loss constant, α .

Next, the explicit solution for the complex Fourier coefficient E_n is derived by considering a lossless system, i.e., $\alpha = 0$. First, multiplying both sides of Equation 2.7 by $e^{[j(2\pi m/p)y]}$ and integrating them over one period, p , yields

$$\int_{-p/2}^{p/2} \mathbf{P}(x, y, z) e^{jm(\frac{2\pi}{p})y} dy = \sum_{n=-\infty}^{\infty} \int_{-p/2}^{p/2} E_n(x, z) e^{j(m-n)\frac{2\pi}{p}y} dy. \quad (2.13)$$

Due to the orthogonality of the complex exponential function,

$$\int_{-p/2}^{p/2} e^{j(m-n)\frac{2\pi}{p}y} dy = \begin{cases} 0, & m \neq n, \\ p, & m = n, \end{cases} \quad (2.14)$$

Equation 2.13 can be rearranged as

$$E_n(x, z) = \int_{-p/2}^{p/2} \mathbf{P}(x, y, z) e^{jn(\frac{2\pi}{p})y} dy. \quad (2.15)$$

Equation 2.15 can be further simplified by multiplying its right-hand side by $e^{\pm j\beta_0 z}$, as follows

$$E_n(x, z) = \int_{-p/2}^{p/2} [\mathbf{P}(x, y, z) e^{-j\beta_0 y}] e^{j(\beta_0 + jn\frac{2\pi}{p})y} dy, \quad (2.16)$$

with β_n reading

$$\beta_n = \gamma_n|_{\alpha=0} = \beta_0 + jn\frac{2\pi}{p}. \quad (2.17)$$

Finally, the explicit solution is obtained for the complex Fourier coefficient E_n by substituting Equation 2.7 into Equation 2.16, which reads

$$E_n(x, z) = \frac{1}{p} \int_{-p/2}^{p/2} \mathbf{E}(x, y, z) e^{+\gamma_n y} dy. \quad (2.18)$$

Equation 2.18 is valid for periodic waves, as in the case in a lossless periodic structure. For the case of lossy or radiative structures, i.e., with a non-zero attenuation constant ($\alpha \neq 0$), the electric field decays by $e^{-\alpha y}$, hence the the positive sign in the propagation term, $e^{+\gamma y}$, in Equation 2.18 allows retrieving the periodicity of the field along the periodic structure. Additionally, the power carried by each space harmonic is proportional to the square of their coefficient, E_n .

2.3.3 Dispersion Diagram

The dispersion diagram is also called $k - \beta$ or $\omega - \beta$ *diagram*, or *Brillouin diagram*, after the French physicist Léon Brillouin, who first applied it to study the wave propagation in periodic crystals [58, 62]. It plots the phase constant, β , and therefore the space harmonics, versus the radial frequency, ω . Moreover, the dispersion diagram is not limited to periodic structures but is also used to study various types of microwave and optical components, such as waveguides, filters, and optical fibers. Furthermore, it is a very useful graphical tool to design P-LWAs and hence, it is of major importance to this thesis.

In the previous section, the derived expression for the phase constant in a periodic structure [Equation 2.17] relates the phase constant of the fundamental spatial harmonic, β_0 , with that of the n th space harmonic. Figure 2.4 plots the dispersion diagram of a P-LWA's unit cell, which is constructed by first calculating the fundamental phase constant, β_0 , and then translating it by $n2\pi/p$ along the horizontal axis of the graph. The dispersion diagram also depicts the direction of wave propagation. For instance, the blue curves indicate that the wave is propagating in the forward

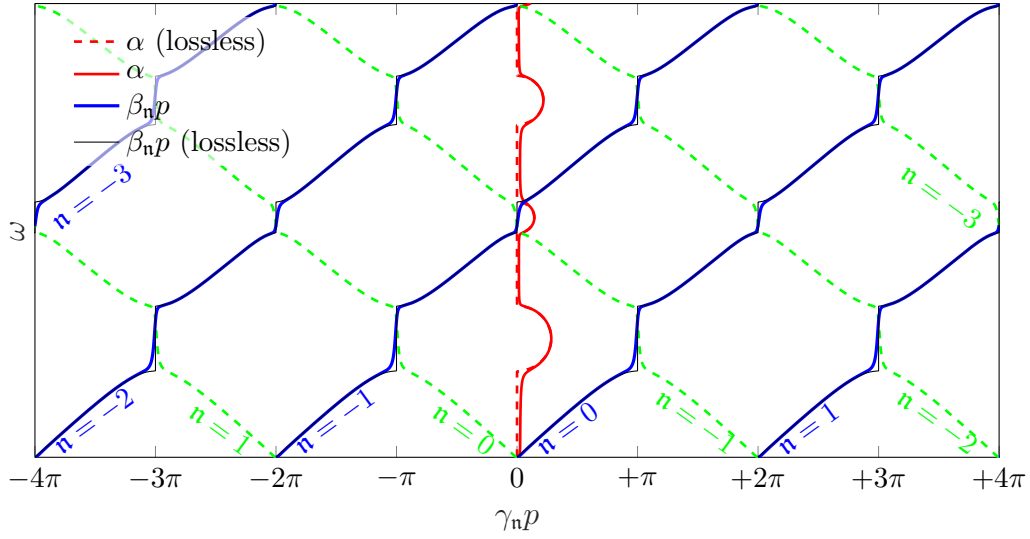


Figure 2.4: Graphical representation of the complex propagation constant, the dispersion diagram. The blue and green lines represent space harmonics of forward (+ y -directed) and backward ($-y$ -directed) propagating waves, respectively.

direction, i.e., + y -direction, while the green dashed curves indicate propagation in the backward direction, i.e., $-y$ -direction. This fact can be verified by calculating the phase velocity of the space harmonics, $v_{p,n}$, which is expressed as follows [63]

$$v_{p,n} = \frac{\omega}{\beta_n} = \frac{\omega}{\beta_0 + 2\pi n/p}. \quad (2.19)$$

The phase velocity can be either positive or negative corresponding to the direction of wave propagation, i.e., *forward* (FW) or *backward waves* (BW), respectively. Additionally, the phase velocity is different for different space harmonics. On the other hand, the group velocity [62, 64],

$$v_{g,n} = \frac{d\omega}{d\beta_n} = \frac{d\omega}{d(\beta_0 + 2\pi n/p)} = v_{g,0}, \quad (2.20)$$

which is the derivative the frequency over the phase constant, is the same for all space harmonics, i.e., $v_{g,n} = v_{g,0}, \forall n$, where $v_{g,0}$ is the group velocity of the fundamental space harmonic. This fact becomes clear when inspecting the slope, i.e., the derivative, of the phase constants in Figure 2.4.

The dispersion diagram shows also the regions of pass and stopbands, which can be seen in Figure 2.4, by the relatively low and high attenuation constant, α , respectively. This effect is even more prominent if a lossless and bounded (non-radiative) structure is considered, as shown by the dashed red line. Also, the phase constant shows these bands, where a stopband occurs when the phase constant is the same for different frequencies within a specific and continuous bandwidth. Since the group velocity is proportional to the real power flow of the wave, then Equation 2.20 may be applied on the dispersion diagram in Figure 2.4, which results in $v_{g,0} = 0$ for the

stopband regions. The real power flow in a lossless periodic structure is given by [60]

$$\Re \left\{ \frac{1}{2} \iint_{n+1-n+1} \mathbf{E}_n \times \mathbf{H}_n^* \cdot \mathbf{e}_y dx dz \right\} = \frac{W_e + W_m}{p} \frac{d\omega}{d\beta_n} = \frac{W_e + W_m}{p} v_{g,0}, \quad (2.21)$$

where $\mathbf{E}_n \in \mathbb{C}$ and $\mathbf{H}_n^* \in \mathbb{C}$ are the cross-sectional electric and magnetic fields at the n th terminal of the periodic structure, respectively. W_e and W_m are the stored electric and magnetic energies, respectively. It is worth noting that Equation 2.21 is only valid for narrowband signals propagating through a lossless structure. Nevertheless, it shows that the power flow of the Bloch wave vanishes for the case of a zero group velocity $v_{g,0} = 0$.

2.4 Radiation Properties of P-LWAs

This section discusses important radiation properties of P-LWAs. Starting from the discussion of periodic structures in the previous section, the dispersion diagram in Figure 2.4 is restricted to only the *Brillouin zone*, $-\pi \leq \gamma_n p \leq +\pi$, without loss of generality, as shown in Figure 2.5.

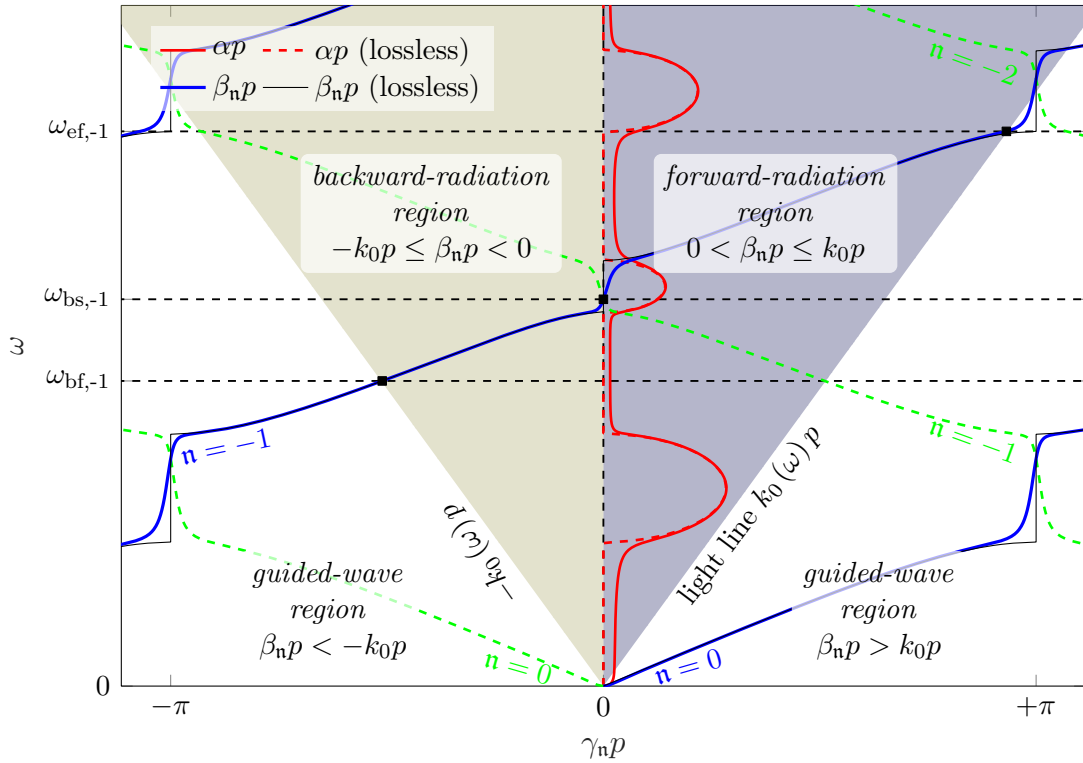


Figure 2.5: Typical dispersion diagram of a P-LWA with its guided and radiation regions.

Figure 2.5 plots the dispersion diagram of a P-LWA radiating with $\mathbf{n} = -1$ and fed from the $-y$ -direction, as depicted in Figure 2.6, which corresponds to the blue curves of the phase constants. Generally, the dispersion diagram for P-LWA can be divided into two regions, *guided* and *radiation* regions, where the boundary of these

regions is divided by the *light line*, i.e., $k_0 = \pm\omega/c$ with c being the speed of light. If $|\beta_n p| > k_0 p$, then the wave will be guided along the structure and called a *slow wave*, since its phase velocity $v_{p,n} = \omega/\beta_n < c$ [Equation 2.19]. In contrast, if $|\beta_n p| < k_0 p$, then leaky-wave radiation will occur and the wave will be called a *fast wave*, since its phase velocity $v_{p,n} > c$. Additionally, the region between the light lines, $\pm k_0 p$, is called the *radiation cone*, which is derived from the dispersion diagram of 2D P-LWAs, where the dispersion diagram is a two-dimensional function in space. Furthermore, a fast wave can either radiate in the forward or backward direction depending on the sign of the phase constant of that space harmonic. It is worth noting, that backward and forward definition is always related to the feed direction, and hence, for the a P-LWA fed from the $+y$ -direction the directions of forward and backward radiation will be flipped.

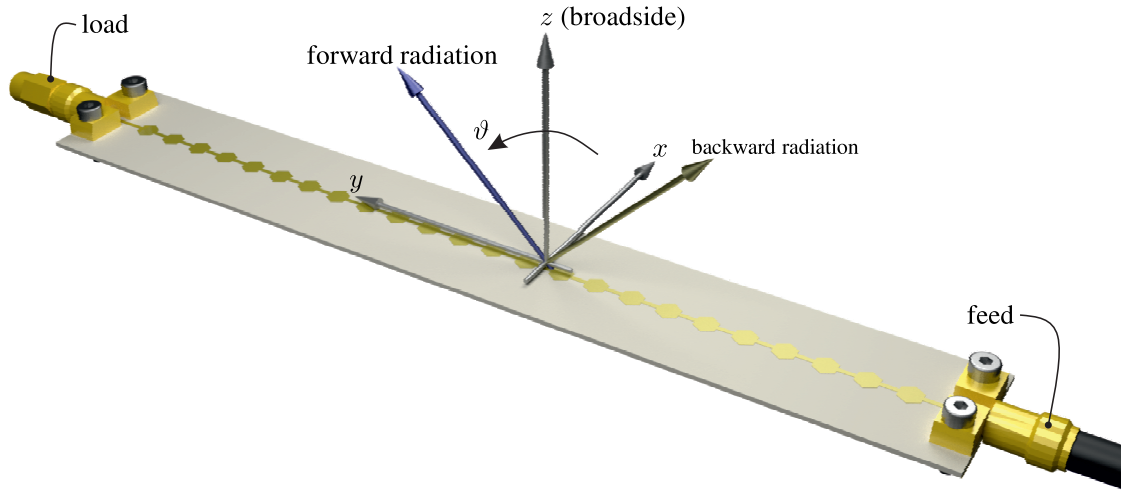


Figure 2.6: Illustration of a planar P-LWA under its normal operational condition. In this thesis, the main lobe scanning with frequency is defined in the principal yz -plane.

2.4.1 Radiation Pattern

The radiation pattern of a P-LWA with a length L can be obtained by transforming the antenna (fields) aperture along the y -axis using the spatial Fourier transform [65]. This transformation will result in a closed-form expression of the array factor of the aperture, which reads [19, 39, 66]

$$AF(\vartheta) = \ell \exp\left(2 \frac{jk_0 \sin \vartheta - \gamma}{\ell}\right) \text{sinc}\left[\frac{\ell}{2}(j\gamma + k_0 \sin \vartheta)\right], \quad (2.22)$$

where $\text{sinc}(x) = \sin(x)/x$ is the unnormalized sinc function and ℓ is the total length of the P-LWA, and ϑ is the angle measured from the z -axis in the yz -plane.

2.4.2 Frequency Scanning

For a fast wave, the main beam direction, $\vartheta_{mb,n}$, of the radiated field can be calculated by the phase constant β_n and k_0 , which are frequency dependent. Hence, the main

beam direction can be expressed by the *scanning law* with the assumption $\beta_n \gg \alpha$ as [39]

$$\vartheta_{\text{mb},n}(\omega) = \arcsin \left[\frac{\beta_n(\omega)}{k_0} \right]. \quad (2.23)$$

For the case $\beta_n = \pm k_0$, i.e., the points where a space harmonic crosses the light lines, the main beam direction will be directed either in the *backfire* ($\beta_n = -k_0$) or in the *endfire* ($\beta_n = +k_0$) directions, respectively, where the backfire direction points towards the feed, i.e. $\vartheta_{\text{mb},n} = -90^\circ$, and the endfire direction points toward the antenna termination, i.e. $\vartheta_{\text{mb},n} = +90^\circ$. For the case $\beta_n = 0$, the main beam points to the broadside direction, i.e., $\vartheta_{\text{mb},n} = 0^\circ$, which corresponds to a phase velocity approaching infinity [Equation 2.19]. Moreover, Equation 2.23 explains the frequency-scanning capability of P-LWA, as the main beam can be scanned from the backfire to the endfire and through the broadside direction solely by changing the frequency. The scanning curve of the main beam over frequency follows a hyperbolic function if the phase constant is linear over frequency, which is only true for TEM transmission lines with vanishingly-small perturbations.

2.4.3 Scanning Range

The scanning range in a P-LWA is not only limited within the boundaries of the radiation cone but also within the frequency bandwidth at which only a single space harmonic is of a fast wave nature. If multiple space harmonics of fast-wave type exist, then the radiation of the P-LWA multiple will radiate multiple beams. In fact, this behavior is very similar to that of phased arrays when a grating lobe enters the visible region [65]. The multiple-beam radiation is considered as a drawback since most applications require single-beam radiation. This section discusses the conditions for a P-LWA to mitigate double-beam radiation.

Figure 2.7(a) shows the vanishingly-small periodic-perturbation version of a dispersion diagram for a homogenous transmission line filled with a dielectric and non-magnetic material.⁹ The dispersion of the transmission line is given by the phase constant of the fundamental space harmonic,

$$\beta_0 = \frac{\omega}{c} \sqrt{\epsilon_r}, \quad (2.24)$$

where ϵ_r is the relative permittivity of the bulk dielectric. Moreover, shifting β_0 by integer multiples of $\pm 2\pi$, as in Equation 2.11 in Figure 2.7(a), constructs the dispersion diagram.

In the case of Figure 2.7(a), the P-LWA radiates in the first space harmonic, $n = -1$, with a single beam from the backfire frequency $\omega_{\text{bf},-1}$ through broadside ($\omega_{\text{bs},-1}$) to the backfire frequency of the second space harmonic $\omega_{\text{bf},-2}$. Between $\omega_{\text{bf},-2}$ and $\omega_{\text{ef},-1}$, two space harmonics radiate, $n = -1$ and $n = -2$, which also can be seen as a grating lobe. Now, to avoid simultaneously radiating with two space harmonics,

⁹A Non-magnetic material implies a relative permeability of 1, i.e., $\mu_r = 1$. Throughout this work, all materials considered are of this type.

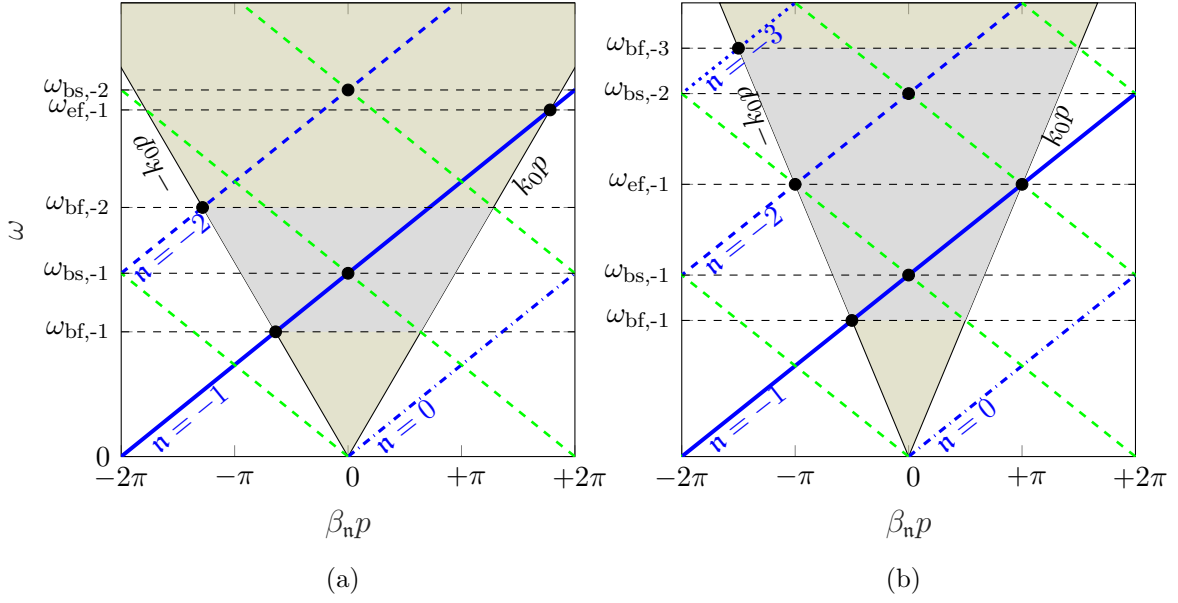


Figure 2.7: Dispersion diagrams of dielectric-filled and homogenous transmission lines. (a) General case ($\epsilon_r < 9$) with double beam space scanning in the frequency range $\omega_{bf,-2} - \omega_{ef,-1}$ and (b) unique case ($\epsilon_r = 9$) with single beam scanning in the frequency range $\omega_{bf,-1} - \omega_{bf,-3}$.

one can satisfy the condition $\omega_{ef,-1} \leq \omega_{bf,-2}$ and solve Equation 2.11 after inserting Equation 2.24 for ϵ_r , which yields [18]

$$\epsilon_r \geq 9. \quad (2.25)$$

Figure 2.7(b) shows the small-perturbation dispersion diagram for $\epsilon_r = 9$, which satisfies Equation 2.25 and results in a single beam radiation from the backfire frequency $\omega_{bf,-1}$ to the endfire $\omega_{ef,-1}$ of $n = -1$. It is important to point out that the single beam condition in the case of a homogenous transmission line can be only satisfied for a single space harmonic [Figure 2.7(b)]. For multiple space harmonics radiation with a single beam, dispersion engineering can be exploited to satisfy the dispersion, as follows

$$\omega_{ef,n-1} \leq \omega_{bf,n-2}, \quad \forall n. \quad (2.26)$$

Although Equation 2.25 results into single beam radiation, it is not practical due to poor radiation efficiency of the antenna. In practice, the required ϵ_r to achieve single beam radiation is lower than that of Equation 2.25 due to the periodic perturbations of the P-LWA. Moreover, ϵ_r may be reduced to $\epsilon_r \geq 4$ by employing phase-reversing elements [67].

Using Equation 2.23 the main beam direction can be calculated for the dispersion diagrams in Figure 2.7. Figure 2.8(a) plots the main beam direction $\vartheta_{mb,n}$ for the dispersion diagram in Figure 2.7(a), which shows the double-beam radiation in the frequency range $\omega_{bf,-2} \leq \omega \leq \omega_{ef,-1}$. In contrast, the dispersion diagram in Figure 2.7(b) results to full-space scanning with a single beam using a single space harmonic ($n = -1$) for the frequency range $\omega_{bf,-1} \leq \omega \leq \omega_{ef,-1}$, as shown in Figure 2.8(b).

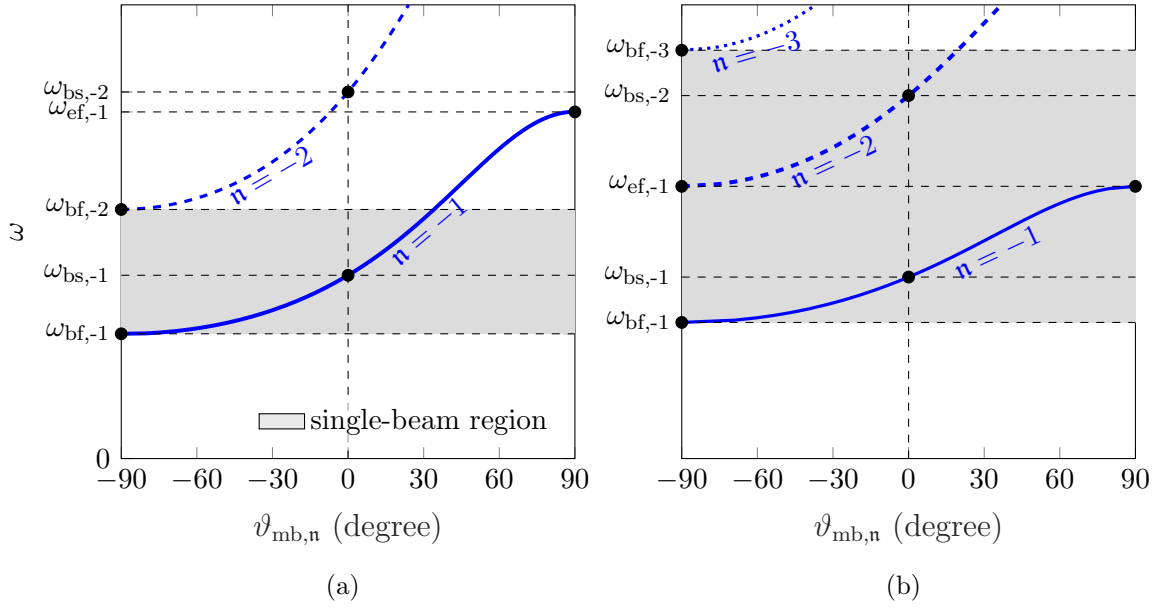


Figure 2.8: Main beam direction calculated from the dispersion diagrams in Figure 2.7. (a) Double-beam scanning for the general case with ($\epsilon_r < 9$) [Figure 2.7(a)] and (b) single-beam scanning for the unique case with ($\epsilon_r = 9$) [Figure 2.7(b)].

2.4.4 Scanning Rate

The scanning rate is the rate at which the main beam of a P-LWA steers over frequency. The higher the rate the more sensitive is the main beam over frequency variation. The scanning rate can be calculated by the derivative of the main beam angle given in Equation 2.23, which yields

$$\xi_n(\omega) = \frac{d\vartheta_{mb,n}}{d\omega}. \quad (2.27)$$

As previously stated in Section 2.4.2, the main beam over frequency follows a hyperbolic path, which means that the scanning rate is not constant over frequency, as can be seen from Figure 2.8, and that the scanning rate is function of frequency similar to the phase constant of a specific space harmonic [Figure 2.7]. Furthermore, the scanning rate is also different for different space harmonics. Nevertheless, many papers give the scanning rate as a single value, which is calculated by the boundaries of the beam scanning range divided by the frequency bandwidth between these boundaries.

2.4.5 Polarization

The polarization of a P-LWA is simply determined by the orientation of the electric field of the perturbation, i.e., radiating element [68]. The P-LWAs shown in Figure 2.3 are all linearly polarized, which is usually the case for 1D P-LWAs. In Chapter 4, the polarization control of the P-LWA radiation will be discussed in great detail.

2.5 Summary

The chapter introduced theory, classification, and radiation properties of the LWAs that will be the foundation for the forthcoming chapters. First, the radiation principle in LWAs has been discussed. Next, the uniform and periodic classes of LWAs have been introduced with historical and modern implementation examples of these classes. Moreover, in this thesis strictly P-LWAs will be considered. Therefore, the theory of periodic structures was analyzed using Floquet's theorem with emphasis on the space harmonics and dispersion diagram, since both govern the radiation properties of P-LWAs. Finally, important formulas for P-LWA radiation are presented.

Symmetry Properties in Periodic Leaky-Wave Antennas

In this chapter, the broadside radiation problem in P-LWAs will be presented and investigated using a generic network model. The important parameters of P-LWA such as the Bloch impedance, propagation constant and radiation efficiency will be introduced and then conditions for resolving the broadside problem will be discussed by utilizing symmetry concepts in the unit cell of P-LWAs.

3.1 Broadside Radiation Problem

The broadside problem, also called the *open-stopband problem*,¹ in P-LWAs is a prominent obstacle that impedes their use for applications where broadside radiation is required. Almost all P-LWAs suffer from this problem if no specific mitigation techniques are considered. The problem arises at the broadside frequency, i.e., $\beta_n(\omega_{bs}) = 0$, where all radiating elements are excited in-phase. At this frequency, the expected traveling wave along the structure turns into a standing wave causing a drop in the attenuation constant, α [18, 19, 53].

To visualize this problem, Figure 3.1 plots important parameters for a P-LWA versus the frequency. Figure 3.1(a) shows the propagation constant, similar to that of Section 2.3.3, where the broadside frequency, f_{bs} is defined at $\beta = 0$. At this frequency, the attenuation constant (α) drops, indicating a standing-wave operation. Moreover, this standing-wave behavior is also visible in the Bloch impedance, Z_B , as shown in Figure 3.1(b), where the peak of the impedance at broadside indicates a resonance in the unit cell. Finally, the open-stopband problem can be summarized in the abrupt drop of the realized gain of the antenna and therefore, the overall antenna efficiency, as shown in Figure 3.1(c).

Within this chapter, network models will be introduced to derive the optimum conditions to analyze and mitigate the open-stopband problem at broadside by exploiting the unit cell's symmetry.

3.2 Network Modeling of an Infinite Cascade of Unit Cells

This section describes two important quantities of the propagating wave in a periodic structure, the complex propagation constant and the Bloch impedance, as they

¹The term *open* refers to the radiation nature of the structure, i.e., non-bounded structure. Other stopbands are not part of the open-stopband problem, see Section 2.3.3 and Figure 2.5, as they lay outside the radiation cone and do not contribute to radiation.

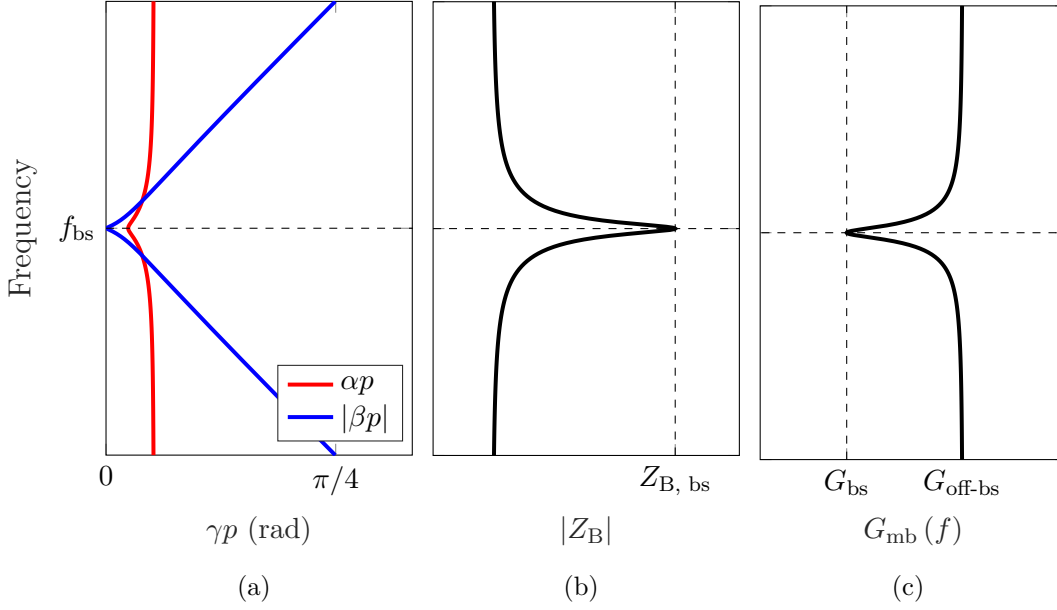


Figure 3.1: Description of the broadside problem for P-LWA's parameters. (a) The propagation constant γ , (b) the Bloch impedance Z_B and (c) the gain of the main beam versus frequency. The open-stopband problem causes an abrupt change across all parameters at the broadside frequency f_{bs} .

directly control the radiation characteristics of P-LWAs. A periodic structure with an infinite number of unit cells can be modeled as a two-port network [69, 70].

3.2.1 Propagation Constant

Figure 3.2(a) depicts a generic planar P-LWA with an infinite number of cascaded unit cells. For a wave propagating in the positive y -direction, the voltage and current at the n^{th} terminal of a unit cell are related to their counterparts at the $n^{\text{th}} + 1$ by the transmission (**ABCD**) matrix, as shown in Figure 3.2(b), as follows

$$\begin{bmatrix} V_n \\ I_n \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_{n+1} \\ I_{n+1} \end{bmatrix}, \quad (3.1)$$

where V and I are the current and voltage, respectively. By applying Floquet's theorem [Section 2.3.1], the voltages and currents at successive terminals can be related to each other using the propagation constant as

$$\begin{bmatrix} V_n \\ I_n \end{bmatrix} = \begin{bmatrix} e^{\gamma p} & 0 \\ 0 & e^{\gamma p} \end{bmatrix} \begin{bmatrix} V_{n+1} \\ I_{n+1} \end{bmatrix}. \quad (3.2)$$

Substituting Equation 3.2 into Equation 3.1 yields

$$\begin{bmatrix} A - e^{\gamma p} & B \\ C & D - e^{\gamma p} \end{bmatrix} \begin{bmatrix} V_{n+1} \\ I_{n+1} \end{bmatrix} = 0, \quad (3.3)$$

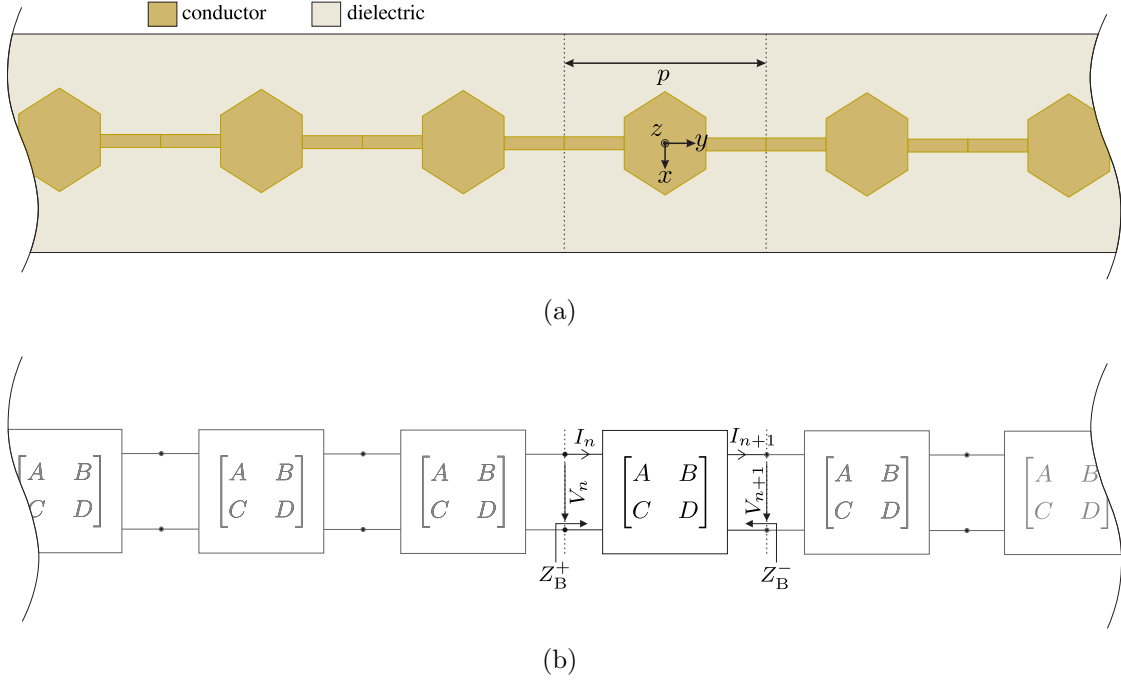


Figure 3.2: Network modeling of cascaded unit cells of P-LWAs. (a) Top view of a generic planar P-LWA and (b) the **ABCD** matrix corresponding to the network parameters of a unit cell.

which provides a nontrivial solution due to the zero determinant,

$$AD - BC - (A + D)e^{\gamma p} + e^{2\gamma p} = 0. \quad (3.4)$$

In this thesis, only reciprocal P-LWAs are considered, i.e., $AD - BC = 1$ [70], then Equation 3.4 simplifies to

$$A + D = e^{\gamma p} + e^{-\gamma p}, \quad (3.5)$$

The solution for the propagation constant can be obtained by solving Equation 3.5 as

$$\gamma p = \text{arccosh} \left(\frac{A + D}{2} \right). \quad (3.6)$$

For a symmetric unit cell with respect to its transversal axis, as shown in Figure 3.3(a) and Figure 3.3(b), the condition $A = D$ is always satisfied. Therefore, Equation 3.6 can be further simplified as

$$\gamma^{\text{sym}} p = \text{arccosh} (A). \quad (3.7)$$

Shifting the terminals of the unit cell by the period p will not affect the propagation constant γ , since the propagation constant is translationally symmetric.

3.2.2 Bloch Impedance

The Bloch impedance Z_B is defined as the characteristic impedance of the Bloch wave, which is equivalent to the input impedance of infinitely-cascaded unit cells.

In practice, the input impedance of an infinite P-LWA converges to the Bloch impedance if the residual power at the termination side is small relative to the input power. In general, a 1D P-LWA has two Bloch impedances depending on the wave propagation direction, i.e., forward and backward waves. The Bloch impedances of forward and backward propagating waves will be denoted by $+$ and $-$ signs, respectively. Mathematically, the forward Bloch impedance Z_B^+ can be calculated from the voltage and current at the unit cell terminals, as

$$Z_B^+ = \frac{V_n}{I_n} = \frac{V_{n+1}}{I_{n+1}}. \quad (3.8)$$

Analogously, the backward Bloch impedance Z_B^- can be written as

$$Z_B^- = \frac{V_n}{-I_n} = \frac{V_{n+1}}{-I_{n+1}}. \quad (3.9)$$

Using Equation 3.3, $(A - e^{\gamma d})V_{n+1} + BI_{n+1} = 0$, the Bloch impedance in Equation 3.8 can be rewritten as

$$Z_B^\pm = \frac{\mp B}{A - e^{\gamma p}}. \quad (3.10)$$

Now, solving Equation 3.4 for $e^{\gamma p}$ results in

$$e^{\gamma p} = \frac{(A + D) \mp \sqrt{(A + D)^2 - 4}}{2}. \quad (3.11)$$

Finally, Equation 3.8 can be rewritten for both forward and backward traveling waves cases as

$$Z_B^\pm = \frac{2B}{A - D \mp \sqrt{(A + D)^2 - 4}}, \quad (3.12)$$

For a symmetric structure with respect to its input and output ports, $A = D$ is valid and one can rewrite Equation 3.12 for the symmetric Bloch impedance ($Z_B^{\text{sym}} = Z_B^+ = -Z_B^-$) as follows

$$Z_B^{\text{sym}} = \sqrt{Z_B^+ Z_B^-} = \frac{B}{\sqrt{A^2 - 1}}. \quad (3.13)$$

3.3 Symmetry Classification

This section defines possible symmetry and asymmetry classes in planar P-LWAs. These symmetries are defined by mirroring the structure along at least one of the two possible axes of the P-LWA with respect to the propagation direction, i.e.,

longitudinal (y) and *transversal* (x) axes, as introduced by Otto [71]. Figure 3.3 depicts four classes of symmetries in a generic *microstrip* (MSL) P-LWA,² which consists of feeding lines and patch-like radiating elements. Figure 3.3(a) shows a *double-symmetric* (DS) P-LWA with its unit cells symmetric with respect to both longitudinal and transversal axes [23]. Figure 3.3(b) shows *longitudinal-asymmetric* (LA) P-LWA with its unit cells asymmetric with respect to the longitudinal axis and symmetric with respect to the transversal axis [24]. Figure 3.3(c) shows *transversal-asymmetric* (TA) P-LWA with its unit cells symmetric with respect to the longitudinal axis and asymmetric with respect to the transversal axis [25]. The last type of asymmetry combines both LA and TA, as shown in Figure 3.3(d). Specifically, *double-asymmetric* (DA) P-LWA consists of unit cells, which are asymmetric with respect to *both* longitudinal and transversal axes [26].

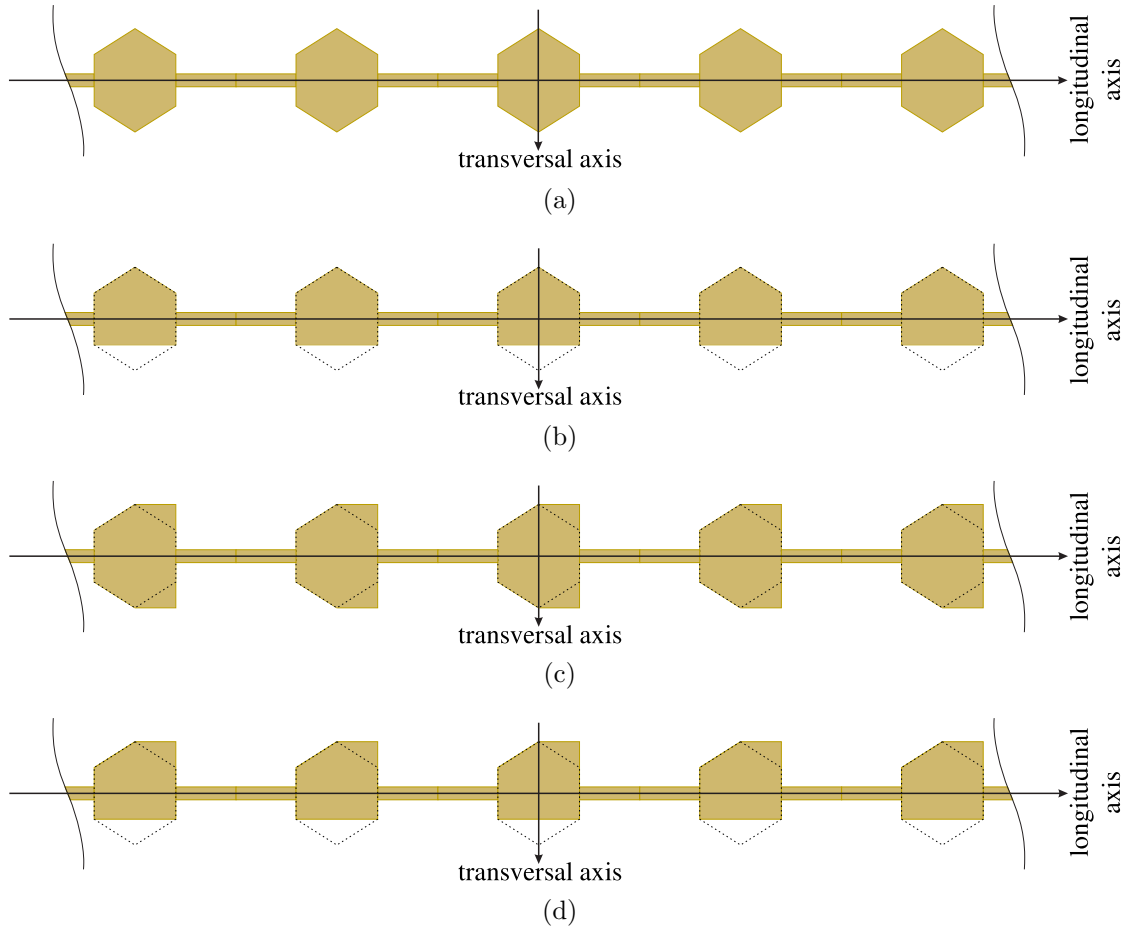


Figure 3.3: Symmetry classifications in P-LWAs. (a) Double symmetric (DS), (b) longitudinal asymmetric (LA), (c) transversal asymmetric (TA) and (d) double asymmetric (DA).

The symmetries/asymmetries, DS, LA, and TA, will be investigated in this chapter for the important parameters of P-LWAs, i.e., the propagation constant (γ) and

²These symmetries are not only restricted to MSL-based P-LWAs but also can be applied to non-TEM transmission lines.

the Bloch impedance (Z_B), whereas DA will be investigated for its polarization properties in Chapter 4 as it combines the properties of LA and TA.

Besides the aforementioned symmetries, further types can be applied to non-planar periodic structures, such as glide and screw symmetries, where glide symmetry utilizes mirror and translation symmetry, and screw symmetry utilizes rotation symmetry [72], respectively. These types of symmetries will not be considered in this work, since all P-LWAs considered here are strictly planar.

3.4 Double-Symmetric and Longitudinally-Asymmetric Unit Cells

A P-LWA with a *double-symmetric* (DS) unit cell is the most common type among the practical implementations of P-LWAs, whereas a *longitudinally-asymmetric* (LA) unit cell is not common, yet it provides unique radiation properties. Here, it will be shown that the former type suffers fundamentally from the open-stopband problem at broadside, while the latter one resolves it if designed properly. Furthermore, the DS and LA types will be investigated using a single generic network model introduced in [22]. Moreover, their radiation characteristics, in particular the polarization, will be discussed in Chapter 4.

The analytical formulas in this section are either taken from or based on the results in [22, 71] and [23, 73].

3.4.1 Odd and Even Mode Decomposition

Since the unit cell of a P-LWA can be considered as a two-ports network, the Bloch wave propagating either in the forward or backward directions, i.e., $\pm y$, may be decomposed into two modes under the odd and even excitations. These modes are called the *series* and *shunt* modes, where the naming is originally derived from circuit modeling of 1D transmission-line metamaterial [20]. To illustrate the nature of the series and shunt modes without loss of generality, a simple series-fed patch (SFP) P-LWA is shown in Figure 3.4. The unit cell consists of three transmission line sections, two feeding lines and a patch with lengths $\lambda_g/4$ and $\lambda_g/2$, respectively, where λ_g is the wavelength of the guided wave.

3.4.1.1 Series Mode

Under the odd excitation, the series mode is isolated. Furthermore, the radiation of the series mode corresponds to the normal $\lambda/2$ -resonant patch mode,³ as can be seen from the radiation pattern and the $\mathbf{E}$ -field seen in Figure 3.4(a). Thus, the series mode contributes to radiation in the broadside direction. Moreover, the series mode can be modeled by series resonator in the series path, as shown in Figure 3.4(c), whose parameters read

³Based on the cavity model of a rectangular patch antenna for an electrically thin substrate, the normal mode is TM_{10} [65].

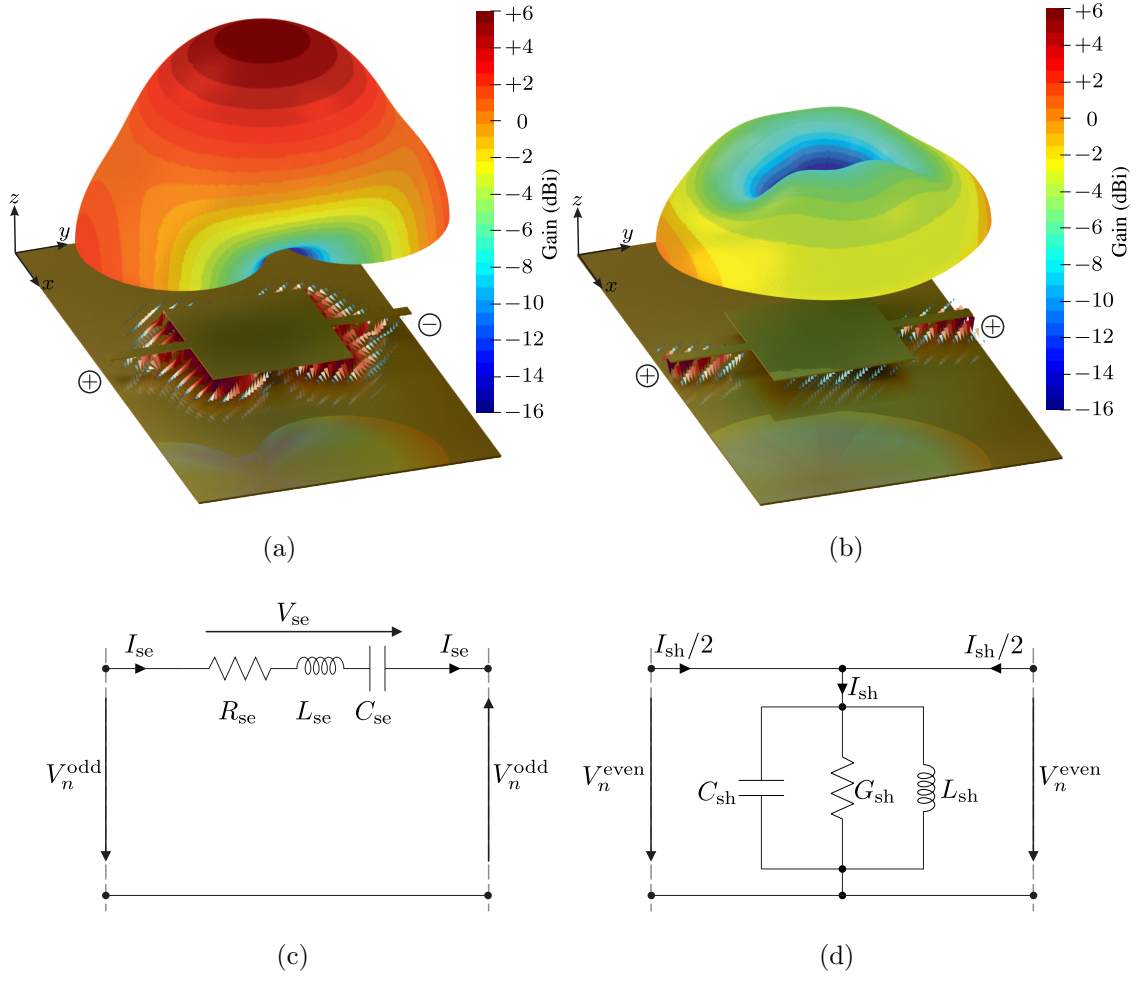


Figure 3.4: Odd and even excitations of a series-fed patch (SFP) P-LWA unit cell suspended in the air over a ground plane. (a)-(b) Unit cell and the resulted fields patterns. (c)-(d) The corresponding circuit models of the resonators, i.e., series and shunt resonators, respectively.

$$R = \Re(Z_{se}), \quad (3.14a)$$

$$L_{se} = \frac{1}{2} \frac{\partial}{\partial \omega} \Im \{Z_{se}\} |_{\omega=\omega_{se}}, \quad (3.14b)$$

$$C_{se} = \frac{1}{\omega_{se}^2 L_{se}}. \quad (3.14c)$$

where ω_{se} is the resonance frequency and R comprises the total loss/radiation of this mode. Using Equation 3.14, the series resonator impedance can be written as

$$Z_{se} = R + j \left(\omega L_{se} - \frac{1}{\omega C_{se}} \right). \quad (3.15)$$

In addition to the RLC parameters, another important parameter of any resonator is its *quality factor*, Q -factor, which reads [70, 74]

$$Q = \omega \frac{\text{average energy stored}}{\text{power loss}} = \omega \frac{W_e + W_m}{P_{\text{loss}}}, \quad (3.16)$$

where $W_e = W_m$ at the resonance frequency, ω_0 , hence, Q in Equation 3.16 can be generally defined for either a series or a shunt resonator as

$$Q = \omega_0 \frac{2W_e}{P_{\text{loss}}} = \frac{\omega_0 L}{R} = \frac{\omega_0 C}{G}. \quad (3.17)$$

As from Equation 3.17, the Q -factor of the series resonator shown in Figure 3.4(c), can be written as

$$Q_{\text{se}} = \frac{\omega_{\text{se}} L_{\text{se}}}{R}. \quad (3.18)$$

3.4.1.2 Shunt Mode

Figure 3.4(b) shows the SFP unit cell under even excitation. In this case, the radiation of the shunt mode does not contribute to broadside radiation due to the $\mathbf{E}$ -field excitation of the patch. It is worth noting that this mode does not correspond to the higher-order mode, TM_{20} , of a simple resonant patch antenna. For the shunt mode, the circuit model consists of a parallel resonator connected in the shunt path, as depicted in Figure 3.4(d). The parameters of the shunt resonator can be written as

$$G = \Re\{Y_{\text{sh}}\}, \quad (3.19a)$$

$$C_{\text{sh}} = \frac{1}{2} \frac{\partial}{\partial \omega} \Im\{Y_{\text{sh}}\}|_{\omega=\omega_{\text{sh}}}, \quad (3.19b)$$

$$L_{\text{sh}} = \frac{1}{\omega_{\text{sh}}^2 C_{\text{sh}}}. \quad (3.19c)$$

where ω_{sh} is the resonance frequency and G comprises the total loss/radiation of the shunt mode. Unlike the series resonator, it is more appropriate to write the admittance of the shunt resonator, Y_{sh} , which reads

$$Y_{\text{sh}} = G + j \left(\omega C_{\text{sh}} - \frac{1}{\omega L_{\text{sh}}} \right). \quad (3.20)$$

Similar to Equation 3.17, the Q -factor of the shunt resonator shown in Figure 3.4(d), can be defined as

$$Q_{\text{se}} = \frac{\omega_{\text{sh}} C_{\text{sh}}}{G} \quad (3.21)$$

The series and shunt mode definitions are general and can be applied to any unit cell with only two resonances under the odd and even excitations.

3.4.2 Linearization of the Series and Shunt Resonators

The impedance Z_{se} and admittance Y_{sh} of the series and shunt resonators usually behave linearly around their resonance frequencies, ω_{se} and ω_{sh} , respectively [22]. In particular, the imaginary part of these immittances are linear at their zero crossing, which allows for the extraction of the series inductor, L_{se} , and the shunt capacitance, C_{sh} , in Equations 3.14 and 3.19. By knowing this fact, the series and shunt immittances may be reduced to

$$Z_{\text{se}} = R + j2L(\omega - \omega_{\text{se}}) = R \left(1 + 2jQ_{\text{se}} \frac{\omega - \omega_{\text{se}}}{\omega_{\text{se}}} \right), \quad (3.22a)$$

$$Y_{\text{sh}} = G + j2C(\omega - \omega_{\text{sh}}) = G \left(1 + 2jQ_{\text{sh}} \frac{\omega - \omega_{\text{sh}}}{\omega_{\text{sh}}} \right), \quad (3.22b)$$

which simplifies the derivation of the P-LWA parameters in the forthcoming sections. For better readability, the series inductance L_{se} and shunt capacitance C_{sh} will be denoted as L and C , respectively.

3.4.3 Lattice Network Model

The concept of odd and even excitation discussed in Section 3.4.1 is useful to isolate and extract the parameters of the series and shunt modes. Nevertheless, it does not describe the normal P-LWA operation, i.e. the traveling-wave regime, where the antenna is excited at one end and terminated at the other end. Therefore, a generic circuit model is needed to combine the series and shunt resonators. Such a model would be essential to derive the P-LWA's parameters described in Section 3.2 concerning series and shunt parameters.

The generic network model must provide two properties. First, under odd and even excitations, the series and shunt resonators are isolated and their parameters can be extracted, as described in Section 3.4.1. Second, it must have an all-pass response, similar to the circuit model of a CRLH P-LWA, where the series and shunt resonators are connected in the series and shunt paths, respectively [20]. There are two-ports networks that provide the latter property, such as T and Π circuit models [70], but they do not provide the isolation property of the series and shunt resonators. From Bartlett's bisection theorem, any symmetrical two-ports network can be modeled by a lattice circuit [75]. Hence, the lattice circuit model was investigated in the context of P-LWA as a generic model in [22, 71]. In contrast to the T and Π networks, the lattice provides the aforementioned properties. Moreover, it provides simple expressions for the propagation constant and Bloch impedance. The lattice circuit model was first used investigated as an all-pass filter and later used for other applications, e.g. phase shifters [76].

Figure 3.5 shows the lattice circuit model, which comprises the series and shunt resonators. In this section, only the final results of the unit cell parameters will be provided and discussed. But for the sake of completeness, however, the derivation of these parameters is provided in Appendix A.1. Using the bisection theory, the series and shunt resonators can be written as [22, 77]

$$Z_{\text{se}} = 2(Z_{11} - Z_{21}), \quad (3.23a)$$

$$Y_{\text{sh}} = 2(Y_{11} + Y_{21}). \quad (3.23b)$$

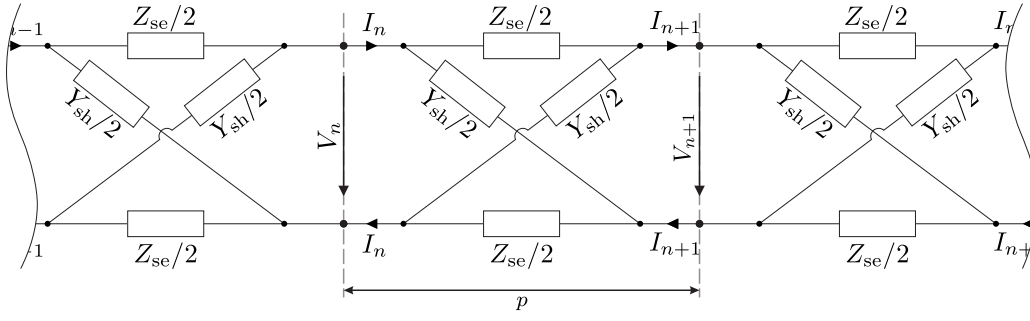


Figure 3.5: Cascaded lattice circuit as a generic model for P-LWAs unit cell. The series and shunt resonators are placed in the series and shunt paths, respectively, providing an all-pass response.

The **ABCD** matrix of the lattice model can be written using Equation 3.23 as

$$\mathbf{ABCD} = \frac{1}{4 - Z_{se}Y_{sh}} \begin{pmatrix} 4 + Z_{se}Y_{sh} & 4Z_{se} \\ 4Y_{sh} & 4 + Z_{se}Y_{sh} \end{pmatrix}. \quad (3.24)$$

3.4.3.1 Propagation Constant

The propagation constant of unit cell can be calculated for the symmetrical two-port network by inserting the corresponding parameters from Equation 3.24 in Equation 3.7. Furthermore, a good approximation can be made to obtain a simple expression for the propagation constant when considering a small bandwidth around the broadside frequency, i.e., $\gamma p \ll \pi/4$ [22]. This expression reads

$$\gamma p \approx \sqrt{Z_{se}Y_{sh}}. \quad (3.25)$$

Now, by inserting Equation 3.22 into Equation 3.25, γ can be obtained in terms of the LCRG parameters of the series and shunt resonators, which yields

$$\gamma p = \sqrt{[R + j2L(\omega - \omega_{se})][G + j2C(\omega - \omega_{sh})]}. \quad (3.26)$$

The above equation allows to calculate the broadside frequency, which is defined as the frequency at which the main beam of a P-LWA points to the broadside direction. From Equation 2.23, the broadside frequency, ω_{bs} , can be derived by evaluating the zero crossing of the phase constant, i.e., $\beta|_{\omega=\omega_{bs}} = 0$. Substituting ω by ω_{bs} and setting the imaginary part of Equation 3.26 to zero as

$$j2LG(\omega_{bs} - \omega_{se}) = -j2CR(\omega_{bs} - \omega_{sh}), \quad (3.27)$$

which allows to calculate a general definition for the broadside frequency, ω_{bs} , which reads

$$\omega_{bs} = \frac{Q_{se} + Q_{sh}}{Q_{se}/\omega_{se} + Q_{sh}/\omega_{sh}}. \quad (3.28)$$

This result is interesting since it shows that the broadside frequency in P-LWAs depends not only on the resonance frequencies of the two resonators but also on

their Q -factors. Moreover, it shows that ω_{bs} is closer to the resonance frequency of that of the resonator with a higher Q -factor, i.e. lower radiation/loss. Therefore, depending on the Q -factors of the two resonators, Equation 3.28 can be rewritten as

$$\omega_{bs} \approx \begin{cases} \omega_{sh}, & \text{for } Q_{se} \ll Q_{sh}, \\ \omega_{se}, & \text{for } Q_{sh} \ll Q_{se}. \end{cases} \quad (3.29)$$

The last equation shows that a P-LWA with only one low Q -factor operates at broadside with low radiation. This fact will be more clear in Section 3.4.3.3 when the radiation efficiency is calculated.

3.4.3.2 Bloch Impedance

Due to symmetry, only one Bloch impedance Z_B is considered since the forward and backward propagating waves are equal for the transversal symmetric case, i.e. $Z_B^+ = -Z_B^- = Z_B$. As mentioned in Section 3.2, the ratio of the voltage and current, V_n/I_n , at the input terminals of the lattice circuit model in Figure 3.5, yields the Bloch impedance

$$Z_B = \frac{V_n}{I_n} = \frac{V_{sh}}{I_{se}} = \sqrt{\frac{Z_{se}}{Y_{sh}}} = \sqrt{\frac{R + j2L(\omega - \omega_{se})}{G + j2C(\omega - \omega_{sh})}}. \quad (3.30)$$

At broadside, the Bloch impedance can be obtained by inserting Equation 3.27 into Equation 3.30 to get

$$Z_{B,bs} = \sqrt{\frac{R}{G}} \sqrt{\frac{1 + j2Q_{se}(\omega_{bs} - \omega_{se})/\omega_{se}}{1 + j2Q_{sh}(\omega_{bs} - \omega_{sh})/\omega_{sh}}}. \quad (3.31)$$

Moreover, the Bloch impedance at the off-broadside direction can be obtained by evaluating the limit in Equation 3.30 of $(\omega - \omega_{se}) \rightarrow \infty$ and $(\omega - \omega_{sh}) \rightarrow \infty$, which yields

$$Z_{B,off-bs} = \lim_{(\omega - \omega_{se}, \omega - \omega_{sh}) \rightarrow (\infty, \infty)} \sqrt{\frac{R + j2L(\omega - \omega_{se})}{G + j2C(\omega - \omega_{sh})}} = \sqrt{\frac{L}{C}}. \quad (3.32)$$

Equation 3.32 is equal to the characteristic impedance of a homogenous transmission line.

3.4.3.3 Radiation Efficiency

The lattice model allows the calculation of the radiation efficiency of the unit cell, which is equal to that of a sufficiently long structure. Generally, the radiation efficiency of an antenna is defined as

$$\eta = \frac{P_{rad}}{P_{total}} = \frac{P_{rad}}{P_{rad} + P_{loss}}, \quad (3.33)$$

where P_{rad} is the radiated and P_{loss} is the loss or dissipated powers. Next, the radiation efficiency can be expressed in terms of Z_B , R and G by

$$\eta = \frac{P_{\text{se,rad}} + P_{\text{sh,rad}}}{P_{\text{se}} + P_{\text{sh}}} = \frac{\eta_{\text{se}}R + \eta_{\text{sh}}G|Z_B|^2}{R + G|Z_B|^2}, \quad (3.34)$$

where R and G represent the radiation and loss contributions of the series and shunt resonators [Section 3.4.1] and reads

$$R = R_{\text{rad}} + R_{\text{loss}}, \quad (3.35a)$$

$$G = G_{\text{rad}} + G_{\text{loss}}. \quad (3.35b)$$

The subscripts “*loss*” and “*rad*” corresponds to the dielectric/conductor loss and radiation, respectively. In the case of a bounded structure, the radiation loss vanishes. Furthermore, the series and shunt radiation efficiencies read

$$\eta_{\text{se}} = \frac{R_{\text{rad}}}{R} = \frac{R_{\text{rad}}}{R_{\text{rad}} + R_{\text{loss}}}, \quad (3.36a)$$

$$\eta_{\text{sh}} = \frac{G_{\text{rad}}}{G} = \frac{G_{\text{rad}}}{G_{\text{rad}} + G_{\text{loss}}}. \quad (3.36b)$$

To further simplify Equation 3.34, the broadside Bloch impedance ($Z_{B,\text{bs}}$) in Equation 3.31 can be expressed in terms of Q_{se} and ω_{se} only using Equation 3.28, as follows

$$Z_{B,\text{bs}} = \sqrt{\frac{R}{G}} \sqrt{\frac{1 + j2Q_{\text{se}}(\omega_{\text{bs}} - \omega_{\text{se}})/\omega_{\text{se}}}{1 - j2Q_{\text{se}}(\omega_{\text{bs}} - \omega_{\text{se}})/\omega_{\text{se}}}}. \quad (3.37)$$

Next, the radiation efficiency at broadside, η_{bs} , is obtained by evaluating the absolute value of Equation 3.37, $|Z_B(\omega_{\text{bs}})| = \sqrt{R/G}$, and inserting its result into Equation 3.34,

$$\eta_{\text{bs}} = \frac{\eta_{\text{se}} + \eta_{\text{sh}}}{2}. \quad (3.38)$$

Equation 3.38 is a general formula for the broadside radiation efficiency in P-LWAs, as it is independent of ω_{se} and ω_{sh} . Also, the equation shows a theoretical 100 % radiation efficiency cannot be achieved if only one resonator has 100 % radiation efficiency. This will be demonstrated analytically in Section 3.4.5. Moreover, Equation 3.38 implies that the series (P_{se}) and shunt (P_{sh}) powers at the broadside frequency coalesce. Finally, inserting Equation 3.32 into Equation 3.34 yields to the off-broadside radiation efficiency, $\eta_{\text{off-bs}}$,

$$\eta_{\text{off-bs}} = \frac{\eta_{\text{se}}Q_{\text{sh}} + \eta_{\text{sh}}Q_{\text{se}}}{Q_{\text{sh}} + Q_{\text{se}}}, \quad (3.39)$$

which shows that the radiation efficiency at the off-broadside frequency depends in addition to the series and shunt efficiencies also on the resonators’ Q -factors. Usually, the radiation efficiency and the Q -factor are coupled, which means a resonator with a high radiation efficiency has a low Q -factor.

3.4.4 Equalization Conditions

In this section, the two conditions to close the open-stopband in P-LWAs are provided. Since the unit cell parameters vary over frequency, as shown in the previous section and the open-stopband problem only occurs at the broadside frequency, equalizing these parameters over the two operation regimes will resolve this problem. The first condition is the *frequency-balancing* or ω -*balancing* condition, where both frequencies of the series and shunt resonator coalesce, it yields

$$\omega_{bs} = \omega_{se} = \omega_{sh}. \quad (3.40)$$

The second condition, the Q -*balancing* condition, is achieved by equalizing the off-broadside [Equation 3.32] and broadside [Equation 3.31] Bloch impedances, i.e.,

$$Z_{B, \text{off-bs}} = Z_{B, \text{bs}}. \quad (3.41)$$

The condition in Equation 3.41 will be satisfied by two conditions, the *frequency-balancing* and Q -*balancing* conditions, i.e., $\omega_{se} = \omega_{sh}$ and $Q_{se} = Q_{sh}$, respectively. Furthermore, formulating these conditions in terms of the LCRG as

$$\frac{R}{L} = \frac{G}{C}, \quad (3.42)$$

shows that the equalization conditions in P-LWAs is equivalent to the Heaviside condition for distortionless transmission lines [70, 78].

3.4.5 Numerical Results

In this section, the equalization conditions presented in Section 3.4.4 is verified using numerical examples. Specifically, plots for the analytical formulas derived in Section 3.4.3 are presented for the propagation constant, Bloch impedance and radiation efficiency. Furthermore, the four operation regimes of P-LWAs correspond to the balanced and unbalanced conditions for the Q -factors and frequencies of the series and shunt resonators. Additionally, some parameters are fixed for all operations regimes for comparability. Here, the series resistance and shunt conductance are set to $R = 10 \Omega$ and $G = 400 \mu\text{S}$, respectively. Moreover, the series and shunt radiation efficiencies are fixed to $\eta_{se} = 0.9$ and $\eta_{sh} = 0.2$, respectively.⁴ Table 3.1 lists the values of the P-LWAs parameters used in this section. The results are plotted over a normalized frequency ω/ω_0 .

3.4.5.1 ω -Balanced and Q -Unbalanced Regime

Most P-LWAs operate with this regime, i.e. $\omega_{se} = \omega_{sh}$ and $Q_{se} \neq Q_{sh}$. Figure 3.6(a) shows the Bloch impedance, propagation constant, and radiation efficiency with the Q -factors $Q_{se} = 15$ and $Q_{sh} = 400$, which are realistic values of a DS or not optimally designed LA P-LWA. The plots show an abrupt change across all P-LWA's parameters over frequency. Specifically, Z_B has a peak and the radiation efficiency

⁴The values of the parameters R , G , η_{se} and η_{sh} are realistic for a double-symmetric unit cell with only one efficient resonator, either series or shunt, as will be shown later in this chapter [see Table 3.2].

Table 3.1: Summary of the used numerical values of P-LWAs parameters in Section 3.4.3.

Condition		series resonator				shunt resonator			
Q -factor	frequency	Q_{se}	$\frac{\omega_{se}}{\omega_0}$	η_{se}	R/Ω	Q_{sh}	$\frac{\omega_{sh}}{\omega_0}$	η_{sh}	$G/\mu S$
Q -unbalanced	ω -balanced	15	1.00	0.9	10	400	1.00	0.2	400
	ω -unbalanced	15	1.12	0.9	10	400	0.90	0.2	400
Q -balanced	ω -unbalanced	15	1.12	0.9	10	15	0.90	0.2	400
	ω -balanced	15	1.00	0.9	10	15	1.00	0.2	400

minimum at the broadside frequency, $\eta_{bs} = 0.55$ [Equation 3.38]. Additionally, the peak of the Bloch impedance indicates a high series and low shunt radiation, since in a real P-LWA the Q -factors and R/G are dependent on each other. Moreover, if $\eta_{se} = 0$ or $\eta_{sh} = 0$ the radiation efficiency of the P-LWA cannot exceed 50 % at broadside, as explained by Equation 3.38, whereas the η_{off-bs} is almost equal to η_{sh} , since it depends also on the Q -factors of the resonators [Equation 3.39].

3.4.5.2 ω -Unbalanced and Q -Unbalanced Regime

This regime may be the start of any P-LWA design without any specific optimization. Thus, it may be applied to DS or LA P-LWAs. Figure 3.6(b) plots the P-LWA parameters for the resonance frequencies of the series and shunt resonators are set to $\omega_{se}/\omega_0 = 1.125$ and $\omega_{sh}/\omega_0 = 0.9$, respectively, which exhibit a large variation over frequency but not symmetric around the broadside frequency. The open-stopband problem is more clear for this regime, as can be observed from the high attenuation constant α at the center frequency. Additionally, it can be seen that η_{bs} is the same for the ω -balancing case in Section 3.4.5.1, which confirms Equation 3.38. Finally, the broadside frequency ω_{bs} is closer to ω_{sh} due to the lower losses in the shunt resonator, i.e., $Q_{sh} > Q_{se}$ [Equation 3.29].

3.4.5.3 ω -Unbalanced and Q -Balanced Regime

This regime can usually be achieved with an LA unit cell [23, 24] but in some cases, it can be achieved by adding loss to the resonator with lower Q -factor, which is not practical due to the overall low radiation efficiency.⁵ Figure 3.7(a) plots the P-LWAs parameters with $Q_{se} = Q_{sh} = 15$ and the same frequencies $\omega_{se}/\omega_0 = 1.125$ and $\omega_{sh}/\omega_0 = 0.9$ as in Section 3.4.5.2. The radiation efficiency at broadside in this case $\eta_{bs} = 0.55$, which confirms that Equation 3.38 is only dependent on η_{se} and η_{sh} . Moreover, it can be observed that the notch in the radiation efficiency is at ω_{sh}/ω_0 while the peak is at ω_{se}/ω_0 [Equation 3.29], where $\eta_{se} > \eta_{sh}$. It is worth noting that for both regimes, the Q -balanced (Section 3.4.5.1) and Q -unbalanced (this section), η_{bs} is the same, since the definition of the broadside frequency was deduced from

⁵The Q -balancing can be also achieved by adding loss, i.e., increasing R or G , to the resonator with high Q -factor to satisfy the equalization condition in Equation 3.42. This has been demonstrated in [79] loading the unit cell of an SFP P-LWA with concentric resistors but leads to an overall poor radiation efficiency.

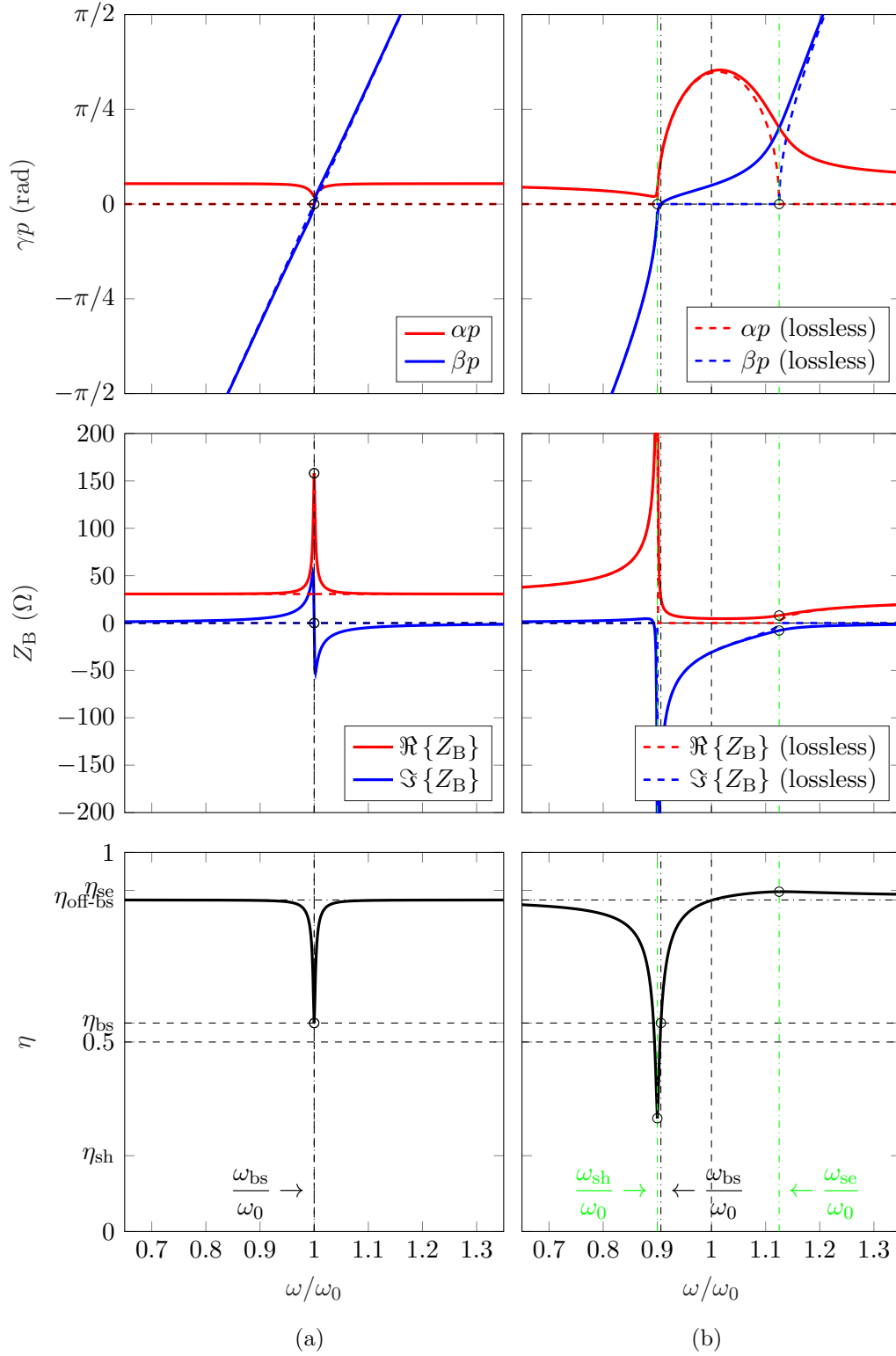


Figure 3.6: Comparison of the propagation constant, Bloch impedance and radiation efficiency for Q -unbalanced and (a) ω -balanced, and (b) ω -unbalanced cases in P-LWA. The lossless Bloch impedance has a singularity at the broadside frequency in (a) according to Equation 3.31.

$\beta = 0$ and is dependent on Q_{se} , ω_{se} , Q_{sh} , and ω_{sh} [Equation 3.28], which are the same for both regimes, as can be found in Table 3.1.

3.4.5.4 ω -Balanced and Q -Balanced Regime

The last regime, the ω -balanced and Q -balanced regime, is shown in Figure 3.7(b). This regime represents an optimally designed P-LWA with its physical parameters equalized over frequency. Furthermore, it can be seen that the broadside and off-broadside radiation efficiencies are equal, $\eta_{bs} = \eta_{off-bs}$, according to Equation 3.38 and Equation 3.39. Although the radiation efficiency (55 %) for this case is rather low, in a physical structure the Q -factor and radiation efficiency of a resonator are not independent of each other, which gives the network model an advantage that cannot be explored with a real structure. Additionally, this regime can be achieved only if the proper amount of LA is applied.

Finally, a P-LWA with optimal LA usually radiates with circular polarization, as will be shown later in Chapter 4, which is not suitable for all applications. Therefore, the next section will discuss the transversally-asymmetric (TA) unit cell, which provides optimal operation conditions with linear polarization.

3.5 Transversally-Asymmetric Unit Cell

Recently, attempts have been made to resolve the broadside problem without affecting the polarization of the unit cell [21, 80]. The approach was based on satisfying a matching condition for the Bloch impedance, i.e., impedance matching for the unit cell. Furthermore, the design was based on a $\lambda/4$ -transformer matching technique, which arises the question on how such matching conditions can be applied to other structures where such a technique cannot be applied to. Hence, a more general approach was introduced in [25], which was based on exploiting the transversal asymmetry (TA) [Figure 3.3(c)] of the unit cell of P-LWAs to achieve equalization of the antenna parameters over frequency. In fact, both matching and asymmetry techniques deliver the same result but the latter is considered as a general approach. Here, it will be shown that TA P-LWA solves the broadside issue without compromising polarization, hence featuring high overall efficiency and linear polarization, by exciting the series and shunt resonators in optimal power ratios, which is allowed by the fact that the TA structure breaks the equal power splitting condition, i.e. $P_{se} \neq P_{sh}$. Furthermore, the formulas for the TA will be first derived based on the circuit model for the TA in P-LWAs and the will be demonstrated by two structures, a non-bounded and bounded, supported by full-wave simulations.

3.5.1 General Network Model for TA Unit Cell

In Section 3.4.3, it was mentioned that the poor radiation efficiency at the broadside frequency is due to the equal power splitting of the series and shunt powers. In order to break the power splitting, mode coupling between the two resonators is needed. Therefore, the general circuit model for TA employs four ideal transformers with equal transformation ratio T to model the coupling. Moreover, the circuit model is based on the lattice circuit model, shown in Figure 3.5 and for case $T = 1$ the circuit

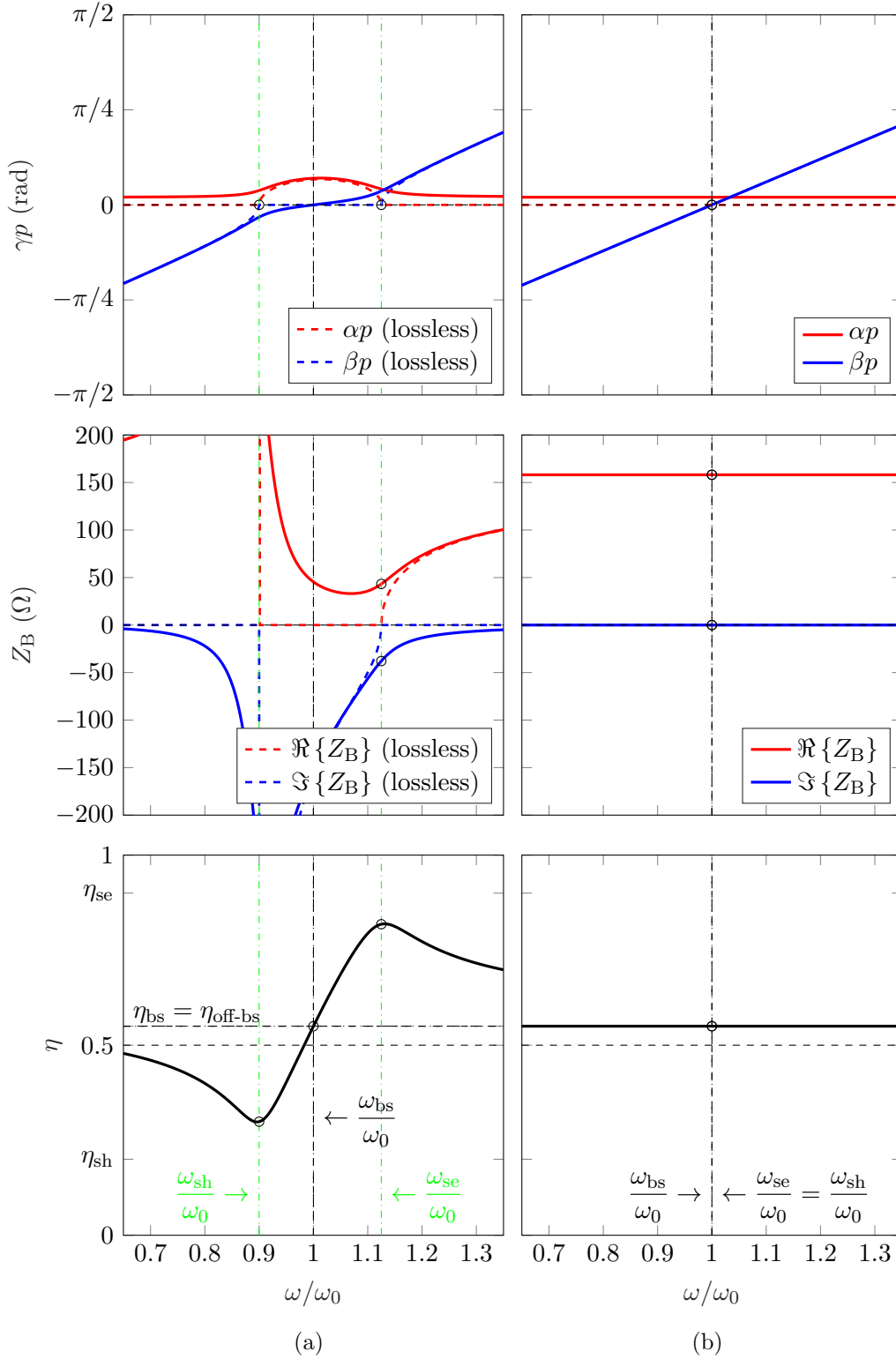


Figure 3.7: Comparison of the propagation constant, Bloch impedance and radiation efficiency for Q -balanced and (a) ω -unbalanced, and (b) ω -balanced cases in P-LWA.

becomes symmetrical. Additionally, under odd and even excitations the series and shunt resonators can be isolated, respectively.

The analytical formulas presented in this section are based on the results in [25]. In this section, only the final results of the unit cell's parameters will be presented. Nevertheless, for the sake of self-consistency, their derivation will be provided in Appendix A.2.

Figure 3.8 shows the circuit model for a TA unit cell with the four ideal transformers. The series and shunt resonators are each connected to two transformers and their resonance frequencies are not affected by the transformation. First, the uncoupled series and shunt resonators read

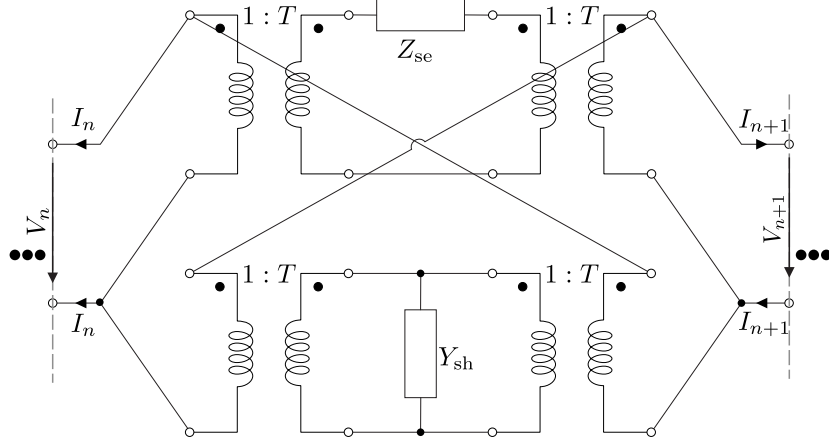


Figure 3.8: Transformer network model for transverse asymmetric unit cell.

$$Z_{se} = \frac{Y_{11} + Y_{22} + 2Y_{21}}{(Y_{11} + Y_{21})(Y_{22} + Y_{21})}, \quad (3.43a)$$

$$Y_{sh} = \frac{Z_{11} + Z_{22} + 2Z_{21}}{(Z_{11} + Z_{21})(Z_{22} + Z_{21})}, \quad (3.43b)$$

which show that for the symmetric case ($T = 1$), they reduce to Equation 3.23. Next, the **ABCD** matrix for the transformer model reads

$$\mathbf{ABCD} = \frac{1}{\left(T + \frac{1}{T}\right)^2 - Z_{se}Y_{sh}} \begin{bmatrix} \frac{1}{T^2} \left(T + \frac{1}{T}\right)^2 + Z_{se}Y_{sh} & \left(T + \frac{1}{T}\right)^2 Z_{se} \\ \left(T + \frac{1}{T}\right)^2 Y_{sh} & T^2 \left(T + \frac{1}{T}\right)^2 + Z_{se}Y_{sh} \end{bmatrix}, \quad (3.44)$$

which again for $T = 1$ reduces to the **ABCD** matrix of the lattice network model [Equation 3.24]. The transformation ratio, T , can be calculated by

$$T = \sqrt{\frac{A+1}{D+1}}. \quad (3.45)$$

Next, the propagation constant, Bloch impedance, and radiation efficiency for the transformer network model will be discussed.

3.5.1.1 Propagation Constant

The propagation constant, γp , for the transformer circuit model can be calculated by inserting the parameters from Equation 3.44 into Equation 3.6. Using the near-broadside operation assumption, i.e., $Z_{se}Y_{sh} \ll \pi/4$, γ can be simplified to

$$\gamma p \approx \sqrt{\frac{1}{4} \left(T - \frac{1}{T}\right)^2 + Z_{se}Y_{sh}} = \sqrt{\tau^2 + Z_{se}Y_{sh}}, \quad (3.46)$$

where τ , for better readability, is defined as

$$\tau = \frac{1}{4} \left(T - \frac{1}{T}\right)^2, \quad (3.47)$$

where for the symmetric case $\tau = 0$ or $T = 1$. Equation 3.46 shows the effect of coupling, where for the broadside case, γ does not depend only on $\sqrt{RG}$ but also on the transformation ratio, T .

3.5.1.2 Bloch Impedance

The Bloch impedance for the transverse-asymmetric case depends on the excitation direction. Using Equation 3.12, the Bloch impedance $Z_B^\pm$ for forward (+) and backward (−) propagating waves can be written as

$$Z_B^\pm = \mp \frac{\tau}{Y_{sh}} + \sqrt{\left(\frac{\tau}{Y_{sh}}\right)^2 + \frac{Z_{se}}{Y_{sh}}}. \quad (3.48)$$

By inserting Equation 3.22 into Equation 3.48, and evaluating the asymptotic result in the off-broadside regime, the off-broadside Bloch impedance ($Z_{B,off-bs}^\pm$) which corresponds to $\omega - \omega_{se} \rightarrow \infty$ and $\omega - \omega_{sh} \rightarrow \infty$, as

$$\begin{aligned} Z_{B,off-bs}^\pm &= \lim_{(\omega - \omega_{se}, \omega - \omega_{sh}) \rightarrow (\infty, \infty)} \mp \frac{\tau}{G + j2C(\omega - \omega_{sh})} \\ &\quad + \sqrt{\left(\frac{\tau}{G + j2C(\omega - \omega_{sh})}\right)^2 + \left(\frac{R + j2L(\omega - \omega_{se})}{G + j2C(\omega - \omega_{sh})}\right)} \\ &= \sqrt{\frac{L}{C}}. \end{aligned} \quad (3.49)$$

The last expression shows that $Z_{B,off-bs}$ does *not* depend on τ , and hence on T . Moreover, it shows that the forward and backward off-broadside Bloch impedances are equal whether the unit cell is symmetric or asymmetric with respect to the transversal axis [cf. Equation 3.32 and Equation 3.49]. Finally, the Bloch impedance at broadside may be derived, which results in

$$Z_{B,bs}^\pm(T) = \mp \frac{\tau}{G} + \sqrt{\left(\frac{\tau}{G}\right)^2 + \frac{R}{G}}, \quad (3.50)$$

Both Equation 3.49 and Equation 3.50 prove that T controls the P-LWA properties only at broadside, which allows for the equalization between the broadside and off-broadside regimes, as it will be shown later in Section 3.5.2.

3.5.1.3 Radiation Efficiency

Finally, the radiation efficiency for the TA can be derived, similar to the symmetric case in Equation 3.34, which reads

$$\eta^{\pm} = \frac{\eta_{\text{se}}R + \eta_{\text{sh}}G |Z_{\text{B}}^{\pm}|^2}{R + G |Z_{\text{B}}^{\pm}|^2}. \quad (3.51)$$

For the TA case, the radiation efficiency depends on the propagation direction, since it depends on the Bloch impedance. Hence, the optimal equalization condition in this case can *only* be achieved for one port.

3.5.2 Equalization Conditions

The equalization condition for TA in P-LWAs will be derived for a forward propagating wave. Furthermore, the same condition for the transversally-symmetric unit cell can be applied, see Equation 3.41, where the symmetric Bloch impedance is replaced by Z_{B}^{+} .⁶ Hence, the equalization condition can be satisfied by equalizing Equation 3.49 and Equation 3.50 then the optimal coupling factor τ_{opt} can be calculated as

$$\tau_{\text{opt}} = \frac{G}{2} \sqrt{\frac{C}{L}} \left(\frac{R}{G} - \frac{L}{C} \right), \quad (3.52)$$

by replacing the LCRG parameters by the corresponding Bloch impedance for the symmetric case results in

$$\tau_{\text{opt}} = \frac{G}{2Z_{\text{B,off-bs}}^{\text{sym}}} \left(Z_{\text{B,bs}}^{\text{sym}2} - Z_{\text{B,off-bs}}^{\text{sym}2} \right). \quad (3.53)$$

This condition implies that the optimal coupling factor can be directly calculated from a *transversally-symmetric* unit cell. In practice, equalizing the unit cell parameters is an iterative process as introducing the asymmetry, i.e. changing the geometry, might change the intrinsic parameters of the series and shunt resonators. Nevertheless, the uncoupled LCRG parameters can be extracted even from a TA unit cell by applying the odd and even excitations, as indicated by Equation 3.43. Finally, the optimal transformation ratio, T , can be calculated using Equation 3.47 as follows

$$T_{\text{opt}} = \sqrt{\tau_{\text{opt}} + \sqrt{\tau_{\text{opt}}^2 + 1}}. \quad (3.54)$$

It is worth noting that τ_{opt} and therefore, T_{opt} can be calculated independently from the ω -balancing condition, which allows for fast convergence of the optimization process. In the forthcoming section, two case studies of periodic structures will be demonstrated and equalized through the broadside frequency using the TA.

⁶The same procedure can be done for a backward propagating wave with Z_{B}^{-} , which would result in $T_{\text{opt}}^{-} = 1/T_{\text{opt}}^{+}$.

3.5.3 Radiative and Bounded Case Studies

In this section, the concept of asymmetry in P-LWAs, and periodic structures in general, will be demonstrated. Up to now, the broadside radiation problem has been resolved for microstrip line-based structures, i.e., *quasi-TEM*, based P-LWAs [21, 67, 80, 81] [25, 82]. Here, the concept of TA will be extended to a *non-TEM* transmission line. Specifically, two substrate-integrated waveguide (SIW) structures operating with the fundamental TE_{10} mode will be presented. First, an SIW P-LWA with longitudinal slots will be investigated. Second, the broadside problem in a bounded SIW structure, i.e., *series-feeding network* (SFN), will be treated, and a solution using the TA will be provided.

3.5.3.1 Slotted Substrate-Integrated Waveguide (SSIW)

The equalization conditions provided in Sections 3.4.3 and 3.5.1 are limited for unit cells that can be modeled by two resonators, the series and shunt under the odd and even excitations, respectively. In slotted waveguides [Figure 2.3(a)] and in their planar counterpart, slotted SIWs [Figure 2.3(e)],⁷ the slots provide an additional resonance in the unit cell. This additional resonance causes an abrupt impedance and gain change over frequency, since the equivalent radiation resistance of the slot changes around the resonance frequency rapidly with frequency. Therefore, in many practical implementations of such antennas, the slot may be operated below resonance [30]. Nevertheless, operating a below-resonance slot leads to a stopband near the broadside frequency due to the reactive part of the slot impedance, which means that the unit cell will be ω -unbalanced. Recently, a method to satisfy the ω -balancing condition has been proposed by adding inductive posts within the unit cell's structure [83]. Although it is claimed that this method resolves the open-stopband problem, it is, in fact, only satisfying the ω -balancing condition and the open-stopband at broadside remains.

The proposed structure in this work leverages the idea of canceling the capacitive effect of the slot with inductive posts by introducing TA to the unit cell to resolve the open-stopband problem completely. Figure 3.9 shows the proposed structure which consists of alternating longitudinal, offset from the longitudinal axis slots, and inductive posts realized by vias placed oppositely to each slot. The TA is achieved by shifting the inductive posts along the y -axis by d . To efficiently analyze the structure, the unit cell with a single slot or a single post will be separately presented first and later merged to achieve the optimized unit cell.

Figure 3.10(a) shows the unit cell of a single slot on the broad wall of an SIW with an offset o_s . Generally, longitudinal slots can be modeled by series resonators in the shunt path [84, 85], as shown in Figure 3.10(b), which inherently works as a bandstop structure at its resonance frequency. Increasing the slot offset, o_s , or width, w_s , increase the radiation and therefore, decrease the slot resistance, R_s . By tuning the slot length ℓ_s , the resonance frequency can be tuned. Hence, two measures may be taken to reduce the effect of the stopband. First, as mentioned before, the slot length can be shortened to operate below resonance and second, by increasing

⁷This type of antenna array is very attractive to radar applications due to the polarization purity of the slot and the shielded nature of WGs.

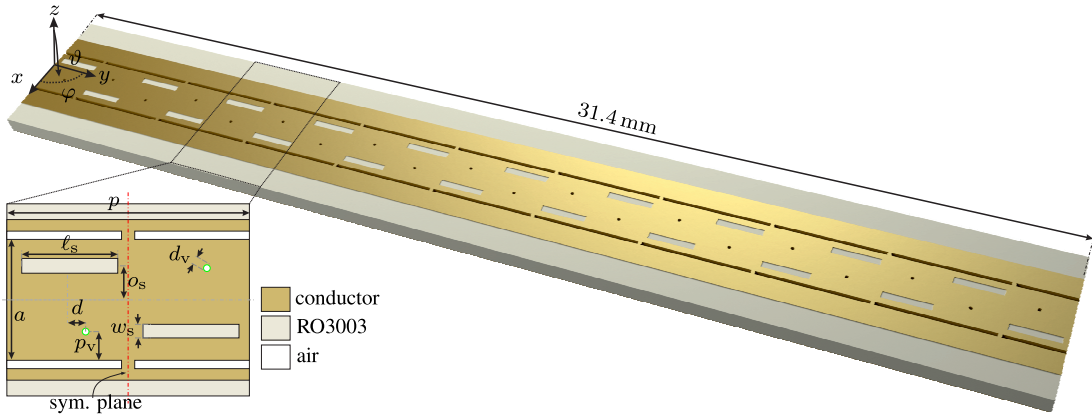


Figure 3.9: SIW P-LWA structure with longitudinal slots consisting of 10 unit cells. The inset shows a single unit cell with its geometric parameters.

the slot resistance to have high value (open circuit). Both measures obviously have issues, i.e., the reactive part of the impedance and low slot efficiency, respectively. For the following analysis, the substrate used to realize the SIW structure is Rogers RO3003 ($\epsilon_r = 2.97$ and $\tan \delta = 0.001$) with thickness $h = 508 \mu\text{m}$ [86].

The ABCD matrix of the slot circuit model can be written as a cascade of three sections, WG-Slot-WG, as follows

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}^{\text{unit cell}} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \times \begin{bmatrix} A & B \\ C & D \end{bmatrix}^{\text{slot}} \times \begin{bmatrix} A & B \\ C & D \end{bmatrix}. \quad (3.55)$$

The slot parameters can be accurately extracted by two sets of full-wave simulations. First, an empty SIW structure is simulated. Second, the slot unit cell is simulated with the slot offset is varying while keeping all other parameters fixed. Finally, the slot matrix is calculated by using matrix inversion of Equation 3.55 of the unit cell in Figure 3.10(a), as follows [70]

$$\text{ABCD}^{\text{slot}} = \sqrt{\text{ABCD}^{-1}} \times \text{ABCD}^{\text{unit cell}} \times \sqrt{\text{ABCD}^{-1}} = \begin{bmatrix} 1 & 0 \\ 1/Z_s & 1 \end{bmatrix}, \quad (3.56)$$

where $Z_s = R_s + jX_s$ is the slot impedance and $X_s = [\omega L_s - 1/(\omega C_s)]$ is the reactance. The proposed structure is intended to operate at a broadside frequency of 78 GHz with the cutoff frequency of the SIW being $f_c = 56.42 \text{ GHz}$. The extracted slot's parameters are plotted in Figure 3.10(c) at the broadside frequency using the time-domain solver of the electromagnetic simulation tool CST MWS [87].⁸ The results show, as expected, that the slot resistance can be controlled by the slot offset o_s . Furthermore, the resonance frequency of the slot, f_s , also changes when varying o_s . To reduce the stopband problem, f_s is chosen at much higher frequency. Specifically, $f_s = 95.5 \text{ GHz}$, which corresponds to a slot offset $o_s = 0.5 \text{ mm}$, $R_s = 38.8 \Omega$, and $X_s = -394 \Omega$.

⁸The time-domain solver in CST uses the finite-integral technique (FIT).

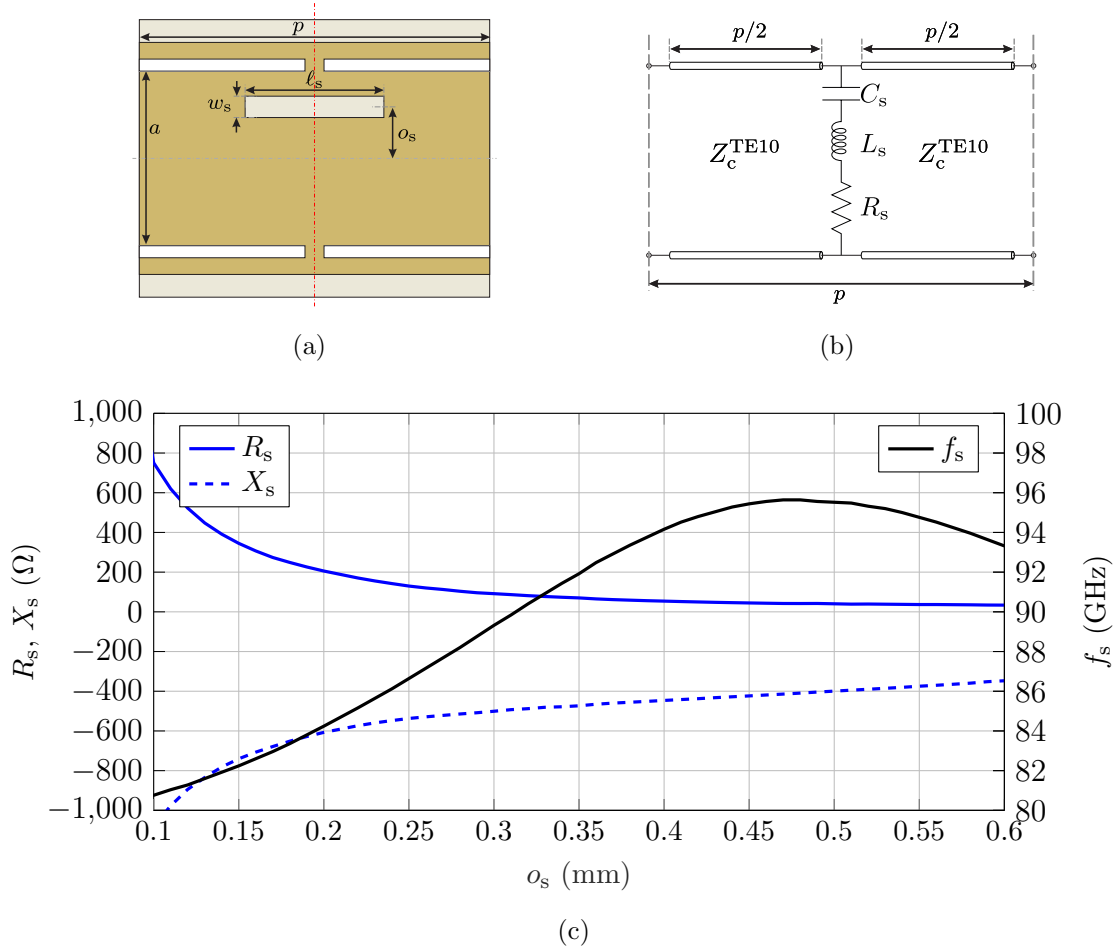


Figure 3.10: Single SIW unit cell for slot parameters extraction for different slot off-sets from the longitudinal axis. (a) Top view of the unit cell geometry, (b) equivalent circuit model, and (c) extracted slot resistance, reactance and resonance frequency.

Now, the capacitive reactance of the slot's impedance at the broadside frequency, $X_s(f_{\text{bs}})$ may be compensated by an equivalent inductance connected either in series or parallel to the slot resonator. From the book of Marcuvitz [47], a parallel inductance connected to the waveguide terminals can be realized by an asymmetrical inductive post with its edges being parallel to the $\mathbf{E}$ -field of the TE₁₀ mode. The proposed SSIW structure here utilizes a via for the inductive post with a via diameter $d_v = 0.1$ mm since it is easy to manufacture, as depicted by the green-highlighted via in Figure 3.11(a). For simplicity, the via is considered as a zero thickness obstacle, since $d_v \ll \lambda_g^{\text{TE10}}$, where λ_g^{TE10} is the guided wavelength of the TE₁₀ mode. Hence, the via post may be modeled solely as an inductor, as shown in Figure 3.11(b). Moreover, the inductive reactance may be calculated analytically using a closed-form expression [47] but for the sake of validation, the reactance value of the inductive post is also extracted from full-wave simulation, similar to the slot, by evaluating the ABCD-matrix of the unit cell in Figure 3.11(a), as follows

$$\text{ABCD}^{\text{post}} = \sqrt{\text{ABCD}^{-1}} \times \text{ABCD}^{\text{unit cell}} \times \sqrt{\text{ABCD}^{-1}} = \begin{bmatrix} 1 & 0 \\ 1/Z_p & 1 \end{bmatrix}, \quad (3.57)$$

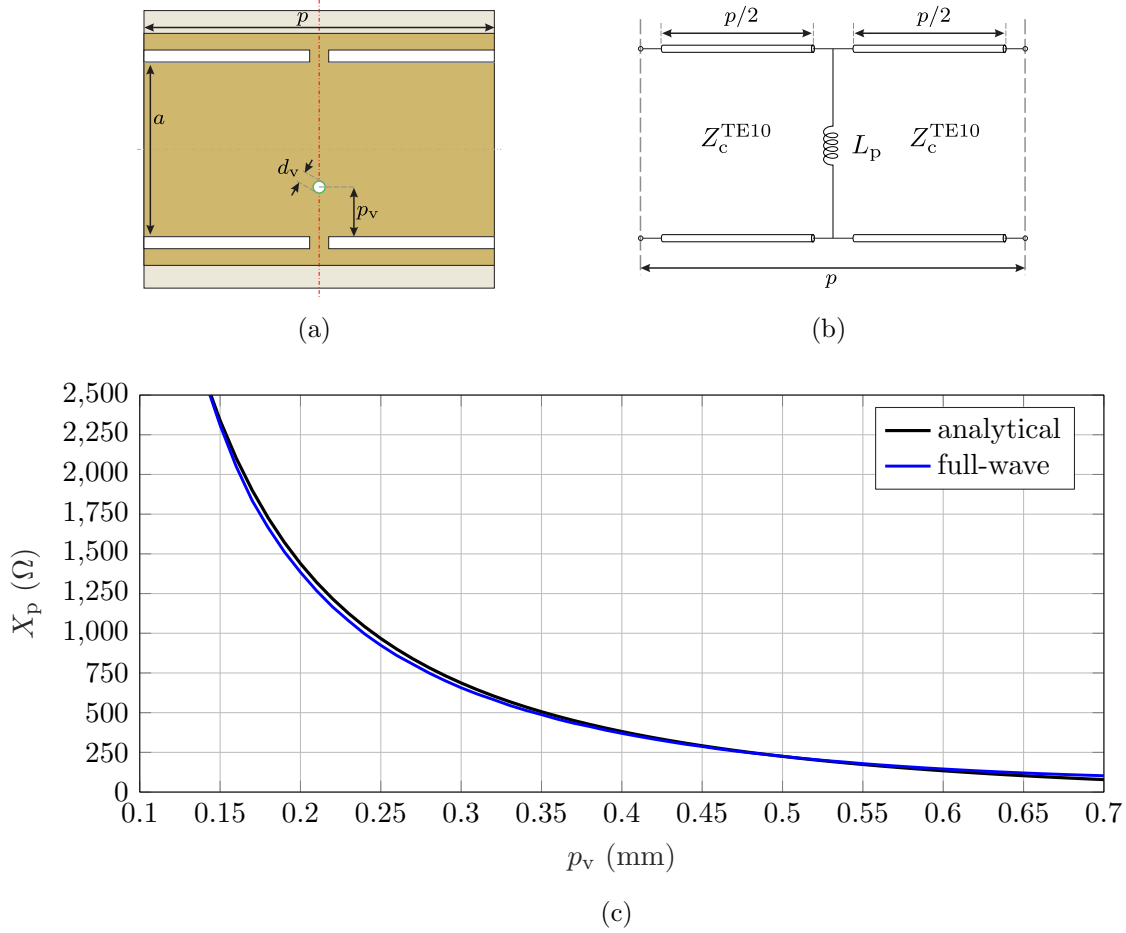


Figure 3.11: Extracted post reactance for different via offsets from the wall of the SIW. (a) Top view of the unit cell geometry, (b) equivalent circuit model of the unit cell, and (c) extracted and analytical via post reactance at the broadside frequency versus p_v .

where $Z_p = jX_p$ and $X_p = \omega L_p$ are the post impedance and reactance, respectively. Figure 3.11(c) compares the analytical and extracted inductive post reactance using Equation 3.57.⁹ The results show a good agreement between the analytical and simulated values for X_p for different p_v values, where p_v is the position of the via from the bottom wall of the SIW. Moreover, it can be clearly seen that the capacitance of the slot can be compensated by the inductive reactance of the via post only by varying p_v .

The structure in Figure 3.10 with two slots in each unit cell with their complementary via post shall now be optimized for the optimal radiation condition at

⁹The analytical expression for the inductive post reactance is lengthy and will not be repeated in this work for the sake of compactness. Nevertheless, it can be found in [47, p. 224].

Table 3.2: Extracted LCRG parameters of the series and shunt resonators for the SSIW P-LWA shown in Figure 3.9.

resonator	parameter				
series	f_{se} (GHz)	R (Ω)	L (nH)	Q_{se}	η_{se}
	78.00	161.64	6.92	20.94	0.85
shunt	f_{sh} (GHz)	G (μS)	C (fF)	Q_{sh}	η_{sh}
	78.00	57.32	42.28	361.61	0.06

Table 3.3: Dimensions in mm of the optimized SSIW P-LWA shown in Figure 3.9.

a	p	l_s	w_s	p_v	d_v	d
1.55	3.14	1.50	0.20	0.23	0.10	0.05

broadside. Additionally, the reason for considering two slots is strictly to reduce the distance between the elements and hence, avoiding grating lobes [65]. Moreover, by alternating the slots along the y -axis is to cut the electric currents on the broad wall constructively, since the distance between two adjacent slots is $\lambda_g/2$. The optimization consists of two steps. First, an initial unit cell is simulated with period $p = \lambda_g$ and the LCRG parameters are extracted using Equations 3.43, 3.14 and 3.19. Second, the unit cell is optimized for the equalization conditions in Section 3.5.2, where the ω -balancing condition can be achieved by tuning the period p and the via position p_v to achieve $f_{bs} = f_{se} = f_{sh} = 78$ GHz. Using the LCRG parameters listed in Table 3.2, the optimal asymmetry can be calculated using Equation 3.54 and achieved by shifting the via along the y -axis. Table 3.3 summarizes the geometrical parameters of the SSIW structure. The optimal transformation ratio was found to be $T_{opt} = 1.098$ for forward wave excitation (+), which corresponds to a via shift $d \approx 0.05$ mm.

Figure 3.12 shows the results of the unit cell full-wave simulation. Figure 3.12(a) shows the Bloch impedance for both the ω -balanced symmetric and optimum TA unit cells. It can be seen that the symmetric Bloch impedance, Z_B^{sym} , abruptly peaks at the center frequency, while the optimized Bloch impedance, Z_B^+ , exhibits the same (dispersive) behavior as the characteristic impedance of a homogeneous waveguide [70], which reads

$$Z_c^{TE10} = \frac{Z_0}{\sqrt{\epsilon_r}} \frac{1}{\sqrt{1 - \left(\frac{f_c}{f}\right)^2}}, \quad (3.58)$$

where $Z_0 = 120\pi \Omega$ is the free-space impedance. Moreover, it can be seen that the TA only affects the Bloch impedance at the broadside frequency. Furthermore, the propagation constant is plotted in Figure 3.12(b), which again shows the equalized transition around the broadside frequency for the TA case. These results confirm that the claimed mitigation technique for SSIW structures reported in [83] cannot be fulfilled unless the TA is utilized. In fact, the reported results in [83] were chosen for a set of parameters so that the losses (dielectric and conductor) in the

structure are equal to the radiation of the slots. In this case, the Q -balancing for a *transversally-symmetric* unit cell [Section 3.4.4] is achieved by either increasing the loss or decreasing the radiation of the slot, as previously investigated in [79].

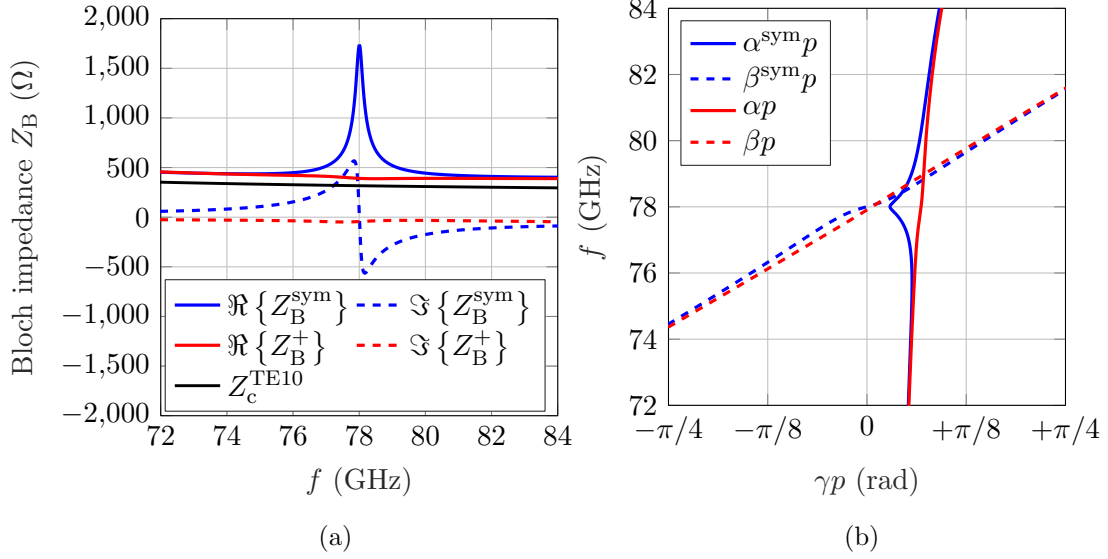


Figure 3.12: Unit-cell simulation of the SSIW structure in Figure 3.9. (a) Bloch impedance and (b) propagation constant for the symmetric and optimal asymmetric cases.

Figure 3.13 plots the full-wave simulated S-parameters normalized to the off-broadside Bloch impedance $Z_{B,\text{off-bs}} = \sqrt{L/C} \approx 405 \Omega$ for 10 cascaded unit cells, which shows the high reflection for the symmetric case and a broadband matching for the TA case, i.e. $|S_{11}| < -10$ dB.

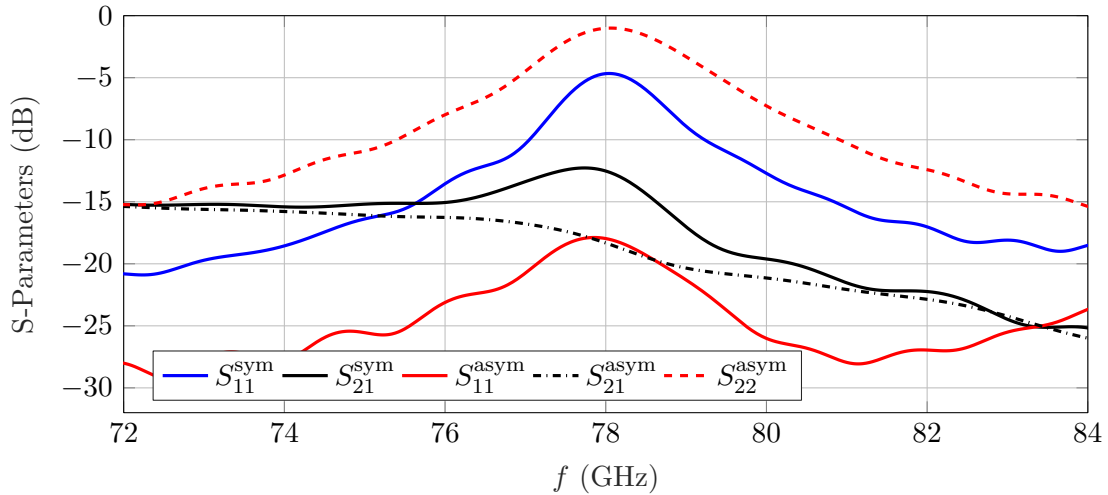


Figure 3.13: Full-wave simulated S-parameters of the SSIW P-LWA with 10 unit cells. S_{22}^{sym} and S_{12}^{asym} are not shown due to symmetry and reciprocity, respectively.

Finally, the radiation performance of the 10 unit cells SSIW P-LWA is shown in Figure 3.14. The realized gain, which includes the reflection coefficient at the input

port, is plotted for the principal yz -plane (scanning plane), i.e. $\varphi = 90^\circ$. A constant gain over frequency can be observed for the TA case (colored lines), whereas the symmetric case (black dashed lines) shows a gain degradation at the broadside frequency 78 GHz. The constant gain can also be seen when plotting the maximum gain, $G(\vartheta_{\text{mb}})$ over frequency, as shown in Figure 3.14(b). A gain improvement of about 2.6 dB is observed when comparing the optimal TA (G^+) and the symmetric case G^{sym} . Moreover, the gain degrades in the symmetric case, not only due to high reflection at the input port but also due to reflections between the unit cell, which can be verified by calculating the radiation efficiency which excludes the reflected power at the input port and the residual power at the output port, i.e., $\eta_{\text{rad}} = \eta_{\text{tot}} / (1 - |S_{11}|^2 - |S_{21}|^2)$. Figure 3.14(c) plots the radiation efficiency for the forward excitation η_{rad}^+ , which shows an almost flat radiation. Moreover, the symmetric radiation efficiency, $\eta_{\text{rad}}^{\text{sym}}$ verifies that the radiation will be degraded regardless of the matching at the input port. Finally, exciting the TA P-LWA in the backward direction results in an even more degraded radiation performance at the broadside frequency since the power of the propagating wave will couple to the shunt resonator, which has a low efficiency [Table 3.2].

A comparison between the full-wave simulated and measured results is provided in Appendix C.1.

3.5.3.2 Series-Feeding Network (SFN)

The contribution of this section was originated from [88]. In order to construct 1D or 2D antenna arrays, feeding networks are essential. While a *parallel-feeding network* (PFN), also called a *corporate-feeding network*, provides constant power division at its outputs, it suffers from narrow frequency bandwidth and high losses if a large number (> 8) of outputs is required [65]. Furthermore, PFNs output ports are restricted to an even number. In contrast, *series-feeding networks* (SFNs) feature broadband frequency bandwidth, an arbitrary number of output ports, and compact size, hence, low loss [89–91]. However, SFNs suffer from the open-stopband problem at the broadside frequency, where all outputs have the same phase. The broadside problem does constitute not only P-LWAs but also all periodic structures if terminated with a matched load. Here, it will be shown that the TA can solve the broadside problem in SFNs, which will be demonstrated by an SIW SFN operating at a broadside frequency $f_{\text{bs}} = 78$ GHz.

Figure 3.15 shows the layout of the proposed SIW SFN, whose dominant mode, as for the SSIW [Section 3.5.3.1], is the TE_{10} mode. The structure uses the same single-layer RO3003 used for the SSIW P-LWA in Section 3.5.3.1. The unit cell is identical to an asymmetrical H-plane WG T-junction [70], whereas the periodic terminals (ports) of the SFN unit cell are located on the collinear axis (y -axis) and the output ports on the T-junction's arms (z -axis). The width of the waveguide was chosen to set the cutoff frequency to $f_c = 56.42$ GHz, and the SFN power split amplitude is controlled by the inductive via posts of length l_p that play the role of an electromagnetic valve. Additionally, the output port of the SFN unit cell is terminated by a perfectly matched layer (PML) boundary condition, which mimics a matched load, as depicted in Figure 3.15.

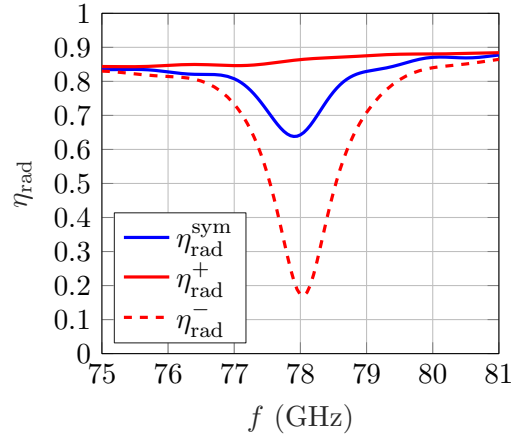
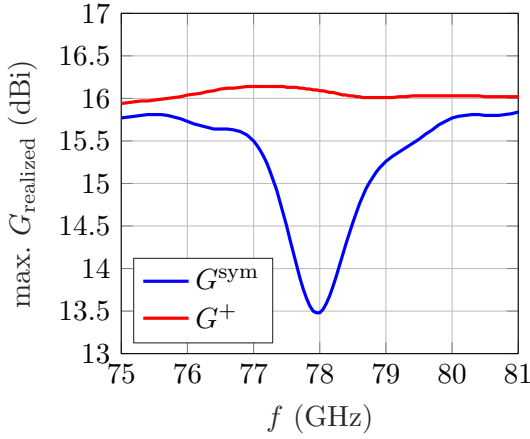
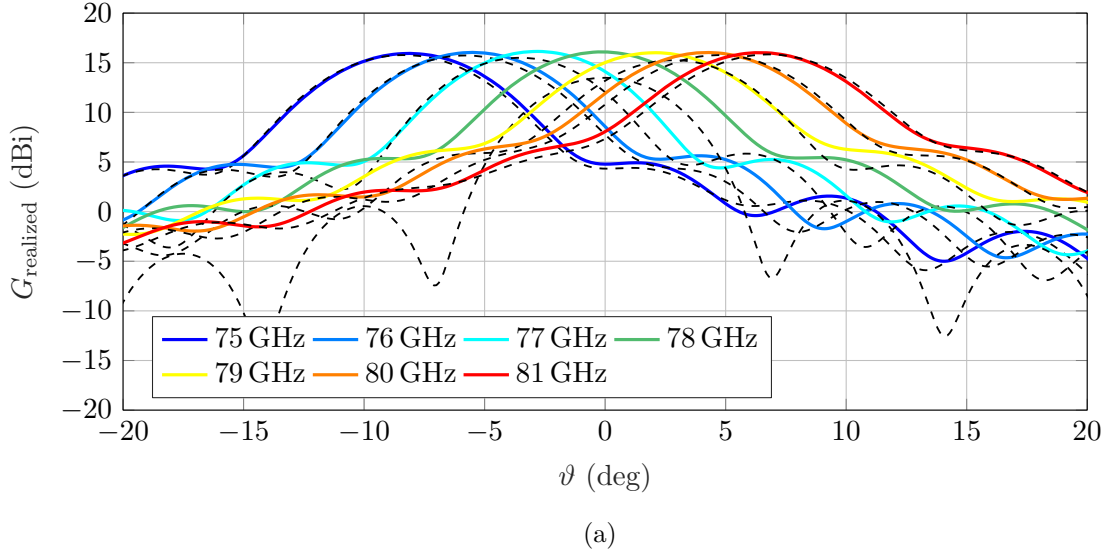


Figure 3.14: Farfield radiation results for the SSIW P-LWA in Figure 3.9 obtained by full-wave simulation. (a) Realized gain in the yz -plane showing the constant scanning over the broadside frequency for the TA case and for comparison the gain of the symmetric structure is depicted by the black dashed lines. (b) Maximum realized gain versus frequency and (c) radiation efficiency for both symmetric and asymmetric cases.

The ω -balancing can be achieved similarly to the SSIW P-LWA by tuning the unit cell period, p , and the offset of the green-highlighted inductive fin to the waveguide wall p_i . Finally, the position of the via d controls the degree of asymmetry in the unit cell. As a result, the optimum dimensions for the unit cell are achieved for the ω -balancing, i.e., $f_{se} = f_{sh} = f_{bs} = 78$ GHz, and for the optimal asymmetry, i.e., transformation ratio $T = T_{opt} = 0.97$ for a wave propagating in the forward direction (port 1). The corresponding dimensions of the optimal unit cell can be found in Table 3.4.

Figure 3.16(a) shows the Bloch impedances of the unit cell for the symmetric and optimal TA cases. It can be seen that the symmetric Bloch impedance, Z_B^{sym} , has

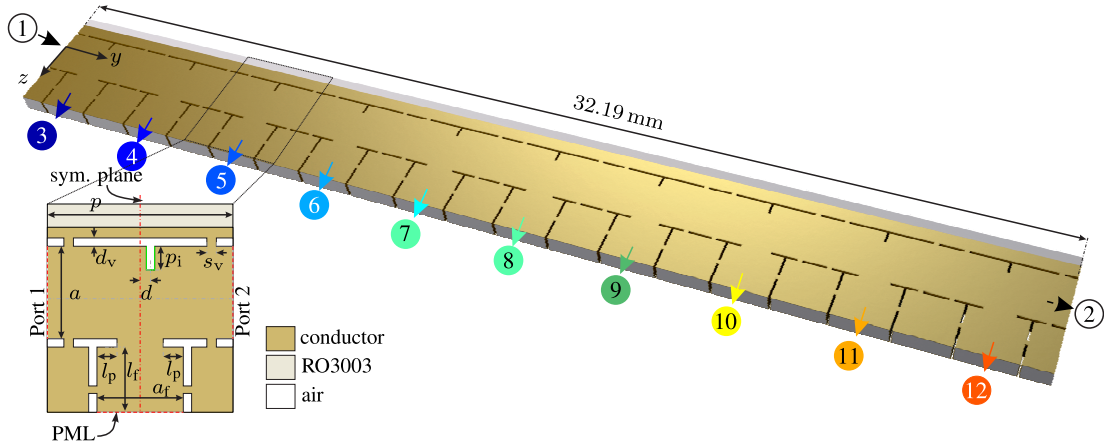


Figure 3.15: Perspective view of the SIW SFN structure with 10 unit cells with the inset showing a detailed view of the unit cell. The asymmetry is achieved by tuning the position of the asymmetric green-highlighted inductive fin, d , along the propagation axis (y -axis).

Table 3.4: Dimensions in mm of the SIW SFN shown in Figure 3.15.

a	p	l_p	a_f	l_f	p_i	d	d_v	s_v
1.54	3.21	0.37	1.54	1.20	0.32	0.075	0.10	0.10

Table 3.5: Extracted LCRG parameters of the series and shunt resonators for the SIW SFN shown in Figure 3.15.

resonator	parameter			
series	f_{se} (GHz)	R (Ω)	L (nH)	Q_{se}
	77.99	6.95	4.24	298.95
shunt	f_{sh} (GHz)	G (μ S)	C (fF)	Q_{sh}
	77.99	459.65	44.74	47.66

a dip at the center frequency, 78 GHz, which is indicative of low losses in the series resonator or $Q_{se} \gg Q_{sh}$. This point may be understood by considering the geometry of the unit cell [Figure 3.4], where the power division port is located at the center of the unit cell. Applying the even and odd excitations, it is clear that only the even mode (shunt) contributes to the power division. Table 3.5 lists the extracted LCRG parameters for the series and shunt resonators. Furthermore, introducing the TA in the unit cell shows that the dip at the broadside frequency can be effectively eliminated, with the same characteristics for the Bloch impedance to the WG impedance [Equation 3.58]. Similarly, the propagation constant is plotted in Figure 3.16(b) using Equation 3.6, which confirms the constant power splitting over frequency for the case of the asymmetric unit cell. As for the Bloch impedance, it can be seen that the attenuation constant for the optimal TA case, αp , is smooth over frequency, meaning that the asymmetry has solved the open-stopband problem observed in the symmetric attenuation constant $\alpha^{\text{sym}} p$.

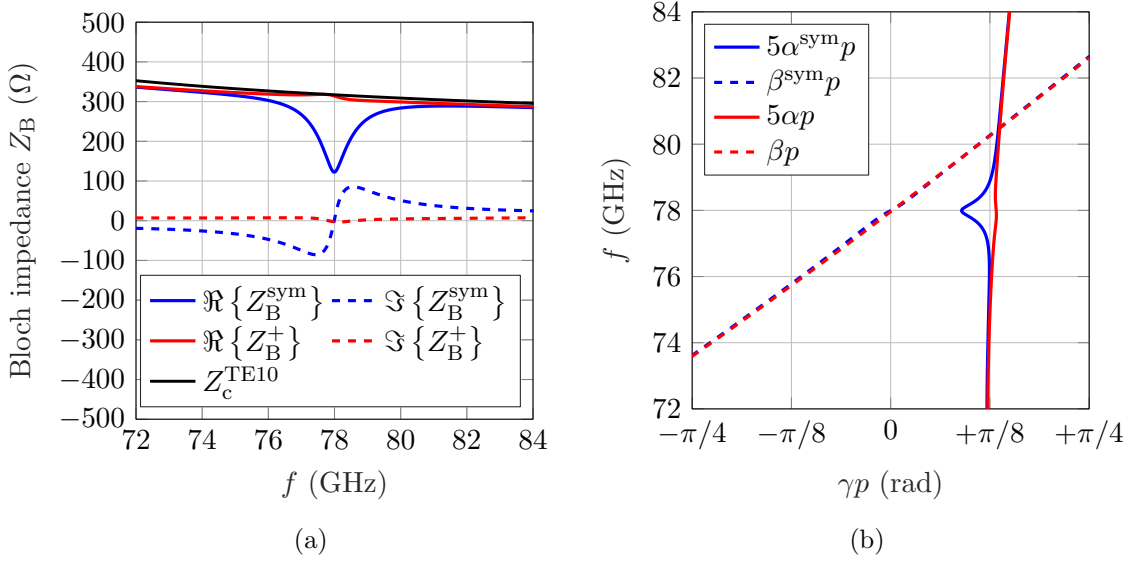


Figure 3.16: Full-wave simulated Bloch impedance and propagation constant of the SIW SFN unit cell for the symmetric and TA cases, i.e., $d = 0$ and $d = 0.075$ mm, respectively. The condition for optimal asymmetry degree $T = T_{\text{opt}} = 0.97$ is satisfied for Z_B^+ . The attenuation constant, αp , is multiplied by a factor 5 for better visibility.

Finally, the unit cell is cascaded to construct the SFN, as shown in Figure 3.15. The S-parameters are plotted in Figure 3.17 for both the symmetric and optimal TA cases when port 1 is excited. The symmetric case [Figure 3.17(a)] shows a high reflection for the input port (S_{11}) at the broadside frequency 78 GHz and an abrupt change for the power division at the output ports (3,4,...,12) at the same frequency. In contrast, the optimal TA achieves a smooth transition through the broadside frequency and excellent matching at the input port with $S_{11} < -30$ dB. Additionally, the amplitude of the power division decays exceptionally across the output ports according to the attenuation constant αp , as shown in Figure 3.16(b). It is worth noting that an equal power division can be achieved if an inverse exponential taper is implemented for successive unit cells. Such a structure can be implemented by starting with a large inductive post l_p on the sidearm (low power division) of the first unit cell and decreasing the length gradually while satisfying the equalization conditions for each unit cell.

3.6 Coupled-Mode Analysis

Section 3.5 proves that the broadside problem can be resolved by introducing the proper amount of TA to the unit cell of a P-LWA. Based on the transformer circuit model [Figure 3.8], the TA shows that the equalization is based on mode coupling between the series and shunt resonators, where an optimal TA corresponding to an optimal coupling results in the bandstop closer at broadside. In this section, an electromagnetic perspective for the mode coupling between the two radiation mechanisms will be provided, and it will be compared to the circuital one. This section is based on the results published in [92, 93] and based on the analytical

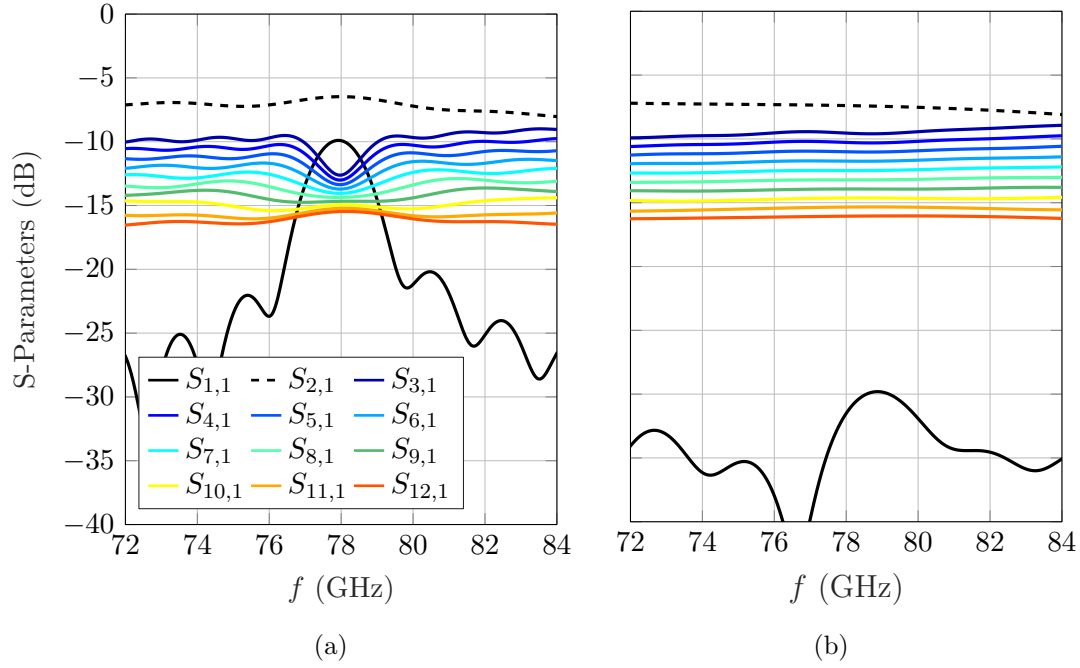


Figure 3.17: S-parameters of SIW SNFs with 10 unit cells for (a) symmetric and (b) asymmetric cases.

approach to map the complex propagation constant to the complex eigenfrequencies published in [25].

3.6.1 Driven-Mode (γ) and Eigenmode (Ω) Approaches

The analysis in the previous sections was performed using the *driven-mode approach*. This approach allows to describe the propagation characteristics of the electromagnetic wave in a periodic structure by the complex propagation constant, $\gamma(\omega) = \alpha(\omega) + j\beta(\omega)$ ($\gamma \in \mathbb{C}$), as

$$\psi(x, y, z, t) = \psi(x, z)e^{-\gamma y}e^{j\omega t}, \quad (3.59)$$

where $\omega \in \mathbb{R}$ is the real-valued angular frequency. In contrast, the *eigenmode approach*, which is derived from the Helmholtz wave equation, describes the electromagnetic wave by a complex eigenfrequency $\Omega \in \mathbb{C}$ and the real-valued phase constant $\beta \in \mathbb{R}$, as follows

$$\psi(x, y, z, t) = \psi(x, z)e^{-j\beta y}e^{j\Omega t}, \quad (3.60)$$

where the real part of the complex frequency, $\Re\{\Omega\}$, corresponds to the oscillation and the imaginary part, $\Im\{\Omega\}$, to the loss (damping) of the wave. In other words, the driven mode describes a temporally oscillating and spatially damping wave, whereas the eigenmode describes a temporally damping and spatially oscillating wave.

Unfortunately, simulating a unit cell with *periodic boundary conditions* (PBCs) is not possible in driven-mode analysis, which means that the driven-mode analysis

does not account for the mutual coupling between the adjacent unit cells. However, a workaround to resolve the issue can be done by simulating multiple unit cells, e.g., N , and then applying the cascading property of the transfer matrix, as $\sqrt[N]{\mathbf{ABCD}}$, to extract the unit cell parameters [94]. Simulating multiple unit cells requires higher computation and therefore takes a long time. On the other hand, PBC is supported by the eigenmode approach, which allows for accurate single unit-cell simulation and hence fast computation. The problem is that the eigenmode parameters do not represent the regular operation of a P-LWA, where it is excited at one end and terminated at the other. However, both approaches may be unified if their respective parameters ($\Omega(\beta) \leftrightarrow \gamma(\omega)$) are mapped to each other, as described for general electromagnetic structures in [95–98]. The mapping procedure will be described in Section 3.6.1.2 in further details.

Furthermore, the Q -factor can be defined for a lossy eigenmode with $\Im\{\Omega\} > 0$,¹⁰ as follows [70, 99],

$$Q = \frac{\Re\{\Omega\}}{2\Im\{\Omega\}}. \quad (3.61)$$

In the case of a *transversally-symmetric* P-LWA, two orthogonal uncoupled eigenmodes exist, which correspond to the series and shunt resonators, whose eigenfrequencies are Ω_{se} and Ω_{sh} , respectively. Moreover, their Q -factors can be written analogously to Equation 3.61 as

$$Q_{\text{se}} = \frac{\Re\{\Omega_{\text{se}}\}}{2\Im\{\Omega_{\text{se}}\}}, \quad (3.62a)$$

$$Q_{\text{sh}} = \frac{\Re\{\Omega_{\text{sh}}\}}{2\Im\{\Omega_{\text{sh}}\}}. \quad (3.62b)$$

In the driven-mode approach [Section 3.4.1], the series and shunt resonators parameters are extracted by applying the odd and even excitations. In the eigenmode approach, the odd and even decomposition corresponds to the *perfect electric conductor* (PEC) and *perfect magnetic conductor* (PMC) boundary conditions, respectively.

3.6.1.1 Coupled-Mode Equation

Here, the coupled-mode theory is applied to calculate the coupled eigenmodes in P-LWAs for the case of a TA unit cell. The modal amplitudes ψ_1 and ψ_2 of two coupled modes can be expressed as a function of the eigenfrequencies of the uncoupled system and the coupling coefficients between them. A system of two coupled modes may be written in the form of an ordinary differential equation, as [100–102]

$$-j\frac{d\psi_1}{dt} = \Omega_{\text{se}}\psi_1 + \kappa_{12}\psi_2, \quad (3.63a)$$

$$-j\frac{d\psi_2}{dt} = \Omega_{\text{sh}}\psi_2 + \kappa_{21}\psi_1, \quad (3.63b)$$

¹⁰In P-LWAs even with the theoretical absence of dielectric and conductor losses, the eigenmodes are lossy (complex) due to radiation.

where κ_{12} and κ_{21} are the coupling coefficients from mode 1 to mode 2 and vice versa, respectively. In the case of two identical waveguides, which is essentially the same case for the P-LWA $\kappa_{12} = \kappa_{12}^* = \kappa$. Moreover, κ accounts for the frequency detuning of the coupled modes. For simplicity, the frequency-balancing condition is only considered, i.e., $\Re\{\Omega_{se}\} = \Re\{\Omega_{sh}\} = \Omega_{bs}$ with $\Omega_{bs} \in \mathbb{R}$, which corresponds to the broadside frequency. The solution for the coupled eigenfrequencies can be found by simply solving the differential equation system in Equation 3.63. To derive the solutions for Ω , first, the coupling matrix is written from Equation 3.63 in the form

$$\mathbf{M} = \begin{bmatrix} \Omega_{se} & \kappa \\ \kappa^* & \Omega_{sh} \end{bmatrix} = \begin{bmatrix} \Omega_{bs} + j\Im\{\Omega_{se}\} & \kappa \\ \kappa^* & \Omega_{bs} + j\Im\{\Omega_{sh}\} \end{bmatrix}, \quad (3.64)$$

where the linear differential equation has a nontrivial solution if $|\mathbf{M} - \Omega\mathbf{I}| = 0$, with $\mathbf{I}$ being the identity matrix. Finally, the two roots for Ω can be found after solving the quadratic equation,

$$(\Omega_{se} - \Omega)(\Omega_{sh} - \Omega) - \kappa^2 = 0, \quad (3.65)$$

whose solutions read

$$\Omega_{1,2} = \Omega_{bs} + j\frac{\Im\{\Omega_{se}\} + \Im\{\Omega_{sh}\}}{2} \pm \sqrt{\kappa^2 - \left(\frac{\Im\{\Omega_{se}\} - \Im\{\Omega_{sh}\}}{2}\right)^2}. \quad (3.66)$$

The above equation shows that the equalization condition for the eigenmode approach, i.e., *complex frequency balancing*, can be achieved by either equalizing the losses of the uncoupled eigenmodes ($\kappa = 0$), $\Im\{\Omega_{se}\} = \Im\{\Omega_{sh}\}$,¹¹ or by introducing coupling (TA) between the eigenmodes so that losses in the *coupled* modes are equalized. From Equation 3.66, the optimal coupling coefficient can be derived so that the radicand disappears. Hence, the optimal coupling coefficient reads

$$\kappa_{\text{opt}} = \frac{\Im\{\Omega_{se}\} - \Im\{\Omega_{sh}\}}{2}. \quad (3.67)$$

Inserting the optimal coupling coefficient in Equation 3.66 coalesces the eigenfrequencies, $\Omega_1 = \Omega_2$, which verifies the complex eigenfrequency balancing condition based on the circuit model of TA P-LWAs [25]. Additionally, the condition $\Omega_1 = \Omega_2$ corresponds to coupled quality factor balancing, i.e., $Q_1 = Q_2$, as

$$Q_{\text{opt}} = \frac{\Omega_{bs}}{\Im\{\Omega_{se}\} + \Im\{\Omega_{sh}\}} = \frac{2}{\frac{1}{Q_{se}} + \frac{1}{Q_{sh}}}. \quad (3.68)$$

Interestingly, the coupled-mode system in Equation 3.63 describes eigenfrequencies of a system that obeys parity-time symmetry, usually denoted by $\mathcal{PT}$ symmetry.¹² Bender and Bottcher first discovered this type of $\mathcal{PT}$ -symmetric system

¹¹This equalization is equivalent to that of Section 3.4.4, where the unit cell parameters satisfy the Heaviside condition, i.e., $Q_{se} = Q_{sh}$.

¹²The topic of $\mathcal{PT}$ symmetry is beyonde the scope of this work and only mentioned due to the intriguing similarity to the P-LWA eigenmode problem.

in 1998 [103],¹³ which led to the discovery of unusual electromagnetic phenomena and motivated various applications in optics such as for sensing and energy transfer [104–107]. However, there are two differences despite the similarity of the coupled $\mathcal{PT}$ symmetry and the problem discussed in this section. First, most $\mathcal{PT}$ -symmetric systems reported are weakly coupled, i.e., $0 < \kappa < \kappa_{\text{opt}}$. Second, the optimal condition involves gain-loss balancing.¹⁴ In contrast, this section showed that the optimal condition for P-LWAs can be achieved by solely loss-loss balancing by utilizing the level of coupling, which is more practical when considering passive structures as P-LWAs.

3.6.1.2 Analytical $\Omega - \gamma$ Mapping

The optimal coupling point [Equation 3.67] derived from the analytical solution of the complex eigenfrequencies may not directly reveal that it corresponds to the optimal operation of a P-LWA. This fact is important because the complex eigenmodes represent a mathematical solution to the periodic boundary problem and may not necessarily represent the physical propagation modes described by the complex propagation constant (γ). Therefore, the mapping between the complex eigenfrequencies (Ω) and complex propagation constant (γ) of P-LWAs will be discussed in this section.

Recently, the mapping between these complex regimes was derived based on the analytic continuity theorem for general electromagnetic structures [97, 98]. Furthermore, the mapping is possible through curve fitting using a generic polynomial expansion when an analytical solution of $\beta(\Omega)$ is unavailable. The analytical mapping procedure can be summarized in two steps: 1) Establish an analytical relationship between phase constant, $\beta(\Omega)$, and the complex eigenfrequency, Ω ; 2) apply the mapping using $\gamma(\omega) \leftrightarrow j\beta(\Omega)$ and $\omega \leftrightarrow \Omega$.

This section provides an analytical solution for the coupled eigenfrequencies and the coupling coefficient κ . Therefore, it is necessary to establish a relationship between κ and β to fulfill the first step of the mapping procedure. From the PBCs and the eigenmode solver, the phase shift, Φ , between the terminals of a unit cell is directly related to β and the period of the unit cell, p , as

$$\Phi = \beta p. \quad (3.69)$$

Moreover, Φ can be considered a detuning parameter, similar to κ , as it enforces an additional phase shift without physically altering p . Hence, Φ also corresponds to the detuning of the structure and, therefore, κ can be expressed by two parameters Φ and κ_{asym} , where κ_{asym} is the detuning due to the transversal asymmetry. Mathematically, κ can be expressed as the distance between the detuning parameters as

$$\kappa = \sqrt{(\Omega_{\text{bs}}\Phi)^2 + \kappa_{\text{asym}}^2} = \sqrt{[\Omega_{\text{bs}}\beta(\Omega)p]^2 + \kappa_{\text{asym}}^2}. \quad (3.70)$$

¹³It is stated that a non-Hermitian Hamiltonian, i.e. lossy system, has real eigenfrequencies and obeys $\mathcal{PT}$ symmetry.

¹⁴The optimal operation point ($\kappa = \kappa_{\text{opt}}$) in a $\mathcal{PT}$ -symmetric system is called the exceptional point of degeneracy [108].

Inserting Equation 3.70 into Equation 3.66 and applying steps of the mapping yields the complex propagation constant, $\gamma(\omega)$, which reads

$$\gamma^\pm(\omega) = \pm \frac{j}{\omega_{\text{bs}} p} \times \sqrt{\left(\omega - \omega_{\text{bs}} - j \frac{\Im\{\Omega_{\text{se}}\} + \Im\{\Omega_{\text{sh}}\}}{2}\right)^2 + \left(\frac{\Im\{\Omega_{\text{se}}\} - \Im\{\Omega_{\text{sh}}\}}{2}\right)^2 - \kappa_{\text{asym}}^2}, \quad (3.71)$$

where $\omega_{\text{bs}} = \Omega_{\text{bs}}$. Equation 3.71 has two solutions that differ only by a $\pm$ sign. These solutions correspond to forward and backward propagating waves. Furthermore, the expression of the mapped γ shows that the $\gamma - \omega$ relationship can be calculated solely by performing a single eigenmode simulation, which provides the complex eigenfrequencies of the uncoupled modes, Ω_{se} and Ω_{sh} . Moreover, the optimal level of transversal asymmetry can be calculated from Equation 3.67, which will be later validated in Section 3.6.2.2 by setting $\kappa_{\text{asym}} = \kappa_{\text{opt}}$.

3.6.1.3 Coupling Efficiency and Overlap Integral

The electromagnetic fields of any two coupled modes can be appropriately expressed by the overlap integral, which calculates the spatial overlap of the fields. In the classical coupled-mode problem of two parallel optical (dielectric) waveguides, the coupling occurs in the transverse direction due to the leakage of the evanescent fields. In contrast, the coupled modes in P-LWAs with TA exchange energy within the same bounded structure, so the coupling must occur along the antenna axis (longitudinal axis). Furthermore, the normalized overlap integral quantifies the coupling efficiency between modes. The normalized coupling efficiency reads thus [109–111]

$$\eta_c = \frac{|\int_V \mathbf{E}_1^* \mathbf{E}_2 dv|^2}{\int_V |\mathbf{E}_1|^2 dv \int_V |\mathbf{E}_2|^2 dv}. \quad (3.72)$$

where $\mathbf{E}_1$ and $\mathbf{E}_2$ are the electric fields of the two coupled modes, and where the superscript $(*)$ denotes the complex conjugate. In the symmetric case, i.e., $\kappa = 0$ or equivalently $T = 1$, the fields correspond to the even and odd modes, i.e., $\mathbf{E}_{\text{sh}}$ and $\mathbf{E}_{\text{se}}$, respectively. Furthermore, the volume integral in Equation 3.72 can be reduced to a surface or even line integral if it is calculated over a full period, p .

3.6.2 Numerical and Full-Wave Validation

In this section, based on the aforementioned coupled-mode analysis, the derived coupled eigenfrequencies will be verified using full-wave simulations. As an example, a series-fed patch (SFP) P-LWA will be studied. The simulation results are taken from [92].

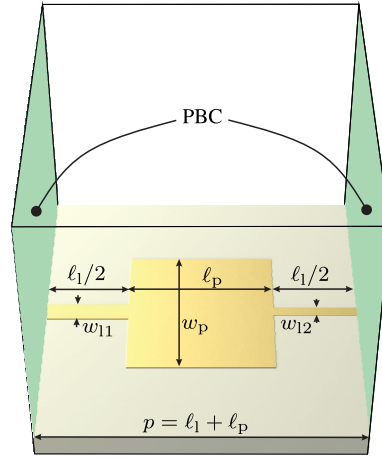
3.6.2.1 Case Study: Series-Fed Patch (SFP) P-LWA

The layout and simulation setup of the SFP P-LWA structure, which operates with the $\mathbf{n} = -1$, is depicted in Figure 3.18. The unit cell consists of three sections, two

Table 3.6: Dimensions in mm of the DS SFP P-LWA shown in Figure 3.18.

ℓ_p	ℓ_1	w_p	w_{11}	w_{12}	w_1	w_d
3.05	3.50	3.30	$w_1 + w_d/2$	$w_1 - w_d/2$	0.3	0

feeding lines connected to two edges of a patch. The conductor of the SFP P-LWA is typically etched on a single core laminate. Here, the substrate core Rogers RO4350B ($\epsilon_r = 3.66$ and $\tan \delta = 0.004$) of height $h = 508 \mu\text{m}$ is used [112]. The period of the unit cell is typically λ_g long at f_{bs} .

**Figure 3.18:** Layout and simulation setup of the SFP P-LWA with PBC at the unit cell's terminals. The TA is tuned by changing the feeding lines widths by w_d .

The unit cell of this P-LWA is first designed as follows. First, tuning the line length (ℓ_1) and patch length (ℓ_p), while fixing all other geometrical parameters, satisfies the frequency-balancing condition ($\Omega_{\text{bs}} = \Re\{\Omega_{\text{se}}\} = \Re\{\Omega_{\text{sh}}\} = 2\pi 24 \text{ GHz}$). Second, tuning the level of transverse asymmetry by varying the feeding line widths, according to $w_{11} = w_1 + w_d$ and $w_{12} = w_1 - w_d$. Furthermore, changing the line width may detune the intrinsic resonance frequencies ($\Re\{\Omega_{\text{se}}\}$ and $\Re\{\Omega_{\text{sh}}\}$), which requires a retuning of the geometries ℓ_p and ℓ_1 by replacing the PBC by PEC and PMC boundary conditions to isolate the intrinsic resonators from the coupled ones.

3.6.2.2 Simulation and Numerical Results

The complex eigenfrequencies are obtained by eigenmode analysis for a single unit cell using the eigenmode solver of the commercial software CST MWS [87]. The simulation is performed by applying *periodic boundary conditions* (PBCs) at the cell's terminals by optimizing the unit cell following the procedure described in Section 3.6.2.1. The simulated uncoupled eigenfrequencies, i.e., $w_d = 0$, are $\Omega_{\text{bs}}/2\pi = 24 \text{ GHz}$, $\Im\{\Omega_{\text{se}}\}/2\pi = 287.44 \text{ MHz}$ and $\Im\{\Omega_{\text{sh}}\}/2\pi = 28.16 \text{ MHz}$. Moreover, the corresponding Q -factors can be calculated using Equation 3.62: $Q_{\text{se}} = 41.75$ and $Q_{\text{sh}} = 426.16$. The dimensions of the optimized unit cell with uncoupled resonators ($w_d = 0$) are given in Table 3.6. Additionally, the coupling coefficient (κ),

which corresponds to the frequency detuning, can be calculated from simulation by rearranging Equation 3.66, as follows

$$\kappa = \sqrt{\left(\Omega_{1,2} - \frac{\Omega_{se} + \Omega_{sh}}{2}\right)^2 + \left(\frac{\Im\{\Omega_{se}\} - \Im\{\Omega_{sh}\}}{2}\right)^2}, \quad (3.73)$$

where $\Omega_{1,2} = \Omega_{sh,se}$ for $w_d = 0$. Figure 3.19 plots the complex eigenfrequency of the two eigenmodes versus the level of TA, i.e., the geometrical parameter w_d and the normalized coupling coefficient κ/Ω_{bs} . The results are plotted for the full-wave simulated and analytical eigenfrequencies, which are obtained by CST and Equation 3.66, respectively. Furthermore, the good agreement between the simulation and analytical results confirms that the coupled-mode theory can accurately model the TA of P-LWAs. Moreover, the optimal asymmetry point can be computed by only knowing the parameters of the uncoupled resonators, similar to the transformer circuit model. By evaluating Equation 3.67, the optimal coupling coefficient yields $\kappa_{opt}/2\pi = 129.63$ MHz or $\kappa_{opt}/\Omega_{bs} = 5.4 \times 10^{-3}$, which corresponds to the optimal asymmetry where the complex frequencies coalesce ($\Omega_1 = \Omega_2$). This point also corresponds to the geometrical parameter $w_d = 22 \mu\text{m}$, as indicated by the green line in Figure 3.19.

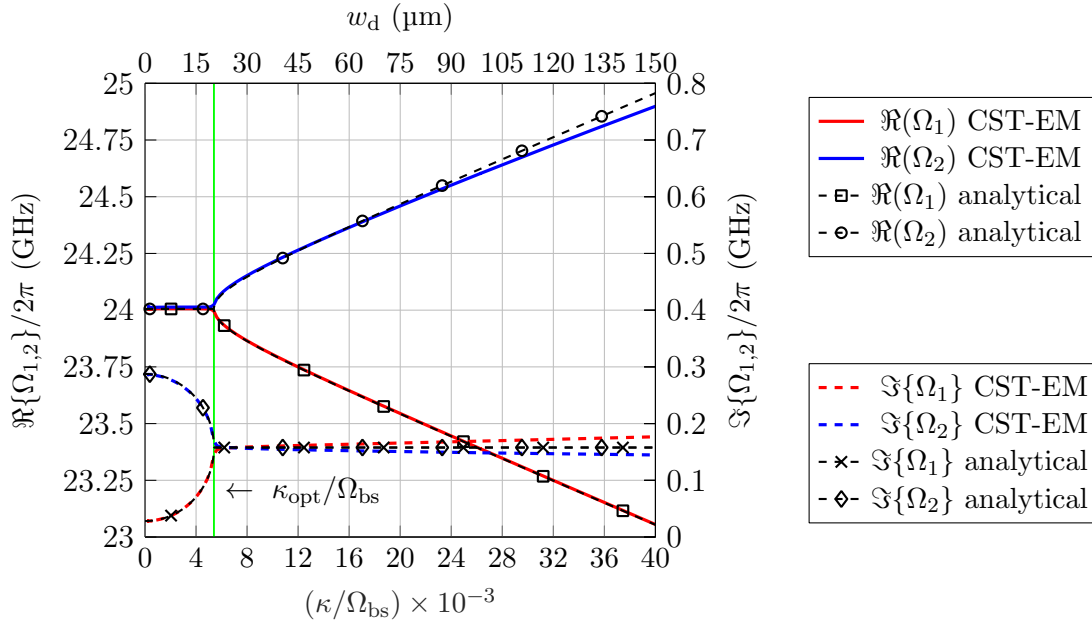


Figure 3.19: Complex eigenfrequencies of the SFP P-LWA unit cell obtained by full-wave simulation using the eigenmode solver in CST (CST-EM) and the analytical expression in Equation 3.66.

Similarly, Figure 3.20 plots the Q -factors, which again confirms that the two Q -factors coalesce at the expected point, i.e., $Q_{opt} \approx 76$ [Equation 3.68]. It is worth noting that the small deviation between the simulation and calculation is due to the detuning of the intrinsic resonance frequencies of the series and shunt resonators at a higher level of asymmetry ($\kappa > \kappa_{opt}$).

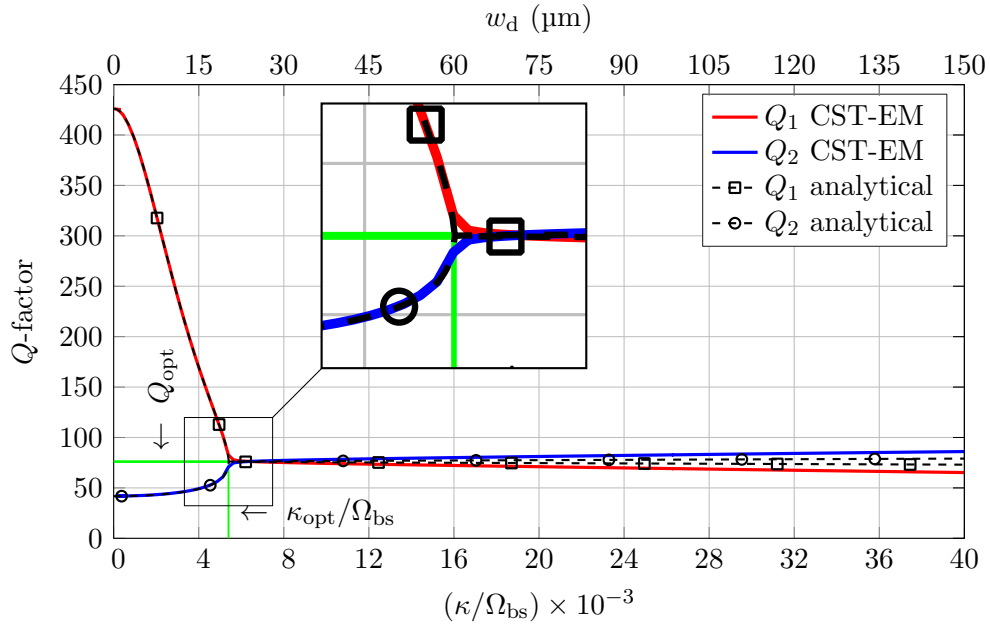


Figure 3.20: Full-wave simulated and analytically calculated coupled Q -factors of the SFP P-LWA unit cell versus the level of TA asymmetry. The inset shows a zoomed view at the optimal asymmetry point.

Next, the dispersion diagram of the SFP unit cell is simulated in the time-domain solver of CST (CST-TD) and compared to the mapped counterpart [Equation 3.71]. To verify the equalization conditions obtained from the transformer network model, T_{opt} , and the coupled-mode analysis, κ_{opt} . The mapped γ is obtained only by calculating the uncoupled modes with the eigenmode solver, whereas the excited structure requires simulating both ports. Figure 3.21(a) compares the dispersion diagrams for the symmetric case, i.e., $w_d = 0$. It can be seen that the analytical mapping delivers excellent agreement with the full-wave simulation. The small divination at the broadside direction can be related to the simulation of a finite structure in CST-TD and without PBCs. In Figure 3.21(b), the optimal level of asymmetry is plotted for the simulated and mapped γ , i.e., $T = T_{\text{opt}}$ and $\kappa_{\text{asym}} = \kappa_{\text{opt}}$. Again, the results show a constant α over frequency, indicating the elimination of the broadside problem. Hence, the unification of both the network-model coupled-mode analysis is established.

Finally, the calculated $\mathbf{E}$ -fields of the two coupled modes are evaluated in the entire calculation domain, i.e., volume integration. Figure 3.22 plots the coupling efficiency, computed by Equation 3.72 for the E_z -component. The integral shows clearly that the optimal asymmetry point corresponds to a 100 % coupling efficiency between the two modes. In other words, at this point, two field patterns of the two modes overlap completely, which is only possible if both modes become degenerate. Furthermore, a 100 % coupling efficiency means the two uncoupled, orthogonal modes, the series, and shunt, couple with increasing transverse asymmetry to the point where they coalesce in magnitude and direction. In the $\mathcal{PT}$ -symmetric representation, this optimal point ($\kappa = \kappa_{\text{opt}}$) is called the *exceptional point of degeneracy* [108].

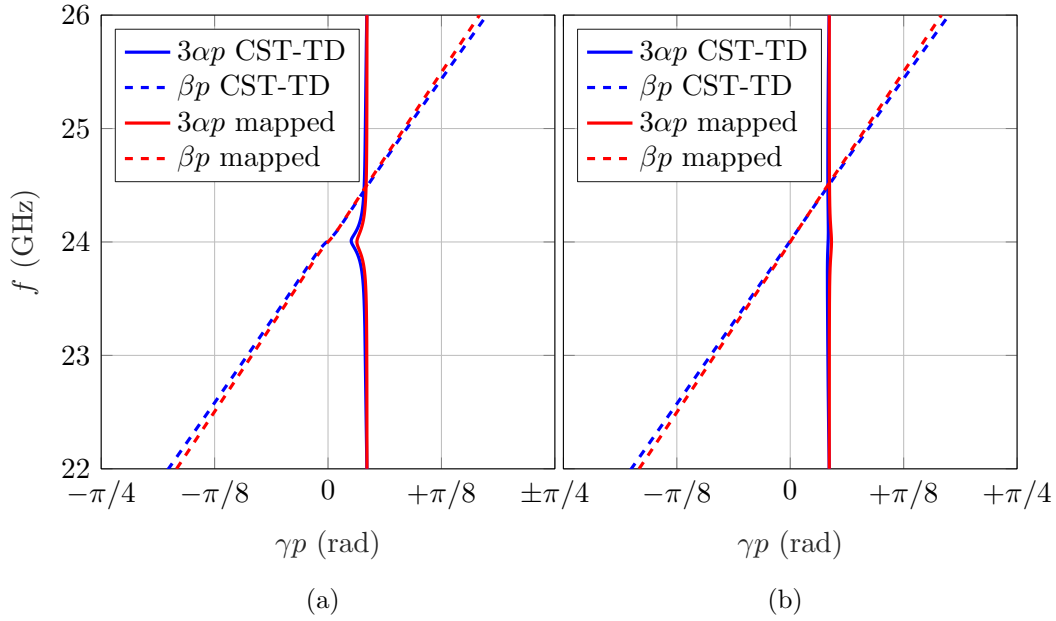


Figure 3.21: Full-wave simulated and analytically mapped [Equation 3.71] dispersion diagrams. (a) Symmetric unit cell with $T = 0$ and $\kappa = 0$. (b) Transversally asymmetric unit cell with $T = T_{\text{opt}}$ and $\kappa_{\text{asym}} = \kappa_{\text{opt}}$. The leakage constant, α , is multiplied by factor 3 for better visibility.

3.7 Summary

In this chapter, the symmetry properties of P-LWAs are discussed based on recently reported general circuit models. First, it is shown that the *double-symmetric* unit cell suffers from gain degradation in the broadside direction due to the open-stopband problem. This stopband is caused by two unfulfilled conditions related to the two radiation mechanisms of P-LWAs, the series and shunt resonators. These conditions are the *frequency-balancing*, and *Q-factor-balancing* conditions, where the former can be satisfied by a double-symmetric unit cell and the latter cannot be satisfied due to the two resonators' unequal loss/radiation contributions, i.e., $Q_{\text{se}} \neq Q_{\text{sh}}$. Furthermore, it is discussed that the asymmetry of the unit cell of a P-LWA is the key to solving the broadside radiation problem. Specifically, the *longitudinally-asymmetric* unit cell satisfies the *Q-factor balancing* condition and hence solves the broadside radiation problem but radiates with elliptical (or circular) polarization due to the orthogonality of the series and shunt radiations. Finally, the *transversal-asymmetric* unit cell resolves the broadside problem by mode coupling and preserves the polarization's linearity. The longitudinally and transversal asymmetries are demonstrated by numerically and two case studies, respectively. Table 3.7 lists the formulas derived from the general circuit models for symmetric and asymmetric unit cells. In the last section, more emphasis is given to the transverse asymmetry. An electromagnetic perspective is provided by applying the coupled-mode theory to understand the coupling mechanism between the series and shunt resonators.

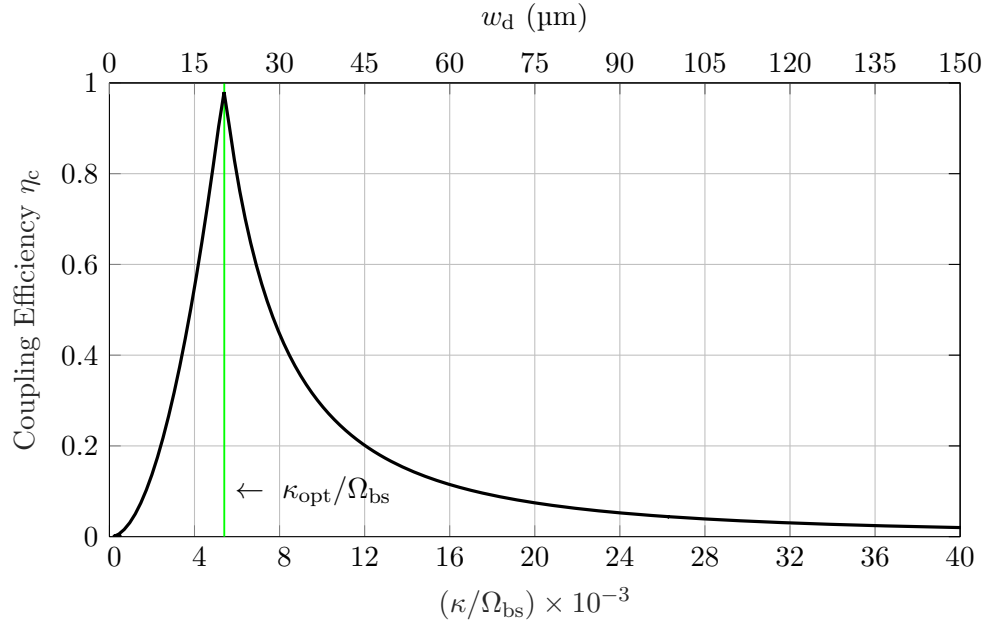
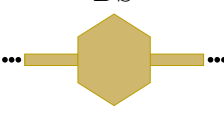
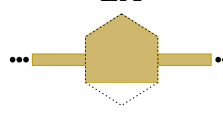
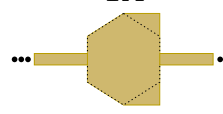


Figure 3.22: Coupling efficiency of the SFP P-LWA unit cell obtained by eigenmode analysis versus asymmetry parameter, κ and w_d .

Table 3.7: Summary of the symmetry types in P-LWA and the corresponding general formulas for the Bloch impedance and radiation efficiency.

symmetry type	DS	LA	TA
$Z_B^\pm$			
$\eta^\pm$	$Z_B^+ = -Z_B^- = Z_B^{\text{sym}}$ $\sqrt{\frac{Z_{se}}{Y_{sh}}}$	$\eta^+ = \eta^- = \eta^{\text{sym}}$ $\frac{\eta_{se}R + \eta_{sh}G Z_B^{\text{sym}} ^2}{R + G Z_B^{\text{sym}} ^2}$	$Z_B^+ \neq Z_B^-$ $\mp \frac{\tau}{Y_{sh}} + \sqrt{\left(\frac{\tau}{Y_{sh}}\right)^2 + \frac{Z_{se}}{Y_{sh}}}$ $\eta^+ \neq \eta^-$ $\frac{\eta_{se}R + \eta_{sh}G Z_B^\pm ^2}{R + G Z_B^\pm ^2}$
Polarization	linear	elliptical or circular	linear

Polarization Diversity

In the previous chapter, it has been shown that the symmetry in the unit cell of a P-LWA constitutes its radiation performance in the broadside direction. Here, it will be shown that the asymmetry also allows the control of the antenna's polarization. First, it will be shown that a P-LWA with optimal LA radiates a circularly-polarized wave with orthogonal handedness depending on the excitation direction, i.e., forward and backward excitation directions. Second, dual linear and dual circular polarizations can be achieved by combining two LA P-LWAs in a mirrored configuration with respect to the longitudinal axis. Finally, orthogonal linearly-polarized radiation can be achieved by combining optimal LA and an extreme level of TA. All three concepts are numerically and experimentally verified.

4.1 Dual Circular-Polarized Radiation with Longitudinal Asymmetric P-LWA

In this section, the polarization of the LA P-LWA will be considered [Figure 3.3(b)]. It will be shown that the series and shunt radiation contributions are orthogonal and excited with a quadrature phase. Furthermore, it will be shown that the radiation of the LA is elliptically or circularly polarized if the Q -balancing condition is satisfied. The investigation in this section will be based on the SFP P-LWA unit cell. The following theoretical derivation is based on the results published in [24] and will be extended to the structure in Section 4.2.

4.1.1 Equivalent Edge Radiation

This section will provide proof that the series and shunt radiations are orthogonal to each other, with the series radiation being transversally polarized and the shunt radiation longitudinally polarized. Additionally, it will be shown that the shunt radiation vanishes for a DS and contributes in the far-field to radiation for the case of LA unit cell. The electromagnetic radiation from antennas occurs by perturbing the electric currents on a conductor. The perturbation of the electric current can be obtained by geometrical discontinuities, such as bending or curving of the conductor edge [113, 114]. Furthermore, an antenna's radiation can be simplified by employing the equivalent principle, which replaces the edge radiation with equivalent sources. There are different ways to define these sources, all derived from the Huygens' principle, which provides the field solution in a defined volumetric region by determining the tangential fields over a surface [113]. Figure 4.1 depicts the unit cell of a DS and LA SFP P-LWA under PBCs, where the radiation from an edge is modeled by *equivalent magnetic surface current density*, $\mathbf{M}$. Furthermore, magnetic surface current

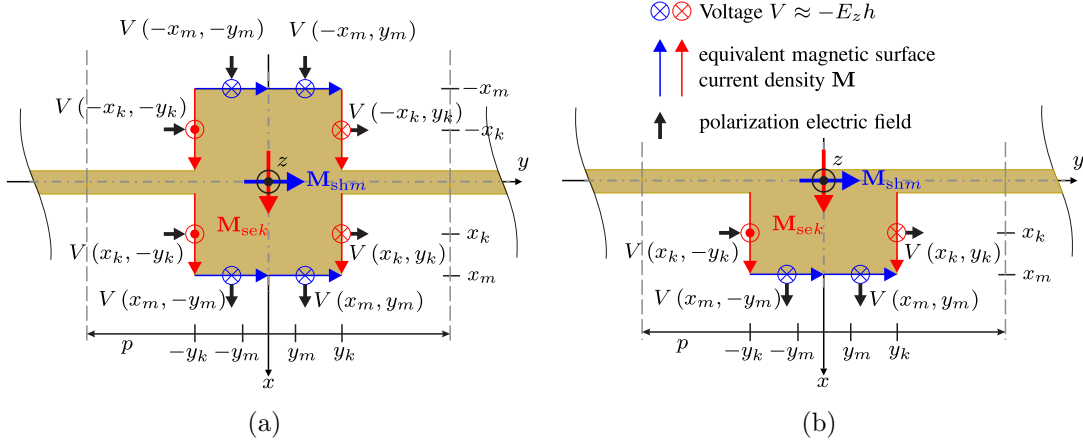


Figure 4.1: Conductor layout of SFP P-LWA unit cells with voltage sampling at the radiation edge corresponding to the series and shunt radiations for (a) double-symmetric (DS) and (b) longitudinally-asymmetric (LA) unit cells. In this chapter, the color codes red and blue will be used for the series and shunt radiation, respectively.

densities corresponding to the series radiation are shown on the vertical edges ($\hat{x}$) by the blue arrows, whereas those to the shunt radiation are depicted on the horizontal edges ($\hat{y}$) by the red arrows. The vertical and horizontal edges are labeled, after [24], by the indices k and m , respectively. Generally, the magnetic surface current can be derived from the transverse electric field [65, 114, 115], as follows

$$\mathbf{M} = \mathbf{E} \times \hat{n} \approx E_z \hat{\mathbf{z}} \times \hat{\mathbf{n}}, \quad (4.1)$$

where $\hat{n}$ is the normal vector to the unit cell's radiation edges (either $\hat{x}$ or $\hat{y}$) and E_z is the z -component of the $\mathbf{E}$ -field between the unit cell's conductor edges and its ground plane. The vectorial surface magnetic current densities can be further simplified by deriving an equivalent scalar quantity, the *voltage*, whose value can be calculated by integrating the electric field between the ground and the top layer conductor (E_z) over the full height, h . For a thin stack, i.e., $h \ll \lambda/10$, E_z is approximately invariant along the z -axis, and thus, the voltage is approximately constant. Mathematically, the voltage can be written as

$$V = \int_0^{-h} \mathbf{E} dz \approx -E_z h, \quad (4.2)$$

where dz is the differential of the integral. Since the voltage is defined from the unit cell's conductor ($z = 0$) to the ground plane ($z = -h$), the voltage quantity in Equation 4.2 is negative.

4.1.2 Polarization in Double-Symmetric Unit Cell

First, the DS unit cell shown in Figure 4.1(a) is considered. For the following derivation, two assumptions are made: 1) The radiation contributions of the series and shunt, in addition to their corresponding equivalent edge voltages, are derived for the far-field radiation case; 2) the loss or radiation across the unit cell is minimal, and therefore, the amplitude taper across the unit cell (y -axis) is neglected for simplicity.

4.1.2.1 Series Radiation

The total magnetic surface current density of the series radiation (red arrows), $\mathbf{M}_{se k}$, is the sum of the four quantities located at the vertical edges, which yields

$$\mathbf{M}_{se k} = \mathbf{M}(-x_k, -y_k) + \mathbf{M}(-x_k, y_k) + \mathbf{M}(x_k, -y_k) + \mathbf{M}(x_k, y_k). \quad (4.3)$$

By using Equations 4.1 and 4.2, the equivalent edge voltage of the series radiation can be calculated by evaluating Equation 4.3 with the normal vectors to the vertical edges, i.e., $\hat{n} = \pm \hat{y}$, as

$$\mathbf{M}_{se k} \cdot \hat{x} = \frac{1}{h} \left[\underbrace{-V(-x_k, -y_k) + V(-x_k, y_k) - V(x_k, -y_k) + V(x_k, y_k)}_{V_{se k}} \right]. \quad (4.4)$$

By symmetry and losslessness, the equivalent edge voltages can be further simplified as

$$V(x_k, y_k) = V(-x_k, y_k), \quad (4.5a)$$

$$V(x_k, -y_k) = V(-x_k, -y_k), \quad (4.5b)$$

which reduces the equivalent edge voltage of the series radiation, $V_{se k}$, in Equation 4.4 to

$$V_{se k} = 2[-V(x_k, -y_k) + V(x_k, y_k)]. \quad (4.6)$$

The series radiation contribution corresponds to the difference of equivalent edge voltages located at the opposite sides of the vertical edges of the patch.

4.1.2.2 Shunt Radiation

Similar to Equation 4.3, the total magnetic surface current density of the shunt radiation [blue arrow in Figure 4.1(a)], $\mathbf{M}_{sh m}$, reads

$$\mathbf{M}_{sh m} = \mathbf{M}(-x_m, -y_m) + \mathbf{M}(-x_m, y_m) + \mathbf{M}(x_m, -y_m) + \mathbf{M}(x_m, y_m), \quad (4.7)$$

which results in the equivalent edge voltage of the shunt radiation after evaluating Equation 4.3 with the normal vectors to the horizontal edges, $\hat{n} = \pm \hat{x}$, as

$$\mathbf{M}_{sh m} \cdot \hat{y} = \frac{1}{h} \left[\underbrace{-V(-x_m, -y_m) - V(-x_m, y_m) + V(x_m, -y_m) + V(x_m, y_m)}_{V_{sh m}} \right]. \quad (4.8)$$

The above equation can be further simplified due to symmetry and the assumption of losslessness, as

$$V(-x_m, -y_m) = V(x_m, -y_m), \quad (4.9a)$$

$$V(-x_m, y_m) = V(x_m, y_m). \quad (4.9b)$$

By inserting Equation 4.9 into Equation 4.8, the equivalent edge voltage from the shunt radiation yields to

$$V_{sh\,m} = 0. \quad (4.10)$$

Equations 4.6 and 4.10 clearly show that $V_{se\,k} \neq 0$ and $V_{sh\,m} = 0$, respectively, which means that in a *double-symmetric* unit cell in P-LWA, the series radiation is the main contributor to radiation at broadside. Furthermore, in the case of the SFP P-LWA, the polarization of the antennas is linear and oriented in the longitudinal direction, i.e., along the y -axis.

4.1.3 Polarization in Longitudinal-Asymmetric Unit Cell

The LA is realized by cutting out the upper part of the patch element with respect to the longitudinal axis, as shown in Figure 4.1(b). In this case, two vertical edges vanishes, i.e., $V(-x_k, -y_k) = V(-x_k, y_k) = 0$. Therefore, the equivalent edge voltage of the series radiation in Equation 4.6 reduces to

$$V_{se\,k} = -V(x_k, -y_k) + V(x_k, y_k). \quad (4.11)$$

Analogously, the horizontal edges in the negative x -direction vanishes, $V(-x_m, -y_m) = V(-x_m, y_m) = 0$, which provides a different result for the equivalent edge voltage of the shunt radiation in Equation 4.10. By re-evaluating Equation 4.8, the new shunt radiation contribution reads

$$V_{sh\,m} = V(x_m, -y_m) + V(x_m, y_m). \quad (4.12)$$

The *longitudinal-asymmetric* unit cell shows that both series and shunt radiations contribute to radiation in the broadside direction. Moreover, in this case, the shunt radiation is orthogonal to the series one and is polarized in the transversal direction. Furthermore, the radiating patch element is cut in half, eliminating two radiating edges (one vertical and one horizontal). In practice, tuning the length of the patch cut allows controlling the shunt radiation accordingly, as it will be shown in Section 4.1.5.

4.1.4 Circular Polarization Conditions

The previous section shows that the series and shunt radiation contributions are orthogonal. Here, it will be shown that they are excited by a phase quadrature, and their amplitude can be equalized by employing LA, ultimately allowing circular polarization. For the following derivation and discussion, the ω -balancing condition ($\omega_{bs} = \omega_{se} = \omega_{sh}$) is assumed since it is the minimum required condition for optimal operation.

4.1.4.1 Quadrature Phase Relation

The phase relation may be derived for the broadside and off-broadside regimes separately. The former will be derived from the simple concept of wave propagation and the latter based on the lattice circuit model, shown in Section 3.4.3.

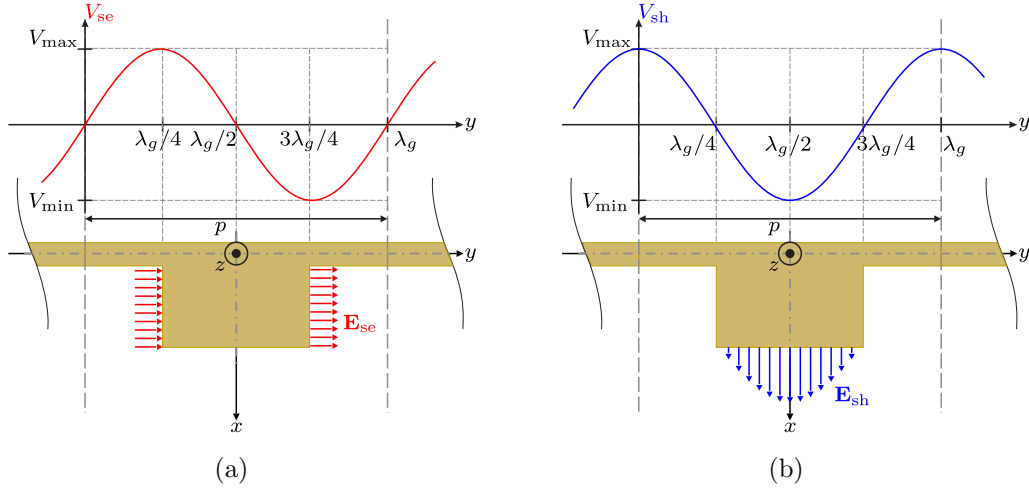


Figure 4.2: Voltage (top) and electric field (bottom) patterns on the unit cell of an SFP P-LWA at the broadside frequency corresponding to the series and shunt radiation. The series and shunt resonators are excited at different time instances due to the feeding line, $\lambda_g/4$. (a) At time $t = 0$, the series voltage excites the vertical edges, (b) whereas after a quarter-time period, $t = T/4$, the shunt voltage excites the horizontal edge of the radiating patch element.

Figure 4.2 shows the series and shunt voltages and electric field distributions along the P-LWA axis (y) on the unit cell of an SFP P-LWA. At a time instance $t = 0$ [Figure 4.2(a)], the vertical edges corresponding to the series radiation are excited, radiating in the longitudinal direction. Since the feeding line length is $\lambda_g/4$ at the broadside frequency, the voltage distribution will propagate by a $\lambda_g/4$ distance after a quarter-time period, i.e., $t = T/4$, as shown in Figure 4.2(b). At this time instance, the horizontal edge corresponding to the shunt radiation is excited, which radiates in the transversal direction. Since the time difference between the excitations of both orthogonal edges is $T/4$, the phase difference must be equal to $\pm 90^\circ$, resulting in an elliptical polarization at the broadside direction. Furthermore, if the amplitudes of both orthogonal polarizations are equal, then the P-LWA will radiate with circular polarization.

The perspective of field distributions along the SFP P-LWA can be visualized and understood for the broadside frequency, whereas at the off-broadside frequency the field distributions cannot provide information about the phase quadrature relation between the series and shunt excitations. Therefore, the phase relation will be derived with the help of the lattice circuit model. The shunt and series voltage ratio, V_{sh}/V_{se} , can be related to the Bloch and the series impedances with the help of Equations 3.30 and 3.32, as follows

$$\frac{V_{sh}}{V_{se}} = \frac{V_{sh}}{I_{se} Z_{se}} = \frac{Z_B}{Z_{se}} = \frac{1}{R + j2L(\omega - \omega_{bs})} \sqrt{\frac{R + j2L(\omega - \omega_{bs})}{G + j2C(\omega - \omega_{bs})}}. \quad (4.13)$$

The phase for the off-broadside frequency can be derived by calculating the limit of Equation 4.13, which yields

$$\left. \frac{V_{\text{sh}}}{V_{\text{se}}} \right|_{\omega=\omega_{\text{off-bs}}} = \lim_{\omega-\omega_{\text{bs}} \rightarrow \infty} \frac{Z_{\text{B}}}{Z_{\text{se}}} = \frac{1}{j2L} \sqrt{\frac{L}{C}} = \frac{-j}{2\sqrt{LC}}, \quad (4.14)$$

which shows that the phase of the series voltage is leading the shunt voltage, i.e., $\angle V_{\text{sh}} - \angle V_{\text{se}} = -90^\circ$. Interestingly, for backward wave propagation, the phase will flip by 180° since the Bloch impedance for the backward propagating wave is negative, i.e., for a transversally symmetric unit cell, $Z_{\text{B}} = Z_{\text{B}}^+ = -Z_{\text{B}}^-$, which yields $\angle V_{\text{sh}} - \angle V_{\text{se}} = +90^\circ$. The result indicates that if the amplitudes of both radiation contributions in the far field are equal, then the radiated wave will be circularly polarized.

4.1.4.2 Amplitude Relation

In the far field, the amplitudes of both orthogonal polarizations are directly proportional to the series and shunt powers. Thus, the power delivered to the series and shunt resonators shall be derived from the lattice circuit model in Section 3.4.3, which can be written as follows

$$P_{\text{se}} = \frac{1}{2} |I_{\text{se}}|^2 R, \quad (4.15a)$$

$$P_{\text{sh}} = \frac{|V_{\text{sh}}|^2}{2G}, \quad (4.15b)$$

and whose ratio is

$$\frac{P_{\text{sh}}}{P_{\text{se}}} = \left| \frac{V_{\text{sh}}}{I_{\text{se}}} \right|^2 \frac{G}{R} = |Z_{\text{B}}|^2 \frac{G}{R} = \left(\frac{|Z_{\text{B}}|}{Z_{\text{B,bs}}} \right)^2. \quad (4.16)$$

The above equation shows that the power ratio between the series and shunt resonators only depends on the Bloch impedance parameters. For the case of a transversally-symmetric unit cell and the broadside operation, the expression in Equation 4.16 can be simplified, using Equation 3.37 to

$$\left. \frac{P_{\text{sh}}}{P_{\text{se}}} \right|_{\omega=\omega_{\text{bs}}} = 1, \quad (4.17)$$

which shows the equal power splitting. Furthermore, at the off-broadside operation, the power ratio can be written as

$$\left. \frac{P_{\text{sh}}}{P_{\text{se}}} \right|_{\omega=\omega_{\text{off-bs}}} = \left(\frac{|Z_{\text{B,off-bs}}|}{Z_{\text{B,bs}}} \right)^2 = \frac{L}{C} \frac{G}{R} = \frac{Q_{\text{sh}}}{Q_{\text{se}}}. \quad (4.18)$$

The last expression clearly indicates that the power splitting at the off-broadside frequency can be equalized, as for the broadside frequency, by fulfilling the Q -balancing condition.

Table 4.1: Dimensions in mm of the optimized SFP P-LWA shown in Figure 4.3.

ℓ_1	ℓ_p	w_1	w_p	d
1.37	1.02	0.30	1.2	0.17

4.1.5 Full-Wave and Experimental Validation

This section will demonstrate an LA SFP P-LWA consisting of 10 unit cells, as shown in Figure 4.3, operating with a broadside frequency at 78 GHz and radiating circular polarization. The antenna structure uses the single-layer laminate from Rogers RO3003 ($\epsilon_r = 2.97$ and $\tan \delta = 0.001$) with thickness $h = 127 \mu\text{m}$ [86]. The forthcoming simulations have been performed with a copper surface roughness of $1 \mu\text{m}$ that has a major impact on the losses and the phase. This value was obtained empirically by first experimentally measuring two through MSL lines with different lengths and then adjusting the surface roughness in the simulation to match the insertion loss and the phase from the measured data. Comprehensive studies measuring the dielectric constant and losses of RF materials around the 78 GHz frequency can be found in the literature [116–118].

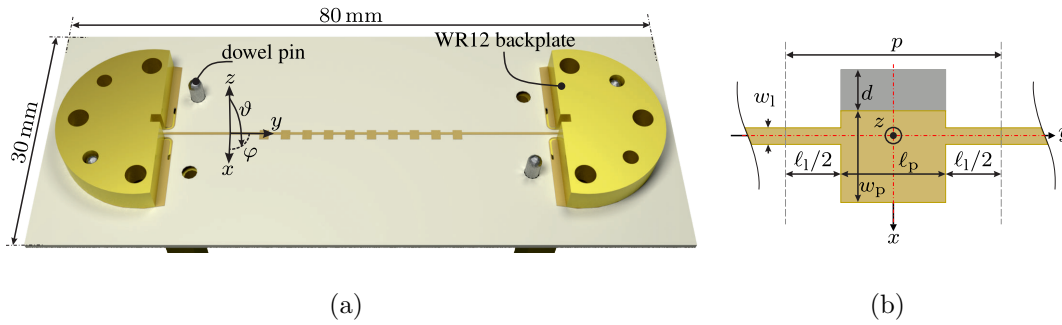


Figure 4.3: SFP P-LWA with LA consisting of 10 unit cell. (a) Perspective view showing the antenna with its feeding structure and (b) top view of a single unit cell including its dimensions.

4.1.5.1 Unit-Cell Simulation

The unit cell layout is shown in Figure 4.3(b). The design procedure starts with optimizing the unit cell for the ω -balancing condition using the procedure in Section 3.5.3 by tuning the patch and feeding lines lengths, ℓ_p and ℓ_1 , respectively. Next, the unit cell is optimized for the Q -balancing condition to equalize the radiation contributions of the series and shunt resonators. The Q -factor equalization is realized by shifting the patch in the x -direction, which mitigates the cancellation of the shunt resonator in the far field direction, as described in Section 4.1.3. The optimized dimensions of the unit cell are listed in Table 4.1.

Figure 4.4 plots the Bloch impedance and propagation constants obtained from the time-domain solver of the commercial simulation software CST MWS [87]. Figure 4.4(a) shows an approximately smooth transition around the broadside frequency 78 GHz. Additionally, the Bloch impedance increases with the frequency, due to an

additional resonance of the patch in the transverse direction at 68.1 GHz. Moreover, Figure 4.4(b) shows that the loss factor, α , has a constant transition around the broadside frequency, which indicates the elimination of the open-stopband problem.

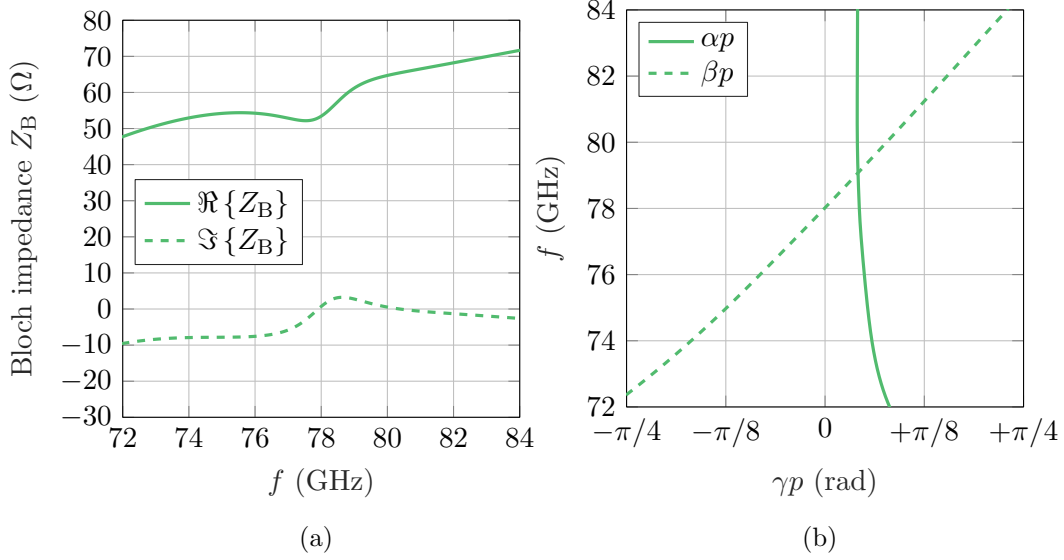


Figure 4.4: Full-wave simulated unit cell of the LA SFP P-LWA. (a) Bloch impedance and (b) propagation constant.

4.1.5.2 10 Unit Cells Structure

A cascade of 10 unit cells will be presented in this section to investigate the circular-polarization radiation performance. First, the effect of the patch shift, d , is studied. Figure 4.5 plots the axial ratio of the SFP P-LWA at the broadside frequency 78 GHz for different patch shifts, d . It can be seen that the antenna is linearly polarized for $d = 0$ and achieves the best circular polarization performance at $d \approx 0.17$ mm, which corresponds to the Q -balancing point in the unit cell simulation. Furthermore, the simulation results provide the allowed tolerance for the axial ratio below 3 dB, approximately ± 25 μ m.

The antenna has been fabricated in-house at the Institute of High Frequency Technology of the RWTH Aachen University. The copper structure has been etched by laser structuring using the LPKF Protolaser S machine [119]. The fabricated prototype of a 10 unit cells structure, including the waveguide-to-MSL transition, is shown in Figure 4.6(a). Further information regarding the transition design is provided in Appendix B.1. Additionally, the waveguide transition does not affect the performance of the P-LWA, due to the shielding provided by the top plate. Furthermore, the S-parameter results are plotted in Figure 4.7 and show good agreement between the simulation and measurement except for a slight frequency shift. Both simulation and measurement show good matching below -10 dB over the frequency operation bandwidth 76 – 81 GHz.

Finally, the far-field measurement was performed in the compact test range test of the institute, as shown in the measurement setup in Figure 4.6(b). E-band mm-wave

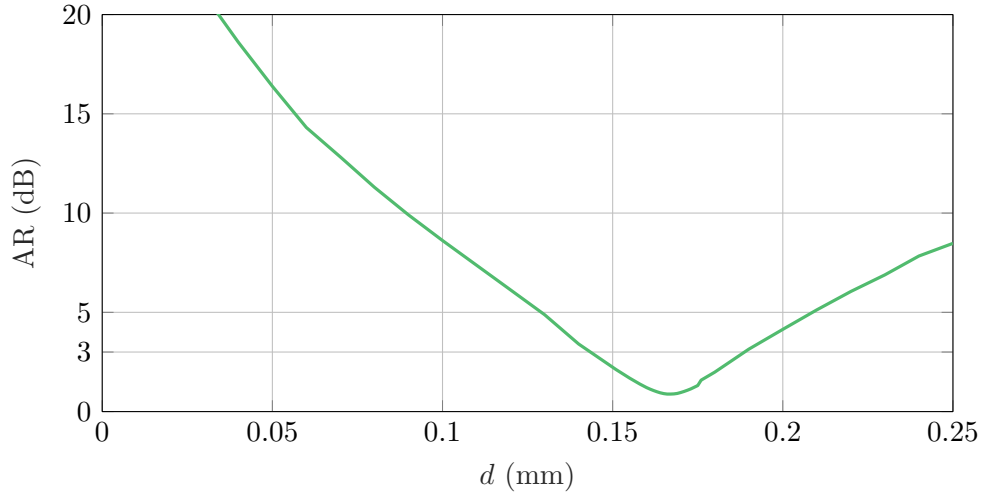


Figure 4.5: Effect of the patch shift in the transverse direction, d , on the axial ratio at broadside (f_{bs}) obtained by full-wave simulation.

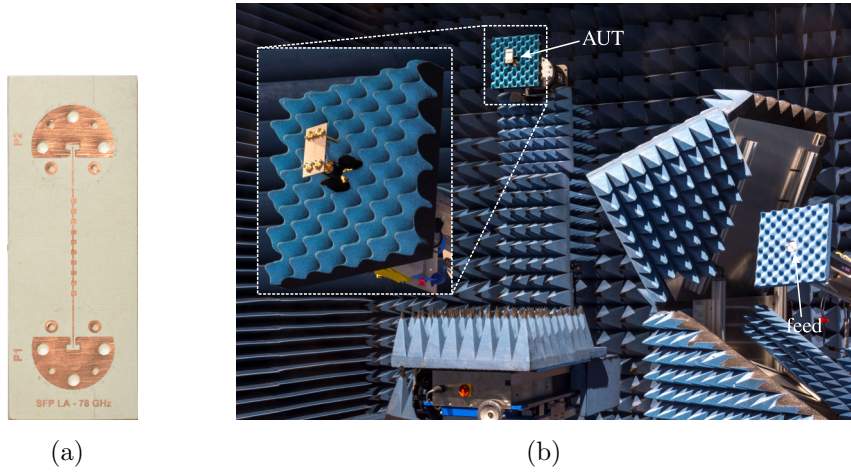


Figure 4.6: Photographs of the fabricated P-LWA SFP antennas and the measurement setup. (a) Top view of the antenna without the transition plates and (b) photograph of the compact antenna test range at the Institute of High Frequency Technology of RWTH Aachen university.

frequency extenders with WR-12 waveguides from Virginia Diodes, Inc. have been used for the measurement. Figure 4.8 shows a comparison between the measured and simulated realized gain. Figure 4.8(a) and Figure 4.8(b) show the radiation patterns for the principal xz -plane ($\varphi = 0^\circ$) and yz -plane ($\varphi = 90^\circ$) [Figure 4.3(a)] at 78 GHz, respectively. Excellent agreement between the simulation and measurement is observed with a maximum LHCP gain of 13.5 dBi and cross-polarization discrimination of 23.2 dB. Additionally, the simulated directivity is about 16.1 dBi at the same frequency, which results in 58 % radiation efficiency. Figure 4.8(c) plots the realized gain for five different frequencies and shows that the antenna exhibits an optimal axial ratio at the center frequency. The cross-polarization discrimination ranges from 12 dB at 76 GHz to 13 dB at 81 GHz.

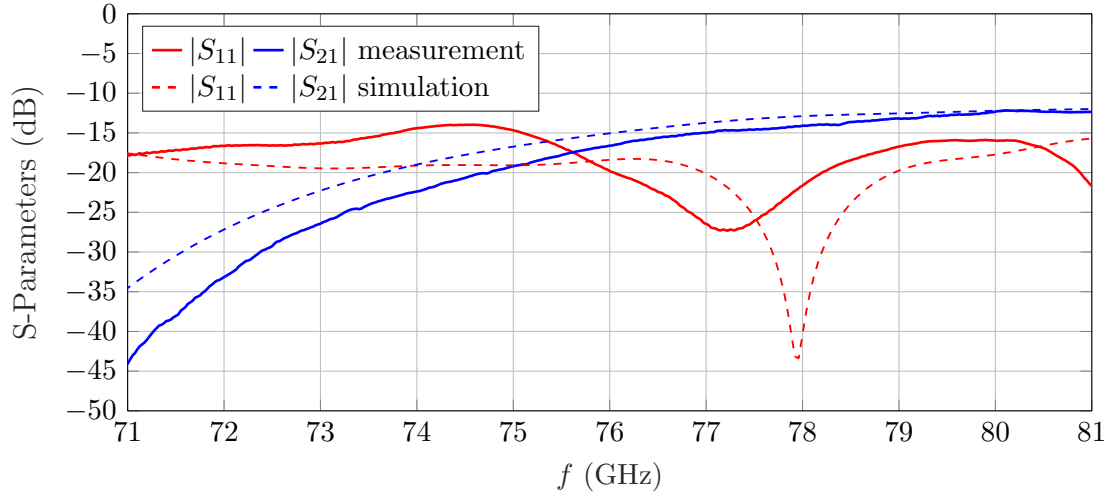


Figure 4.7: Full-wave simulation and measurement comparison of the S-parameters for a single SFP P-LWA with longitudinal asymmetry.

Figure 4.9 plots the axial ratio at the main beam direction, ϑ_{mb} , over frequency, for better interpretation of the circular polarization performance. It can be observed that the axial ratio remains below 3 dB from 76.6 GHz to 81 GHz in the simulation result. In comparison, the measured axial ratio is shifted towards higher frequencies and exhibits a slightly worse axial ratio (>3 dB) above 79.5 GHz. The disagreement between the simulation and measurement can be attributed to the fabrication tolerances. Nevertheless, the prototype verifies the circular polarization radiation in P-LWA with longitudinal asymmetry in the 76 – 81 GHz frequency band. Furthermore, the concept of circular polarization radiation with LA was also implemented in a non-TEM transmission line, i.e., dielectric image line with all-dielectric radiating elements, as reported in [120].

4.1.6 Orthogonal Circular Polarization

The previous section showed that exciting the P-LWA in the forward direction, i.e., $+y$ -direction will result in LHCP. For the case of backward excitation ($-\hat{y}$), the series radiation [Figure 4.2(a)] will radiate in the opposite direction with respect to the forward direction. Since the phase relation of the series and shunt excitations does not change with the excitation direction due to the transverse symmetry of the structure, the handedness of the radiated CP will flip. Therefore, the excitation in the backward direction will be RHCP for SFP P-LWA in Figure 4.3. Due to the symmetry of the antenna structure, the radiation performance is identical to the forward port (port 1). Thus, the results for port 2 will not be shown to avoid redundancy, but they can be derived from those in Section 4.1.5, except for the handedness of the polarization (RHCP).

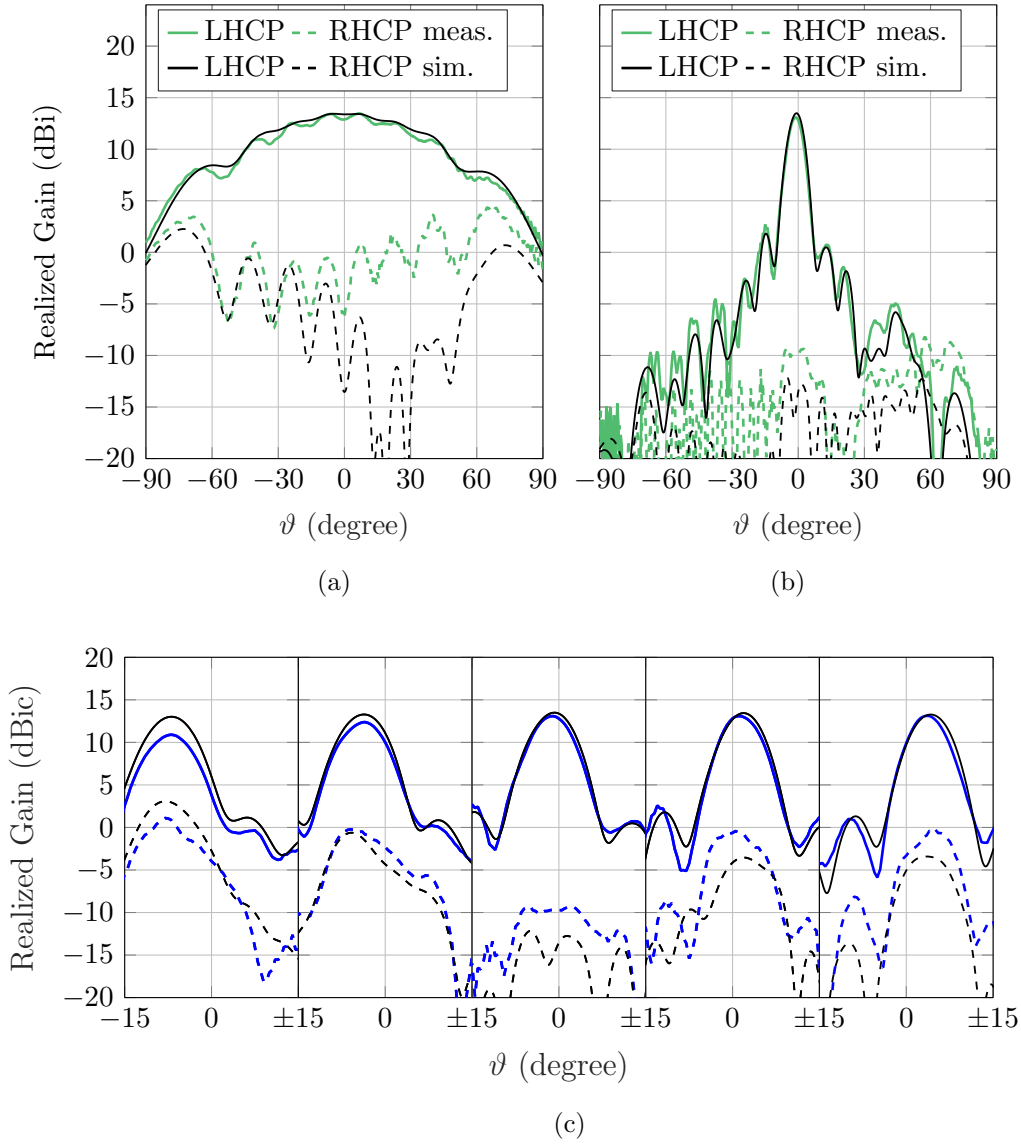


Figure 4.8: Full-wave simulated and measured realized gain of the LA SFP P-LWA with 10 unit cells. (a) Cuts in the principal xz -plane ($\varphi = 0^\circ$) and (b) in the yz -plane ($\varphi = 90^\circ$) at 78 GHz. (c) Radiation patterns in the yz -plane at 75, 76.5, 78, 79.5 and 81 GHz.

4.2 Quad-Polarized Radiation with Parallel P-LWAs

In Section 4.1, it has been shown that a P-LWA radiates an elliptical or circular polarization by tuning the level of the LA. Here, it will be shown that combining two P-LWAs with LA in a mirrored configuration with respect to the longitudinal axis will result in a P-LWA with four ports capable of radiating two circular and two linear polarizations. The concept for this antenna is based on the results published in [121].

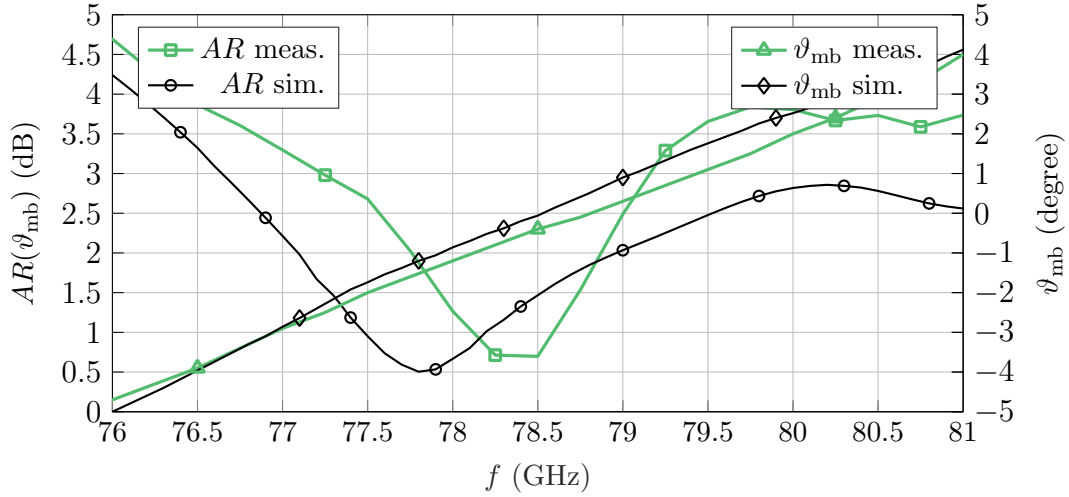


Figure 4.9: Comparison between the simulated and measured axial ratio and main beam direction for the 10 unit cells SFP LA P-LWA shown in Figure 4.3.

4.2.1 Concept

Figure 4.10 illustrates the aforementioned mirrored configuration of two LA SFP P-LWAs with four possible excitation pairs. For simplicity, it is assumed that the P-LWA pair is placed in small proximity to each other and that the coupling between them is negligible. Figure 4.10(a) illustrates a single excitation with the normal of the horizontal patch edge pointing in the $+x$ -direction while terminating the other ports with a matched load. The radiated polarization is LHCP, the same as the P-LWA presented in Section 4.1.5 since the patch orientation in both P-LWAs is identical. In contrast, exciting the top P-LWA [Figure 4.10(b)] will radiate an RHCP wave since the shunt radiation from the horizontal edges is directed in the $-x$ -direction. The scan direction with frequency is maintained for both single excitations since the LW propagates strictly in the $+y$ -direction. Now, suppose both P-LWAs are excited in-phase (even) simultaneously. In this case, the radiation contributions from the horizontal edges of the shunt resonator will cancel each other by symmetry in the far field, as previously proved in Section 4.1.2.2. In fact, this excitation will produce radiation similar to that of a double-symmetric P-LWA in which the produced polarization is longitudinal, as shown in Figure 4.10(c). Finally, Figure 4.10(d) shows the out-phase (odd) simultaneous excitation of both P-LWAs, which results in the cancellation of the vertical edge contributions from the series resonator in the far field. Therefore, the polarization, in this case, is transverse. Interestingly, similar to their single excitation counterparts, the simultaneous excitations do not affect the main beam scanning with frequency. Thus, these four excitation pairs can produce two orthogonal circular and two orthogonal linear polarizations while the main beam scans by changing the frequency.

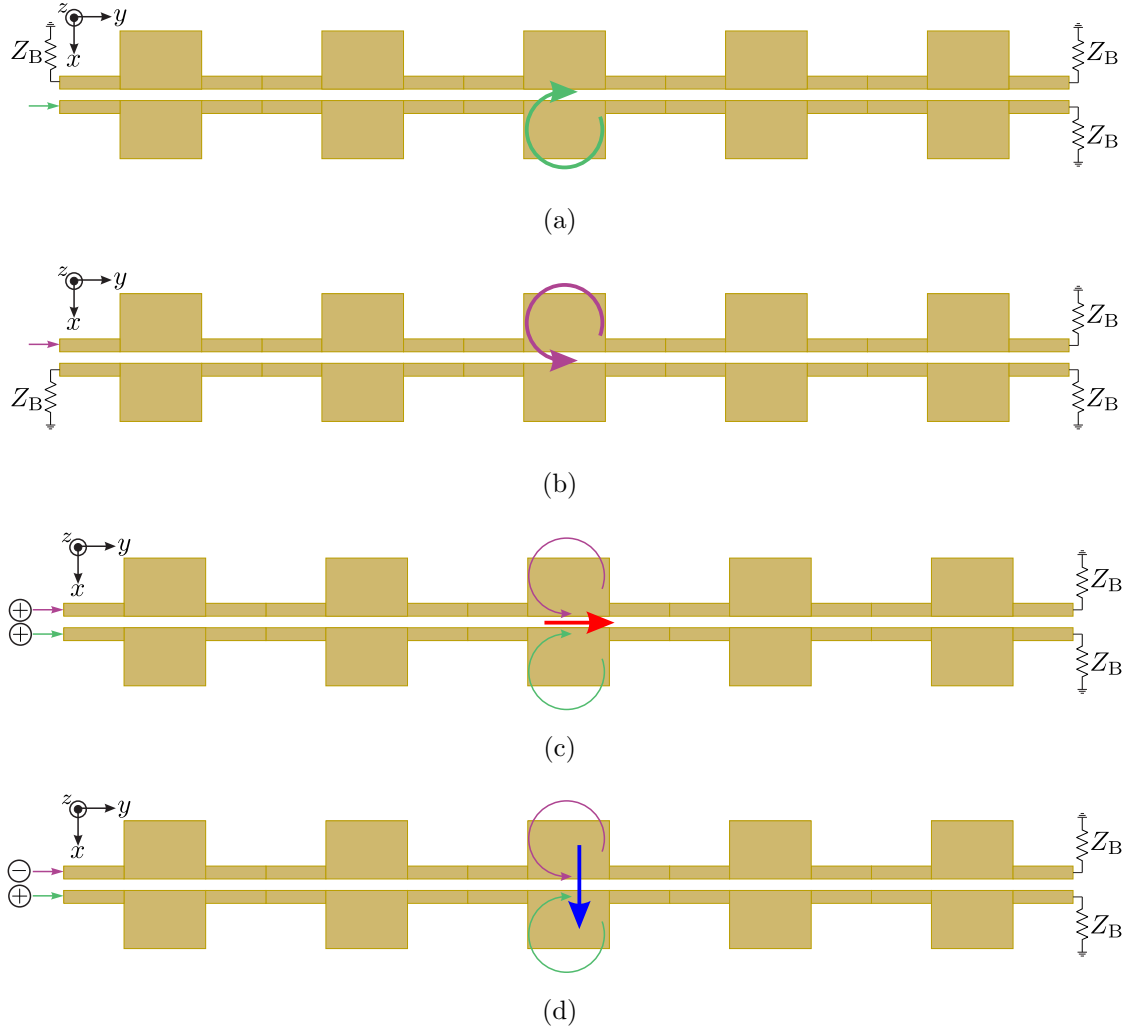


Figure 4.10: Layout of the proposed quad-polarized parallel P-LWAs, consisting of two parallel circularly-polarized series-fed patch P-LWAs with different excitation pairs. (a) Single excitation of the bottom P-LWA with LHCP. (b) Single excitation of the top with RHCP. (c) Even excitation of both P-LWAs with vertical LP (VLP). (d) Odd excitation of both P-LWAs with horizontal LP (HLP).

4.2.2 Edge Radiation from Parallel P-LWAs

The polarization produced by the excitation pairs will be proved theoretically by inspecting the equivalent magnetic surface current densities, as has been done in Section 4.1.1.

4.2.2.1 Single Excitation: Circular Polarization

Under the assumption that the coupling between both P-LWAs can be neglected, the derivation for a single P-LWA with LA is identical to that of a single P-LWA in Section 4.1.3 and the phase/amplitude relations in Section 4.1.4.

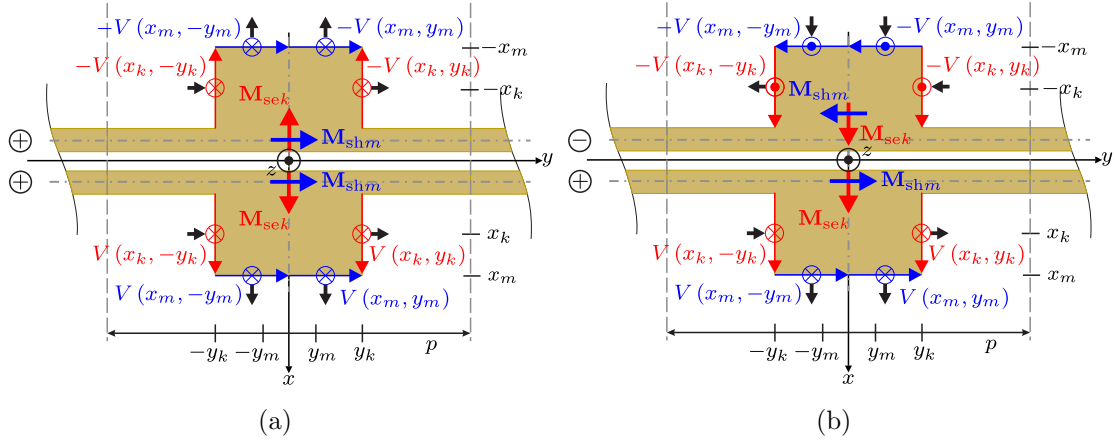


Figure 4.11: Conductor layout of parallel SFP P-LWAs unit cells with voltage sampling at the radiation edges corresponding to the series and shunt radiations for (a) even and (b) odd excitations.

4.2.2.2 Even Excitation: Longitudinal Polarization

Figure 4.11(a) shows the even excitation of the unit cell of the proposed parallel SFP P-LWAs and their corresponding equivalent surface magnetic densities for the series ($\mathbf{M}_{se}$) and shunt ($\mathbf{M}_{sh}$) radiation contributions. The equivalent magnetic surface density of the series radiation, $\mathbf{M}_{se}^{even}$, can be written by evaluating the voltage sampling using Equation 4.2 at the vertical edges and by calculating Equation 4.1 with the normal vectors to the vertical edges, i.e., $\hat{n} = \pm\hat{y}$, as

$$\mathbf{M}_{se}^{even} \cdot \hat{x} = \frac{1}{h} \left[\underbrace{-V(-x_k, -y_k) + V(-x_k, y_k) - V(x_k, -y_k) + V(x_k, y_k)}_{V_{se k}^{even}} \right]. \quad (4.19)$$

Similar to the derivation of Section 4.1.2, the assumption of transversal symmetry and losslessness is made here also. Hence, the equivalent edge voltages can be further simplified using,

$$V(x_k, y_k) = V(-x_k, y_k), \quad (4.20a)$$

$$V(x_k, -y_k) = V(-x_k, -y_k), \quad (4.20b)$$

to

$$V_{se k}^{even} = 2[-V(x_k, -y_k) + V(x_k, y_k)]. \quad (4.21)$$

Next, the equivalent edge voltage of the shunt radiation is derived. Analogously, the equivalent magnetic surface density of the shunt radiation, $\mathbf{M}_{sh}^{even}$, can be calculated using Equations 4.2 and 4.1 with the normal vectors to the horizontal edges, i.e., $\hat{n} = \pm\hat{x}$, as

$$\mathbf{M}_{sh m}^{even} \cdot \hat{y} = \frac{1}{h} \left[\underbrace{-V(-x_m, -y_m) - V(-x_m, y_m) + V(x_m, -y_m) + V(x_m, y_m)}_{V_{sh m}^{even}} \right]. \quad (4.22)$$

By utilizing the same procedure for Equation 4.20, the equivalent edge voltages can be written as

$$V(-x_m, -y_m) = V(x_m, -y_m), \quad (4.23a)$$

$$V(-x_m, y_m) = V(x_m, y_m), \quad (4.23b)$$

which after inserting them in Equation 4.22, the total equivalent edge voltage of the shunt radiation yields

$$V_{shm}^{\text{even}} = 0 \quad (4.24)$$

Equations 4.21 and 4.24 clearly indicate that the polarization's longitudinal (horizontal) component exists while the transverse (vertical) polarization component vanishes, since $V_{sek}^{\text{even}} \neq 0$ and $V_{shm}^{\text{even}} = 0$, respectively. This fact confirms the predicted operation mentioned in Section 4.2.1. Furthermore, the even excitation exhibits the same radiation characteristic, as for P-LWAs with DS unit cells and the same propagation perplexities, as it will be later shown in Section 4.2.3.

4.2.2.3 Odd Excitation: Transversal Polarization

The proof of the polarization for the odd excitation can be followed in the same manner, as in Section 4.2.2.2. Figure 4.11(b) shows the conductor layout of the parallel SFP P-LWA pair with odd excitation and its corresponding equivalent magnetic surface densities. First, using Equations 4.2 and 4.1 with the normal vectors to the vertical edges (index k), i.e., $\hat{n} = \pm\hat{y}$, the equivalent magnetic surface density $\mathbf{M}_{sek}^{\text{odd}}$ can be written as

$$\mathbf{M}_{sek}^{\text{odd}} \cdot \hat{x} = \frac{1}{h} \left[\underbrace{-V(-x_k, -y_k) - V(-x_k, y_k) + V(x_k, -y_k) + V(x_k, y_k)}_{V_{sek}^{\text{odd}}} \right]. \quad (4.25)$$

By symmetry and losslessness, the vertical edges on the left and right sides of the patch can be simplified to

$$V(x_k, y_k) = V(-x_k, y_k), \quad (4.26a)$$

$$V(x_k, -y_k) = V(-x_k, -y_k). \quad (4.26b)$$

By inserting Equation 4.26 into Equation 4.25, the equivalent edge voltage from the series radiation reads

$$V_{sek}^{\text{odd}} = 0 \quad (4.27)$$

In a similar fashion, the equivalent edge voltage from the shunt radiation can be derived by firstly evaluating Equation 4.1 with the normal vectors to the horizontal edges (index m), i.e., $\hat{n} = \pm\hat{x}$, as follows

$$\mathbf{M}_{shm}^{\text{odd}} \cdot \hat{y} = \frac{1}{h} \left[\underbrace{-V(-x_m, -y_m) + V(-x_m, y_m) - V(x_m, -y_m) + V(x_m, y_m)}_{V_{shm}^{\text{odd}}} \right]. \quad (4.28)$$

Then applying the symmetry and losslessness conditions to the top and bottom patch edges [Figure 4.11(b)], as

$$V(-x_m, -y_m) = V(x_m, -y_m), \quad (4.29a)$$

$$V(-x_m, y_m) = V(x_m, y_m), \quad (4.29b)$$

which allows the derivation of the total equivalent edge voltage from the shunt radiation for the odd excitation, which yields

$$V_{shm}^{\text{odd}} = 2[-V(x_m, -y_m) + V(x_m, y_m)]. \quad (4.30)$$

Equations 4.27 and 4.30 show that the polarization is purely directed in the longitudinal direction, i.e., $\hat{x}$, since $V_{sek}^{\text{odd}} = 0$ and $V_{shm}^{\text{odd}} \neq 0$. It is important to note that the linear polarization purity obtained from the even or odd excitations at broadside is irrespective of the CP of the single LA P-LWA. Nevertheless, a perfect CP ensures that the amplitudes of the produced linear polarizations corresponding to the even or odd excitations are equal to each other.

4.2.2.4 Other Operation Modes

The even and odd excitations considered in the previous sections assume that both P-LWA pairs are excited from one end, while the other end is either extended to infinity or terminated with a matched load. Other excitation configurations may be considered, such as exciting the top P-LWA from the left ($-y$ -direction) and the bottom P-LWA from the right ($+y$ -direction). However, this type of excitation configuration has two drawbacks. First, the opposite excitation will generate two oppositely directed propagating waves, producing a standing wave that omits the main beam scanning feature with frequency. Second, the opposite excitation might be impractical due to the need for long feeding lines, which cause parasitic radiation and higher loss, especially at higher frequencies. Therefore, such excitation configurations will not be considered in this work.

4.2.3 Full-Wave and Experimental Validation

In this section, the concept of the quad-polarized P-LWA will be verified numerically and experimentally. Figure 4.12(a) shows a 3D view of the proposed P-LWA consisting of two mirrored SFP antennas with respect to the longitudinal axis. The SFP P-LWA is the same as in Section 4.1.5, which operates at the broadside frequency $f_{bs} = 78$ GHz. Furthermore, the four ports of the P-LWA are connected to the WR12-to-MSL transitions [Appendix B.1].

4.2.3.1 Unit-Cell Simulation

The unit cell dimensions are identical to the LA SFP P-LWA, which are listed in Table 4.1. For the parallel LA P-LWA pair, the spacing between them [Figure 4.12(b)], s , is optimized such that the coupling between the P-LWAs and the effect on the CP is minimum. The coupling between the P-LWA pairs is reflected in the frequency shift for the series and shunt resonators. This frequency shift is different for the even and odd excitations, as the former couples inductively and the latter capacitively.

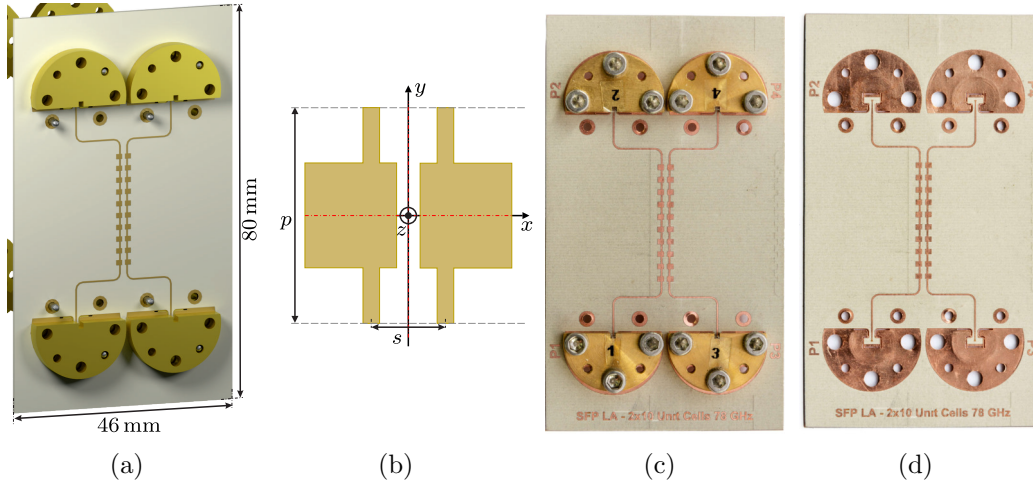


Figure 4.12: Quad-polarized SFP P-LWA with LA. (a) Perspective view of the antenna, (b) zoomed view of the unit cell, (c) and (d) photographs of the fabricated P-LWA SFP antennas, without and with transition plates, respectively. The SFP P-LWA pair is separated by the spacing s .

Figure 4.13 shows the full-wave simulated Bloch impedance and propagation constants for even and odd excitations for an antenna spacing $s = 1.5$ mm. For the line width $w_1 = 0.3$ mm and the set separation, the characteristic impedances for the even and odd modes of the coupled lines are $Z_c^{\text{even}} = 26.4 \Omega$ and $Z_c^{\text{odd}} = 107 \Omega$, respectively. These values are also reflected in the Bloch impedance, as shown in Figure 4.13(a). Furthermore, the Bloch impedance of the odd excitation shows a slight variation over frequency, indicating a non-perfectly balanced unit cell, which is caused by the coupling between the P-LWA pair. Additionally, Figure 4.13(b) shows the leakage factor, α , over frequency, with equal magnitude for both even and odd excitations. In contrast, the phase constant of the odd excitation, β^{odd} , has a transition frequency that is shifted to 78.4 GHz. Nevertheless, the phase constants show that the beam scanning with frequency is still possible since no bandgaps (standing wave) are visible in the dispersion diagram. Finally, the scanning rate given by the slope of the phase constant is the same for both even and odd excitations and also the same for single port excitation, as shown in Figure 4.4(b).

4.2.3.2 10 Unit Cells Structure

The fabricated prototype is shown in Figure 4.12(d) and was fabricated using laser structuring, as described in Section 4.1.5.2. Due to the size of the WR12 flange, the antenna size is 46 mm $\times$ 80 mm. The S-parameters measurement was performed using a two-port network set up so that only two ports are measured simultaneously, and the other ports are connected to a matched load.¹ Accordingly, the measurement is repeated for the other ports, which requires six measurements to obtain the full 4×4 S-matrix. Figure 4.14 shows the measured S-parameters for port 1 compared

¹The measurement setup utilizes two E-band mm-wave frequency extenders from Virginia Diodes Inc. [cf. Section 4.1.5.2]. Simultaneous measurement is not possible due to the hardware limitation, which it would require four mm-wave frequency extenders.

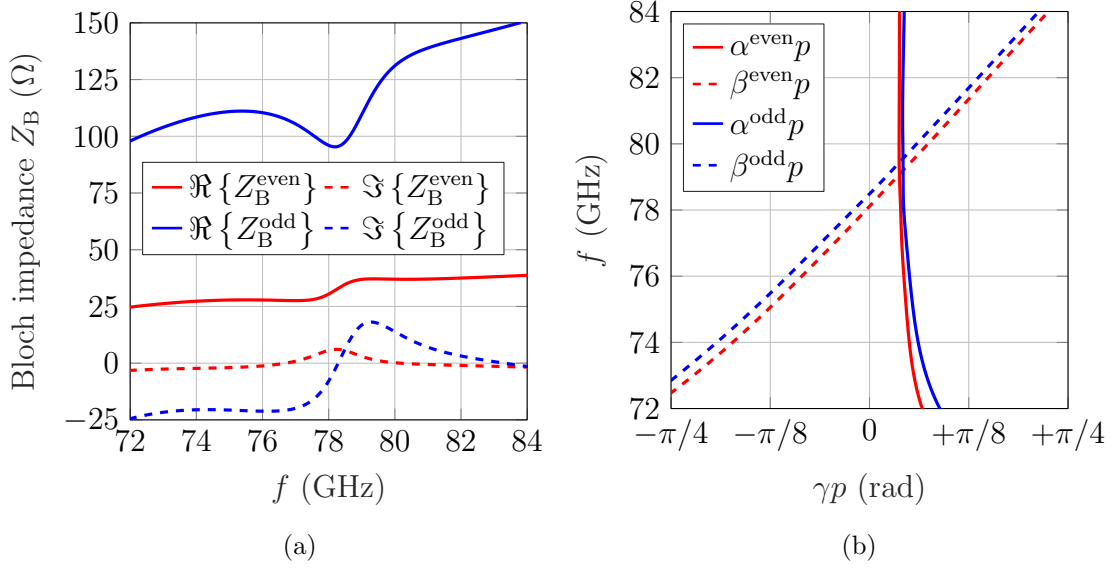


Figure 4.13: Full-wave simulated unit cell of the quad-polarized SFP P-LWA pair shown in Figure 4.12 for both even and odd excitations. (a) Bloch impedance and (b) propagation constant.

to the simulation. It can be seen that the return loss is better than 10 dB for the frequency bandwidth 76 – 81 GHz. Furthermore, a good agreement can be observed between the simulation and measurement for the transmission coefficients S_{21} , S_{31} , and S_{41} . Additionally, the coupling to port 2 and port 4 is relatively higher than for port 3, $S_{21} \approx -20$ dB, $S_{41} \approx -25$ dB, and $S_{31} \approx -40$ dB, respectively. However, these coupling levels will not affect the operation of the P-LWA pair since port 2 and port 4 are permanently terminated with a matched load. In contrast, port 3 is essential for the even and odd excitations. It is worth noting that the coupling levels will be much higher if the antenna is directly connected to a radar transceiver, as the WR12-flanges and the extra feeding lines/bends are omitted. Finally, the other S-parameter sets (for port 2, port 3, and port 4 excitations) are not shown due to symmetry and reciprocity.

Figure 4.15 plots the far-field radiation results obtained from full-wave simulations and measurements for the four operation modes of the LA P-LWA pair at the center frequency 78 GHz. Figure 4.15(a) shows the radiation pattern for a single port excitation from port 1. For this excitation, the polarization is RHCP, which corresponds to the operation illustrated in Figure 4.10(b). Furthermore, the realized gain and the cross-polarization discrimination are about 12.3 dBi and 11.77 dB, respectively. The reason for the decreased gain and increased cross-polarization discrimination or axial ratio are extra feeding line length and the increased substrate width that affected the transversal (horizontal, HLP) polarization, respectively. Figure 4.15(b) shows the result for port 3 excitation that is almost identical to port 1 except for the flipped polarization, as predicted in Figure 4.10(a). Figure 4.15(c) shows the far-field results for the even excitation. It can be observed that the radiation pattern is linearly polarized in the longitudinal (vertical, VLP) direction, whereas Figure 4.15(d) shows the radiation patterns for the odd excitation with horizontal polarization.

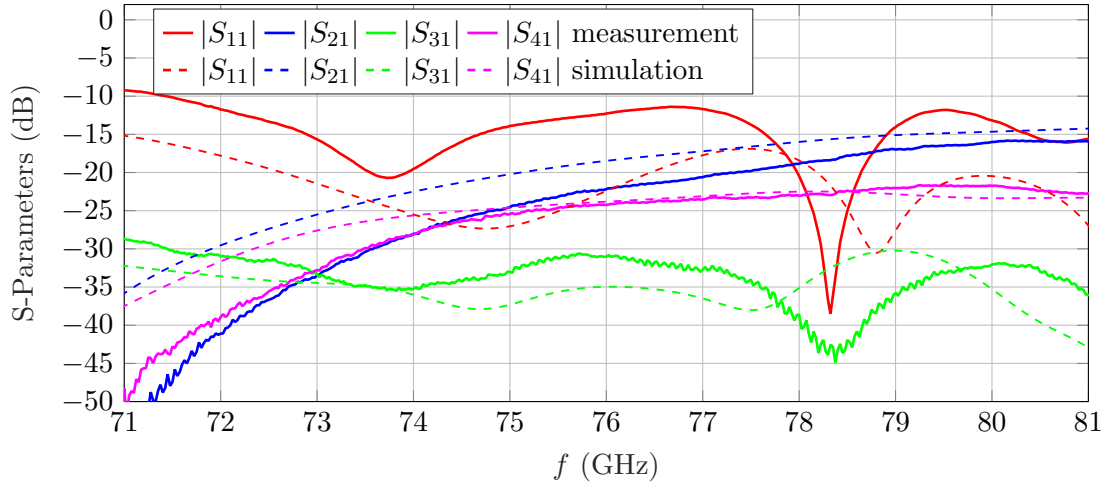


Figure 4.14: Full-wave simulation and measurement comparison of the quad-polarized P-LWA shown in Figure 4.12.

The cross-polarization discrimination obtained from the even and odd excitations measurement is better than 35 dB in the broadside direction. In contrast, the simulated cross-polarization discrimination is better than 80 dB (not visible in plots). Moreover, the odd and even excitations are calculated experimentally by three steps. First, the far-field patterns for port 1 and port 3 are separately measured. Second, the phase of the far-field patterns is corrected due to the fixed separation (20 mm) of the waveguide ports. Third, the measured patterns are then superimposed in-phase (even) and out-of-phase (odd) to obtain the combined radiation patterns.

4.3 Dual Linearly-Polarized Radiation with Double Asymmetric P-LWA

This section introduces the last asymmetry configuration that combines the asymmetries in Section 3.4 (LA) and Section 3.5 (TA), as shown in Figure 4.16. It will be shown that this double asymmetric (DA) configuration combines properties inherited from both LA and TA, where LA provides high efficiency in both the series and shunt resonators. At the same time, TA controls the series and shunt power ratios at the broadside.

The results in this section are based on and taken from [26, 122].

4.3.1 Concept

From the transformer circuit model for the TA unit cell presented in Section 3.5 and shown in Figure 3.8, it was shown that the degree of TA can be modeled solely by the transformation ratio T , which controls the level of coupling between the series and shunt resonators. Additionally, the Bloch impedance at the broadside frequency ($\omega - \omega_{bs} = 0$) as seen from opposite directions, i.e., forward (+) and backward (−) directions, are proportional to the corresponding T -dependent Bloch impedance [Equation 3.50]. Moreover, the shunt-to-series power ratios at broadside

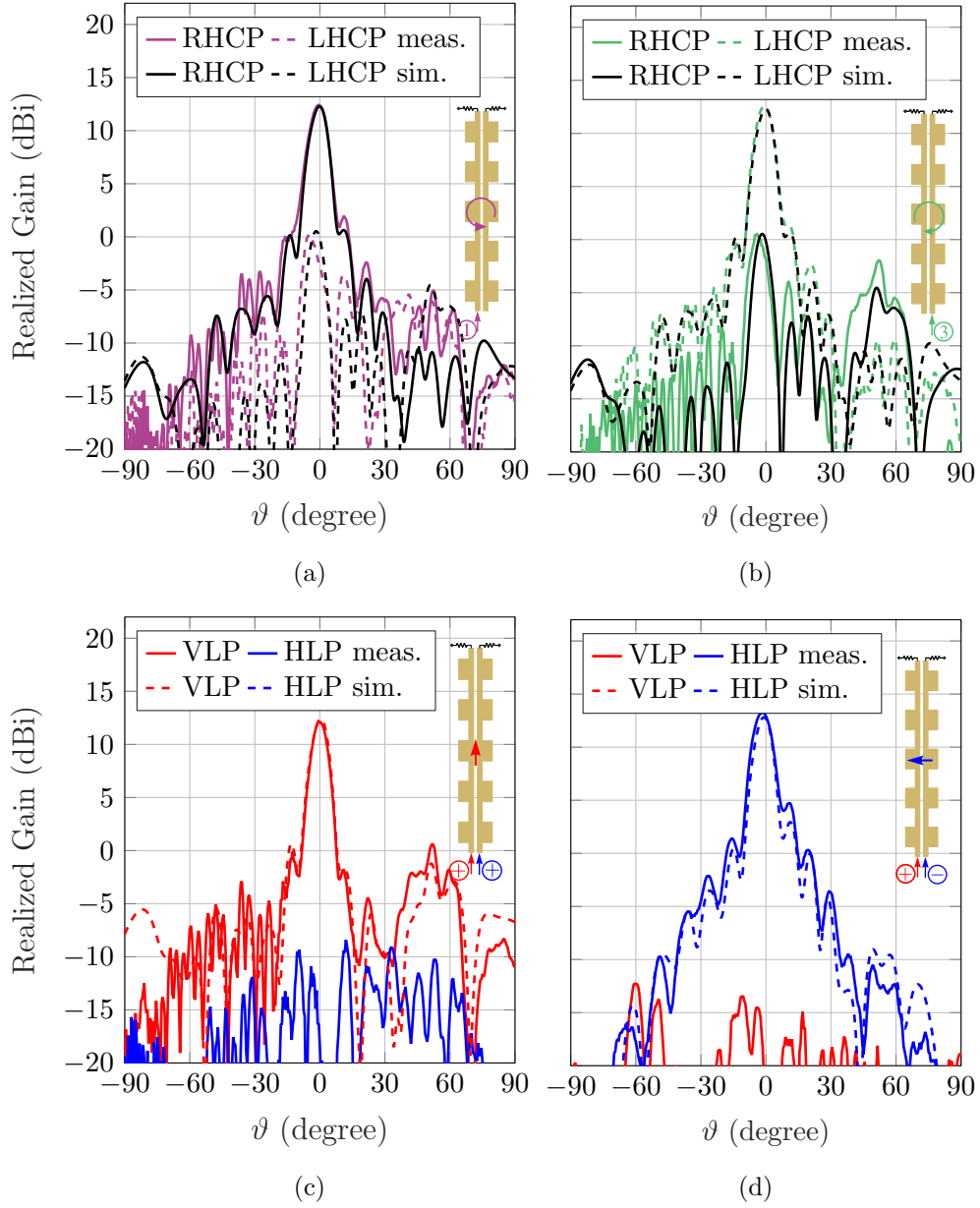


Figure 4.15: Full-wave simulated and measured realized gain in the yz -plane ($\varphi = 90^\circ$) for the four operation modes of the parallel SFP P-LWA pair with LA consisting of 10 unit cells. (a) Port 1 excitation with RHCP, (b) port 3 excitation with LHCP, (c) even excitation with VLP, and (d) odd excitation with HLP. All results are for the broadside frequency 78 GHz.

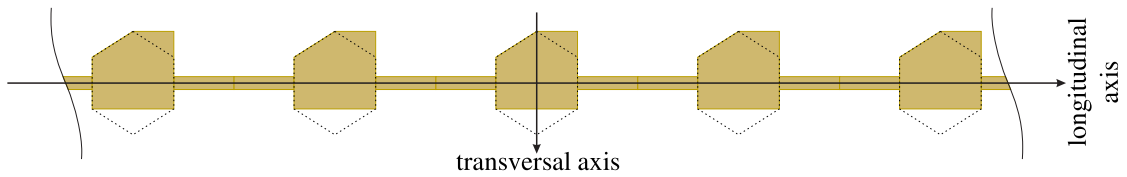


Figure 4.16: Generic representation of the double-asymmetric (DA) P-LWA, with its unit cell asymmetric with respect to both the transversal and longitudinal axes.

correspond directly to the squared Bloch impedance, and therefore, they depend on T . These direction-dependent power ratios can be written as

$$\left(\frac{P_{\text{sh}}}{P_{\text{se}}}\right)_{\text{bs}}^{\pm} = \frac{G}{R} \left| \frac{V_{\text{sh}}}{I_{\text{se}}} \right|^2 = \frac{G}{R} |Z_{\text{B,bs}}(T)^{\pm}|^2. \quad (4.31)$$

where for a transversally-symmetric unit cell, i.e., T is unity, or equivalently $\tau = 0$, then Equation 3.50 yields $Z_{\text{B,bs}}^{\pm}(T = 1) = \pm\sqrt{R/G}$, and therefore, the power equally splits between the series and shunt resonators, corresponding to the DS or LA cases. Furthermore, Equation 4.31 shows that the forward and backward Bloch impedances control the corresponding power ratios, which will ultimately determine the corresponding radiation efficiencies given as in Equation 3.51.

Since a P-LWA with LA has high efficiency, then the DA P-LWA will have high efficiency as well irrespective of whether the unit cell is transversally symmetric or asymmetric. In fact, TA does not affect the intrinsic series and shunt resonators, but only their coupling, as has been discussed in Section 3.5. Concurrently, the Bloch impedance strongly depends on the direction of propagation for high TA ($T \gg 1$), as seen from Figure 4.17(a). Moreover, the power splitting provided by the forward and backward Bloch impedances is shown in Figure 4.17(b). By setting the limits for a high degree of TA, i.e., ($T \gg 1$), the direction-dependent Bloch impedance yields

$$Z_{\text{B,bs}}^{+}|_{T \gg 1} \rightarrow 0, \quad (4.32a)$$

$$Z_{\text{B,bs}}^{-}|_{T \gg 1} \rightarrow \infty. \quad (4.32b)$$

Now, by inserting Equation 4.32 into Equation 3.51, the radiation efficiencies of the forward (η_{bs}^{+}) and backward (η_{bs}^{-}) propagating waves for extreme degree asymmetry, i.e., $T \gg 1$ can be written as

$$\eta_{\text{bs}}^{+}|_{T \gg 1} \approx \eta_{\text{se}}, \quad (4.33a)$$

$$\eta_{\text{bs}}^{-}|_{T \gg 1} \approx \eta_{\text{sh}}, \quad (4.33b)$$

which reveals that exciting a DA P-LWA with high TA corresponds to selectively exciting the series resonator in the forward direction and the shunt resonator in the backward direction. Moreover, from LA, both resonators can reach maximal efficiency (100 % in the absence of dissipation loss), which means that both ports can achieve maximal efficiency. Finally, since the series and shunt resonators provide orthogonal field radiation, as discussed in Section 4.1, the polarizations from the two ports will be orthogonal to each other. Specifically, the forward excitation will generate longitudinal (vertical) polarization, while the polarization from the backward excitation will be transverse (horizontal).

4.3.2 Full-Wave Simulation

The concept of radiating two orthogonal linear polarizations with DA P-LWA is demonstrated in this section by full-wave simulations of two P-LWA examples with

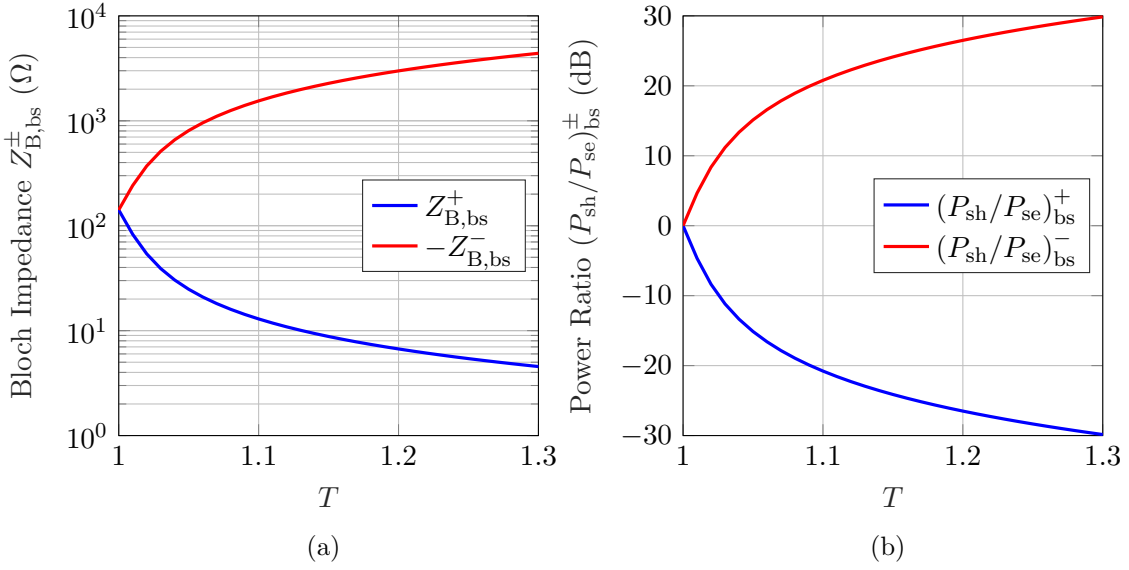


Figure 4.17: Numerically calculated (a) Bloch impedance [Equation 3.50] and (b) series-to-shunt power ratio [Equation 4.31] at broadside versus the degree of TA, T , for $R = 10 \Omega$ and $G = 500 \mu S$.

different geometrical parameters, a series-fed patch (SFP) P-LWA and a series-fed coupled patch (SFCP) P-LWA. Both are designed for a transition frequency of 24 GHz.

4.3.2.1 Series-Fed Patch (SFP) P-LWA

The SFP P-LWA with DA is shown in Figure 4.18. The P-LWA is structured on a single-layer Rogers RO4350 substrate with height $h = 508 \mu m$ [112]. The dimensions of the unit cell are listed in Table 4.2.

Three design steps are performed to achieve the direction-dependent orthogonal polarizations. First, starting from the DS configuration, the unit cell is optimized to satisfy the ω -balancing condition, $\omega_{se} = \omega_{sh} = \omega_{bs} = 24 \text{ GHz}$, by tuning the length parameters ℓ_1 (feeding line length) and ℓ_p (patch length), as has been done in Section 4.1.5.1. Second, Q -balancing is achieved by tuning the patch shift in the transverse direction ($\hat{x}$), d , which leads to an optimal LA with circular polarization. It is important to note that the parameter d does not influence the transversal asymmetry and therefore does not affect T . Finally, TA, and thereby realizing DA, is introduced by changing the widths of the feeding lines, w_{l1} and w_{l2} , different from each other. The frequency balancing condition is typically altered through these steps, which are iteratively repeated until the design is frequency balanced. The design is considered complete when the cross-polarization discrimination level, controlled by tuning T , is sufficiently high and when the broadside directivities at the two ports are sufficiently close, which is the former is ensured by high TA, i.e., $T \gg 1$ and the latter by Q -balancing intrinsic series and shunt resonators.

Figure 4.19 plots the backward (Z_B^-) and forward (Z_B^+) Bloch impedances for $d = 1.2 \text{ mm}$ or $T = 1$. The forward Bloch impedance Z_B^+ exhibits a low impedance at broadside, which indicates the behavior of a series resonator connected in the

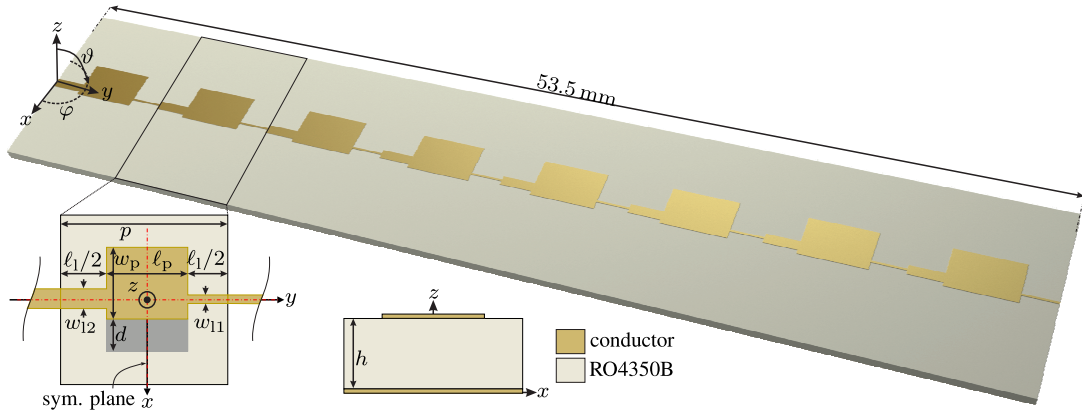


Figure 4.18: Series-fed patch (SFP) DA P-LWA with the unit cell dimensions shown in the inset. Symmetry with respect to the longitudinal axis (y -axis) is broken by shifting the patch by a quantity d , while symmetry with respect to the transverse axis (x -axis) is broken by changing the feeding lines widths $w_{11} \neq w_{12}$.

Table 4.2: Dimensions in mm of the DA SFP P-LWA shown in Figure 4.18.

ℓ_1	ℓ_p	p	w_{11}	w_{12}	w_p	d
3.41	3.28	6.69	0.10	0.50	2.10	0.60

series path, while Z_B^- exhibits a high impedance corresponding to a shunt resonator connected in the shunt path, as understood from Equation 3.22 with $R \rightarrow 0$ and $G \rightarrow 0$, respectively. Equation 4.31 shows and Figure 4.17(b) illustrates that the different Bloch impedances lead to unequal series-to-shunt power splitting, i.e., $P_{se} \neq P_{sh}$ at broadside [Figure 4.17(b)]. Specifically, for $T \gg 1$, $Z_B^+ \rightarrow 0$ [Equation 4.32a], and therefore $P_{sh} \rightarrow 0$ [Equation 4.33a], which means that only the series resonator is excited. Conversely, again for $T \gg 1$, $Z_B^- \rightarrow \infty$ [Equation 4.32b], and therefore $P_{se} \rightarrow 0$ [Equation 4.33b], which means that only the shunt resonator is excited.

To provide more insight into the operation of the DA P-LWA, Figure 4.20 plots the electric surface current density, J , on the top metallization layer of the SFP P-LWA at the time instants $t = 0$, $t = T/4$, $t = T/2$, and $t = 3T/4$, where T represents here the time period of the wave. First, Figure 4.20(a) shows the case of the forward excitation. In this case, the surface currents are approximately zero at the time instances $t = 0$ and $t = T/2$. In contrast, they are nonzero and directed in the longitudinal direction at the long edges of the patch at the time instances $t = T/4$ and $t = 3T/4$, which corresponds to transversal polarization as discussed in Section 4.3.1. Next, Figure 4.20(b) shows the case of backward excitation. In this case, the surface currents are approximately zero at $t = T/4$ and $t = 3T/4$, whereas they are nonzero and directed in the transverse direction at the short edges of the patch at the time instances $t = 0$ and $t = T/2$, which corresponds to longitudinal polarization. Interestingly, the quadrature relation between the series and shunt resonators introduced in Section 4.1.4.1 can be clearly observed by comparing the electric surface current densities J_{se} in Figure 4.20(a) to J_{sh} in Figure 4.20(b), since the series and shunt surface currents are excited with a quarter-time period delay, i.e., 90° .

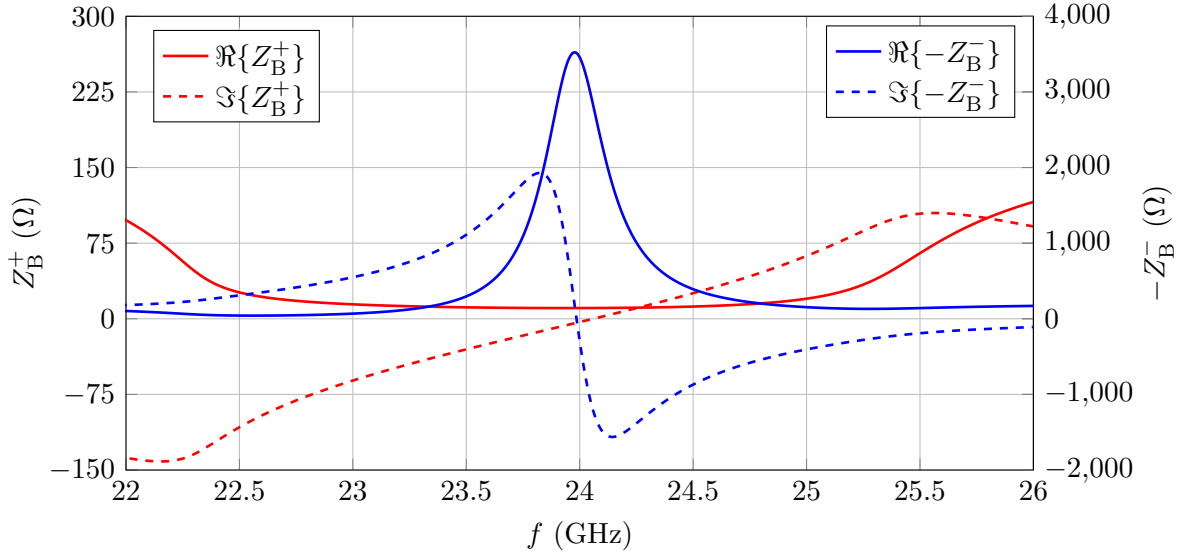


Figure 4.19: Bloch impedance of the forward and backward propagating waves, Z_B^+ and Z_B^- , respectively, for $T \approx 1.3$. Z_B^+ exhibits the behavior of a series resonator (low impedance) and Z_B^- of a shunt resonator (high impedance).

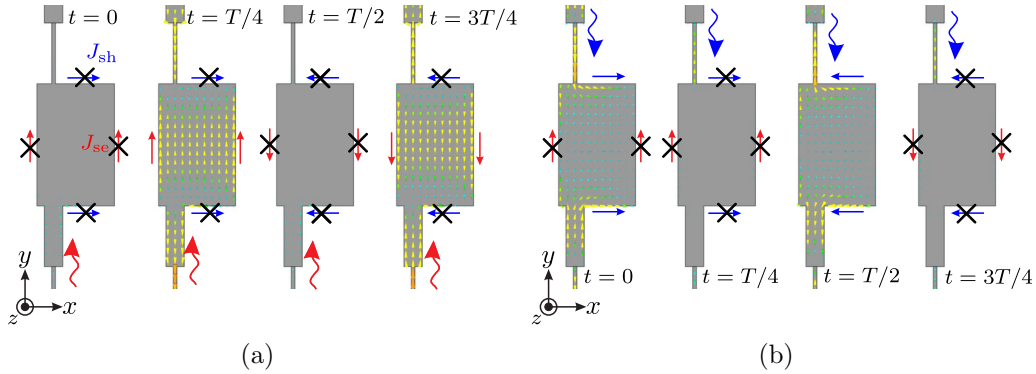


Figure 4.20: Full-wave simulated electric surface current density (J) on the top metallization of the DA SFP LWA unit cell. (a) Forward excitation corresponding to the field distribution of the series mode for $t = T/4$ and $t = 3T/4$ and (b) backward excitation corresponds to the field distribution of the shunt mode for $t = 0$ and $t = T/2$.

The broadside directivity dependency on the level of TA (T) is shown in Figure 4.21(a) for 8 unit cells of the proposed DA SFP P-LWA, for the P-LWA being excited in the forward direction ($+\hat{y}$), i.e., port 1, and in the backward direction ($-\hat{y}$), i.e., port 2. For $T = 1$, the unit cell is longitudinally asymmetric and transversally symmetric, which is confirmed by the equal amplitudes, circular polarization, of the vertical (VLP) and horizontal (HLP) far-field components for both ports [Section 4.1]. By progressively increasing T , the series-to-shunt power ratio is no longer unity, as shown in Figure 4.17(b), which is achieved by varying the widths of the transmission lines ($w_{11} \neq w_{12}$), while maintaining the double balancing condition, i.e., $\omega_{se} = \omega_{sh}$ and $Q_{se} = Q_{sh}$. Thus, the amplitudes of the linear far-field com-

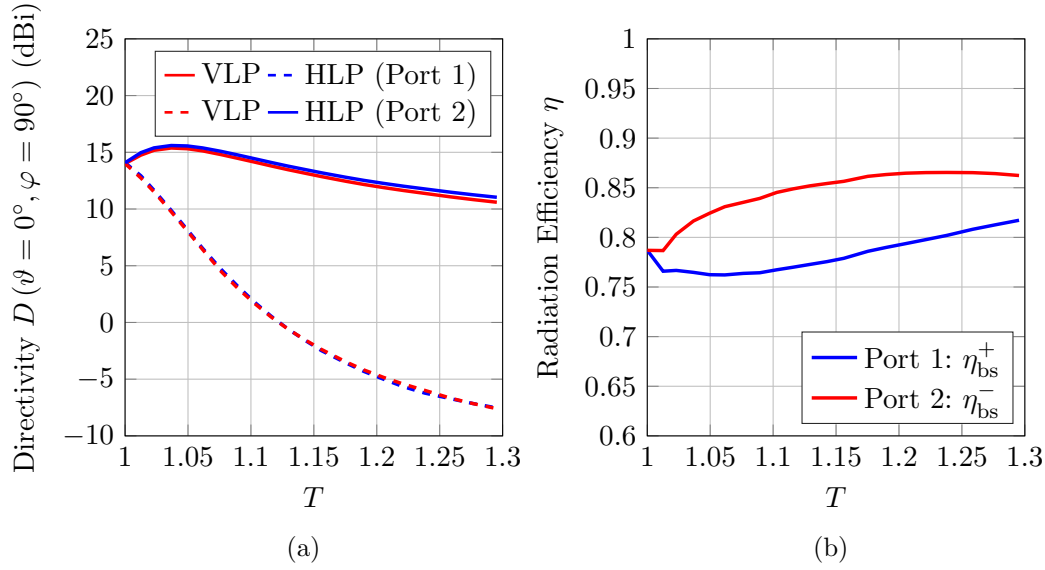


Figure 4.21: Radiation parameters of the SFP P-LWA shown in Figure 4.18. (a) Broadside directivity (at $\vartheta = 0^\circ$ and $\varphi = 90^\circ$) and (b) two-ports radiation efficiencies versus of degree of transversal asymmetry (T) obtained by full-wave simulation for 8 unit cells. In the case $T = 1$, the unit cell is transversally symmetric and therefore elliptically polarized and for high degree of TA, i.e., $T \gg 1$, the radiation efficiencies in the forward and backward directions correspond to the efficiencies of the series η_{se} and shunt resonators η_{sh} , respectively.

ponents will not be equal. This fact can be observed in Figure 4.21(a), where the circular polarization transforms into elliptic polarization with progressively increasing ellipticity until the polarization becomes essentially linear. Therefore, the degree of TA directly controls the cross-polarization level, with higher T corresponding to lower cross-polarization. Additionally, at $T \approx 1.3$, the antenna's polarization for port 1 excitation is vertical (transverse), corresponding to the series radiation, while for port 2 the polarization is horizontal (transversal), corresponding to the shunt radiation, with the cross-polarization discrimination reaching about 19 dB. When increasing the level of TA T , the directivities seen from both ports decrease due to the increase of the leakage factor α at broadside, which reduces in the effective aperture of the P-LWA. This fact is also observed in the vanishing zeros and sidelobes in the radiation patterns caused by the high exponential decay rate of the antenna aperture [65].

Figure 4.21(b) plots the broadside radiation efficiencies η_{bs}^+ and η_{bs}^- for the forward and backward excitations, respectively. For $T = 1$ (LA), the two radiation efficiencies are equal, as is always the case when the unit cell is transversally symmetric. Furthermore, the two radiation efficiencies increase for increasing transversal asymmetry. For $T \approx 1.3$ the radiation efficiencies are $\eta_{bs}^+ = 86\%$ and $\eta_{bs}^- = 81\%$. Finally, Figure 4.22 shows the radiation patterns of the DA SFP P-LWA in the principal yz -plane ($\varphi = 90^\circ$) at 24 GHz, which confirms a good level of the cross-polarization discrimination over a wide range of radiation angles for both forward [Figure 4.22(a)] and backward [Figure 4.22(b)] directivities.

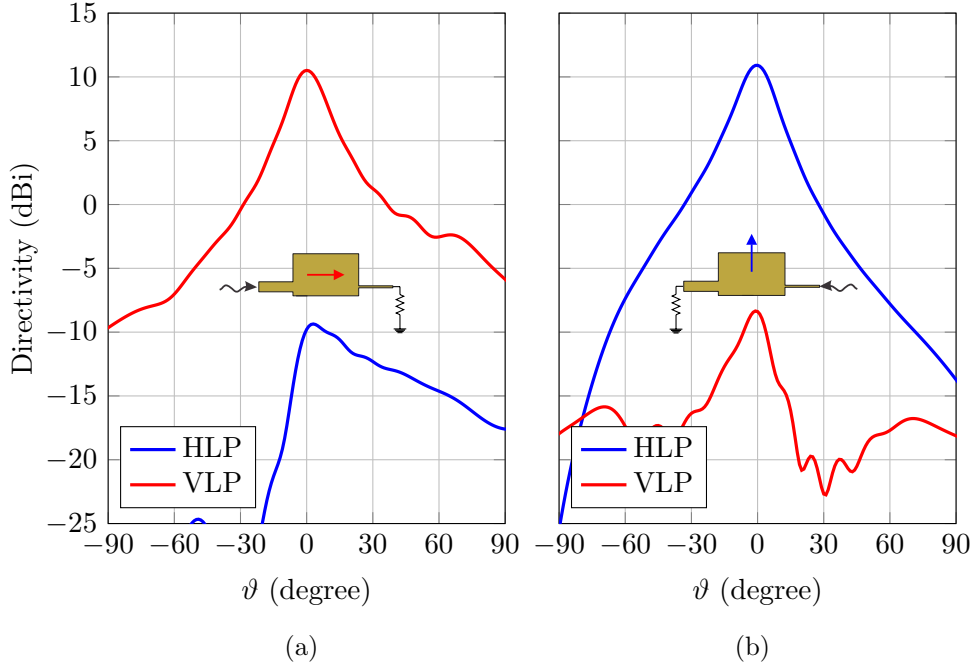


Figure 4.22: Full-wave simulated directivity (8 unit cells as in Figure 4.21(a)) for $\varphi = 90^\circ$. (a) Longitudinal (horizontal) and (b) transversal (vertical) polarizations corresponding to the series and shunt radiation mechanisms, respectively.

The Bloch impedances in Figure 4.19 show a high variation over frequency. The electric surface current densities in Figure 4.20 illustrate an apparent standing wave operation, caused by the high level of TA, i.e., $T \gg 1$. As mentioned in Section 3.5, the TA affects the broadside operation of the P-LWA; therefore, the series-to-shunt power splitting decreases with diverting frequency from broadside until the power splitting becomes equal to the off-broadside frequency, i.e., $\omega - \omega_{\text{off-bs}} \rightarrow \infty$. For this reason, the proposed DA P-LWA design is only intended to operate at the broadside frequency.

4.3.2.2 Series-Fed Coupled Patch (SFCP) P-LWA

The SFCP P-LWA radiates in the $\mathbf{n} = -1$ space harmonic and its layout is shown in Figure 4.23. The unit cell of the SFCP P-LWA consists of three metallization layers, a patch on the top layer, feeding lines with a gap in the center on the middle layer, and a ground plane on the bottom layer. The unit cell is essentially based on the proximity-coupled patch antenna design reported in [123] and later extended to a P-LWA in [82]. The structure includes two Rogers RO4350 substrate laminates with heights $h_1 = 254 \mu\text{m}$ and $h_3 = 168 \mu\text{m}$ [112], respectively. Furthermore, the two substrates are stacked together with a Rogers RO4450F prepreg laminate of height $h_2 = 101 \mu\text{m}$ [124]. The design procedure follows the same steps as the DA SFP P-LWA in Section 4.3.2.1. The line length ℓ_1 and patch length ℓ_p are tuned to obtain the ω -balancing condition $\omega_{\text{se}} = \omega_{\text{sh}} = \omega_{\text{bs}}$, and the Q -balancing condition is obtained by shifting the patch on the top metallization layer by a quantity w_s with respect to the longitudinal axis. In contrast to the SFP unit cell, the TA here is

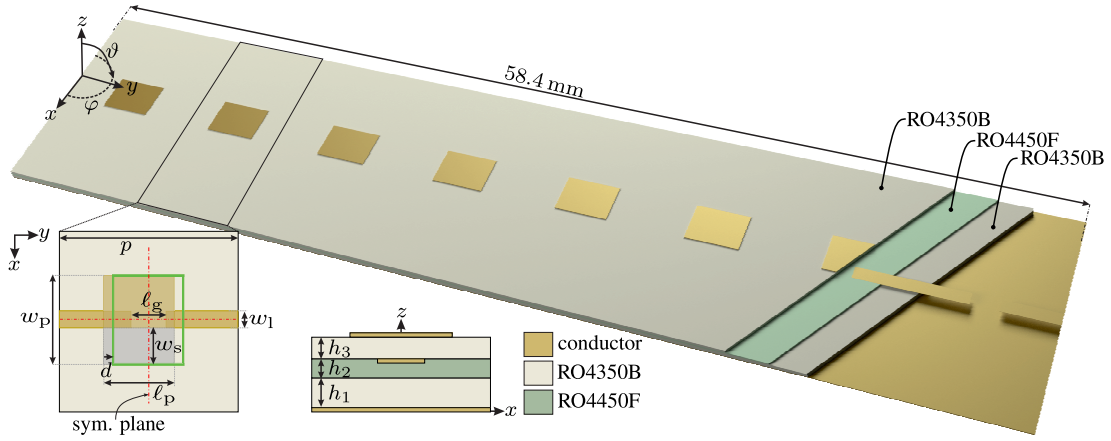


Figure 4.23: Series-fed coupled patch (SFCP) double asymmetric P-LWA with the unit cell dimensions shown in the inset. Symmetry with respect to the longitudinal axis is broken by vertically cutting out the patch by a quantity w_s , while symmetry with respect to the transverse axis has been broken by shifting the patch in the longitudinal direction by a quantity d .

Table 4.3: Dimensions in mm of the DA CSFP P-LWA shown in Figure 4.23.

l_1	l_g	p	l_p	d	w_1	w_p	w_s
2.93	1.44	7.30	2.95	0.65	0.70	3.60	1.45

obtained by shifting the patch in the longitudinal direction by a quantity d , which has the same effect on the Bloch impedance and directivity at broadside as the variation of the parameters w_{11} and w_{12} in the SFP P-LWA of Section 4.3.2.1. The final dimensions of the SFCP unit cell are listed in Table 4.3.

Figure 4.24 plots the forward and backward directivities for the DA SFCP P-LWA with 8 unit cells obtained by full-wave simulation. The results are qualitatively identical to those obtained in Figure 4.22 for the DA SFP P-LWA. The achieved cross-polarization discrimination is about 12 dB, as shown for the forward [Figure 4.24(a)] and backward [Figure 4.24(b)] excitations.

4.3.3 Matching Network

The proposed DA P-LWA operates at the broadside frequency. Hence, a matching network design is presented to operate at 24 GHz. Furthermore, the matching network is designed only for the DA SFP P-LWA, presented in Section 4.3.2.1, which will be experimentally demonstrated in the next section.

The forward and backward Bloch impedances of the SFP P-LWA [Figure 4.19] are purely real at broadside with very low impedance seen from port 1 and very high impedance seen from port 2. So, a simple quarter-wavelength transmission-line transformer was not appropriate for matching. Instead, the matching is accomplished using open-ended stubs, as shown in Figure 4.25(a).

The first step of the design is to transform the normalized impedance seen from the backward direction z_B^- , with $z_B^- = Z_B^- / Z_{B,bs}^{\text{sym}}$, into the normalized impedance seen from the forward direction z_B^+ . The symmetric and asymmetric Bloch impedances

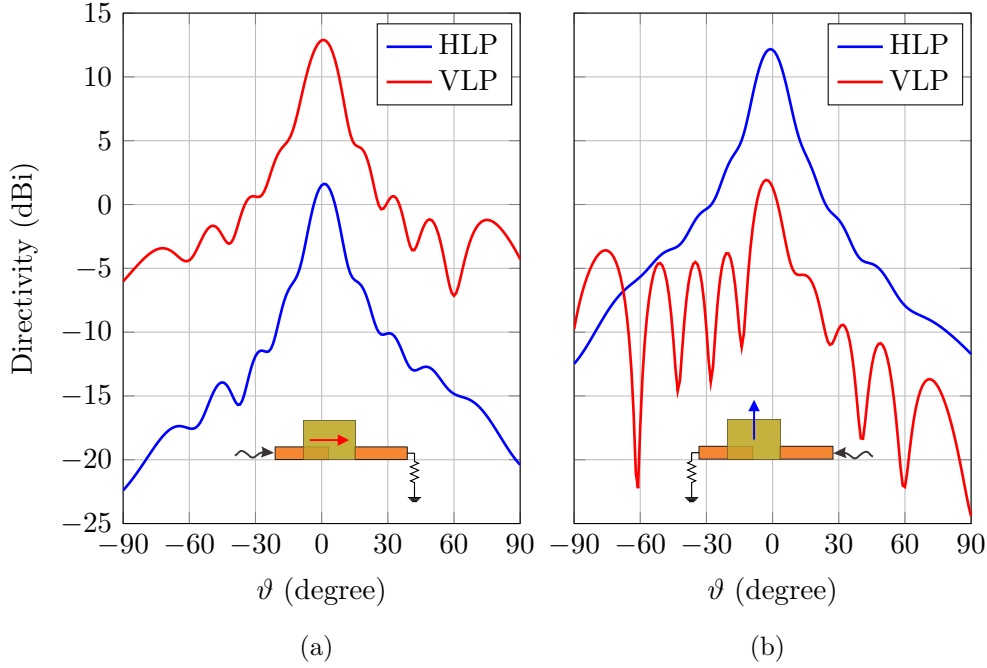


Figure 4.24: Full-wave simulated directivity of the DA SFCP P-LWA shown in Figure 4.23 with 8 unit cells for $\varphi = 90^\circ$. (a) Longitudinal (horizontal) and (b) transverse (vertical) polarizations corresponding to the series and shunt radiation mechanisms, respectively.

are related by $Z_{B,bs}^{\text{sym}} = \sqrt{Z_B^+ Z_B^-}$ [Equation 3.13], which is the same for a quarter-wavelength transformer [70]. Therefore, a quarter-wavelength transformer with $Z_B^{\text{sym}} = 130 \Omega$ is appropriate to transform z_B^- to z_B^+ , as shown in the Smith chart in Figure 4.25(b). Since both impedances are equalized, an identical matching network section may be used for both ports. Therefore, two sections are selected and calculated to match the ports to 50Ω . First, a 75Ω transmission line section is used. Then, a double-sided open-ended stub with 53Ω impedance and length of 0.75 mm is added. Finally, the double-sided stub [Figure 4.25(a)] is used to reduce the effect of parasitic radiation on the overall radiation pattern of the DA P-LWA.

4.3.4 Measurement Results

The fabricated prototype of the DA SFP P-LWA is shown in Figure 4.26. The feeding networks are placed at the back of the structure and connected to the antenna with metallized via transitions. Two 2.92 mm -coax end-launch connectors from Southwest Microwave, Inc. (type 1092-04A-6) are used to feed the SFP P-LWA.

The measured and full-wave simulated S-parameters are compared in Figure 4.27. The results verify that good matching is achieved at the design frequency around 24 GHz . The slight deviations in the experiment are attributed to fabrication tolerances. Furthermore, the low level of the transmission coefficient, S_{21} , clearly indicates a high radiation level and also contributed to the standing wave operation of the DA P-LWA. Additionally, a relative frequency bandwidth of the P-LWA can be evalu-

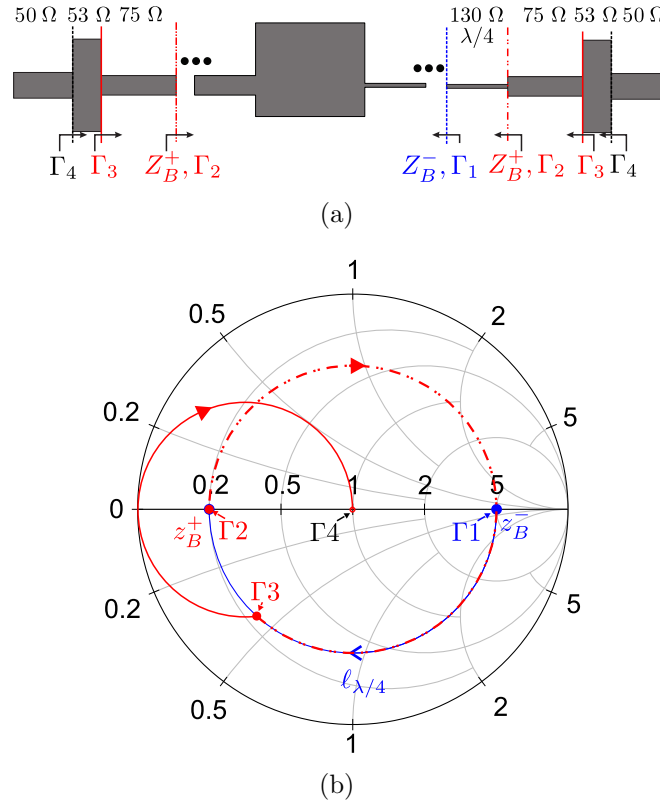


Figure 4.25: Matching network for the SFP P-LWA. (a) Geometry and parameters of the forward and backward matching networks. (b) Smith Chart showing the path of impedance transformations.

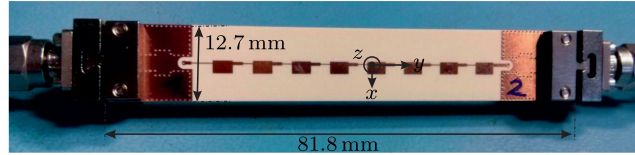


Figure 4.26: Photograph of the fabricated prototype connected to end-launch connectors. The antenna is fed by the matching network at the back by a via transition.

ated from the reflection coefficients, S_{11} and S_{22} , which reads approximately 1.3 % for both.

Figure 4.28 plots the measured and full-simulated realized gain for the forward excitation at port 1 [Figure 4.28(a)] and for the backward excitation at port 2 [Figure 4.28(b)], while port 2 and port 1 are terminated with a match load, respectively. The realized gains of port 1 and port 2 at broadside are 8.5 dBi, and 9.5 dBi, and the cross-polarization discrimination levels are 13.75 dB and 14 dB, respectively. Excellent agreement is observed between the simulation and measurement results, both confirming the dual linear-polarization operation of the P-LWA.

Although the concept of the DA P-LWA provides an interesting method to achieve dual-linearly polarized radiation, it faces two major drawbacks that are related to the high leakage factor, α . First, the antenna's gain cannot be increased by increasing the length of the antenna, i.e., cascading more unit cells, because of the low power left

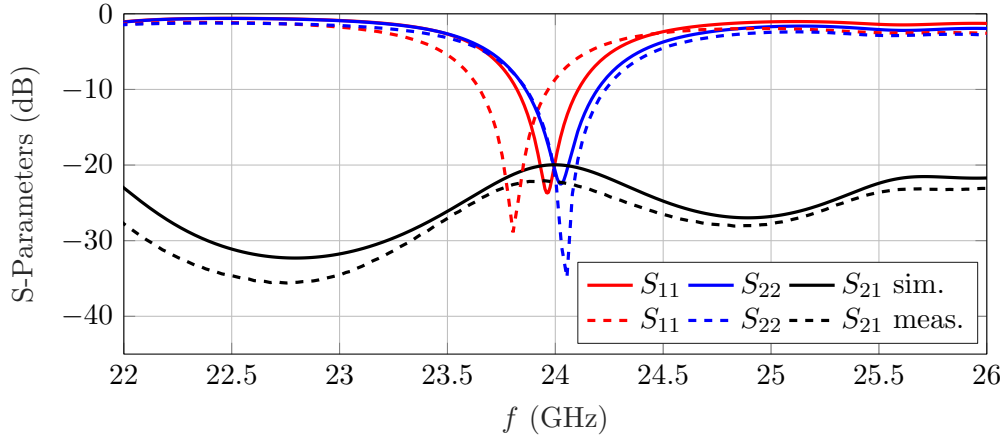


Figure 4.27: Measured and full-wave simulated S-parameters of the DA SFP P-LWA with 8 unit cells. S_{12} is not shown due to reciprocity, i.e., $S_{12} = S_{21}$.

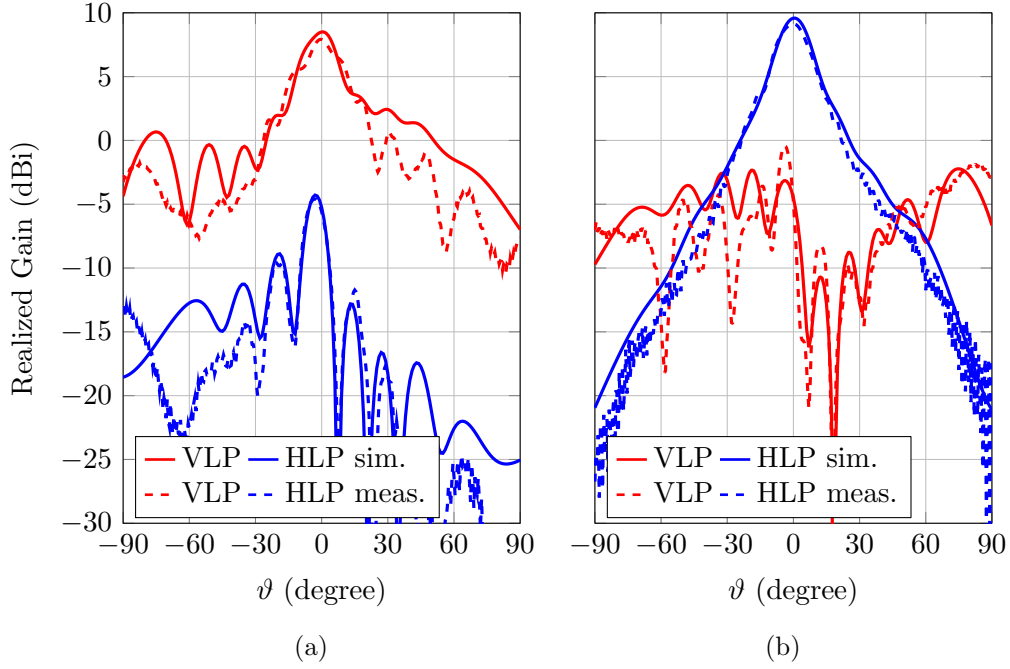


Figure 4.28: Measured and full-wave simulated realized gain of the DA SFP P-LWA with 8 unit cells in the principal yz -plane ($\varphi = 90^\circ$) at 24 GHz. (a) Excitation at port 1 and (b) excitation at Port 2.

at the end of the structure. Second, the relative frequency bandwidth is relatively low, which may not suit specific applications. However, these drawbacks may be overcome by designing a unit cell with a low leakage factor.

4.4 Summary

This chapter provided three different approaches to achieve polarization diversity in P-LWAs. First, it has been shown that the series and shunt radiation contributions

can be equalized by introducing the proper level of *longitudinal asymmetry* (LA). Furthermore, the phase relation between the series and shunt resonators was found to be in quadrature, which is confirmed by the formulas obtained from the generic network model of P-LWAs, shown in Section 3.4.3. The circular polarization in LA P-LWA is numerically and experimentally demonstrated by a series-fed patch (SFP) P-LWA operating at the center frequency 78 GHz. Second, the LA P-LWA concept is then extended by adding an identical LA P-LWA, which is mirrored with respect to its axis. This parallel pair can be fed from its two ports at one end with four excitation configurations, providing dual circular (RHCP and LHCP) and dual linear (vertical and horizontal) polarizations. Again, this concept is verified by the SFP P-LWA, operating at 78 GHz. Finally, *longitudinal asymmetry* (LA) and *transversal asymmetry* (TA), forming a *double asymmetry* (DA), are combined to achieve a direction-dependent orthogonal linear polarization, where LA provides circular polarization and TA allows the selective excitation of either the series or shunt resonator depending on the excitation directions. Two case studies are presented, an SFP and a series-fed coupled patch (SFCP) P-LWA, operating at 24 GHz to verify the DA operation in P-LWAs.

Frequency-Scanning Concepts

In the previous chapters, it has been shown theoretically and analytically that the open-stopband problem at the broadside frequency can be solved by utilizing the symmetry of the unit cell. Furthermore, it was shown that symmetry also constitutes the polarization of P-LWAs. This chapter investigates frequency-scanning concepts for 1D and 2D planar P-LWAs to improve the scanning sensitivity over frequency and to unlock 2D space scanning, respectively. First, methods will be introduced to enhance the dispersion and applied them to two 1D P-LWAs, scanning in one plane. Second, these methods will be used with the series-feeding network (SFN) presented in Section 3.5.3.2, to realize 2D P-LWAs capable of frequency scanning in two planes.

5.1 One-Dimensional Frequency-Scanning Concepts

In many practical applications, the available frequency bandwidth is usually limited by hardware front-end and regulatory requirements. Therefore, controlling the scanning rate over the frequency of the P-LWA's main beam is essential. Frequency scanning is an inherent property of P-LWAs, which solely depends on the $\omega - \beta$ relation. The phase constant, β , is a product of the medium in which the traveling wave is propagating. Changing the medium's properties, e.g., its permittivity, as described in Section 2.4, allows controlling the dispersion diagram of P-LWAs. However, this approach is not always practical since changing the dielectric constant comes with the drawback of efficiency reduction [65]. Furthermore, the fundamental space harmonic (β_0) in P-LWA can be dispersive due to periodic perturbations. In this section, the wave dispersion will be exploited to enhance the frequency scanning in P-LWAs.

5.1.1 Scanning-Rate Enhancement

Generally, two methods may be defined to control the frequency scanning. First, are meandered lines, which control the period of the propagating wave, p , independently of the spacing between the radiating elements, ℓ . Second are dispersive lines, which allow controlling the $\omega - \beta$ slope of the fundamental space harmonic. The first method will be considered only theoretically in this section and will be implemented later in a P-LWA, whereas the second method will be validated here by two structures at 24 GHz and 78 GHz by full-wave simulations and measurements, respectively.

5.1.1.1 Serpentine Lines

Figure 5.1(a) shows the concept of using serpentine lines in P-LWAs. By increasing the period of the guided wave, p , by a multiple of n , where $n \in \mathbb{N}$, the slope of the dispersion curve can be increased. In a typical P-LWA, the period increase, ($p = \ell$), would usually lead to a large spacing between elements and produces grating lobes. This is equivalent to a shift of the radiating space harmonic inside the radiation region ($|\beta_n| < k_0$). To remedy this problem, ℓ has to be decreased by employing *serpentine* or meandered lines in the unit cell, which would break the relation $p = \ell$ [125]. This technique dates back to the 1960s when first implemented using microstrip lines [126, 127]. Due to the high parasitic radiation of the line bends, different waveguide implementations were proposed [29, 30, 128–130] and, just recently, SIWs [131, 132]. The frequency-scanning law in Equation 2.23 considers $p = \ell$. Hence, for the sake of generality, the equation is modified here, which accounts for $p \neq \ell$, as follows

$$\vartheta_{\text{mb},n}(\omega) = \arcsin \left[\frac{\beta_n(\omega) p}{k_0 \ell} \right]. \quad (5.1)$$

For the case of an unperturbed transmission line, $\beta_0/k_0 = \sqrt{\epsilon_r}$.¹ From Equation 5.1, it becomes evident that the ratio p/ℓ controls the slope of the main beam scanning. Furthermore, the condition $p = n\lambda_g$ must be satisfied to maintain the same broadside frequency f_{bs} as for the case $p = \ell$. Moreover, grating lobes may be avoided by meeting the condition $\ell < \lambda_0/2$.

Figure 5.1(b) plots an analytical dispersion diagram of a TEM propagating wave in a P-LWA with a unit cell's period $p = n\lambda_g$, $n = 1$ to 5, and the guided wavelength $\lambda_g = \lambda_0/\sqrt{\epsilon_r}$, with $\epsilon_r = 3$. The spacing between the radiating elements is set to $\ell = \lambda_0$. The analytical results show that the slope of β_n is proportional to n . Furthermore, it can be seen that increasing the period shifts higher space harmonics into the radiation region resulting in $p = n\lambda_g$ into radiation from the space harmonic $\mathbf{n} = -n$ at the same broadside frequency $f_{\text{bs}} \forall n$. Finally, the main beam direction can be calculated by evaluating Equation 5.1, which is plotted in Figure 5.1(c). Again, it can be seen that the increase in the period length enhances the scanning rate. In fact, evaluating the scanning-rate expression introduced in Equation 2.27 shows that the scanning rate increase with n . In this illustrative example, the scanning range also increased, and the frequency bandwidth required to cover the entire upper hemisphere decreased from about 60 % to about 20 % relative frequency bandwidth. It is worth noting that the spacing, ℓ , may be theoretically further reduced to achieve a steeper light line ($k_0 \ell$), but a P-LWA with smaller ℓ would require more unit cells to achieve the same directivity as for large ℓ . Although the method of serpentine lines is simple and very effective, it comes with two significant disadvantages. First, the loss in a single unit cell increases with a larger period by $\exp(-\alpha p)$. Second, the P-LWA increase in size in the transversal direction limits the feasibility of realizing planar 2D P-LWA. Nevertheless, this method will be applied in Section 5.2 to an SFN to construct a 2D P-LWA capable of scanning multiple planes with a pencil beam.

¹ ϵ_r is the effective permittivity of the medium for unshielded transmission lines [70].

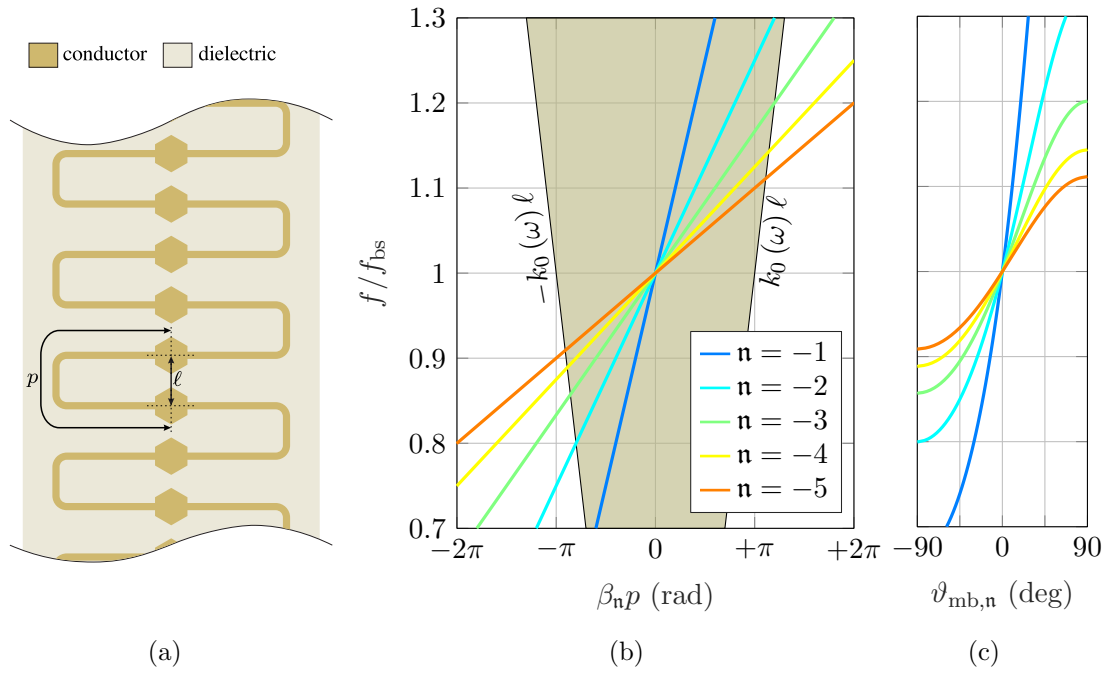


Figure 5.1: Scanning rate enhancement with serpentine lines. (a) Top view of a generic layout of a P-LWA with serpentine lines, (b) the dispersion diagram for $p = n\lambda_g$ with $n = 1$ to 5 corresponding to a shift of the space harmonics $\mathbf{n} = -n$ along the horizontal axis, and (c) the corresponding main beam scanning direction.

5.1.1.2 Dispersive Lines

The dispersion diagram of P-LWAs usually exhibits a dispersive behavior, i.e., non-linear phase constant over frequency, due to the periodic perturbation within the structure. Moreover, the group velocity [Equation 2.20] in the case of a dispersive line is not constant over frequency, $v_g(\omega)$. Additionally, non-TEM transmission line types, e.g., TE or TM modes in metallic waveguides, are inherently dispersive. Generally, the dispersion curve becomes nonlinear as a function of frequency in the vicinity of stopbands or cutoff frequencies, where usually these stopbands are controlled by reactive loading (capacitive or inductive) of the unit cell. The increased dispersive behavior can be exploited to increase the scanning rate of a P-LWA. Multiple P-LWAs with unit cell loading designs have been reported [133–137]. In general, the techniques of transmission line loading are adapted from filter designs, where high dispersion occurs at the edge of the filter bands. In this chapter, two novel implementations of P-LWA will be presented in Section 5.1.3 with enhanced scanning rate.

5.1.2 Scanning-Range Enhancement

The methods presented in the previous section allow the enhancement of the scanning rate. On the other hand, the scanning range does not always increase accordingly since, for example, in the case of line loading, the stopbands move inside the radiation region ($k_0 > |\beta_n|$), limiting the scanning range of the main beam. Further-

more, the available scanning range might not always be applicable to applications with a specific frequency bandwidth. Hence, another technique may be used to increase the scanning range further. Since 1D P-LWAs have two ports, the opposite port can be excited, resulting in a wave propagating in the backward (−) direction, which radiates a beam scanning in the opposite direction. Therefore, employing the two ports allows for doubling the scanning range within the same frequency bandwidth. One problem when exciting the two ports is that the port must be excited sequentially to avoid radiating two beams. Additionally, the open-stopband closure at the broadside does not allow the opposite excitation to radiate at the broadside, as shown in Section 3.5. Therefore, the broadside frequency can be avoided since it provides a redundant beam from the opposite side.

The grating lobes also limit the scanning range in P-LWAs. Other methods than the ones mentioned here and in Section 5.1.1 can be used to manipulate the radiation pattern at the unit cell level (element factor). One method is proposed by adding parasitic radiating elements around the unit cell resulting in a null in the grating lobe's direction, as reported in [138, 139]. Such a method will not be investigated in this work.

5.1.3 Implementation of Scanning-Enhanced Structures

This section demonstrates two P-LWAs with enhanced scanning rates. The First P-LWA is based on transmission loading and operates at 24 GHz. In contrast, the second structure is based on tuning the cutoff frequency of the SFN in Section 3.5.3.2 with Vivaldi radiation elements operating at 78 GHz.

5.1.3.1 Stub-Loaded SFCP P-LWA

The SFCP P-LWA is similar to that presented in Section 4.3.2.2, except for a *longitudinally-symmetric* unit cell [82]. In addition, the multilayer stack includes the same Rogers RO4450F prepreg laminate of height $h_2 = 101 \mu\text{m}$ [124], as for the structure in Section 4.3.2.2, whereas the height of the two Rogers RO4350 laminates changed to $h_1 = 508 \mu\text{m}$ and $h_3 = 254 \mu\text{m}$ [112].

Figure 5.2 shows the modified SFCP P-LWA with stub-loaded unit cells operating at 24 GHz. The stubs are placed on the middle layer and connected to the feeding lines. Furthermore, the stubs are periodically repeated with the stub period $p_s = w_s + \ell_o$ resulting in a total unit cell's period $p = 2n_s p_s + \ell_g$ with n_s being the number of stubs within a unit cell. Additionally, their resonance frequency is higher than the broadside frequency, which determines the maximum stub's length $\ell_s \ll \lambda_g/2$.² Whereas for $\ell_s < \lambda_g/4$, the impedance of the stub is purely capacitive, with its reactance equal to $-jZ_s \cot(\beta_s \ell_s)$, where Z_s and β_s are the characteristic line impedance and phase constant of the stub, respectively [70]. If the stub period is very small, i.e., $p_s \ll p$, then the phase constant of the P-LWA's unit cell can be calculated as $\beta = \sqrt{L(C + C_s)}$, where L and C are the inductance and capacitance of the unloaded unit cell, respectively. The additional stub capacitance, C_s , serves two purposes. First, the slope of β over ω increases with C_s (or ℓ_s), and second,

²It is worth noting that the stubs do not affect the patch mode due to field cancellation, as they are symmetric with respect to the longitudinal axis ($\hat{x}$).

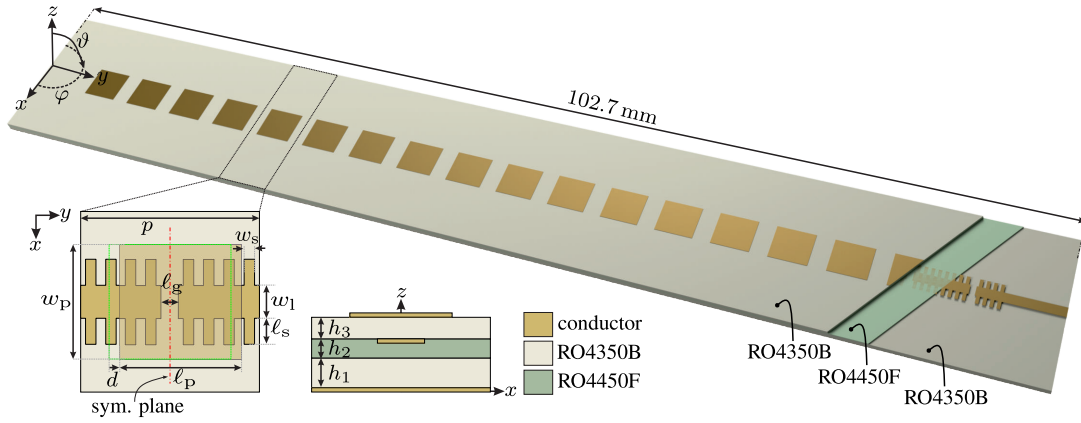


Figure 5.2: SFPP P-LWA structure consisting of 18 stub-loaded unit cells. The insets show a single unit cell with its geometric parameters and the multilayer stackup.

Table 5.1: Dimensions in mm of the stub-loaded CSFP P-LWA shown in Figure 5.2.

unit-cell type	tuned parameters					fixed parameters				
	p_s	ℓ_s	p	ℓ_p	d	n_s	w_s	w_p	w_1	ℓ_g
unloaded	0.759	0.0	6.57	2.87	0.235	4	0.3	3.3	0.95	0.5
loaded	0.575	1.5	5.10	3.57	0.066					

the period of the unit cell, p , decreases. Hence, the scanning rate can be enhanced without facing the problem of grating lobes. Finally, the equalization conditions [Section 3.5.2] can be achieved after setting a particular number of stubs, n_s , and ℓ_s for the broadside radiation as follows: 1) The frequency balancing condition can be obtained by tuning the length of the patch ℓ_p and the stub period w_s while keeping the other parameters fixed, and 2) the Q -balancing condition is obtained by tuning the patch shift, d , along the antenna axis (y).

The effect of the stubs on the scanning rate will be compared to the conventional SFPP P-LWA, i.e., the unloaded P-LWA. Table 5.1 lists the dimensions of the optimized unit cell for the balancing-condition pair at broadside. Figure 5.3 plots the full-wave simulated Bloch impedance and dispersion diagram for the optimized TA loaded and unloaded unit cells. The Bloch impedance [Figure 5.3(a)] shows the reduced frequency bandwidth for the stub-loaded unit cell at $f > f_{bs}$, where the resonance of the open-ended stubs occurs at 25.26 GHz. Furthermore, it can be seen that the off-broadside Bloch impedance of the loaded unit cell is much lower than its unloaded counterpart. The lower impedance is due to the additional capacitance of the stubs, where the off-broadside Bloch impedance can be approximately calculated from the line impedance as $Z_{B, \text{off-bs}} \approx \sqrt{L/(C + C_s)}$. Figure 5.3(b) shows that for $f > 25.26$ GHz, the attenuation constant α increases dramatically, indicating a stopband. Moreover, the phase constant for the loaded unit cell varies more over frequency when compared to the unloaded case near the stopband, which indicates the increased scanning rate. However, the slope of β is similar for both unit cells for $f < f_{bs}$, the scanning rate is increased for the loaded unit cells due to decreased

unit cell's period [Table 5.1]. The shaded blue and red areas in Figure 5.3(b) show the radiation region for the unloaded and loaded unit cells, respectively, where the loaded unit cell has a much smaller radiation region accounted for by the smaller period.

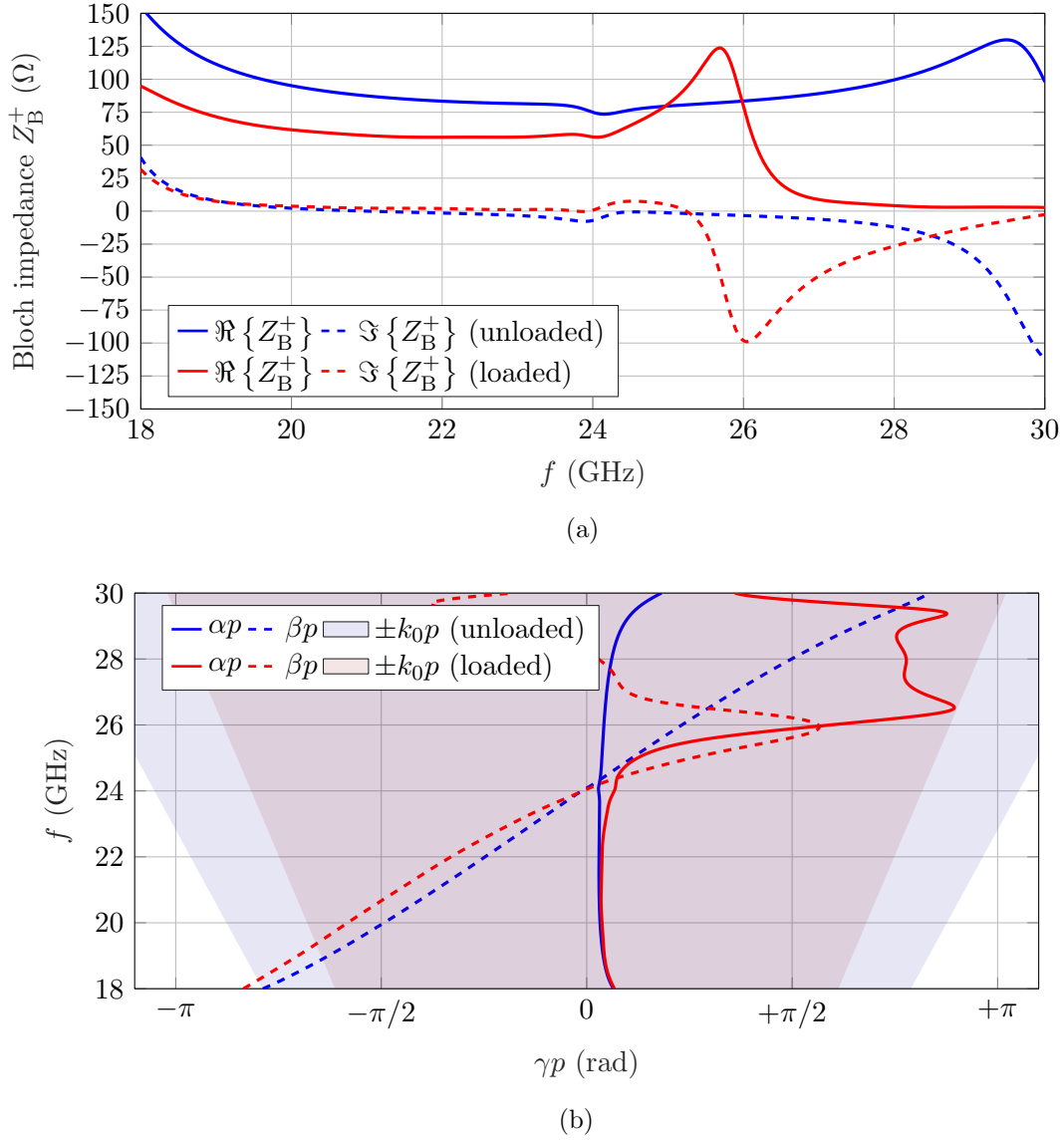


Figure 5.3: Full-wave simulated stub-loaded and unloaded (conventional) SFCP P-LWA unit cells. (a) Bloch impedance and (b) propagation constant with the blue and red shaded areas indicate the radiation regions for the unloaded and loaded unit cells, respectively.

In order to verify the scanning-rate enhancement, two P-LWAs are constructed by cascading 14 unloaded and 18 loaded unit cells, corresponding to a total antenna length of 91.8 mm and 91.9 mm, respectively. Figure 5.4(a) plots the directivity radiation patterns for the loaded (solid lines) and unloaded (dashed lines) structures for the frequency range of 19 – 25 GHz. The plot shows, as expected, that the scanning range is enhanced for the stub-loaded P-LWA. At 19 GHz, the main beams point to

-80° and -47° , and at 24 GHz, they point to -4.3° and 0° , which correspond to scanning rates of $15.14^\circ/\text{GHz}$ and $9.3^\circ/\text{GHz}$ for the loaded and unloaded P-LWAs, respectively. Additionally, it can be noticed that the directivity of the loaded P-LWA is lower when compared to the unloaded P-LWA. The steeper slope of the $\beta - f$ curve or, equivalently, the decrease in the group velocity [Equation 2.20] explains the increasing losses caused by the increased group delay. The increased losses lead to an electrically shorter structure and, therefore, reduced directivity. Figure 5.4(b) summarizes the scanning-rate enhancement performance of the loaded versus unloaded P-LWAs, where the loaded P-LWA capable of scanning to the start of the stopband at 25.26 GHz. Moreover, the proposed structure scans the same scanning range as the conventional structure but with smaller frequency bandwidth. For example, the scanning range from -75° to $+25^\circ$ can be achieved only with 6 GHz instead of 9.5 GHz bandwidth. It is worth noting that the main-beam scanning angle, ϑ_{mb} , obtained from the far-field radiation patterns [Figure 5.4(a)] deviates from the theoretical one obtained from the dispersion diagram [Figure 5.3(b)] in the back-fire direction. This deviation can be explained by the cosine taper of the radiating patches, which is not considered in the theoretical main beam angle expression in Equation 2.23.

Although the proposed method of unit-cell loading increases the scanning rate, this comes at the cost of reduced directivity, which is generally a trade-off that must be considered when designing P-LWAs with enhanced scanning rates.

5.1.3.2 High Dispersive SIW SFN P-LWA with Vivaldi Elements

This section will investigate the implementation of a P-LWA with enhanced scanning rate and single-beam scanning. The design of the SIW SFN in Section 3.5.3.2 will be utilized since it is based on a waveguide transmission line. As mentioned before, dispersive transmission lines inherently have a nonlinear $\beta - \omega$ curve near the respective propagating mode's cutoff frequency (f_c) of the respective propagating mode, e.g., TE_{10} mode in RWG. For the SFN [Section 3.5.3.2] operating with the TE_{10} mode, β reads [70]

$$\beta^{\text{TE}_{10}} = \frac{2\pi\sqrt{\epsilon_r}}{c_0} \sqrt{f^2 - f_c^2}. \quad (5.2)$$

From the above equation, it can be seen that β is real for $f > f_c$ and imaginary for $f < f_c$, meaning that the modes within the waveguide are *propagating* or *evanescent*, respectively. Additionally, for $f \gg f_c$, Equation 5.2 is approximately linear, whereas for frequencies near f_c , the phase constant, β , changes rapidly with f . Unfortunately, increasing f_c will result in a longer unit cell's period ($p > \lambda_0$ at f_{bs}), which limits the frequency bandwidth of the single-beam radiation and, therefore, the scanning range. Furthermore, at $f < f_{\text{bs}}$, the phase velocity of the fundamental space harmonic is $v_{p,0} = \omega/\beta_0 > c_0$ [Equation 2.19], which means in waveguide-based P-LWAs, the fundamental space harmonic ($\beta_0 \approx \beta^{\text{TE}_{10}}$) can also be of a fast wave nature³ and consequently, radiates as a second beam in the backward direction. In fact, the single-beam condition in a dielectric-filled waveguide P-LWA is more restrictive than

³In waveguides, the starting point of the phase constant, i.e. $\beta = 0$, is shifted by the cutoff frequency, f_c .

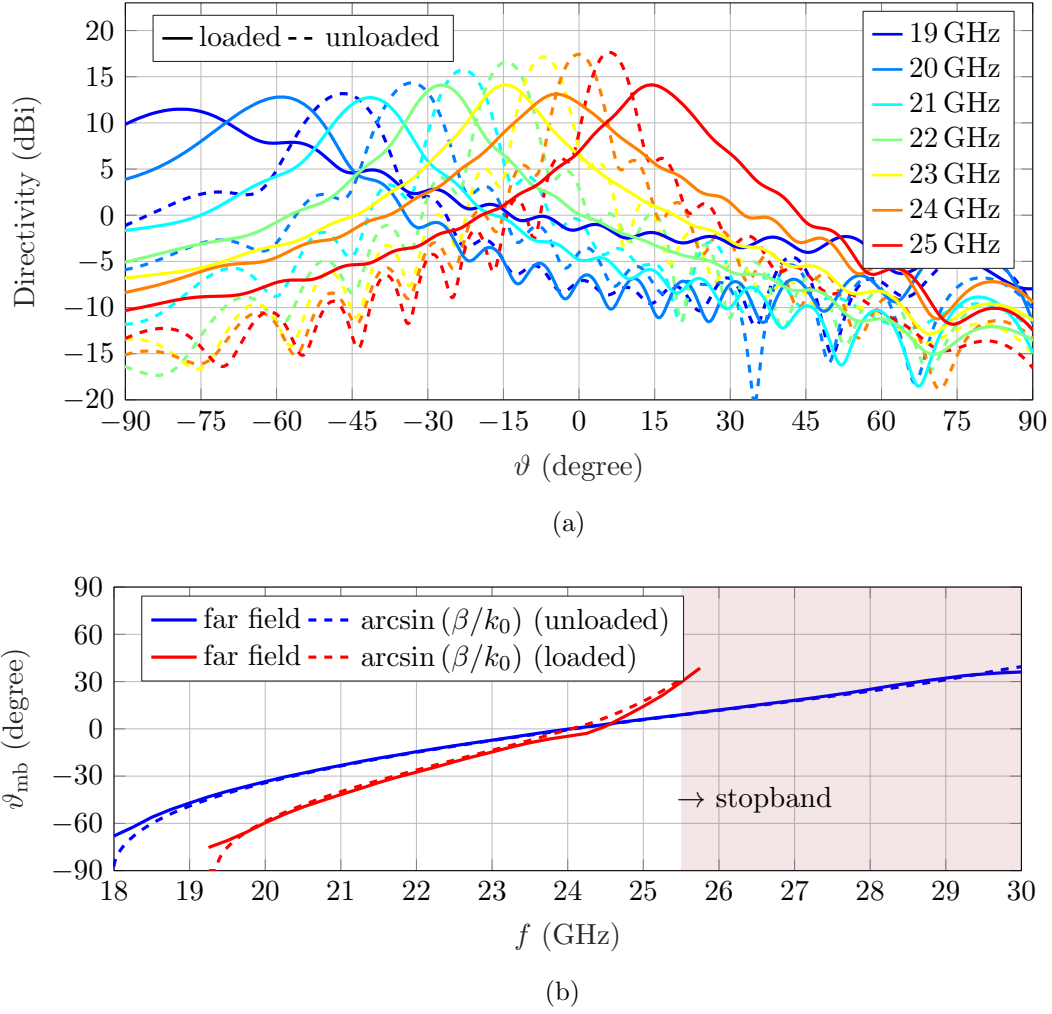


Figure 5.4: Scanning-rate enhancement performance obtained by full-wave simulated 18 unit-cell stub-loaded SFCP P-LWA compared to an unloaded 14 unit-cell SFCP P-LWA. (a) Radiation patterns in the principle yz -plane and (b) main beam direction obtained from the far-field results and the dispersion diagrams in Figure 5.3(b).

that of a TEM-based P-LWA [Section 2.4.3] where the dielectric constant must be $\epsilon_r > 9 + (p/a)^2$ compared to $\epsilon_r > 9$, respectively [18, 140]. However, this condition can be relaxed if phase-reversing elements are realized in waveguide-based P-LWAs [67], which will be demonstrated next.

Figure 5.5(a) shows the SIW SFN from Section 3.5.3.2, whose output ports are terminated with *antipodal Vivaldi* radiating elements. The period of the unit cell is $p = \lambda_g$ at the broadside frequency f_{bs} , which corresponds to 360° phase shift, i.e., $\beta p = 2\pi$. In the following, this structure will be denoted as *phase-maintaining* P-LWA. Furthermore, there are two reasons to choose Vivaldi radiating elements: 1) Antipodal Vivaldi antennas have broadband bandwidth due to their non-resonant radiation mechanism [141], and 2) the inverted polarity of the fields at the top and bottom layers may be exploited to achieve the aforementioned phase-reversing unit cells. To increase the scanning rate, f_c can be increased by decreasing the SIW width a , which

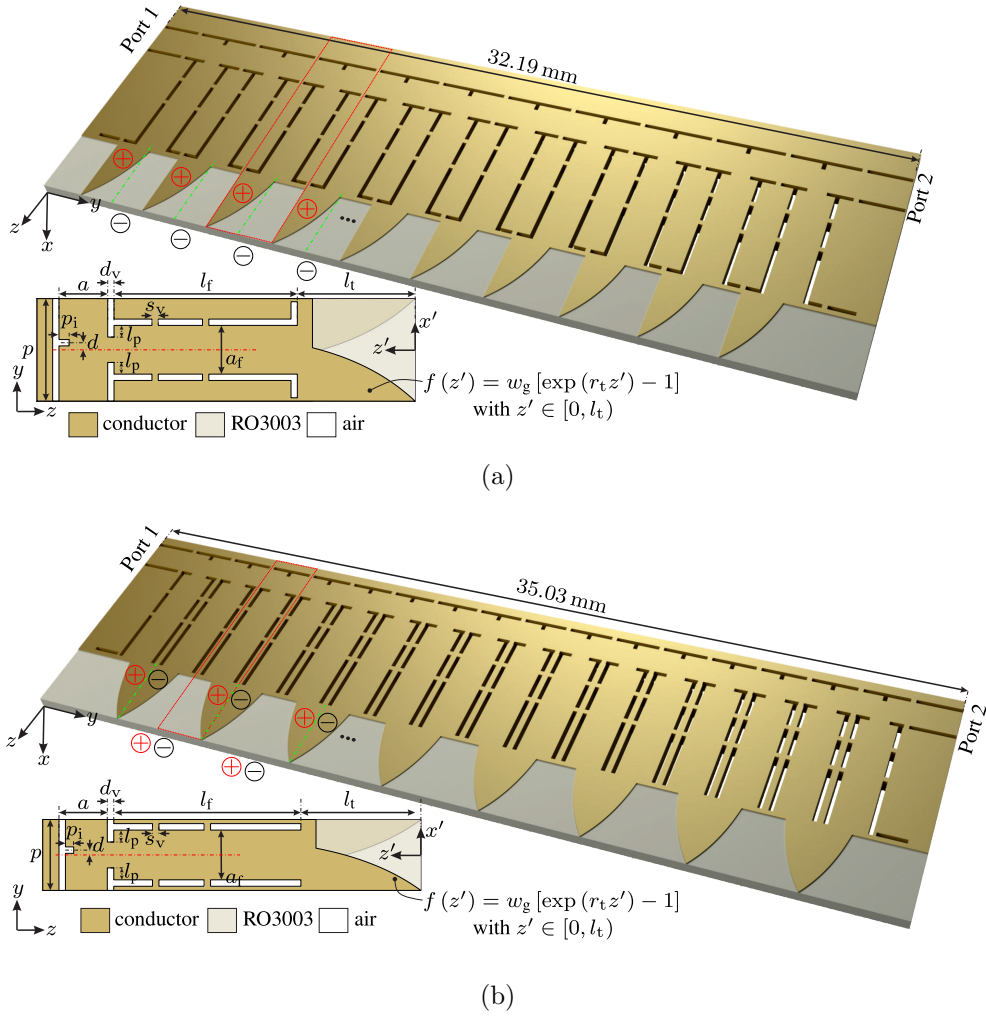


Figure 5.5: Implementation of the scanning-rate enhanced P-LWA based on an SIW SFN with Vivaldi radiating elements. (a) Phase-maintaining P-LWA with limited single-beam scanning range and (b) phase-reversing P-LWA with enhanced scanning rate and single-beam scanning range. For both P-LWAs the unit cell is shown in the inset.

results in a longer guided waveguide length at the same broadside frequency f_{bs} as for the phase-maintaining P-LWA [Figure 5.5(a)]. Hence, the period of the unit cell is shortened to $p = \lambda_g(f_{bs})/2$, which results in a phase shift of 180° across the unit cell, i.e., $\beta p = \pi$. Subsequently, the radiating elements of successive unit cells must provide a phase reversal (π) to retain the equal phase excitation for all unit cells and achieve constructive radiation in the main beam direction, ϑ_{mb} . Figure 5.5(b) shows the implementation of the phase-reversing elements, which is realized by inverting the top and bottom layers of the Vivaldi antennas for each successive unit cell. The design procedure for the SIW SFN is described in Section 3.5.3.2. First, the cutoff frequency for the phase-maintaining P-LWA is set to $f_c = 56.42$ GHz, and then the ω -balancing and Q -balancing conditions are optimized for the broadside frequency $f_{bs} = 78$ GHz. Second, the cutoff frequency for the phase-reversing unit cell is set to $f_c = 65.5$ GHz, and then the same conditions are optimized. Table 5.2 lists the

Table 5.2: Dimensions in mm of the SFN SIW P-LWAs with Vivaldi radiating elements shown in Figure 5.5.

unit-cell type	tuned parameters				fixed parameters				
	a	p	d	p_i	l_f	l_p	a_f	s_v	d_v
Figure 5.5(a)	1.54	3.24	0.04	0.32	5.8	0.37	1.54	0.2	0.2
Figure 5.5(b)	1.30	2.73	0.05	0.25					

optimized dimensions of both structures corresponding to the parameters defined in Figure 5.5. The period of the phase-reversing unit cell is not exactly half of that of the phase-maintaining unit cell due to the different cutoff frequencies.

After optimizing the unit cell, the Vivaldi antenna is tuned for low input reflection seen from the power splitting branch of the SFN unit cell. A set of four parameters achieves the tuning. First, the element's width is defined by the period of the optimized unit cell, p . Second, the length l_t of the taper is tuned along with w_g and r_t .

The simulation results of both unit cells are shown in Figure 5.6. The dispersion diagram of the phase-maintaining unit cell in Figure 5.6(a) clearly shows that not only the $\mathbf{n} = -1$ is a fast wave but also $\mathbf{n} = 0$ and -2 . The corresponding main-beam direction calculated by Equation 2.23 is plotted in Figure 5.6(b) and confirms the limited single-beam scanning for the $\mathbf{n} = -1$. Specifically, at 68 GHz, the $\mathbf{n} = -1$ and 0 radiate a double beam pointing to -26.4° and 65.7° , respectively. The $\mathbf{n} = 0$ exits the radiation cone at 70.7 GHz, and only the $\mathbf{n} = -1$ radiates towards -17° . Again, above f_{bs} at 82.4 GHz, the P-LWA radiates double beams from the $\mathbf{n} = -2$ (backward) and -1 (forward) at -90° and 6.9° , respectively. Finally, at 90 GHz, both beams scan the space with $\mathbf{n} = -2$ points to -45.3° and $\mathbf{n} = -1$ to 18.3° . In the frequency range of 70.7 – 82.4 GHz, the P-LWA features a single-beam scanning range of only 23.9° ($-17^\circ - +6.9^\circ$).

Conversely, Figure 5.6(c) plots the dispersion diagram of the phase-reversing unit cell. Due to the phase-reversing elements, the dispersion curves are shifted along the horizontal line by an additional $+\pi$, as described in the MSL implementation of phase-reversing P-LWA [67]. Specifically, to obtain the dispersion diagram [Figure 5.6(c)], the definition of β_n in Equation 2.17 must be modified as follows

$$\beta_n = \beta_0 + \frac{(2\mathbf{n} + 1) \pi}{p}. \quad (5.3)$$

Equation 5.3 shifts the harmonics spectrum by $+\pi$, and therefore, the $\mathbf{n} = 0$ outside the radiation region, while the shorter period, p , [see Table 5.2] provides a flat radiation region ($\pm k_0 p$) when compared to Figure 5.6(a). These two reasons result in a single-beam scanning for the entire frequency range of 68 – 90 GHz. Additionally, Figure 5.6(d) shows the main-beam direction over frequency, where it can be observed that the main beam scans from -53.8° at 68 GHz to 23.5° at 90 GHz. Moreover, the proposed implementation achieves a total scanning range of about 77.3° . It is worth noting that the beam scans rapidly near the cutoff frequency, whereas at higher frequencies ($f > f_{bs}$), the $\vartheta_{mb} - f$ relationship becomes approximately linear.

The Bloch impedances of both structures are not shown to avoid redundancy, but they have a similar trend to that of the SFN SIW structure shown in Figure 3.16(a).

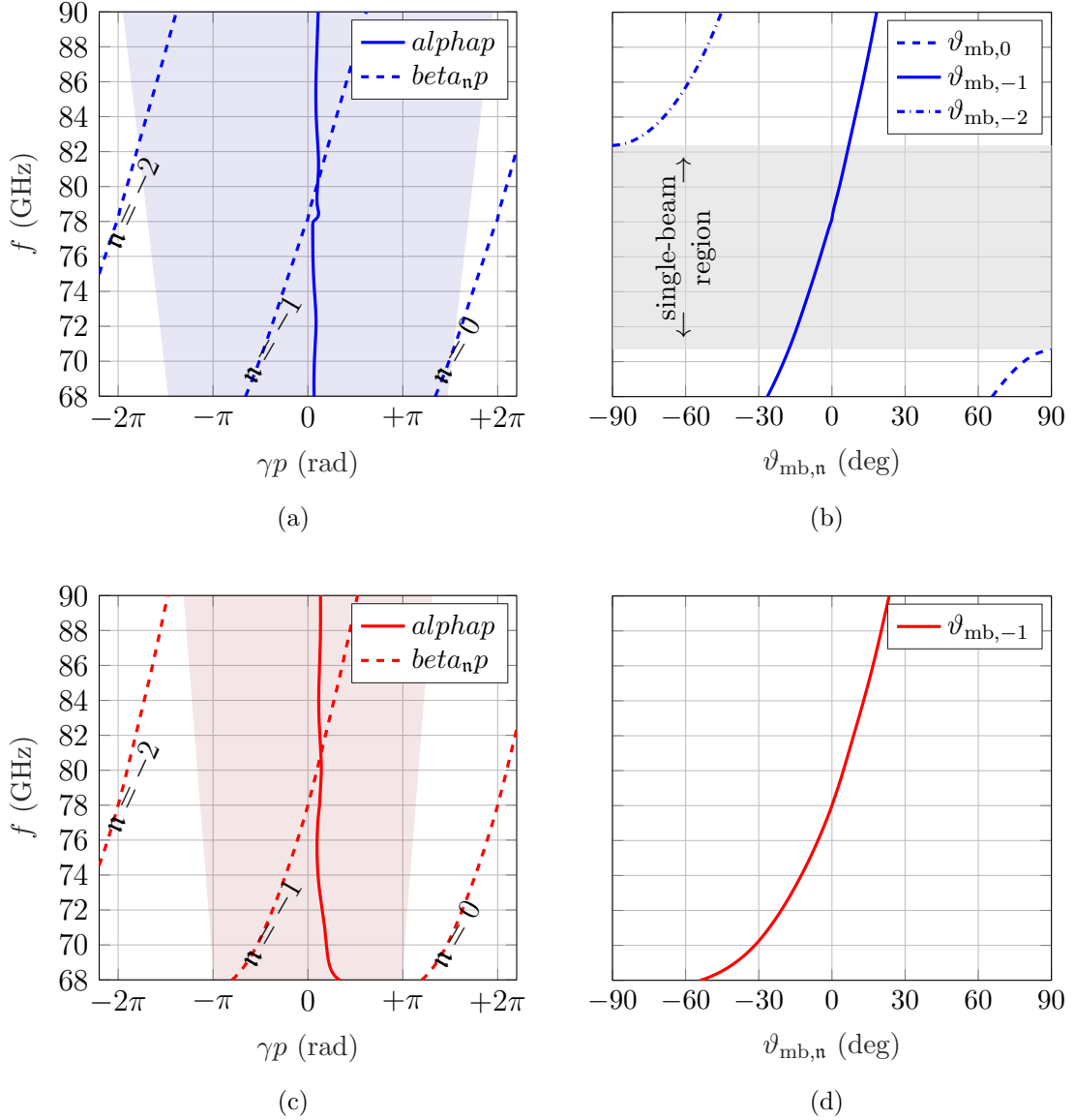


Figure 5.6: Unit-cell simulation of the SFN SIW P-LWAs with Vivaldi elements in Figure 5.5. (a-b) Dispersion diagram and main-beam direction of the phase-maintaining unit cell. (c-d) Dispersion diagram and main-beam direction of the phase-reversing unit cell.

Finally, a cascade of 10 and 14 unit cells are constructed for the phase-maintaining and phase-reversing P-LWAs, which corresponds to a total length of 32.19 mm and 35.05 mm, respectively. Figure 5.7 compares the directivities for both P-LWAs at six different frequencies: 68, 70, 76, 78, 81, and 90 GHz. As expected, the phase-maintaining P-LWA [Figure 5.7(a)] suffers from grating lobes at lower and higher frequencies. In contrast, the phase-reversing P-LWA [Figure 5.7(b)] overcomes double-beam radiation and achieves a larger scanning range.

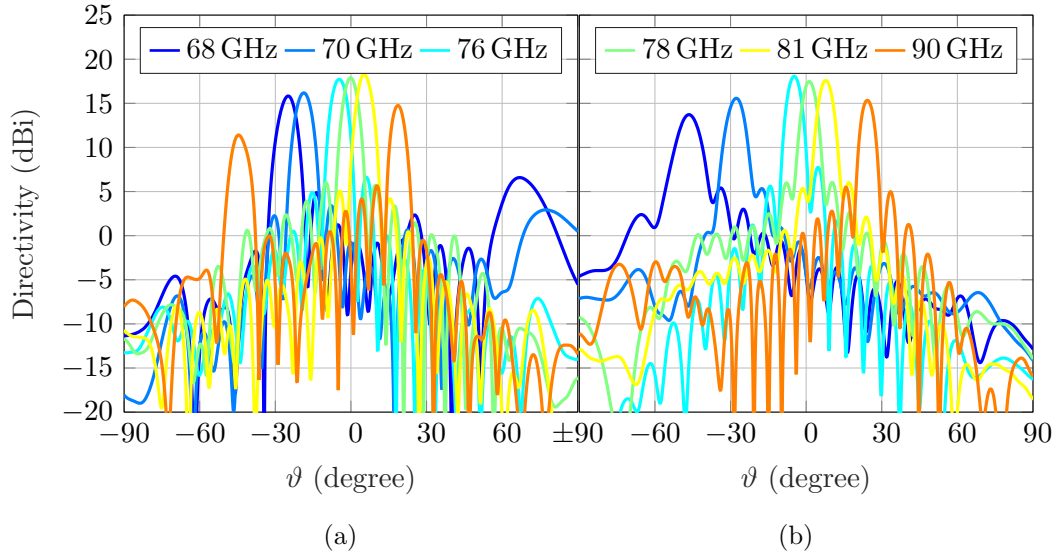


Figure 5.7: Full-wave simulated directivities of the SFN SIW P-LWAs in the principle yz -cut. (a) Phase-maintaining [Figure 5.5(a)] and (b) phase-reversing [Figure 5.5(b)] P-LWAs.

The phase-reversing technique suppresses the radiation from $\mathbf{n} = 0$, but at higher frequencies (> 90 GHz), the $\mathbf{n} = -2$ will eventually become fast and also will contribute to radiation. Therefore, two measures may be considered for this particular case to eliminate the double-beam radiation. First, choosing a substrate with a higher dielectric constant or, second, employing the serpentine-line technique, as described in Section 5.1.1.1.

5.2 Two-Dimensional Frequency-Scanning Concepts

A 1D P-LWA is typically restricted by the problem of beam scanning in one direction with a fan beam. For some applications, such as radar and point-to-point communication, it may be necessary to have a pencil beam pointing in an arbitrary plane or direction. On the other hand, 2D P-LWAs operating with either TEM waves or surface waves propagating in two directions can have a pencil beam at the broadside frequency, i.e., in one direction or a conical beam at other frequencies [142–144].

In contrast, the concept of this section is based on combining an SFN and 1D P-LWAs to achieve a pencil beam capable of scanning two planes with frequency, which means that waves are propagating only along the direction of the 1D P-LWAs. Thus, the pencil-beam radiation is maintained for the full-scanning range. Arguably, the proposed concept may be considered as 1.5D P-LWA since the wave propagates along orthogonal directions belonging to the SFN and 1D P-LWAs. Nevertheless, the concept will be called 2D P-LWA. Part of the results of this section originated from [145, 146], where an MSL-based design of a stub-load SFN and 1D SFN P-LWA is proposed. Here, the implementation will utilize the designs of the SFN and 1D SSIW P-LWA presented in Sections 3.5.3.2 and 3.5.3.1, respectively.

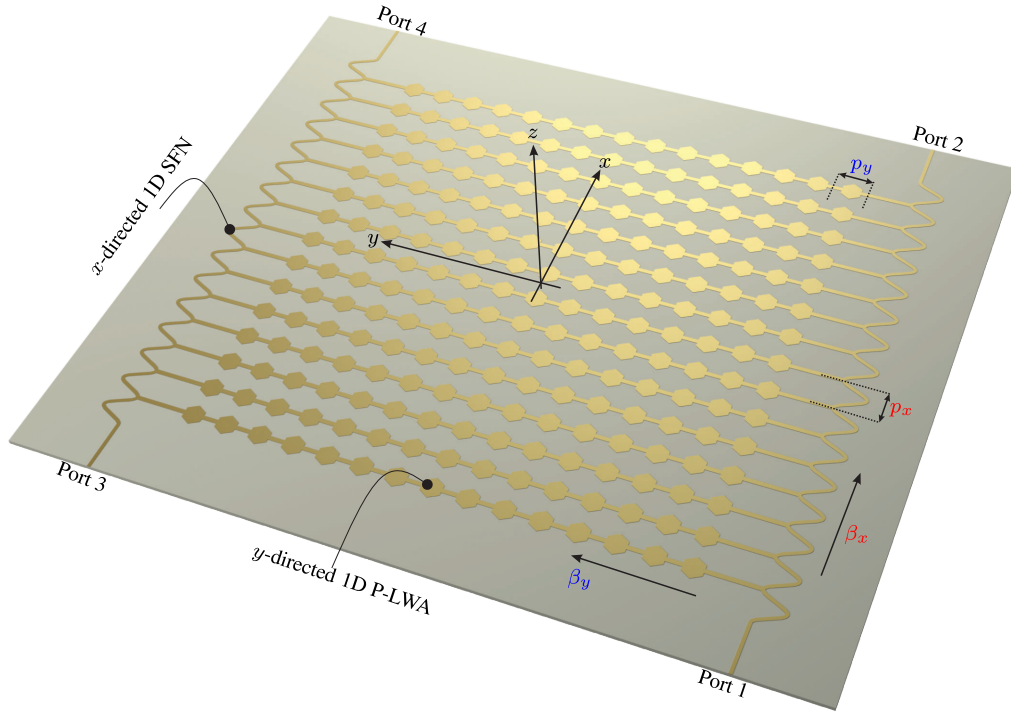


Figure 5.8: Proposed 2D P-LWA, consisting of an array of a y -directed array of 1D P-LWAs and x -directed SFN, capable of frequency scanning the main beam in multiple planes. The overall antenna consists of four ports located at its four corners.

5.2.1 Concept

The 2D P-LWA concept consists of an array of 1D P-LWAs terminated by an SFN. The former provides radiation and scanning in the axis of the 1D P-LWAs, and the latter provides scanning in the perpendicular direction. The 1D P-LWAs and the SFN may have the same f_{bs} , i.e., pointing to the broadside. The dispersion diagram of each structure can be engineered individually so that an oblique frequency-scanning plane or multiple planes can be achieved. Figure 5.8 shows the proposed 2D P-LWA concept, which consists of the array of 1D P-LWAs, extending along the y -direction and fed at both ends by two SFNs extending in the x -direction. Thus, the structure has four ports placed at its four corners. Since all P-LWAs are identical and have the same propagation axis direction if excited at one corner, then polarization can be maintained throughout the scanning bandwidth.

As the 2D P-LWAs are excited at a given port, e.g., port 1, the wave propagates along the corresponding SFN along the x -direction. The SFN progressively feeds the different 1D P-LWAs, with decreasing power due to uniform power division [Section 3.5.3.2]. Furthermore, the SFN provides an x -directed phase gradient versus frequency, which combines with the y -directed dispersion characteristics of the P-LWAs to produce oblique scanning. Multiple scanning planes can be achieved by employing serpentine lines [Section 5.1.1.1], which correspond to radiation with higher space harmonics [147]. Alternatively, additional scanning planes can be achieved by exciting the other three ports, which lead to different x and y -phase combinations and hence independent beam steering in four different directions of space.

The following section will describe the operation principle in more detail. However, for a better presentation of the results, it is more convenient to use a different coordinate system to represent the 3D radiation pattern produced by the 2D P-LWA. Specifically, *azimuth-over-elevation* ($Az - El$ or $\phi_{Az} - \theta_{El}$) angles are chosen instead of the so-far used *roll-over-azimuth* ($\varphi - \vartheta$) angles. The former definition has the pole of the coordinate system's sphere along the y -axis and the latter along the z -axis, as defined in [148, 149]. Hence, in the following, the new coordinate system transforms the radiation patterns by $z \rightarrow y'$ only, whereas the definition of the structure's coordinate system is kept as before.

5.2.2 Analysis of 2D Frequency Scanning

The operation principle of the 2D P-LWA is presented in two steps: 1) a 1D P-LWA array excited with a corporate-feeding network, and 2) a 1D P-LWA array excited with SFN. The foundation of the discussion is based on theoretical dispersion diagrams and the array factor theory. Generally, the radiation pattern of a 2D phased array or 2D P-LWAs of $N \times M$ unit cells can be written as [65, 150]

$$AF(u, v) = \left[\frac{1}{N} \sum_{n=1}^N e^{jnk_0 p_x (u + \Phi_x)} \right] \times \left[\frac{1}{M} \sum_{m=1}^M e^{jmk_0 p_y (v + \Phi_y)} \right], \quad (5.4)$$

where p_x and p_y are the periods of the SFN and P-LWA unit cells, respectively, and u and v are the uv mapping of the radiation pattern, and they are defined for the azimuth-over-elevation convention as follows

$$u = \cos(\theta_{El}) \sin(\phi_{Az}), \quad (5.5a)$$

$$v = \sin(\theta_{El}), \quad (5.5b)$$

Furthermore, $\Phi_x = \cos(\theta_{mb,n}) \sin(\phi_{mb,n})$ and $\Phi_y = \sin(\phi_{mb,n})$ are scanning angles provided by the SFN (x direction) and the P-LWA (y direction), where $\theta_{mb,n}$ and $\varphi_{mb,n}$ are the main-beam directions in the elevation and azimuth directions, respectively. These main-beam directions can be calculated from the scanning law [Equation 2.23] of their respective structures as

$$\theta_{mb,n} = \arcsin\left(\frac{\beta_{y,n}}{k_0}\right), \quad (5.6a)$$

$$\phi_{mb,n} = \arcsin\left(\frac{\beta_{x,n}}{k_0}\right). \quad (5.6b)$$

The above equation shows that the frequency-scanning plane can be controlled by the phase constants of the SFN ($\beta_{x,n}$) and the P-LWA ($\beta_{y,n}$).

5.2.2.1 Single-Plane Scanning

First, the single-plane scanning is considered by a 2D P-LWA consisting of 1D P-LWAs arrayed in the x -direction and excited uniformly with equal phase ($\beta_x = 0$), e.g., with a corporate-feeding network. Then, the 1D P-LWAs dispersion diagram (β_y) fulfills the single-beam scanning condition [Equation 2.25]. Figure 5.9 plots

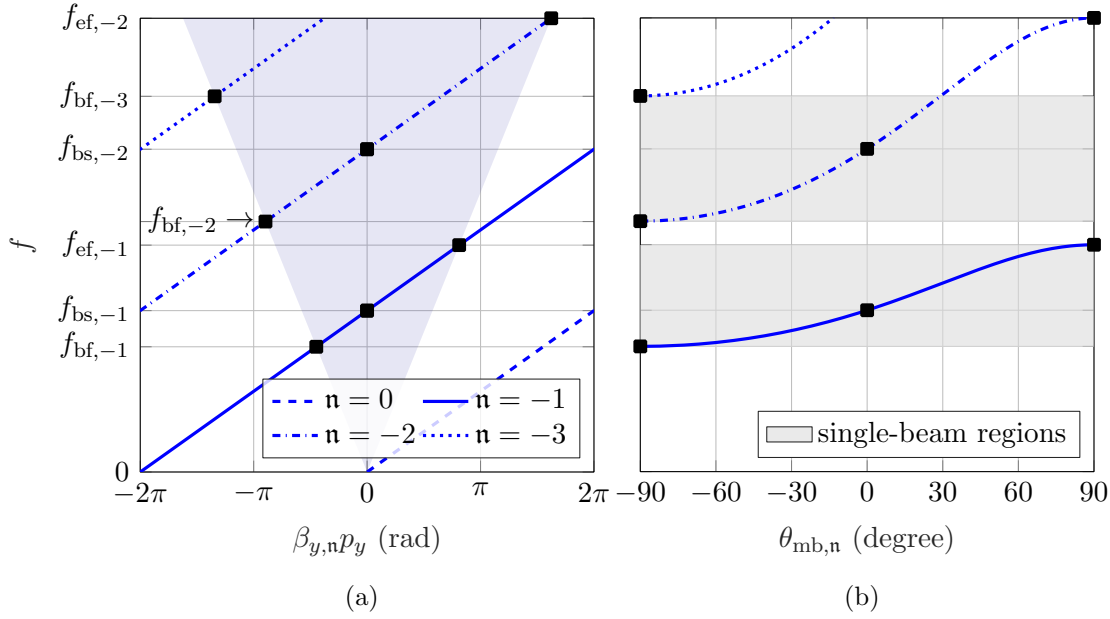


Figure 5.9: Full-space scanning with 1D P-LWAs. (a) Dispersion diagram for a TEM transmission line with $\epsilon_r = 12$, and (b) main-beam direction versus frequency showing the single-beam radiation capability with the $n = -1$ space harmonic.

the dispersion diagram [Figure 5.9(a)] of four space harmonics ($n = -3, -2, -1, 0$) and the corresponding main-beam direction versus frequency, $\theta_{\text{mb},n}$ [Figure 5.9(b)], which are obtained by considering an unperturbed transmission line with $\epsilon_r = 12$. From Figure 5.9(a), the $n = 0$ is a slow wave and, therefore, does not radiate. In contrast, the $n = -1, -2$, and -3 space harmonics become fast when they enter the radiation cone at their respective backfire frequencies, $f_{\text{bf},-1}$, $f_{\text{bf},-2}$, and $f_{\text{bf},-3}$, respectively. It can also be seen that the single-beam scanning is achieved for the $n = -1$ space harmonic, while the $n = -2$ single-beam scanning range is limited by the simultaneous radiation from $n = -3$ space harmonics, i.e., from $f_{\text{bf},-2}$ to $f_{\text{bf},-3}$. The gray-shaded areas in Figure 5.9(b) show the single-beam radiation regions of $n = -1$ and -2 .

Figure 5.10 shows the array factor for a phased array with 20×20 elements and $p_x = p_y = \lambda_0(f_{\text{bf},-3})/2$, corresponding to the arrayed 1D P-LWAs, excited uniformly in the x -direction and by the dispersion diagram shown in Figure 5.9(a) at a frequency between $f_{\text{bs},-1}$ and $f_{\text{ef},-1}$. The plot shows the amplitude of the array factor [Equation 5.4] versus the azimuth (ϕ_{Az}) and elevation (θ_{El}) angles. For the following discussion in this chapter, the radiation pattern for the 2D frequency scanning will be visualized by the -3 dB-contour plot, as shown by the black curve in Figure 5.10. The single-plane frequency scanning is plotted in Figure 5.10(b) for the single-beam radiation, i.e., ($f < f_{\text{bf},-3}$). Each contour in the plot corresponds to a different frequency. As expected, this 1D P-LWA scans the entire upper hemisphere with the $n = -1$ space harmonic from backfire to endfire through the broadside and the half upper hemisphere with the $n = -2$ space harmonic. Moreover, the uniform excitation in x , i.e., $\beta_x = 0$, produces redundant scanning planes at $\phi_{\text{Az}} = 0^\circ$. The

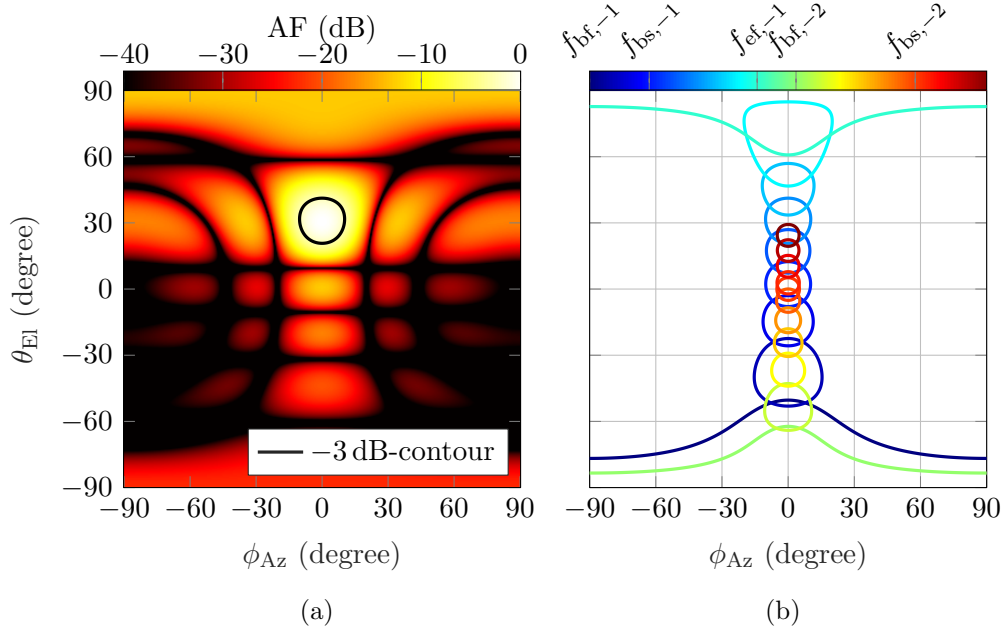


Figure 5.10: Array factor results of an array of 1D P-LWAs excited uniformly in the x -direction and with the dispersion diagram shown in Figure 5.9 in the y -direction. (a) Amplitude of the array factor at a single frequency $f_{bs,-1} < f < f_{ef,-1}$ and (b) single-plane scanning visualized by the -3 dB-contours with each contour corresponding to a different frequency.

wide -3 -dB-beamwidth observed at f_{bf} and f_{ef} ($\theta_{El} = \pm 90^\circ$) is due to the main beam pointing toward the poles of the radiation-pattern sphere.

5.2.2.2 Multi-Plane Scanning

This section presents the analysis of the proposed 2D P-LWA shown in Figure 5.8 to achieve multiple-plane scanning with frequency [145]. From the discussion above, the structure consists of 1D P-LWAs arrayed in the y -direction and terminated with x -oriented SFNs at their ends [Figure 5.8]. Furthermore, the dispersion diagram depicted in Figure 5.11(a) is considered, where the space harmonics $\mathbf{n} = -2, -1, 0$ for the P-LWA and SFN are shown. The dispersion diagram shows that the SFN has phase-reversing unit cells since $\beta_{x,-1}(f=0) = -\pi$, similar to the SFN presented in Section 5.1.3.2. Additionally, the 2D P-LWA has two broadside frequencies (f_{bs}^y and f_{bs}^x) corresponding to zero crossings of the phase constants of the P-LWA (β_y) and SFN (β_x), respectively. From Figure 5.11(a), both the P-LWA and SFN have the same broadside frequency for $\mathbf{n} = -1$, i.e., $f_{bs,-1}^y = f_{bs,-1}^x$, where it is expected that the main beam at this frequency points to $(\phi_{Az}, \theta_{El}) = (0^\circ, 0^\circ)$. Although the $\mathbf{n}_x = -1$ space harmonic of the SFN becomes fast at lower frequencies than that of the P-LWA ($\mathbf{n}_y = -1$), i.e., $f_{bf,-1}^x < f_{bf,-1}^y$, the 2D P-LWA will not radiate, since only the 1D P-LWAs contribute to radiation. Conversely, if only a space harmonic of the P-LWAs is fast, then the SFN will guide the wave without feeding the P-LWAs.

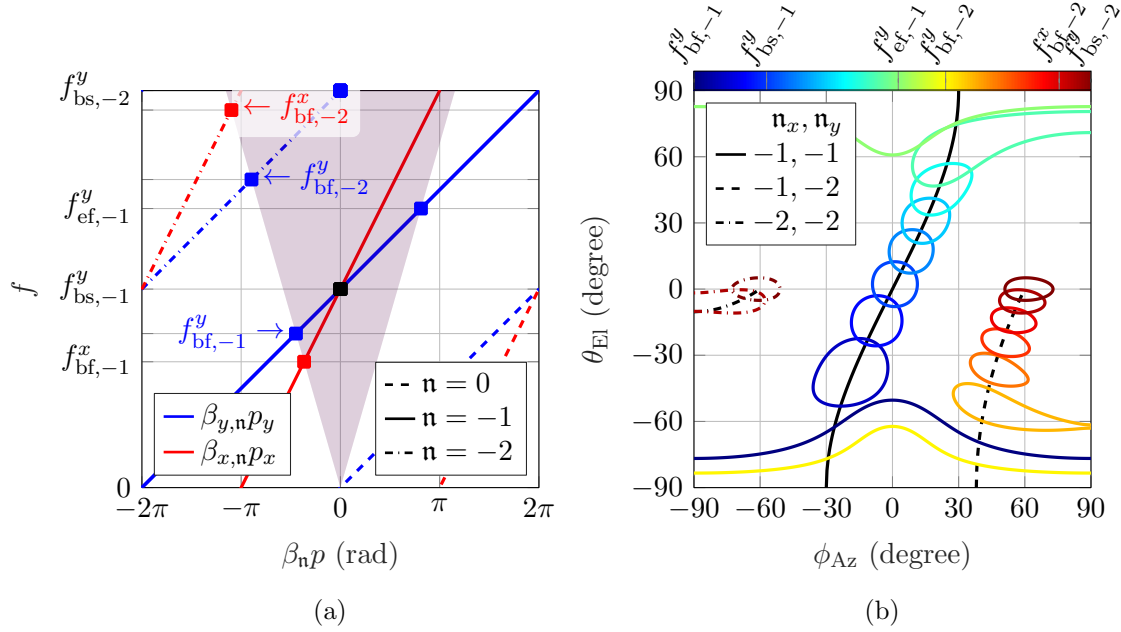


Figure 5.11: Multi-plane scanning with the proposed 2D P-LWA. (a) The dispersion diagram of the 1D P-LWAs ($\beta_{y,n}$) and the SFN ($\beta_{x,n}$) with equal broadside frequencies for both structures (black marker), and (b) the corresponding radiation patterns depicted by -3 dB-contours for an array of 20×20 elements excited by the respective dispersion curve in (a).

Figure 5.11(b) shows the frequency scanning of the proposed 2D P-LWA, which is obtained as in Section 5.2.2.1 by the array factor theory for 20×20 elements for $f_{bf,-1}^y \leq f \leq f_{bs,-2}^y$ [Figure 5.11(a)]. First, the 1D P-LWAs and SFN are considered individually, and their respective main-beam directions are obtained by Equation 5.6 for the different space harmonics corresponding to the excitation of port 1 [Figure 5.8]. It can be clearly seen that the scanning planes follow a path that varies in both ϕ_{Az} and θ_{El} -directions, which are depicted by the solid, dashed, and dotted-dashed black lines. Furthermore, it can be seen that the main beam crosses the $\theta_{El} = 0^\circ$ twice at $\phi_{Az} = 0^\circ$ and $\phi_{Az} = 60^\circ$, which corresponds to $f_{bs,-1}^y$ with $\beta_{x,-1} = 0$ and to $f_{bs,-2}^y$ with $\beta_{x,-1} \neq 0$, respectively. Additionally, for $f > f_{bf,-2}^y$, the SFN has two fast space harmonics, which contribute to radiation as the main beam for $n = -1$ and a grating lobe for $n = -2$ in the ϕ_{Az} -direction. Hence, Figure 5.11(b) also depicts the grating lobes by the dotted-dashed contours. The amplitude of the second beam or grating lobe depends on its scanning angle and the radiation pattern of the 1D P-LWA's unit cell. Increasing the frequency further will produce radiation with multiple beams as the beam scans with frequency. It is also possible to radiate with four beams (two in elevation and two in azimuth directions) corresponding to four space harmonics. This case will not be considered in this thesis.

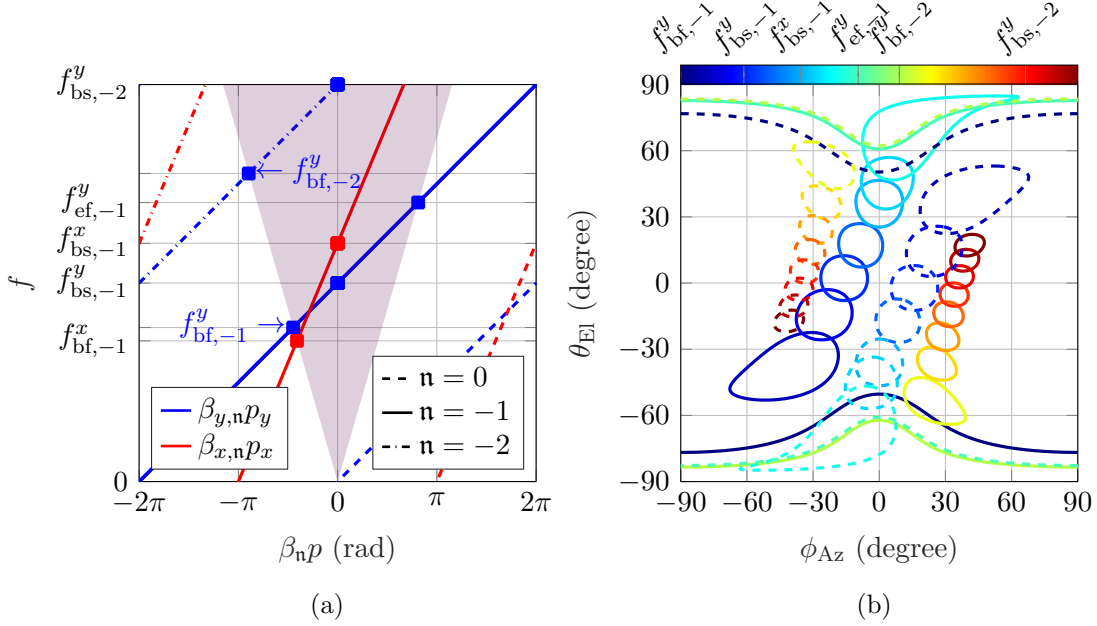


Figure 5.12: Multi-port multi-plane scanning with a 2D P-LWA. (a) The dispersion diagram of the 1D P-LWAs ($\beta_{y,n}$) and the SFN ($\beta_{x,n}$) with different broadside frequencies for both structures, and (b) the corresponding radiation patterns visualized by -3 dB-contours. The solid and dashed contours correspond to the excitation of port 1 and port 4 [Figure 5.8], respectively.

5.2.2.3 Multi-Port Multi-Plane Scanning

In the previous section, the 2D P-LWA covers the upper hemisphere partially since only two space harmonics support single-beam scanning. Therefore, to overcome this limitation, port 4 [Figure 5.8] of the structure may be employed to increase the frequency-scanning coverage further, as described in [146]. Here, the dispersion diagram is slightly modified by shifting the broadside frequency of the SFN to a higher frequency $f_{bs,-1}^x > f_{bs,-1}^y$, as shown in Figure 5.12(a). Figure 5.12(b) plots the -3 dB-contours of the array factor for excitations of port 1 (solid lines) and port 4 (dashed lines) for the single-beam frequency range $f_{bf,-1}^y - f_{bs,-2}^y$. First, the excitation of port 1 (at $-x$ and $-y$) results in a beam that does not cross the $(0^\circ, 0^\circ)$ point due to the different broadside frequencies ($f_{bs,-1}^x \neq f_{bs,-1}^y$). Instead, at these broadside frequencies, the main beam will cross $\theta_{El} = 0^\circ$ and $\phi_{Az} = 0^\circ$ at $f_{bs,-1}^y$ ($\beta_{y,-1} = 0$) and $f_{bs,-1}^x$ ($\beta_{x,-1} = 0$), respectively. In contrast, the excitation of port 4 (at $+x$ and $+y$) will excite a backward wave in the $-x$ and $-y$ -directions and, therefore, will mirror the scanning planes around the elevation and azimuth axes, as shown by the dashed lines in Figure 5.12(b). The excitation of ports 2 and 3 will generate orthogonal scanning planes to those of 1 and 4.

5.2.3 Full-Wave Simulations and Measurement Verification

Two implementations of the proposed 2D P-LWA will be presented in this section. The implementations are based on the SIW designs of the SFN in Section 5.1.3.2

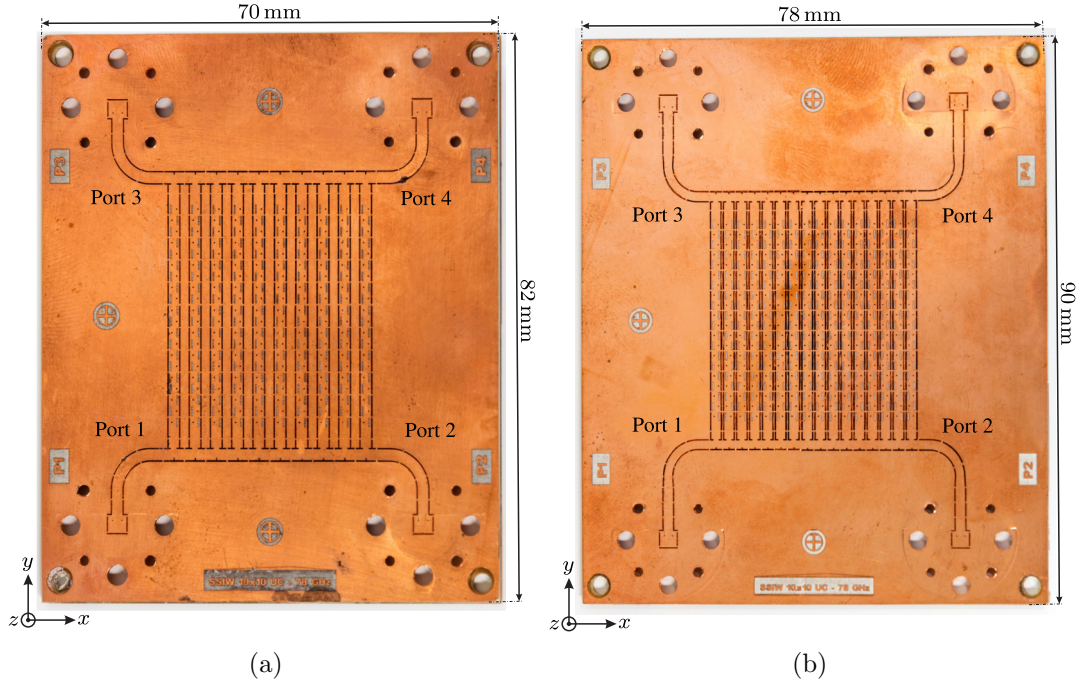


Figure 5.13: Top view of the fabricated prototypes of the 2D P-LWA with an x -directed SFN and y -directed SSIW fed by a WG-to-SIW transition. Implementation of SFNs with (a) phase-maintaining, and (b) phase-reversing unit cells.

and the slotted SIW (SSIW) in Section 3.5.3.1 operating at the broadside frequency 78 GHz. The substrate used to realize the SIW structures for the following two sections is Rogers RO3003 ($\epsilon_r = 2.97$ and $\tan \delta = 0.001$) with thickness $h = 508 \mu\text{m}$ [Section 3.5.3]. In addition, the antenna PCB is backed by an FR-4 substrate with a thickness of 1 mm to improve mechanical stability.

5.2.3.1 2D P-LWAs with SFN and SSIW

Two designs will be presented in this section to verify the proposed concept and the analytical analysis presented in Section 5.2.2. Additionally, for both designs, a y -directed array of the SSIW 1D P-LWA [Section 3.5.3.1] will be used without any modification. Its geometry dimensions are listed in Table 3.3. Furthermore, two different SFNs designs are utilized along the x -direction to terminate the SSIW array. First, the SFN with phase-maintaining unit cells will be used, but with limited frequency scanning. Second, the phase-reversing SFN will also be used to verify the frequency-scanning enhancement in 2D P-LWAs. The dimensions of these SFNs are the same used for the 1D P-LWAs with Vivaldi radiation elements, which can be found in Table 5.2.

Figure 5.13 shows the fabricated prototypes of both designs, where the phase-reversing unit cells are connected to the SSIWs with alternating slots along the y -direction, as shown in Figure 5.13(b). The number of unit cells is 10×10 ($N \times M$) and 16×10 for the phase-maintaining and phase-reversing structures, respectively. Figure 5.14 compares the normalized radiation pattern for the structure shown in Figure 5.13(a) at 78 GHz. Since the broadside frequency for both structures is at

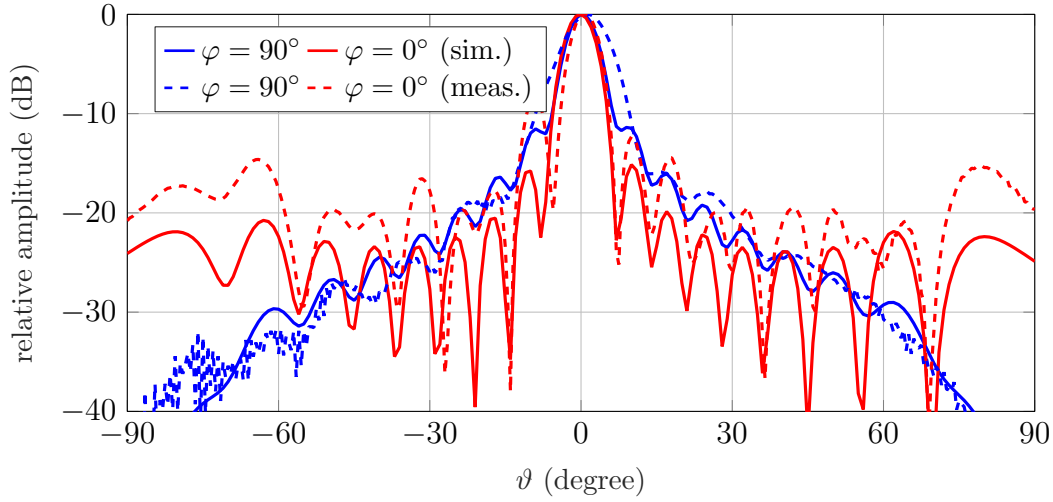


Figure 5.14: Full-wave simulation and measurement comparison of the normalized radiation patterns of the 10×10 SFN SSIW P-LWA [Figure 5.13(a)] at 78 GHz for the principle xz and yz -planes.

78 GHz, it can be seen that the main-lobe direction points to the boresight for both plane cuts ($\varphi = 0^\circ$ and 90°). Also, good agreement between the simulation and measurement is observed except for slight disagreement for $|\vartheta| > 60^\circ$.

Figure 5.15 depicts -3 dB-contours of the radiation pattern for different frequencies (color-coded plot) for both structures. It is worth noting that the radiation-pattern measurements were performed using a *roll-over-azimuth* positioning system. Thus, the measured results (only presented for the phase-maintaining structure shown in Figure 5.13(a)) are transformed from $\varphi - \vartheta$ to $\phi_{Az} - \theta_{El}$ coordinate system, which is done as a post-processing step. Moreover, this type of measurement takes a long time due to the large number of angular and spectral samples. Therefore, only a limited number of frequencies were measured. Figures 5.15(a) and 5.15(c) show the measured and simulated results for the frequency scanning of the phase-maintaining structure. The measured frequency range spans from 70 to 83 GHz, while the simulated data are shown for the frequency range of 68 – 90 GHz since the measurement requires a long time due to the sampling in both φ and ϑ . Nevertheless, the results show good agreement between the simulation and measurement and verify the oblique scanning proposed in Section 5.2.1. Additionally, grating lobes appear for $f > 83$ GHz, which is expected due to the $n_x = 2$ of the feeding network [cf. Figure 5.6(a)]. On the other hand, Figures 5.15(b) and 5.15(d) show the frequency scanning for the phase-reversing structure, which mitigates grating lobes and achieves a high-scanning rate. Finally, exciting the other ports, e.g., port 2, will achieve a scanning plane orthogonal to that shown in Figure 5.15(b). Such excitation has to be frequency selective since the 2D P-LWA is optimized for port 1 excitation only. For the specific case of the structures presented here, the broadside frequency (78 GHz) has to be avoided due to the degradation in the antenna's gain.

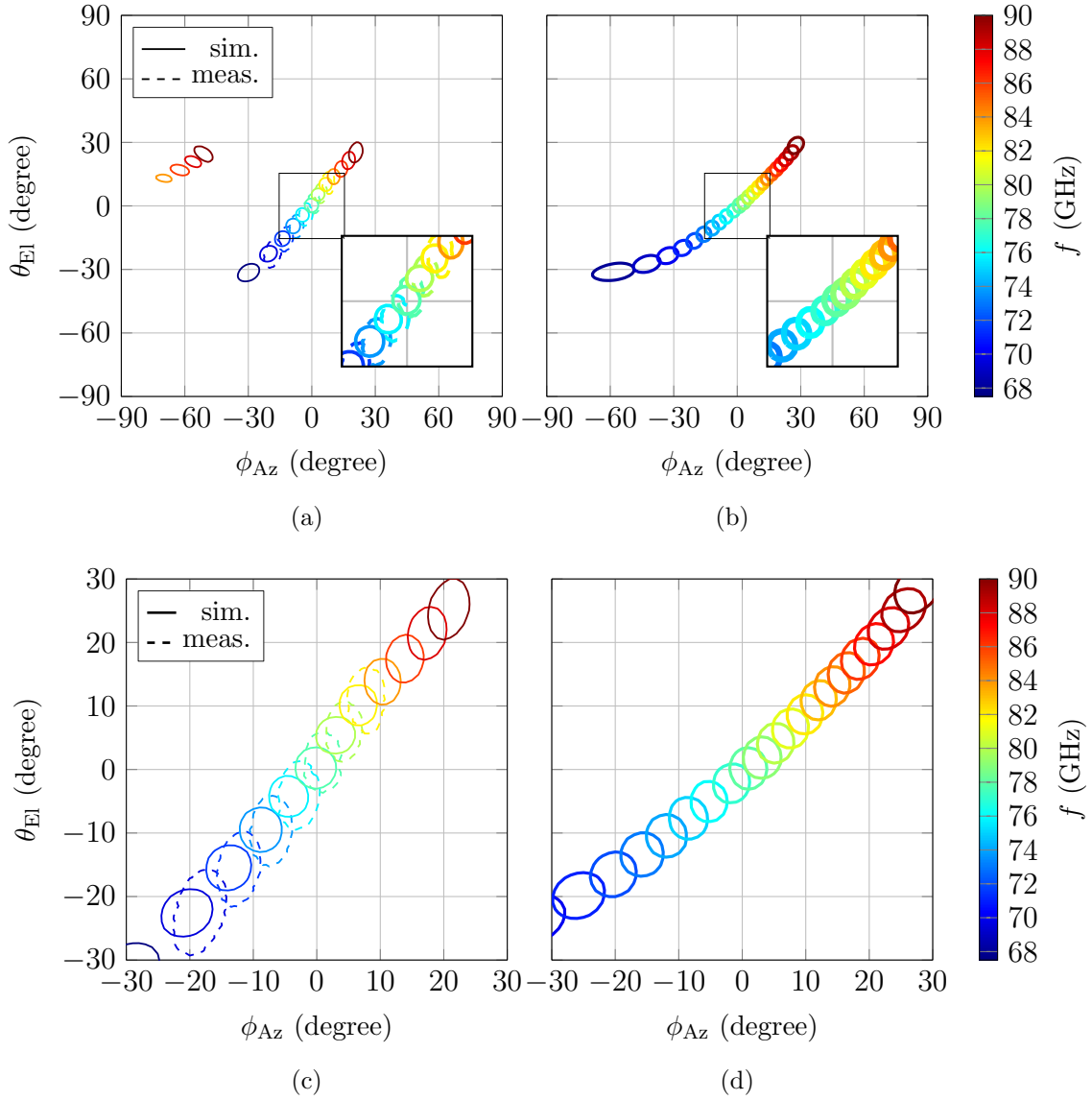


Figure 5.15: -3dB -contour radiation patterns obtained by full-wave simulation and measurement of the structures with (a) phase-maintaining [Figure 5.13(a)], and (b) phase-reversing [Figure 5.13(b)] unit cells. (c) and (d) Zoomed view of (a) and (b), respectively.

5.3 Summary

This chapter presented concepts for frequency scanning in 1D and 2D P-LWAs. First, general methods for frequency-scanning rate enhancement in 1D P-LWAs were discussed in Section 5.1, and later two different designs were presented. The first design [Figure 5.1.3.1] is based on unit-cell loading with reactive elements to increase the dispersion curve, which is demonstrated by the SFCP P-LWA operating at 24 GHz presented in Section 4.3.2.2 loaded with short stub. This design achieves a scanning rate of $15.14^\circ/\text{GHz}$ compared to $9.3^\circ/\text{GHz}$ of its unloaded counterpart. The second design utilized the SIW SFN, discussed in Section 3.5.3.2, where its outputs are

terminated by Vivaldi radiating elements, which operate at the center frequency of 78 GHz [Section 5.1.3.2]. The novelty of the design lies in the realization of phase-reversing elements to suppress radiation from the fundamental $\mathbf{n} = 0$ and the $\mathbf{n} = -2$ space harmonics due to the 180° phase shift across the unit cells and the short period of the unit cell, respectively. Here also, the design achieves single-beam scanning for the frequency range of 68 – 90 GHz. Second, Section 5.2 presented a concept of a 2D P-LWA based on combining an SFN in the x -direction terminated with an array of 1D P-LWAs in the y -direction. Hence, the P-LWA provides frequency scanning in one direction and the SFN in the other, which results in beam scanning in multi-plane with frequency. This concept radiates with a pencil beam over a wide frequency bandwidth while maintaining polarization. Furthermore, the concept was first explained by theoretical dispersion diagrams and later validated by two designs of SSIW terminated with SFNs operating again at 78 GHz.

FMCW-Radar Demonstrator

The main focus of this work is the theory and development of new concepts for frequency-scanning P-LWA. Nevertheless, a demonstration of the P-LWA functionalities is presented in this chapter by a frequency-modulated continuous wave (FMCW) radar operating in the frequency range of 76 – 81 GHz.

6.1 Concept of Frequency Scanning with FMCW Radar

This section introduces the principle of combining frequency scanning with FMCW radars.

6.1.1 Principle of FMCW Radars

A simplified block diagram of an FMCW radar is shown in Figure 6.1, which consists of a frequency-chirped signal transmitted by the transmitter (Tx) antennas. For a chirp with a sawtooth waveform, the frequency of the transmitted signal increases with time [151, 152], as

$$f_{Tx}(t) = f_{start} + \delta_c t, \quad (6.1)$$

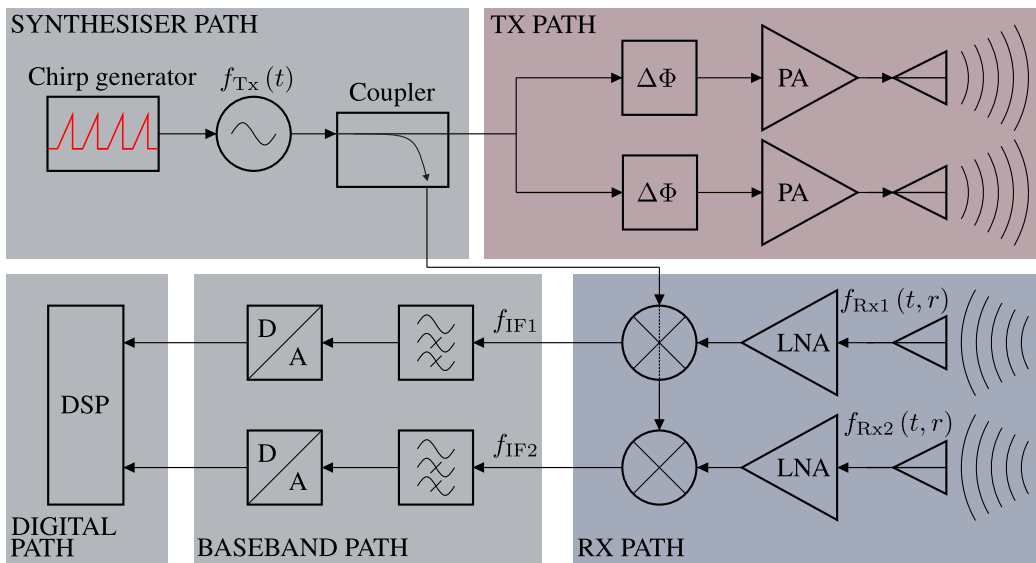


Figure 6.1: Simplified block diagram of an FMCW radar consisting of two Tx and two Rx antennas.

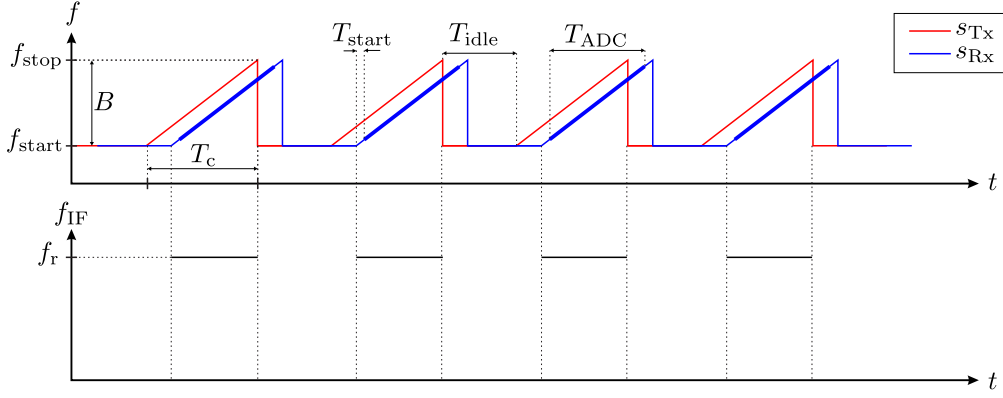


Figure 6.2: Transmitted and received chirps from an FMCW radar. The received chirp is delayed by a target located at a range R , where R can be determined from the beat frequency f_r of the mixed-down signal.

where f_{start} is the start frequency and $\delta_c = B/T_c$ is the chirp slope rate, which can be calculated from the transmitted chirp bandwidth $B = (f_{\text{stop}} - f_{\text{start}})$ and the total chirp time T_c . Thus, the real-valued Tx signal can be expressed as follows

$$s_{\text{Tx}}(t) = A_{\text{Tx}} \sin[2\pi f_{\text{Tx}}(t)t] = A_{\text{Tx}} \sin[2\pi (f_{\text{start}} + \delta_c t)]. \quad (6.2)$$

At the receiver antennas, the received (Rx) signal of the reflected wave from a single target, as depicted in Figure 6.2, reads

$$s_{\text{Rx}}(t, \tau_r) = A_{\text{Rx}} \sin \left\{ 2\pi \left[f_{\text{start}}(t - \tau_r) + \frac{\delta_c}{2} (t - \tau_r)^2 \right] \right\}, \quad (6.3)$$

where τ_r is the time of flight of the signal from the transmitter to the target and back. τ_r includes both the range (R) and velocity of the target. For the sake of simplicity and compactness, a scenario of only a single static target will be assumed, which yields $\tau_r = 2R/c_0$. The Rx signal [Equation 6.3] is then mixed down with the Tx signal [Equation 6.2] to obtain the *intermediate frequency* (IF) of the signal,

$$s_{\text{IF}}(t, \tau_r) = A_{\text{IF}} \cos \left[2\pi \left(f_{\text{start}} \tau_r + \delta_c \tau_r t - \frac{\delta_c}{2} \tau_r^2 \right) \right]. \quad (6.4)$$

Note that A_{Tx} , A_{Rx} , and A_{IF} are the amplitudes of the Rx, Tx, and IF signals, respectively, which are determined by the amplification and attenuation of their respective signals. Furthermore, the beat frequency of a static target located at a range R away from the radar can be calculated as [Figure 6.2]

$$f_r = \frac{2BR}{T_c}, \quad (6.5)$$

which can be determined after applying the *Fourier transformation* on the IF signal, s_{IF} . Finally, the range resolution of a radar is directly related to the bandwidth, B , of the chirp, as

$$\Delta r = \frac{c_0}{2B}. \quad (6.6)$$

The discussion above can be extended to multiple moving targets due to the linearity of the received signals [151].

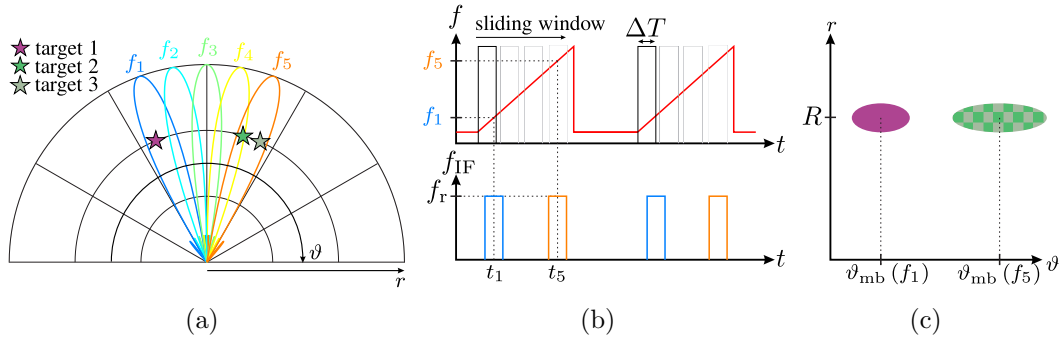


Figure 6.3: Range-angle signal processing of an FMCW radar with P-LWAs. (a) Frequency-scanned radiation patterns illuminating 3 targets located at the same range R and different angles. (b) Processing fraction of the received chirp using STFT and (c) the corresponding range-angle profile.

6.1.2 Frequency-Scanning Radar

The concept of utilizing LWAs with FMCW radars was first introduced by Tospann and Grüner, which used a P-LWA with serpentine lines [Section 5.1.1.1] in an airborne FMCW radar system [29], providing range and 1D angular resolutions. For the following discussion, it will be assumed that two P-LWAs are connected to the Tx and the Rx paths with the same radiation characteristics.

Figure 6.3 illustrates the principle of range-angle radar processing. A scenario of three targets located at the same range, R , and two different angles is shown in Figure 6.3(a), where the frequency-scanned beams of the P-LWA illuminate the targets. Furthermore, target 1 is located in the backward direction ($\vartheta < 0$) and illuminated by a beam at f_1 , whereas target 2 and 3 ($\vartheta > 0$) are close to each other, and a single beam illuminates them at f_5 . In order to obtain both the range and angle information, the short-time Fourier transformation (STFT) can be used [153]. Figure 6.3(b) sketches the principle of using STFT on the received chirp, where the Rx signal $s_{\text{IF}}(t)$ is multiplied by a sliding window of width ΔT (time gate). For each shift of the sliding window, a range-angle-dependent beat frequency $f_r(r, \vartheta)$ is calculated. In Figure 6.3(c) shows a range-angular profile of the scenario shown in Figure 6.3(a), where it can be seen that the objects illuminated by the beam at f_5 cannot be distinguished from each other. Since the chirp time (T_c) is not available for the range processing, it is evident that there will be a trade-off between the range and angular resolution. This trade-off will be discussed in the next section.

6.1.2.1 Range Resolution

Due to the beam-scanning feature of P-LWAs, the power transmitted by the Tx antenna and received by the Rx antenna is direction-dependent, reducing the frequency bandwidth available for detection and, therefore, the range resolution of the radar [Equation 6.6]. An effective bandwidth, B_{eff} , can also be derived from the processing window ΔT , yielding a modified range resolution relation for P-LWAs. Furthermore, the window width ΔT corresponds to an effective beamwidth $\vartheta_{3\text{dB}}(B_{\text{eff}})$. Assuming a constant beamwidth ($\vartheta_{3\text{dB}}$) and a constant beam-scanning range ($\Delta\vartheta_{\text{mb}}$) over the

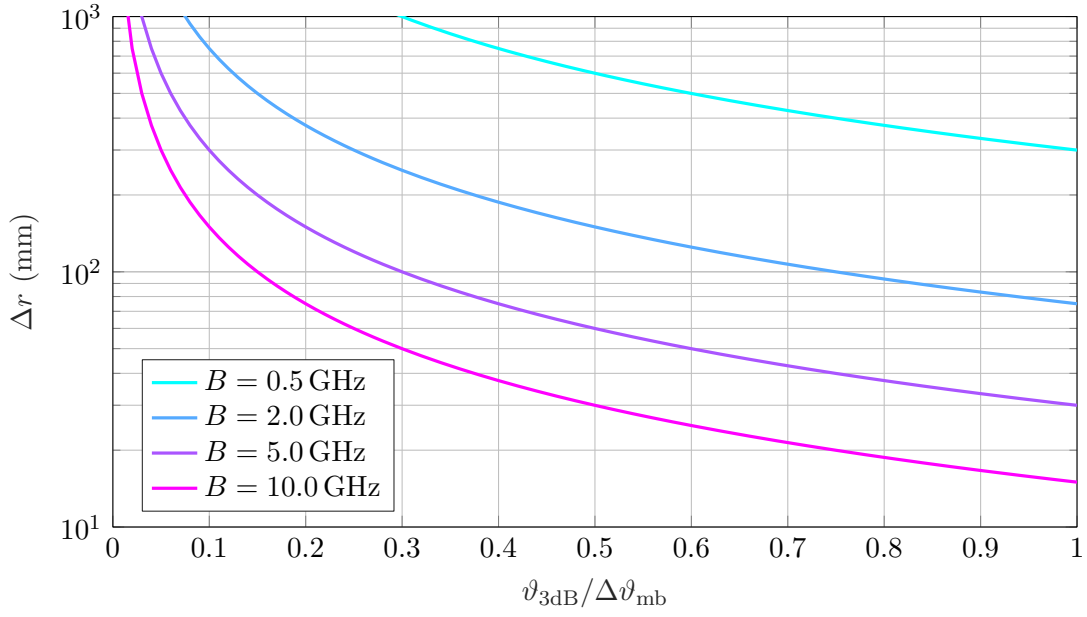


Figure 6.4: Trade-off between the range resolution (Δr) and beamwidth-to-scan range ratio ($\vartheta_{3\text{dB}}/\Delta\vartheta_{\text{mb}}$) for different chirp's bandwidths B .

entire bandwidth, B , the effective bandwidth reads [31]

$$B_{\text{eff}} = \frac{\Delta T}{T_c} B = \frac{\vartheta_{3\text{dB}}}{\Delta\vartheta_{\text{mb}}} B. \quad (6.7)$$

By inserting Equation 6.7 in Equation 6.6, the new range resolution relation yields

$$\Delta r = \frac{c_0}{2B} \frac{\Delta\vartheta_{\text{mb}}}{\vartheta_{3\text{dB}}}, \quad \text{for } \vartheta_{3\text{dB}} \leq \Delta\vartheta_{\text{mb}}. \quad (6.8)$$

The above equation is only valid if the scanning range is larger than the beamwidth ($\vartheta_{3\text{dB}}/\Delta\vartheta_{\text{mb}} \leq 1$) since B_{eff} cannot be larger than B . To illustrate the trade-off between the range and angle resolutions, Figure 6.4 plots Equation 6.8 for different B values. For a better comparison, the horizontal axis is normalized by the scanning range $\Delta\vartheta_{\text{mb}}$. It can be seen that Δr increases dramatically when decreasing the ratio $\vartheta_{3\text{dB}}/\Delta\vartheta_{\text{mb}}$. Moreover, at $\vartheta_{3\text{dB}}/\Delta\vartheta_{\text{mb}} = 1$, the range resolution, Δr , converges to that of the conventional FMCW radar [Equation 6.6].

6.1.2.2 Angular Resolution

The beamwidth of the antenna determines the angular resolution of the radar. Since Equation 6.8 relates the range resolution (Δr) to the angular resolution ($\vartheta_{3\text{dB}}$), then for an antenna with a fixed scanning range, ΔT is the only parameter available to control the trade-off between the range and angular resolution. Moreover, if ΔT is chosen to be large, then $\vartheta_{3\text{dB}}$ is large as well, which results in a large angular resolution. On the other hand, if ΔT is too small, then the received signal power will be degraded since the STFT essentially integrates the reflected energy over the time ΔT .

6.2 Radar Polarimetry

Chapter 4 presented concepts to control the polarization of P-LWAs while maintaining the frequency-scanning feature. Furthermore, since the quad-polarized SFP P-LWA shown in Section 4.2 will be demonstrated in Section 6.4, it is necessary to discuss the principle of polarimetric radars briefly.

6.2.1 Polarization Scattering Matrix

In general, the scattered field from a target, $\mathbf{E}^s$, can be decomposed into two orthogonal polarizations depending on the polarization of the incident field, $\mathbf{E}^i$. Furthermore, suppose the Tx and Rx antennas of the radar system shown in Figure 6.1 have two orthogonal polarizations. In that case, such a system is called *fully polarimetric* radar. Its *polarization scattering matrix*, $\mathbf{S}$, can be calculated from the incident and scattered field vectors $\mathbf{E}^s$ and $\mathbf{E}^i$, respectively. For linearly-polarized antennas, it is convenient to express $\mathbf{S}_{LP}$ in the conventional orthogonal basis $(\hat{h}, \hat{v})$ [154], as

$$\begin{bmatrix} \mathbf{E}_h^s \\ \mathbf{E}_v^s \end{bmatrix} = \begin{bmatrix} S_{hh} & S_{hv} \\ S_{vh} & S_{vv} \end{bmatrix} \begin{bmatrix} \mathbf{E}_h^i \\ \mathbf{E}_v^i \end{bmatrix}, \quad (6.9)$$

where h and v correspond to the horizontal and vertical polarizations, respectively. Additionally, the diagonal entries of $\mathbf{S}_{LP}$ represent the co-directional polarization of the Tx and Rx antennas, while the anti-diagonal entries represent the cross-polarization. If the full 2×2 linear matrix, $\mathbf{S}_{LP}$, is obtained, then its circular counterpart, $\mathbf{S}_{CP}$, can be calculated by a simple transformation, as

$$\begin{bmatrix} S_{ll} & S_{lr} \\ S_{rl} & S_{rr} \end{bmatrix} = \frac{1}{2} \begin{bmatrix} 1 & -j \\ 1 & +j \end{bmatrix} \begin{bmatrix} S_{hh} & S_{hv} \\ S_{vh} & S_{vv} \end{bmatrix} \begin{bmatrix} 1 & 1 \\ -j & +j \end{bmatrix}, \quad (6.10)$$

where r and l denote RHCP and LHCP, respectively. It is worth noting that the left and right matrices from $\mathbf{S}_{LP}$ correspond to the polarization transformations from circular to linear and vice versa, respectively. Although the transformation is simple, in practice, such transformation requires non-negligible computational effort since the scattering matrix also contains information on the range and velocity of the target. Additionally, in the case of P-LWAs, it also includes the angular direction of the target, as explained in Section 6.1.2. Therefore, it would be more efficient to directly calculate the matrix based on the antennas' inherent Tx and Rx polarization.

6.2.2 Canonical Scattering Mechanisms

The scattering mechanisms caused by a target can be discriminated by a set of basic scattering, as described by Huynen [155]. In general, there are three symmetric canonical scatters: plate, dihedral, and dipole. Starting with a plate, its polarization scattering matrices in both linear and circular polarization bases for the monostatic radar-cross section (RCS) case read

$$\mathbf{S}_{LP}^{\text{plate}} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \quad (6.11a)$$

$$\mathbf{S}_{\text{CP}}^{\text{plate}} = \begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix}. \quad (6.11b)$$

A plate is an orientation-independent scatterer, representing targets causing a single or an odd number of reflections. Furthermore, such a scatterer flips the handedness of the circular polarization [Equation 6.11b], whereas linear polarizations are not affected [Equation 6.11a]. A plate is usually not preferred when conducting radar measurements due to its pencil-beam RCS pattern, which is challenging to align with the radar. For such measurements, a trihedral is a preferable option with a similar polarization characteristic to the plate.

Combining two plates with orthogonal orientation (right angle) is referred to as dihedral. Similar to the plate, its polarization scattering matrices read

$$\mathbf{S}_{\text{LP}}^{\text{dihedral}} = \begin{bmatrix} \cos(2\psi_s) & \sin(2\psi_s) \\ \sin(2\psi_s) & -\cos(2\psi_s) \end{bmatrix}, \quad (6.12a)$$

$$\mathbf{S}_{\text{CP}}^{\text{dihedral}} = \begin{bmatrix} \exp(j2\psi_s) & 0 \\ 0 & \exp(-j2\psi_s) \end{bmatrix}, \quad (6.12b)$$

where ψ_s is the orientation angle. For $\psi_s = 0^\circ$, the vertical polarization component of $\mathbf{E}_v^i$ is parallel to the intersection lines between the plates [156, 157]. The dihedral is useful for polarimetric radar measurements since the $\mathbf{S}_{\text{LP}}$ contains information on all polarization combinations of the incident and reflected waves. For example, at $\psi_s = 0^\circ$, the anti-diagonal entries of $\mathbf{S}_{\text{LP}}$ are zero. In contrast, for $\psi_s = 45^\circ$, the diagonal entries are null, which means that a dihedral maintains and flips the linear polarizations at $\psi_s = 0^\circ$ and $\psi_s = 45^\circ$, respectively. Furthermore, $\mathbf{S}_{\text{CP}}$ [Equation 6.12b] indicates that the dihedral represents scatters with an even number of reflections since the handedness of the circular polarization does not change.

The final canonical scatterer is the dipole, whose polarization scattering matrices read

$$\mathbf{S}_{\text{LP}}^{\text{dipole}} = \begin{bmatrix} \cos^2(\psi_s) & \sin(\psi_s) \cos(\psi_s) \\ \sin(\psi_s) \cos(\psi_s) & \sin^2(\psi_s) \end{bmatrix}, \quad (6.13a)$$

$$\mathbf{S}_{\text{CP}}^{\text{dipole}} = \frac{1}{2} \begin{bmatrix} \exp(j2\psi_s) & 1 \\ 1 & \exp(-j2\psi_s) \end{bmatrix}. \quad (6.13b)$$

Similar to the dihedral, $\mathbf{S}_{\text{LP}}$ depends on the orientation angle ψ_s . For $\psi_s = 0^\circ$, only S_{hh} is scattered back, which means that the horizontal component of the incident electric field is parallel to the dipole for this angle. In Section 6.4, targets with only plate and dihedral scattering mechanisms will be used. Additionally, the amplitudes of Equations 6.11, 6.12, and 6.13 are normalized to unity. These amplitudes practically represent each scatterer's RCS level, which can be determined analytically or by RCS measurement techniques [158, 159].

6.2.3 Polarimetric Decomposition

Polarimetric radar measurements are widely used in synthetic aperture radar (SAR) to obtain colored-like images representing different polarizations. To do so, polarimetric decomposition techniques are usually utilized to describe the information obtained from the polarization scattering matrix. Furthermore, the decomposition

techniques are categorized into *coherent* and *incoherent* decompositions [160]. Here, only the first class will be considered, which is restricted to point scatterers. The most used scattering decomposition technique is the *Pauli* decomposition, which measures the $\mathbf{S}_{\text{LP}}$ in the Pauli basis $(\mathbf{S}_a, \mathbf{S}_b, \mathbf{S}_c, \mathbf{S}_d)$, given by the following matrices

$$\mathbf{S}_a = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \quad (6.14a)$$

$$\mathbf{S}_b = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 0 \\ 0 & -1 \end{bmatrix}, \quad (6.14b)$$

$$\mathbf{S}_c = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix}, \quad (6.14c)$$

$$\mathbf{S}_d = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}. \quad (6.14d)$$

Due to the reciprocity of the monostatic radar, the anti-diagonal entries of $\mathbf{S}_{\text{LP}}$ are identical, i.e., $S_{hv} = S_{vh}$. Therefore, $\mathbf{S}_c$ and $\mathbf{S}_d$ do not provide any additional information. Hence, $\mathbf{S}_{\text{LP}}$ in Equation 6.9 can be rewritten in terms of the Pauli basis as [160]

$$\mathbf{S}_{\text{LP}} = a\mathbf{S}_a + b\mathbf{S}_b + c\mathbf{S}_c, \quad (6.15)$$

where a , b and c are complex coefficients and can be calculated from the components of the (h, v) basis, as

$$a = \frac{S_{hh} + S_{vv}}{\sqrt{2}}, \quad (6.16a)$$

$$b = \frac{S_{hh} - S_{vv}}{\sqrt{2}}, \quad (6.16b)$$

$$c = \sqrt{2}S_{hv}. \quad (6.16c)$$

These coefficients represent the scattering behavior of the canonical scatters discussed in Section 6.2.2. Specifically, $\mathbf{S}_a$ represents scatterers with odd-bounce reflections, such as a plate or trihedral. In comparison, $\mathbf{S}_b$ represents scatterers with even-bounce reflections, such as a dihedral with $\psi_s = 0^\circ$. Furthermore, $\mathbf{S}_c$ represents scatterers with volume-type reflections, which correspond to a dihedral with $\psi_s = 45^\circ$. Finally, each matrix [Equation 6.16] is assigned to a color, usually, *red-green-blue* (RGB), representing the corresponding scattering mechanism of a target. This representation will be shown in Section 6.4.2 for a two-target measurement scenario.

6.3 System Description

The FMCW radar used for the demonstration in the next section is based on the AWR2243 automotive transceiver chip from Texas Instruments, Inc., operating in the frequency range of 76 – 81 GHz [161]. The chipset has three Tx and four Rx channels, but the following demonstration requires only two Tx and two Rx channels. A custom board is fabricated to connect the antennas to the transceiver chip with the

Table 6.1: Chirp parameters of the FMCW radar used in the demonstration in Section 6.4.

parameter	value	unit	description
N_{ADC}	512	—	number of ADC samples
N_{chirp}	128	—	number of chirps
f_s	6	MHz	sampling rate
δ_c	50	MHz/ μs	chirp slope rate
f_{start}	76.00	GHz	start frequency
f_{stop}	80.99	GHz	stop frequency
T_c	99.80	μs	total chirp time
T_{start}	5.7	μs	ADC start time
T_{ADC}	85.33	μs	ADC sampling time
T_{idle}	150	μs	idle time

help of a CPW-to-waveguide (WR-12) transition. Additionally, the chip's internal ADCs digitize the captured baseband data with 16-bit resolution, where the IF data are transferred to a host PC via an acquisition FPGA board. The host PC also controls the radar board and sets the radar parameters according to [162]. Table 6.1 lists the radar parameters used for the demonstration. Some parameters are not related to the measurement with the P-LWA but only to the hardware. For example, the idle time, T_{idle} , is necessary to allow time for the chip to transfer the data from the ADC buffer to the FPGA board. Additionally, the frequency bandwidth programmed is $B = \delta_c T_c = 4.99 \text{ GHz}$. Still, practically the bandwidth is smaller since the ADCs need time at the start of the chirp to achieve the best quality of the signal envelope settling. Hence, $T_{\text{start}} = 5.7 \mu\text{s}$ was chosen, as recommended in [162]. For each measurement, 128 chirps are configured and superimposed coherently in the digital domain to improve the signal-to-noise ratio (SNR).

6.4 Show Cases

This section demonstrates the measurement results with the FMCW radar and the quad-polarized P-LWAs [Section 4.2]. The used FMCW radar has a limited bandwidth of 5 GHz, which leads to a limited scan range of the antenna of about $\Delta\vartheta_{\text{mb}} \approx 6^\circ$. However, it will be shown that the angular resolution can be enhanced by leveraging the polarimetric information. For the measurements, three targets are used: a dihedral, a trihedral, and a substrate-based grid polarizer backed by a plate. The last target is similar to a dihedral, except for the symmetric RCS patterns, irrespective of the target's orientation. The principle of design can be found in [163], where such a target is used to calibrate a full-polarimetric automotive radar. The window width of the STFT is chosen to time gate 64 of the 512 captured samples, which corresponds to an effective bandwidth of $B_{\text{eff}} = 620 \text{ MHz}$ or $\Delta r \approx 24 \text{ cm}$. A shift of 4 samples or about 33 MHz is used for each adjacent STFT window. To obtain the orthogonal polarizations, the measurement is conducted by applying the time-division multiplexing scheme, where a pair of chirps are transmitted corresponding either to (h, v) or (r, l) . The orthogonal linear polarizations are ob-

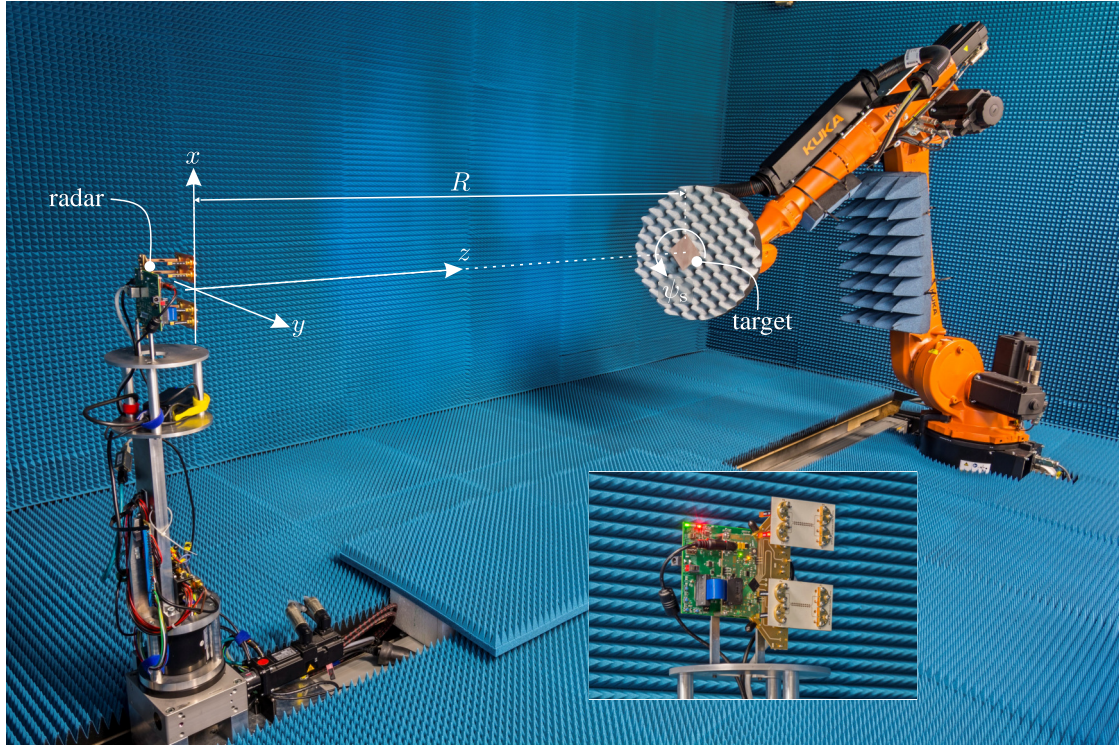


Figure 6.5: Measurement setup with the FMCW radar and the quad-polarized P-LWA [Section 4.2]. The setup consists of a single static target mounted on a robotic arm located at a distance R . A zoomed view of the radar is shown in the inset.

tained by combining the Tx and Rx channels in-phase and out-of-phase according to the discussion [Section 4.2]. The demonstration will be divided into two parts. First, static single-target measurements will be shown. Second, combined static- and dynamic-target measurements will be presented to show the combined angular and polarimetric capabilities of the quad-polarized P-LWAs. The measurements are performed inside the anechoic chamber of the IHF at RWTH Aachen University with the robotic system described in [164]. It should be noted that for all the following measurement results, background subtraction is applied to eliminate unwanted reflections from the robotic arm and surrounding objects.

6.4.1 Polarimetric Radar Measurements

Figure 6.5 shows the measurement setup of a single target located at a distance $R = 2.3$ m from the FMCW radar. Furthermore, the orientation angle of the target, ψ_s , can be rotated using the robotic arm. Moreover, the P-LWAs scan the main beam over frequency in the yz -plane, as described in the previous chapters. The first measurement utilizes the circular polarization of the P-LWAs. Therefore, a dihedral (even-bounce) and a trihedral (odd-bounce) are used both with the orientation $\psi_s = 0^\circ$. The range Fourier transform is processed for the sliding window with the frequency (78.5 ± 0.31) GHz. Figure 6.6(a) shows the range profile of the dihedral measurement. As expected, the reflection from the dihedral corresponds to even-bounce reflections [see Equation 6.12b], which can be seen at $r = R$. The dis-

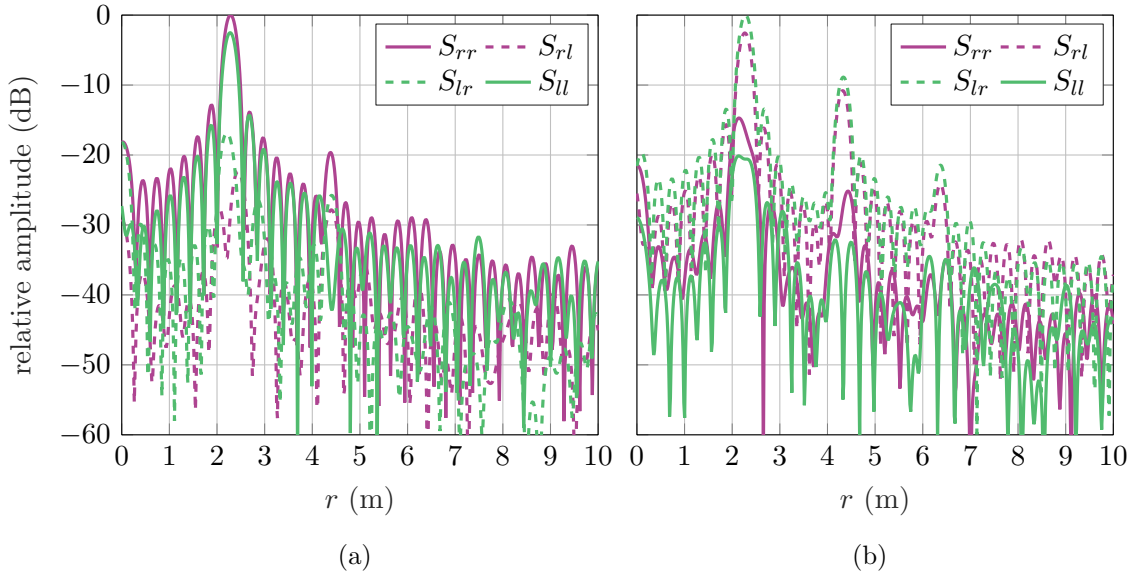


Figure 6.6: Full-polarimetric radar measurement with the quad-polarized P-LWAs obtained in the circular-polarization basis $(\hat{r}, \hat{l})$. (a) Dihedral and (b) trihedral.

crimination between the diagonal and anti-diagonal entries is about 15–20 dB, which corresponds to the circular-polarization purity of the antennas [cf. Figure 4.15]. It can also be observed that the amplitudes of the r - and l -channels are unbalanced. To correct these amplitudes, a calibration measurement has to be performed, which is omitted without loss of generality in this work. The measurement of the trihedral is plotted in Figure 6.6(b). Again, the measurement shows the expected results of an odd-bounce reflection from the trihedral, which is indicated by the high amplitude of the anti-diagonal entries of S_{CP}^{plate} [Equation 6.11b]. It is worth noting that the peaks at multiples of R , i.e., $2R$ and $3R$ correspond to the multiple reflections between the target and the radar.

The second measurement utilizes the linear polarizations of the P-LWAs, which are obtained using the even and odd excitations. For this measurement, the grid polarizer is used. As mentioned before, this target is similar to a dihedral, which means that for $\psi_s = 0^\circ$ or $\psi_s = 45^\circ$ the incident linear polarizations are either maintained or flipped when reflected from the grid polarizer, respectively. Figures 6.7(a) and 6.7(b) show the measurement results for $\psi_s = 0^\circ$ or $\psi_s = 45^\circ$, respectively. The results clearly show that the radar is able to distinguish the different polarization components depending on the orientation. Additionally, one can be noticed that discrimination of the diagonal to anti-diagonal components for $\psi_s = 45^\circ$, which is attributed to the cross-polarization discrimination of the grid polarizer. Specifically, the cross-polarization level of the grid polarizer for this orientation is about 13 dB, which is obtained from full-wave simulations.

6.4.2 Hybrid Angular-Polarimetric Measurement

To demonstrate the range-angle measurement of the FMCW radar with P-LWAs, two targets are used and located at the same range, $R \approx 1.65$ m. Figure 6.8 shows

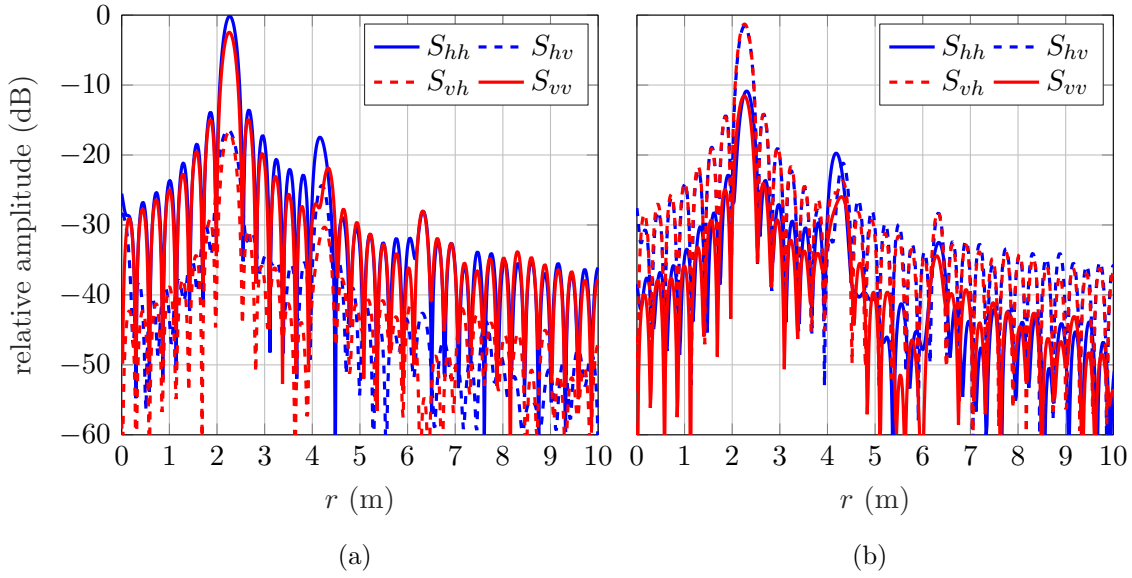


Figure 6.7: Full-polarimetric radar measurement with the quad-polarized P-LWAs obtained in the linear-polarization basis $(\hat{h}, \hat{v})$. Measurement with a grid polarizer with (a) $\psi_s = 0^\circ$ and (b) $\psi_s = 45^\circ$.

a photograph of the measurement setup, where a dihedral with $\psi_s = 0^\circ$ is mounted on a tripod, and a grid polarizer with $\psi_s = 45^\circ$ is mounted on the robotic arm. Furthermore, the robotic arm is moved on a circle around the P-LWA while keeping R constant, which results in an angle ϑ_s . For $\vartheta_s = 0^\circ$, the grid polarizer is aligned to the broadside direction of the antenna. The robotic arm is moved in the range from $\vartheta_s = -25^\circ$ to $\vartheta_s = +25^\circ$. The same radar processing parameters used for the single-target measurements in Section 6.4.1 will be used here.

Figure 6.9 shows the measurement results represented by the Pauli decomposition for five positions of the grid polarizer, i.e., $\vartheta_s = -10^\circ, -5^\circ, 0^\circ, 5^\circ$, and 10° . Figure 6.9(a) plots the RGB-colored range-angle profile of the radar measurement. It can be seen that for $\vartheta_s = \pm 10^\circ$, a green-coded target is located at the expected range ($r = 1.65$ m). The green color means that an even-bounce reflection occurs at the target [Section 6.2.3], which is expected since only the dihedral with $\psi_s = 0^\circ$ is seen by the radar. As the grid polarizer moves closer to the radar's field of view, another target starts to appear. This time the object is blue-coded, which means a reflection from a volume-like scatterer occurs. A reflection from the dihedral or the grid polarizer with $\psi_s = 45^\circ$ is considered as a volume-like scatterer, as described in Section 6.2.3. To compare the results quantitatively, Figure 6.9(b) plots the angle profile at the maximum range bin, i.e., $r = 1.65$ m. For the case of $\vartheta_s = \pm 10^\circ$, it can be seen that the maximum relative amplitude is about -15 dB when compared to the maximum at $\vartheta_s = 0^\circ$. The reason for the different reflection amplitudes of both targets is the different heights of the targets, as seen in Figure 6.8. Due to the narrow RCS patterns of the dihedral with $\psi_s = 0^\circ$, the manual alignment of the target is difficult, especially at close distance. Nevertheless, the results show the target moving in the corresponding direction according to ϑ_s . Two measures can be considered to improve the angular resolution further: First, a P-LWA with enhanced frequency scanning

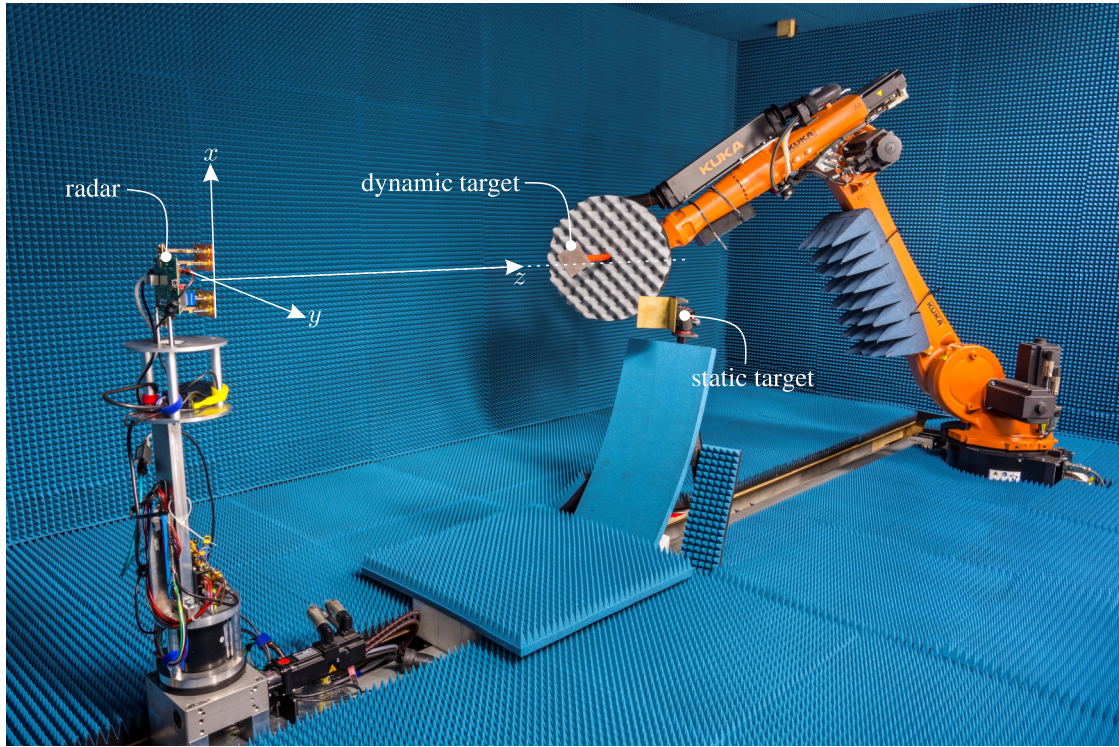


Figure 6.8: Measurement setup with the FMCW radar and the quad-polarized P-LWA [Section 4.2]. The setup consists of two targets located at range $R \approx 1.65$ m: a dihedral mounted on a static tripod and a grid polarizer mounted on the moving robotic arm with circular motion around the radar.

can be used, as discussed in Section 5. Second, a longer P-LWA resulting in a smaller beamwidth can be utilized.

6.5 Summary

This chapter demonstrated the quad-polarized P-LWA presented in Section 4.2. First, the principle of the frequency-scanning radar has been discussed, emphasizing the trade-off between the angle and range resolution. Since the polarization diversity of the proposed antenna design has been utilized, a brief introduction to the polarization scattering matrix of symmetric canonical targets has been recalled. Additionally, the Pauli polarimetric decomposition technique has been presented. The demonstration was performed with an FMCW radar operating at 76 – 81 GHz inside an anechoic chamber with a robotic system. Three targets were considered: a dihedral, a trihedral and a planar grid polarizer. Furthermore, two showcases were investigated. First, a measurement scenario with a single static target was considered, where the radar is configured to send and receive the four possible polarization combinations, dual linear and dual circular. It could be shown that the radar with the developed P-LWAs could discriminate the targets' polarization scattering with better than 15 dB, corresponding to the axial ratio measurement. Second, a measurement scenario was performed with two static and moving targets mounted on the robotic arm and moved in a circular direction around the radar. Although the P-

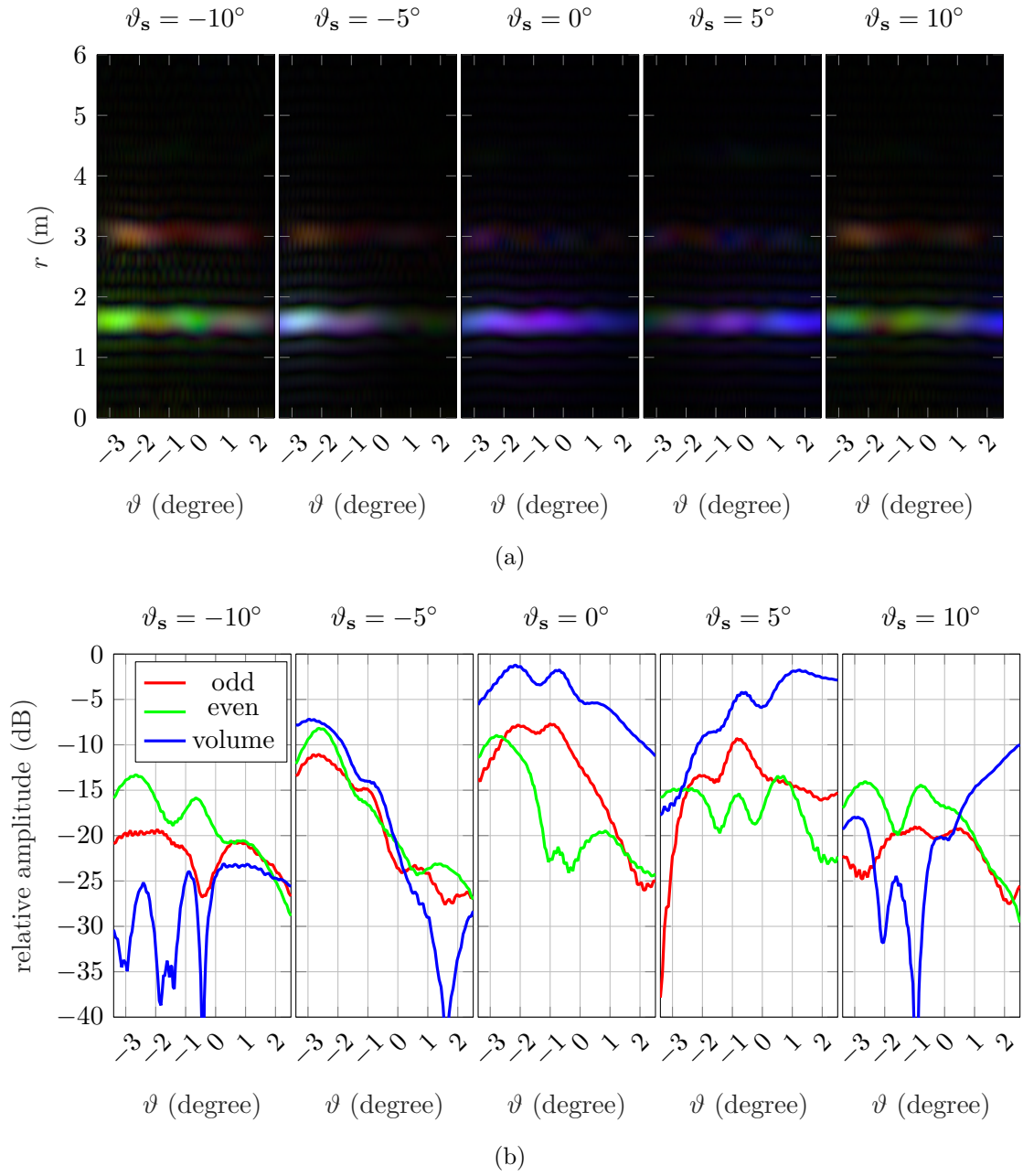


Figure 6.9: Measurement results of the setup shown in Figure 6.8 with different positions of the grid polarizer ϑ_s . (a) Colored range-angle profile and (b) the corresponding angular cuts at $r = 1.65$ m.

LWA has a limited scanning range within the radar's frequency bandwidth, it could be shown that combining the polarization information and the frequency scanning enhances the angular resolution.

Conclusion and Outlook

The main research contributions are summarized in this chapter. Additionally, an outlook for future works and developments is provided.

7.1 Conclusion

This thesis focuses on the theoretical description and solution aspects of the broadside problem in periodic leaky-wave antennas (P-LWAs). These aspects aim for the practical implementations of efficient P-LWAs with polarization-diverse radiation and enhanced frequency scanning, which are suitable for both low-cost and advanced radar applications.

First, the theory of LWAs and their subclass, P-LWAs, was introduced in Chapter 2. Then Chapter 3 described the *open-stopband* problem in P-LWAs, where the antenna's parameters, Bloch impedance, propagation constant, and radiation efficiency, exhibit an abrupt change when the main beam scans through the broadside. Furthermore, formulas for these parameters were provided with the help of an established lattice network model. The lattice model showed that the stopband closure is only possible if two conditions are satisfied for the normal modes of a P-LWA, the series and shunt modes. These conditions are *frequency-balancing* (ω -balancing) and *quality-factor-balancing* (Q -balancing). Numerical examples were given for the four operation regimes of P-LWA: 1) ω -unbalancing/ Q -unbalancing, 2) ω -balancing/ Q -unbalancing, 3) ω -unbalancing/ Q -balancing, and 4) ω -balancing/ Q -balancing. The fourth operation regime ensures optimal operation for P-LWA, which cannot be achieved with a double-symmetric unit cell. Hence, another established network model for *transversally-asymmetric* unit cells is presented. It consists of ideal transformers with their transformation ratio corresponding to the coupling between the series and shunt modes. The optimal transformation ratio, and therefore, the optimal coupling, mitigates the stopband problem at the broadside. Radiative and bounded periodic structures, whose unit cells are transversally asymmetric, were investigated. Both structures are based on a non-TEM transmission line, i.e., substrate-integrated waveguide operating around the broadside frequency 78 GHz. The first structure is the well-known waveguide-based P-LWA with longitudinal slots, where the transverse asymmetry is implemented by a simple via that is shifted along the longitudinal axis of the P-LWA. The second structure is a series-feeding network whose outputs correspond to the radiating elements in a P-LWA. Also, for this structure, it has been shown that the broadside problem can be solved using transverse asymmetry, which was implemented using asymmetric inductive posts. Finally, an electromagnetic perspective is provided by utilizing the coupled-mode theory to describe the coupling between the series and shunt modes. First, the coupled-mode equations were de-

rived, and an analytical expression for the coupled eigenfrequencies in P-LWAs was introduced. Second, the propagation constant for transversally-asymmetric unit cells was derived analytically by mapping the derived expression for the coupled eigenfrequency. Moreover, it could be shown that a specific amount of coupling corresponds to the optimal operation of the P-LWA, similar to the transformer network model. In Chapter 4, it has been shown that the symmetry of the unit cell in P-LWAs governs its polarization. Specifically, an optimal *longitudinal asymmetry* can achieve perfect circular polarization with the unity of the axial ratio defined by the *Q*-balancing condition. Furthermore, orthogonal circular-polarization handedness can be achieved when exciting the P-LWA's ends sequentially. A design of a series-fed patch P-LWA operating at the broadside frequency 78 GHz was fabricated and measured, confirming the circular polarization. Additionally, a more practical implementation of polarization-diverse P-LWA based on the longitudinal-asymmetry concept was presented by combining two identical circularly-polarized P-LWAs in a mirrored configuration, allowing the radiation of quad polarizations. Dual circular polarizations when exciting either P-LWAs or dual-linear polarization when exciting both P-LWAs in-phase or out-of-phase. Again, the same fabricated longitudinally-asymmetric series-fed patch structure is used to construct the proposed quad-polarized P-LWA. The measurement confirmed the quad-polarization radiation. Finally, the combination of longitudinal and transversal asymmetries, resulting in *double asymmetry*, achieves orthogonal-linear polarization corresponding to the polarizations of the series and shunt modes. The longitudinal asymmetry provides equal radiation contributions from the series and shunt modes, and the transverse asymmetry provides mode coupling to selectively excite each mode individually, depending on the excitation side of the P-LWA. For this concept, two P-LWAs, a series-fed patch, and a series-fed coupled patch, operating with a broadside frequency of 24 GHz, were fabricated and measured.

Chapter 5 is divided into two parts. The first part provided concepts to systemically enhance the frequency scanning rate and range in 1D P-LWAs. Two general techniques were identified: 1) the serpentine-line technique, which provides a delay between the radiation elements, and 2) the dispersive-line technique, which controls the slope of the phase constant and, therefore, the scanning rate. For the latter technique, two structures were investigated, a loaded series-fed coupled patch operating at 24 GHz and a series-feeding network with Vivaldi radiation elements operating at 78 GHz. The second part presented a concept of a 2D P-LWA consisting of a series-feeding network terminated by an array of P-LWAs, allowing for frequency scanning in two directions. The series-feeding network and the P-LWA provide two independent and orthogonal scanning planes. Unlike conventional 2D P-LWAs with conical-beam radiation, the presented concept achieves pencil-beam radiation while maintaining the polarization of the unit cells. Furthermore, the concept was analyzed using the array-factor theory and analytical dispersion diagrams. Finally, a 2D P-LWA structure based on the slotted substrate-integrated waveguide is fabricated and measured, confirming the proposed concept.

Finally, Chapter 6 demonstrated the quad-polarized P-LWA with an FMCW radar operating in the frequency range of 76 – 81 GHz. Furthermore, the concept of frequency-scanning radars was introduced and followed by the definition of the polarization scattering matrix for polarimetric radars. The FMCW radar was fabricated

using commercial off-the-shelf components and modular connections for the transmitter and receiver antennas. The FMCW radar system with the quad-polarized antennas was measured in an anechoic chamber with a robotic arm. Although the scanning range of the quad-polarized antenna is limited; however, combining the polarimetric and frequency information enhances angular detection.

7.2 Outlook

The outlook for future work is divided into three parts: general P-LWA theory and modeling, advanced radar implementations, and other emerging applications.

The expression for the coupled eigenfrequencies was derived in Chapter 3 assuming coupling in time between the series and shunt modes using the *eigenmode approach*. However, this approach does not represent the normal operation of the P-LWA, where it is excited at one end and terminated with a matched load at the other. Hence, the coupled-mode analysis can be extended to the *driven-mode approach*, which would require the formulation of the counterpart problem of coupling in space. Furthermore, the formulation of the problem could require the description of the coupling factor as a function in space, which is much more complicated than the coupling in time but would allow capturing the physics of the coupling due to the transversal asymmetry.

Conventional methods to achieve radiation with low sidelobe level, e.g., Chebyshev tapering, are challenging to implement in P-LWA since the tapering of the radiating elements alters the unit cell's parameters. Hence, even with the individual tuning of the unit cells, there will be an inner mismatch between successive unit cells. However, the series-feeding network based on the substrate-integrated waveguide might be an exception since an optimally designed unit cell has approximately the same impedance as the unperturbed waveguide. Therefore, optimizing a set of unit cells with different power-splitting ratios and similar Bloch impedances is possible, which results in a P-LWA design with a low-sidelobe level for wideband applications.

The combined concept of 2D P-LWA and serpentine lines presented in Chapter 5 achieves asymmetric scanning planes when the structures are excited at opposite sides of the series-feeding network. Hence, one potential implementation would be maximizing the scanning rate from one port and achieving a squintless beam from the other, which is very attractive for both high angular and high-range resolution.

The radar demonstration presented in this work is based purely on amplitude discrimination (except for the Pauli decomposition) over angle. If the phase center of the P-LWA is known, then conventional angle-of-arrival algorithms, such as MUSIC, can be applied to enhance the angular resolution further. For the case of the FMCW radar, the phase of the transmitted signal over frequency has to be considered as well.

Furthermore, the polarimetric radar implementation with the quad-polarized P-LWA with elevation frequency scanning requires two transmitter and receiver channels. Such implementation does not allow the scanning in the azimuth direction

using conventional beamforming techniques (phased array) since cross-polarized antennas receive the reflected signal. One way to overcome this limitation is to apply a hybrid polarization scheme, e.g., transmitting linear polarization and receiving orthogonal circular polarizations. Two steps are required to implement the hybrid polarization, a calibration procedure is required for the phases of the received signals, and the formulation of the scattering polarization matrix has to be established.

In the coming *sixth-generation standard for mobile communication*, proposals are reported to use reconfigurable intelligent surfaces, which allow circumventing obstacles at high frequencies when a direct line-of-sight communication link between a transmitter and a receiver is not possible. The term “intelligent” implies that the surface has a priori knowledge of the relative locations of both transmitter and receiver. Since the proposed surfaces must be low-cost, two 1D P-LWAs in an orthogonal configuration can be used to determine the angle of arrival in both azimuth and elevation directions.

Finally, *terahertz communication and sensing* is another emerging technology with an extreme frequency bandwidth available. At these frequencies, the free-space path loss significantly impacts a system’s power budget. Additionally, on-chip antennas are also needed at the terahertz frequency range to eliminate the losses of the interconnections between the chip and antennas. However, such antennas suffer from limited frequency bandwidth and low input matching. Hence, P-LWAs can be utilized for these applications, providing high-gain pencil beams and broadband frequency matching.

Appendix A

Network Model Derivation of P-LWA Parameters

This appendix provides the derivation of the Bloch impedance and radiation efficiency from the lattice and transformer network models.

A.1 Lattice Circuit Model

Figure A.1 shows the lattice network model presented in Section 3.4.3 with voltage and current notations to help with the derivation of the Bloch impedance and radiation efficiency. The derivation is taken from [23].

A.1.1 Bloch Impedance

Generally, the Bloch impedance can be derived by calculating the voltage-to-current ratio at the unit cell's terminal [70]. The forward Bloch impedance Z_B^+ seen from the n^{th} terminal reads

$$Z_B^+ = \frac{V_n}{I_n}. \quad (\text{A.1})$$

First, the current flowing through the n^{th} terminal can be calculated by applying Kirchhoff's current law at the node k_1 as

$$I_n = I_{se} + I_{sh}/2, \quad (\text{A.2})$$

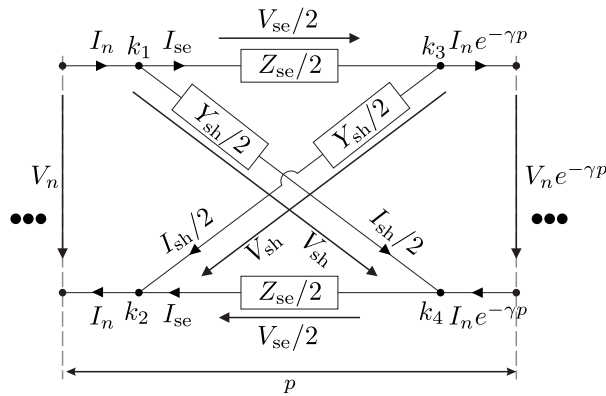


Figure A.1: Lattice network model Transformer network model [Figure 3.5] with voltage, current and reference point notations required for the derivation of the Bloch impedance and radiation efficiency.

and similarly at the node k_3 as

$$I_{n+1} = I_{se} - I_{sh}/2. \quad (\text{A.3})$$

The sum of Equations A.2 and A.3 leads to a relation between the series current I_{se} and the current at the n^{th} terminal, which reads

$$I_{se} = \frac{I_n + I_{n+1}}{2} = \frac{I_n (1 + e^{-\gamma p})}{2}. \quad (\text{A.4})$$

Additionally, the voltage drop the the n^{th} terminal can be calculated by evaluating the current loop $k_1 - k_4 - k_3 - k_2 - k_1$, which yields

$$V_{sh} - V_{n+1} + V_{sh} - V_n = 0, \quad (\text{A.5})$$

where the above expression can be simplified further by substituting V_{n+1} by $V_n e^{-\gamma p}$ as

$$V_{sh} = \frac{1}{2} (V_n + V_{n+1}) = \frac{V_n}{2} (1 + e^{-\gamma p}). \quad (\text{A.6})$$

Finally, the Bloch impedance is obtained by diving Equation A.6 by Equation A.4 as follows

$$Z_B^+ = \frac{V_{sh}}{I_{se}} = \frac{I_n (1 - e^{-\gamma p})}{V_n (1 - e^{-\gamma p})} \frac{Z_{se}}{Y_{sh}}. \quad (\text{A.7})$$

Replacing V_n/I_n by Z_B^2 [Equation A.1] yields

$$Z_B^+ = \frac{1}{Z_B^+} \frac{Z_{se}}{Y_{sh}} = \sqrt{\frac{Z_{se}}{Y_{sh}}}. \quad (\text{A.8})$$

Using the linearized series impedance and shunt admittance from Equation 3.22 yields the final result for the Bloch impedance

$$Z_B^+ = \sqrt{\frac{R + j2L(\omega - \omega_{se})}{G + j2C(\omega - \omega_{sh})}}. \quad (\text{A.9})$$

Since the lattice network model is symmetric, the derived forward Bloch impedance differs only from its backward counterpart by a sign, i.e., $Z_B^+ = -Z_B^-$.

A.1.2 Radiation Efficiency

The radiation efficiency [Equation 3.33] can be derived by first calculating the total and radiated power from the series and shunt paths. The total power is the sum of the series and shunt powers, which mathematically reads

$$P_{rad} = P_{se} + P_{sh}, \quad (\text{A.10})$$

and similarly, the total radiated power reads

$$P_{rad} = P_{se,rad} + P_{sh,rad}. \quad (\text{A.11})$$

Furthermore, Equation 3.33 can be rewritten using Equations A.10 and A.11 as

$$\eta = \frac{P_{\text{se,rad}} + P_{\text{sh,rad}}}{P_{\text{se}} + P_{\text{sh}}}, \quad (\text{A.12})$$

where the series and shunt powers can be expressed by the series current flowing through R and shunt voltage drop across G , as follows

$$P_{\text{se}} = \frac{1}{2} |I_{\text{se}}|^2 R, \quad (\text{A.13a})$$

$$P_{\text{sh}} = \frac{1}{2} |V_{\text{sh}}|^2 G. \quad (\text{A.13b})$$

Analogously, the radiated series and shunt power read

$$P_{\text{se,rad}} = \frac{1}{2} |I_{\text{se}}|^2 R_{\text{rad}}, \quad (\text{A.14a})$$

$$P_{\text{sh,rad}} = \frac{1}{2} |V_{\text{sh}}|^2 G_{\text{rad}}. \quad (\text{A.14b})$$

Finally, the radiation efficiency can be calculating by inserting Equations A.13 and A.14 in Equation A.12, which yields

$$\eta = \frac{|I_{\text{se}}|^2 R_{\text{rad}} + |V_{\text{sh}}|^2 G_{\text{rad}}}{|I_{\text{se}}|^2 R + |V_{\text{sh}}|^2 G} = \frac{R_{\text{rad}} + G_{\text{rad}} |V_{\text{sh}}/I_{\text{se}}|^2}{R + G |V_{\text{sh}}/I_{\text{se}}|^2}. \quad (\text{A.15})$$

Substituting $V_{\text{sh}}/I_{\text{se}}$ by Z_{B}^+ [Equation A.7] yields

$$\eta = \frac{\eta_{\text{se}} R + \eta_{\text{sh}} G |Z_{\text{B}}^2|^2}{R + G |Z_{\text{B}}^2|^2}. \quad (\text{A.16})$$

As discussed in Section 3.4.3.3, Equation A.16 shows that the radiation efficiency is impacted by the Bloch impedance and that the equalization condition at the broadside can only be achieved by equalizing the Bloch impedance.

A.2 Transformer Network Model

This section recalls the derivation of the formulas for the Bloch impedance and the radiation efficiency using in Section 3.5.1 in terms of the the transformation ratio, T . Figure A.2 shows the transformer network model consisting of the series impedance, shunt admittance and four ideal transformers with transformation ratio T . The transformation ration models the level of the transversal asymmetry in the unit cell. Furthermore, Figure A.2 provides the voltage, current and reference point notations, which are required for the derivation. The derivation is taken from [25, 26].

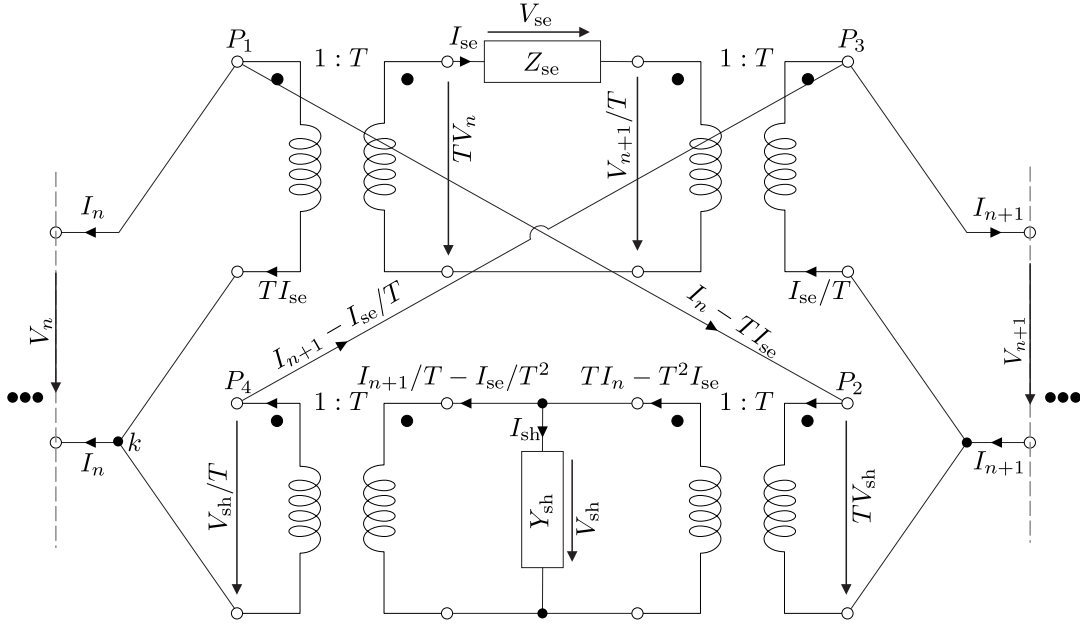


Figure A.2: Transformer network model [Figure 3.8] with voltage, current and reference point notations required for the derivation of the Bloch impedance.

A.2.1 Bloch Impedance

Following the same procedure used in Section A.1, the Bloch impedance can be derived by calculating the voltage and current at the n^{th} terminal of the unit cell. Applying the Kirchhoff's voltage law to the loop $P_1 - P_2 - P_3 - P_4 - P_1$ and by using the Bloch theorem yield

$$V_{\text{sh}} \left(T + \frac{1}{T} \right) = V_n (1 + e^{-\gamma p}), \quad (\text{A.17})$$

where V_{sh} is the voltage drop across the shunt resonator. Next, applying the Kirchhoff's current law to node k and using the Bloch theorem similarly yields

$$I_{\text{se}} \left(T + \frac{1}{T} \right) = I_n (1 + e^{-\gamma p}). \quad (\text{A.18})$$

Similar to Equation A.7, the Bloch impedance is therefore simply obtained by dividing Equation A.17 by Equation A.18, which yields

$$Z_{\text{B}} = \frac{V_n}{I_n} = \frac{V_{n+1}}{I_{n+1}} = \frac{V_{\text{sh}}}{I_{\text{se}}}, \quad (\text{A.19})$$

To express $V_{\text{sh}}/I_{\text{se}}$ in terms of T , Z_{se} and Y_{sh} , the Kirchhoff's voltage law is applied to the Z_{se} loop and Kirchhoff's current law to the Y_{se} node, resulting into

$$I_{\text{se}} Z_{\text{se}} = TV_n - \frac{V_{n+1}}{T}, \quad (\text{A.20})$$

and

$$V_{\text{sh}} Y_{\text{sh}} = TI_n - \frac{I_{n+1}}{T} - 2I_{\text{se}} \tau, \quad (\text{A.21a})$$

where

$$\tau = \frac{1}{2} \left(T^2 - \frac{1}{T^2} \right). \quad (\text{A.21b})$$

In Equation A.21a, I_n and I_{n+1} may be expressed in terms of Z_B , V_n and I_{n+1} using Equation A.8, which yields

$$V_{\text{sh}} Y_{\text{sh}} = \frac{1}{Z_B} \left(T V_n - \frac{V_{n+1}}{T} \right) - 2 I_{\text{se}} \tau. \quad (\text{A.22})$$

It is worth noting that the bracket term in this expression is exactly the right-hand side of Equation A.20, this relation transforms to

$$V_{\text{sh}} Y_{\text{sh}} = \frac{I_{\text{se}} Z_{\text{se}}}{Z_B} - 2 I_{\text{se}} \tau, \quad (\text{A.23})$$

Upon division by I_{se} , Equation A.23 provides the required $V_{\text{sh}}/I_{\text{se}}$ ratio in Equation A.8, as

$$Z_B^2 + \frac{2\tau}{Y_{\text{sh}}} Z_B - \frac{Z_{\text{se}}}{Y_{\text{sh}}} = 0, \quad (\text{A.24})$$

which is a quadratic equation in Z_B , with the two solutions

$$Z_B^{\pm} = \mp \frac{\tau}{Y_{\text{sh}}} + \sqrt{\left(\frac{\tau}{Y_{\text{sh}}} \right)^2 + \frac{Z_{\text{se}}}{Y_{\text{sh}}}}. \quad (\text{A.25})$$

In contrast to Equation A.8, the forward and backward Bloch impedances are different for the case of asymmetric unit cell, i.e., $T \neq 1$. As discussed in Section 3.5.2, either forward or backward Bloch impedance can be equalized through broadside by tuning the unit cell to have the optimal TA level, T_{opt} .

A.2.2 Radiation Efficiency

The derivation of the radiation efficiency for transversally-asymmetric unit cells can be done using the same procedure applied in Section A.1.2. Since the radiation and loss powers in the series and shunt elements are identical to those derived in Section A.1.2, the radiation efficiency for an asymmetric unit cell is identical to the symmetric one, as provided in Equation A.16. It is worth noting that the asymmetry information is provided by the Bloch impedance, and therefore, the radiation efficiency of an asymmetric P-LWA depends on the excitation side (forward or backward).

Appendix B

Feeding Structures

The design description and full-wave simulation results of two feeding transitions used in this thesis are presented. The transitions convert a TE₁₀ mode from a WR-12-waveguide mode either to a quasi-TEM mode (MSL) or to another TE₁₀ mode (SIW), as will be shown in Section B.1 and Section B.2, respectively. The WR-12-waveguide utilizes a round flange with the flange designation UG-387/U.

B.1 MSL-to-Waveguide Transition

The design of the MSL-to-waveguide transition is shown in Figure B.1. The transition is design on RO3003 with height $h = 127\text{ }\mu\text{m}$ and it is used to feed the SFP P-LWA design presented in Section 4.1 and Section 4.2. The design principle is similar to that of the pin-fed waveguide, where the MSL inside the cavity acts as an electric coupler [Figure B.1(c)]. A machined brass backplate with a cavity is used to fix the waveguide to the PCB via three screws, where it is aligned to the transition using dowel pins. The cavity is used to improve the matching and it has a height of 0.5 mm. Additionally, a shorted stub on the top layer is used to increase the frequency bandwidth.

The simulated S-parameters are obtained from CST MWS and are shown in Figure B.2. It can be seen that an excellent matching ($< -15\text{ dB}$) is achieved for the frequency range 65 – 90 GHz with a very low loss $S_{21} = -0.5\text{ dB}$.

B.2 SIW-to-Waveguide Transition

The fabricated P-LWAs based on SIW in this work uses an SIW-to-waveguide transition. The design of the transition is illustrated in Figure B.3. It is designed on RO3003 substrate with the height $h = 508\text{ }\mu\text{m}$. The transition couples the TE₁₀ mode from the air-filled WR-12-waveguide to the TE₁₀ mode in the SIW (dielectric-filled) with the help of a cavity within the substrate. Furthermore, two vias are used to eliminate unwanted resonant modes within the cavity. The backplate is used for this transition to mechanically fix the waveguide to the PCB.

Figure B.4 shows the full-wave simulated S-parameters of the transition obtained from CST MWS. Similar to the MSL-to-waveguide transition, the SIW transition shows an excellent matching ($< -15\text{ dB}$) for the frequency range 67 – 90 GHz. The high reflection at $f \approx 64\text{ GHz}$ is due to a resonant mode in the cavity.

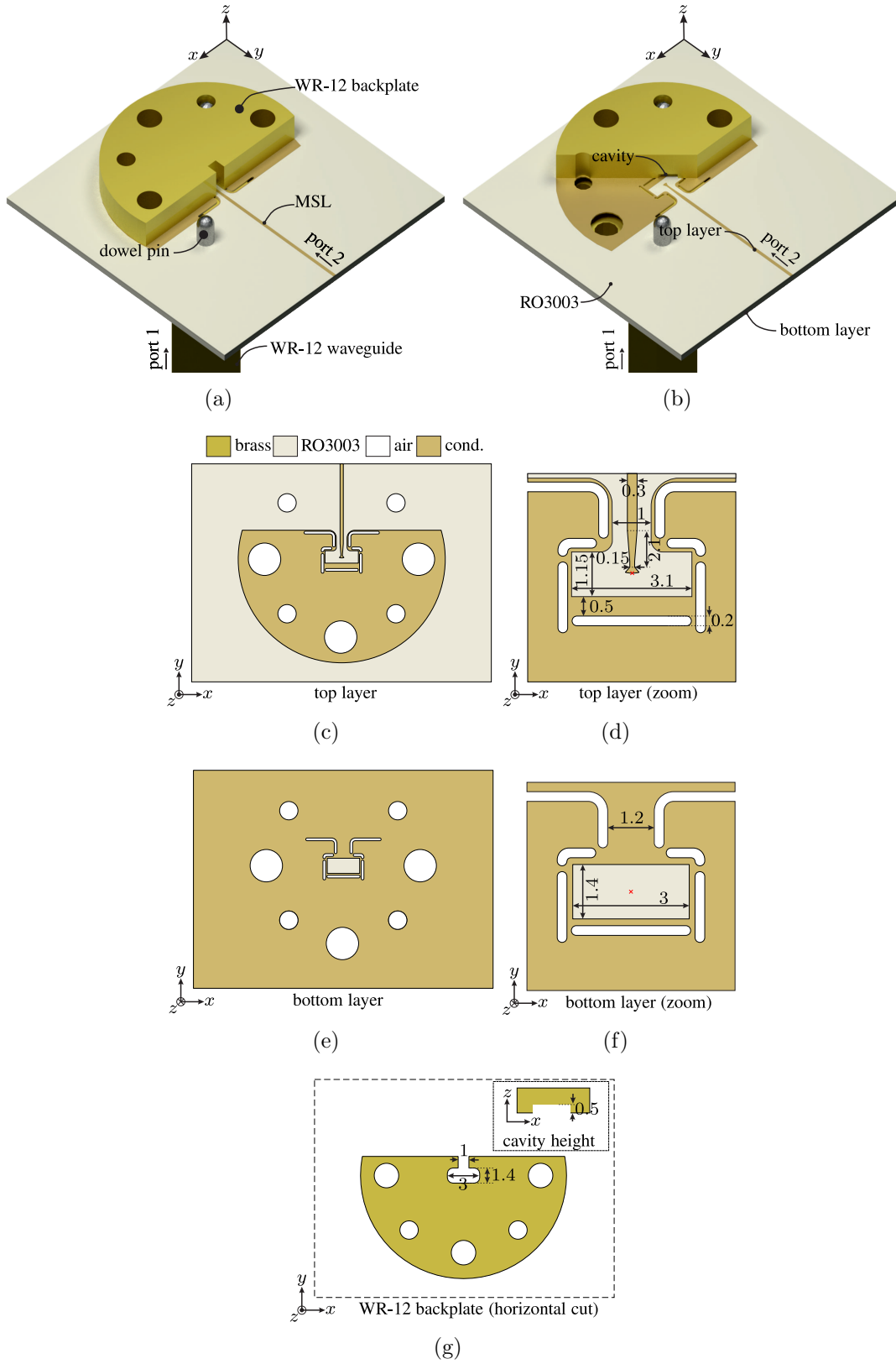


Figure B.1: Design and dimensions (in mm) of the MSL-to-waveguide transition. 3D perspective view of the design (a) with the brass backplate and (b) cut-off view. (c-d) Top and (e-f) bottom views of the transition. (g) Horizontal cut through the cavity in the brass backplate (inset showing the height of the cavity).

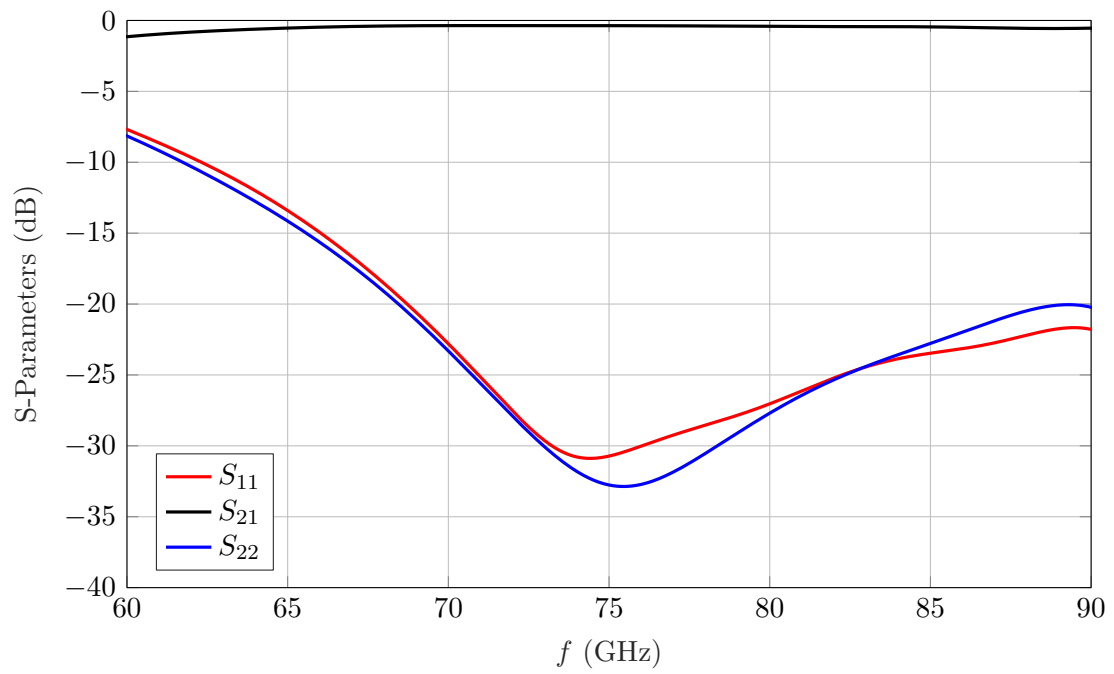


Figure B.2: Full-wave simulated S-parameters of the MSL-to-waveguide shown in Figure B.1. The S_{12} parameter is omitted due to reciprocity.

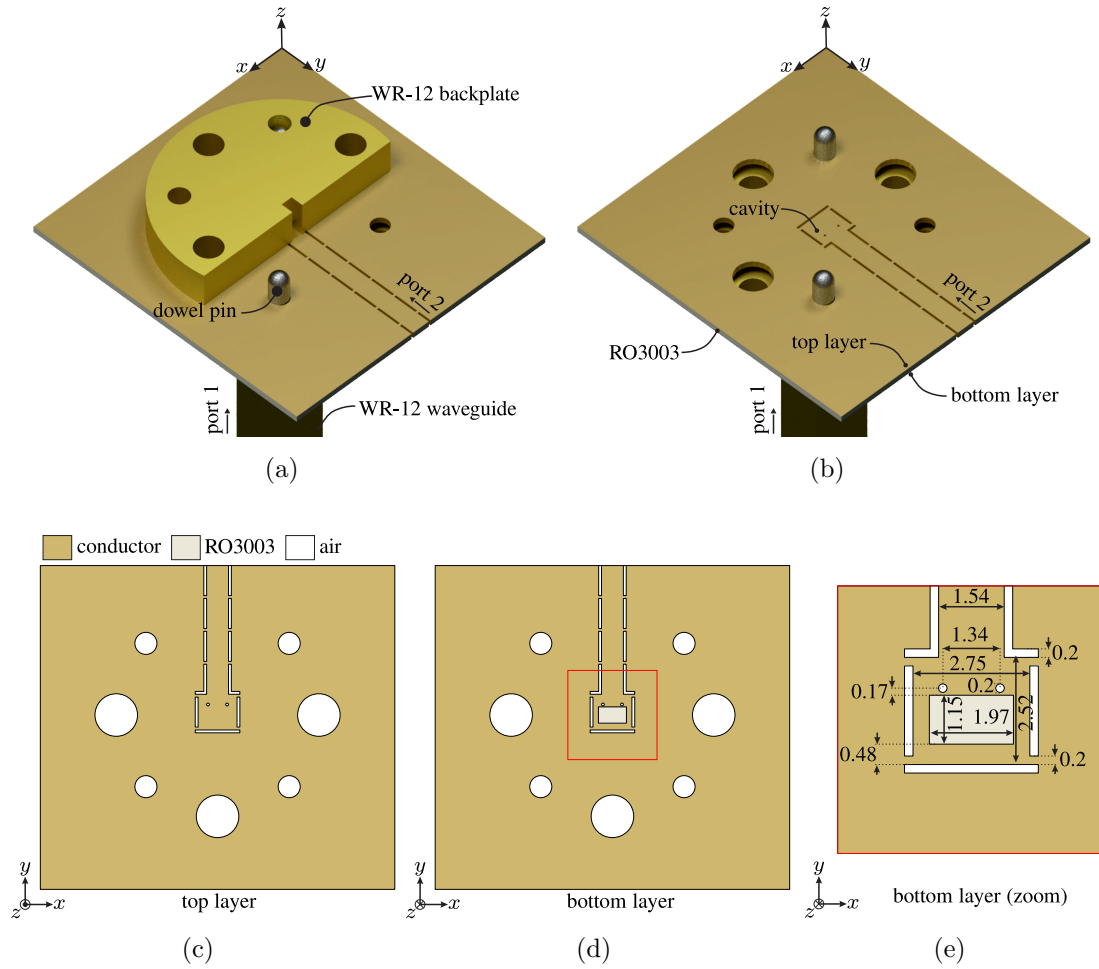


Figure B.3: Design and dimensions (in mm) of the SIW-to-waveguide transition. 3D perspective view of the design (a) with and (b) without the brass backplate. (c) Top and (d) bottom views of the transition. (e) Zoom view of the bottom layer.

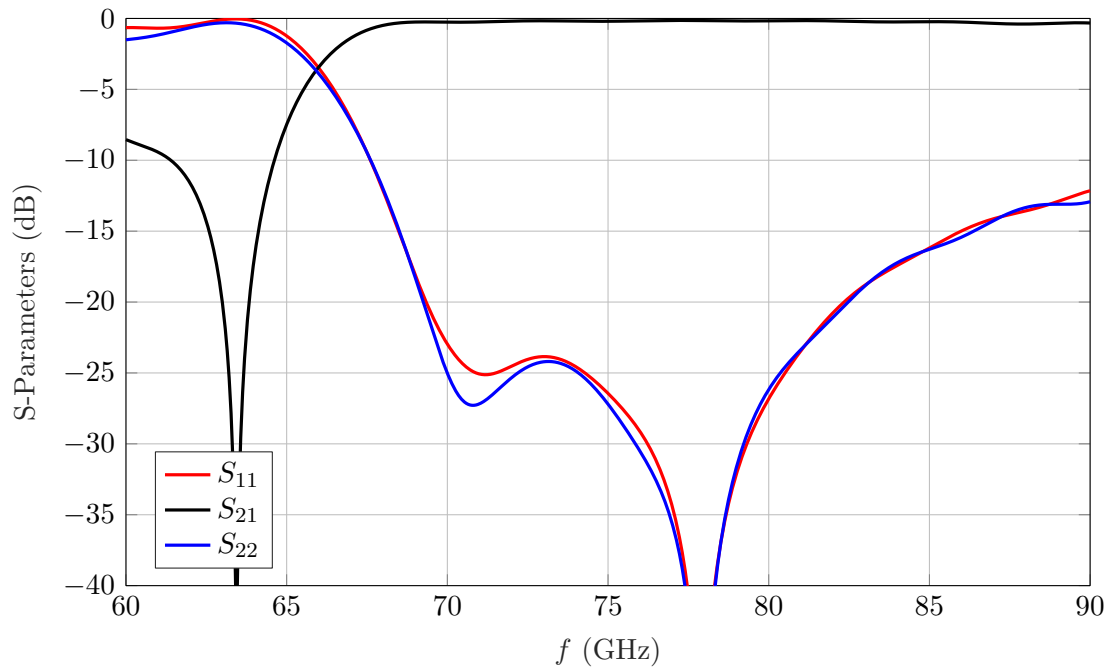


Figure B.4: Full-wave simulated S-parameters of the SIW-to-waveguide shown in Figure B.3. The S_{12} parameter is omitted due to reciprocity.

Additional Simulation and Measurement Results

In some parts of the work, results were omitted for the sake of compactness and better readability. This appendix shows additional simulation and measured results without detailed discussions.

C.1 Simulation and Measurement of SSIW P-LWA

This section presents a comparison between the full-wave simulated and measured results of the SSIW P-LWA shown in Section 3.5.3.1. Figure C.1 shows the fabricated prototype, which is fed by two SIW-to-waveguide transition [Appendix B.2]. The prototype is fabricated using laser structuring with the LPKF Protolaser S machine [119]. A comparison between the simulated and measured S-parameters is shown in Figure C.2. Additionally, the radiation patterns of the fabricated prototype were measured in the CATR of IHF and are shown in Figure C.3. Finally, the realized gain over frequency is plotted in Figure C.4. A good agreement between the simulation and measurement results can be observed.

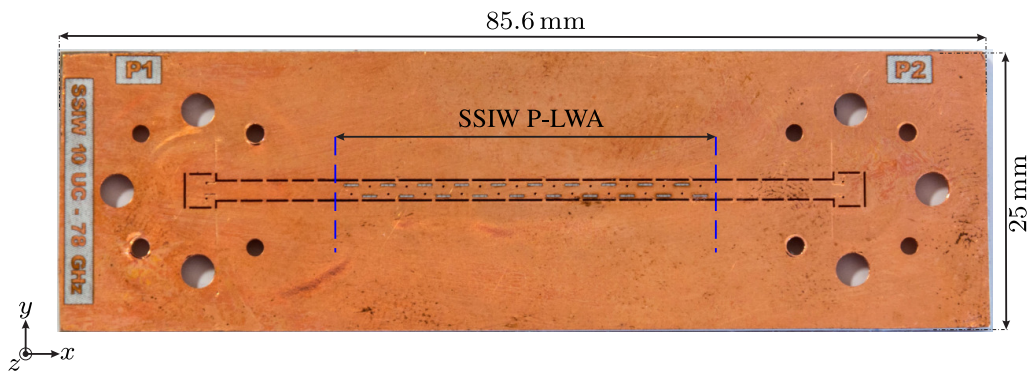


Figure C.1: Top view of the fabricated prototype of the SSIW P-LWA with 10 unit cells fed by SIW-to-waveguide transitions.

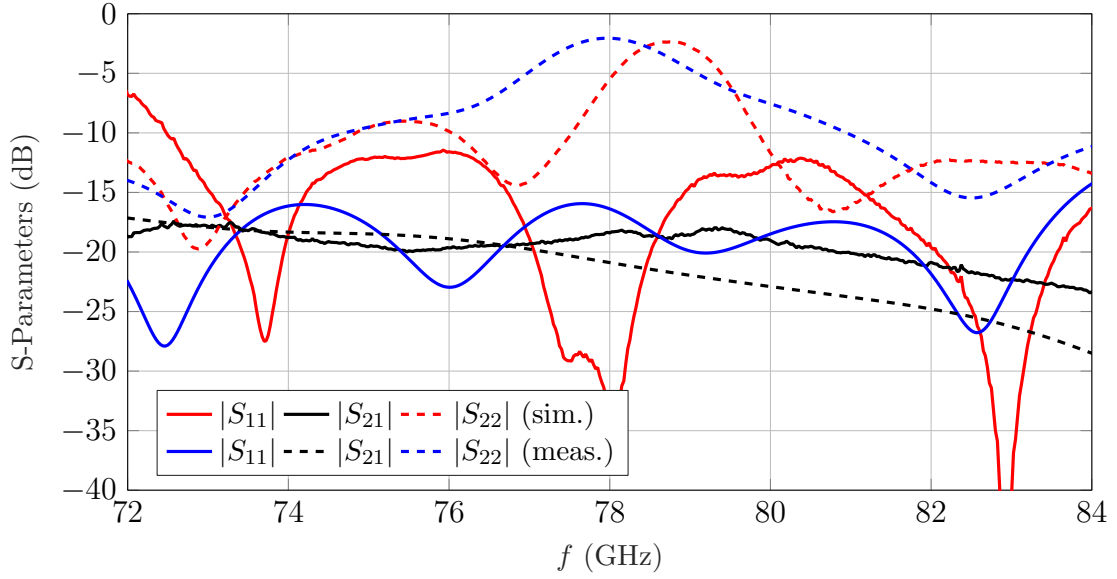


Figure C.2: Full-wave simulated and measured S-parameters of the SSIW P-LWA with 10 unit cells. S_{12} is not shown due to reciprocity.

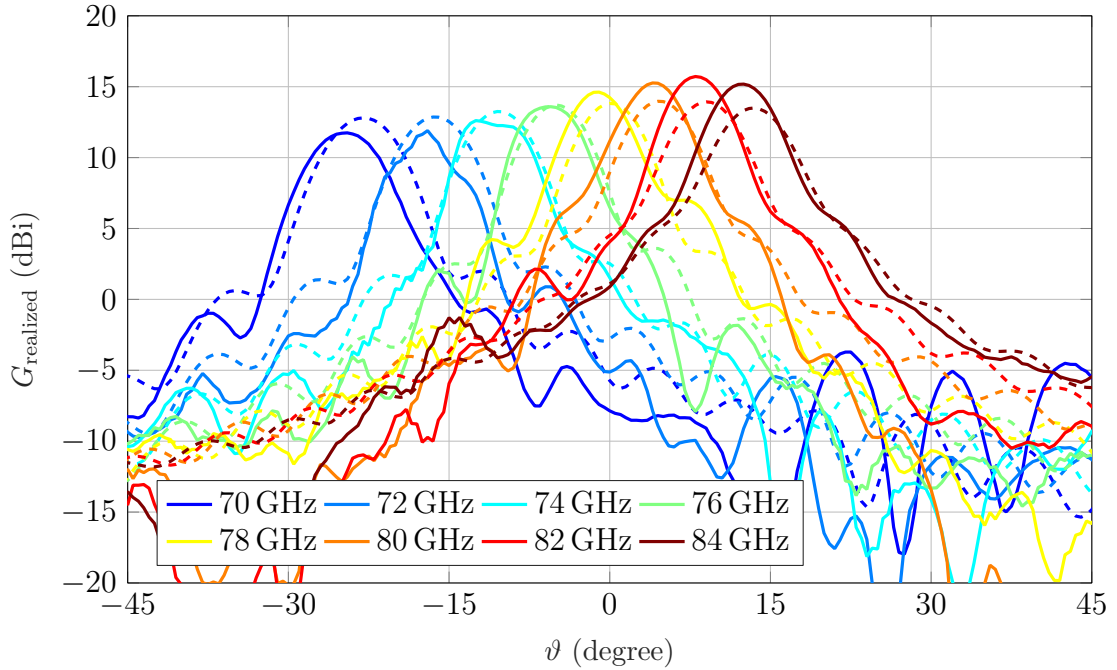


Figure C.3: Full-wave simulated and measured realized gain pattern of the SSIW P-LWA with 10 unit cells.

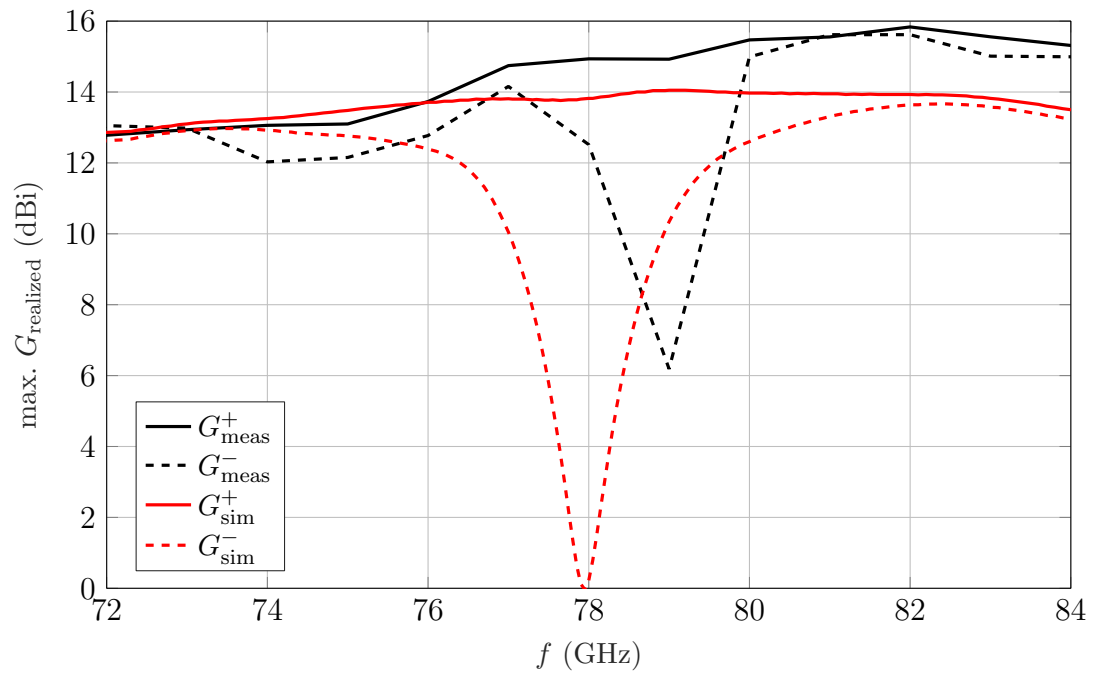


Figure C.4: Realized gain of the SSIW P-LWA versus frequency obtained by full-wave simulation and measurement.

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List of Acronyms and Symbols

All acronyms used in the thesis are listed below. Symbols may be reused for clarity and better readability reasons. Especially, symbols corresponding to geometrical parameters have been reused and are omitted in this section.

Acronyms

1D	one-dimensional
2D	two-dimensional
3D	three-dimensional
6G	the sixth generation mobile system standard
ADC	analog-digital converter
BW	backward wave
CATR	compact antenna test range
CP	circular polarization
CPW	coplanar waveguide
CRLH	composite right/left hand
CST MWS	computer simulation technology microwave studio
CST-EM	CST eigenmode solver
CST-TD	CST time-domain solver
DA	double asymmetry
DS	double symmetry
DSP	digital signal processor
FIT	finite integral technique
FMCW	frequency-modulated continuous wave
FPGA	field-programmable gate array
FW	forward wave
HLP	horizontal linear polarization
IDC	inter-digital capacitor
IEEE	Institute of Electrical and Electronics Engineers
IHF	Institute of High Frequency Technology
JCRS	joint communication and radar sensing
LA	longitudinal asymmetry
LHCP	left hand circular polarization
LNA	low noise amplifier
LP	linear polarization
LWA	leaky-wave antenna
MIM	metal-insulator-metal
MIMO	multiple-input and multiple-output
MSL	microstrip line
PA	power amplifier
PBC	periodic boundary condition
PC	personal computer

PEC	perfect electric conductor
P-LWA	periodic leaky-wave antenna
PMC	perfect magnetic conductor
RCS	radar-cross section
RHCP	right hand circular polarization
RWG	rectangular waveguide
Rx	receiver
SAR	synthetic-aperture radar
SFCP	series-fed coupled patch
SFN	series-feeding network
SFP	series-fed patch
SIW	substrate-integrated waveguide
SNR	signal-to-noise ratio
SSIW	slotted substrate-integrated waveguide
STFT	short-time Fourier transformation
TA	transversal asymmetry
TDM	time-division multiplexing
TE	transverse electric
TEM	transverse electromagnetic
TM	transverse magnetic
Tx	transmitter
VLP	vertical linear polarization

Symbols

General Symbol Notation

$\square_{bs}$	subscript describing a quantity at the broadside frequency
$\square_{off-bs}$	subscript describing a quantity at the off-broadside frequency
$\square_{bf}$	subscript describing a quantity at the backfire frequency
$\square_{ef}$	subscript describing a quantity at the endfire frequency
$\square^{+}$	superscript describing a quantity of the forward wave
$\square^{-}$	superscript describing a quantity of the backward wave
$\square^{sym}$	superscript describing a quantity of a transversally-symmetric unit cell
$\square^{asym}$	superscript describing a quantity of a transversally-asymmetric unit cell
$\square_{se}$	subscript describing a quantity of the series mode
$\square_{sh}$	subscript describing a quantity of the shunt mode
$\square^{odd}$	superscript describing a quantity of the odd excitation

$\square^{\text{even}}$	superscript describing a quantity of the even excitation
$\hat{\square}$	unit vector

Latin Symbols

a	waveguide width
$ABCD$	transmission matrix
A, B, C, D	entries of the ABCD matrix
AF	array factor
AR	axial ratio
B	bandwidth
B_{eff}	effective bandwidth
c_0	speed of light in vacuum
C	capacitance
C_{se}	series capacitance
C_{sh}	shunt capacitance
$\mathbf{E}$	electric field
$\mathbf{E}_1, \mathbf{E}_2$	electric fields of mode 1 and mode 2
$\mathbf{E}_{\text{se}}, \mathbf{E}_{\text{sh}}$	electric fields of the series and shunt modes
$\mathbf{E}^i$	incident electric field
$\mathbf{E}^s$	scattered electric field
E_n	Fourier coefficient of the n^{th} space harmonic
f	frequency
f_{bf}	backfire frequency
f_{bs}	broadside frequency
f_{ef}	endfire frequency
f_s	sampling frequency
f_{start}	start frequency of the FMCW chirp
f_{stop}	stop frequency of the FMCW chirp
G	conductance
G_{realized}	realized gain

h	substrate height
$\mathbf{H}$	magnetic field
$\mathbf{I}$	identity matrix
I_n, I_{n+1}	currents flowing through the terminals of the n^{th} and $(n+1)^{\text{th}}$ unit cells
$I_{\text{se}}, I_{\text{sh}}$	series and shunt currents
$J_{\text{se}}, J_{\text{sh}}$	series and shunt electric surface current densities
k_0	wavenumber in vacuum
ℓ	spacing between of unit cells
L	inductance
L_{se}	series inductance
L_{sh}	shunt inductance
M	number of unit cells in $\hat{x}$
$\mathbf{M}$	coupling matrix
$\mathbf{M}_{\text{se}}, \mathbf{M}_{\text{sh}}$	series and shunt magnetic surface current densities
$\mathbf{n}$	index of the space harmonic
N	number of unit cells in $\hat{y}$
N_{ADC}	number of ADC samples
N_{chrip}	number of chirps
p	period
p_x, p_y	periods in x and y
$\mathbf{P}$	periodic function
P	power
P_{loss}	loss power
P_{rad}	radiation power
$\mathcal{PT}$	parity-time
Q	quality factor
Q_1, Q_2	mode 1 and mode 2 coupled quality factors
$Q_{\text{se}}, Q_{\text{sh}}$	series and shunt quality factors

r	range
R	range of a target
R	resistance
R_{loss}	loss resistance
R_{rad}	radiation resistance
$\mathbf{S}_a, \mathbf{S}_b, \mathbf{S}_c, \mathbf{S}_d$	polarization scattering matrices in the Pauli decomposition basis
$\mathbf{S}_{\text{CP}}$	polarization scattering matrix in the circular polarization basis
$\mathbf{S}_{\text{LP}}$	polarization scattering matrix in the linear polarization basis
$s_{\text{Tx}}, s_{\text{Tx}}, s_{\text{IF}}$	transmitted, received and down-mixed radar signals
t	time
T	transformation ratio, or time period
T_{ADC}	ADC sampling time
T_{c}	total chirp time
T_{idle}	idle time
T_{start}	ADC start time
v_{p}	phase velocity
v_{g}	group velocity
V	voltage
V_n, V_{n+1}	voltages at the terminals of the n^{th} and $(n + 1)^{\text{th}}$ unit cells
$V_{\text{se}}, V_{\text{sh}}$	series and shunt voltages
W	energy
$W_{\text{e}}, W_{\text{m}}$	electric and magnetic stored energies
x, y, z	Cartesian coordinates system
X	reactance
Y	admittance
Y_{sh}	shunt admittance
Z	impedance
Z_{B}	Bloch impedance
Z_{bs}	Bloch impedance at the broadside frequency

$Z_{\text{off-bs}}$	Bloch impedance at the off-broadside frequency
$Z_{\text{c}}^{\text{TE10}}$	characteristic impedance of the TE ₁₀ mode in a rectangular waveguide
$Z_{\text{B}}^{\text{sym}}$	Bloch impedance of a double-symmetric unit cell
$Z_{\text{B}}^{+}, Z_{\text{B}}^{-}$	Bloch impedances corresponding to the forward and backward wave propagation
Z_{se}	series impedance

Greek Symbols

α	attenuation or leakage constant
β	phase constant
β_{n}	phase constant of the n^{th} space harmonic
β^{TE10}	phase constant of the TE10 mode in dielectric-filled waveguide
γ	complex propagation constant
γ_{n}	complex propagation constant of the n^{th} space harmonic
δ_{c}	slope rate of the FMCW chirp
Δr	range resolution
ΔT	possessing width of the STFT
$\Delta \vartheta_{\text{mb}}$	beam-scanning range
ϵ	permittivity
ϵ_{r}	relative permittivity or dielectric constant
η_{c}	coupling efficiency
η_{rad}	radiation efficiency
η_{tot}	total antenna efficiency
η^{+}, η^{-}	radiation efficiency corresponding to the forward and backward wave propagation
$\eta_{\text{bs}}, \eta_{\text{off-bs}}$	at-broadside and off-broadside radiation efficiency.
ϑ, φ, r	spherical coordinates with roll-over-azimuth angles
ϑ_{3dB}	3 dB beamwidth or angular resolution
ϑ_{mb}	main beam direction in the yz -plane
ϑ_{s}	orientation angle of a scatterer with respect to the radar

$\theta_{\text{El}}, \phi_{\text{Az}}, r$	spherical coordinates with elevation-over-azimuth angles
κ	coupling coefficient
κ_{12}	coupling coefficient from mode 1 to mode 2
κ_{21}	coupling coefficient from mode 2 to mode 1
κ_{asym}	coupling coefficient corresponding to geometrical asymmetry
λ_0	wavelength in vacuum
λ_g	guided wavelength
λ_g^{TE10}	guided wavelength of the TE10 mode in dielectric-filled waveguide
μ	permeability
ξ_n	scanning rate of the n^{th} space harmonic
Φ	phase shift between the unit cells
Φ_x	phase shift between the unit cells in $\hat{x}$
Φ_y	phase shift between the unit cells in $\hat{y}$
ψ_s	orientation of a scatterer
ω	angular frequency
ω_0	resonance angular frequency
ω_{bs}	broadside angular frequency
ω_{bf}	backfire angular frequency
ω_{ef}	endfire angular frequency
Ω	complex temporal eigenfrequency
Ω_1, Ω_2	coupled temporal eigenfrequencies of mode 1 and mode 2
Ω_{bs}	eigenfrequency at the broadside
$\Omega_{\text{se}}, \Omega_{\text{sh}}$	uncoupled eigenfrequencies of the series and shunt modes

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List of Publications

- [P1] **A. Al-Bassam**, S. Otto, Z. Chen, A. Rennings, K. Sobach, and C. Caloz, “A capacitively-coupled series-fed patch leaky-wave antenna and optimization concepts for efficient broadside radiation,” in *German Microwave Conference (GeMiC)*, Ilmenau , Germany, Mar. 2012.
- [P2] S. Otto, **A. Al-Bassam**, A. Rennings, K. Sobach, and C. Caloz, “Q-balancing in periodic leaky-wave antennas to mitigate broadside radiation issues,” in *German Microwave Conference (GeMiC)*, Ilmenau , Germany, Mar. 2012.
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- [P6] S. Otto, Z. Chen, **A. Al-Bassam**, A. Rennings, K. Solbach, and C. Caloz, “Circular polarization of periodic leaky-wave antennas with axial asymmetry: Theoretical proof and experimental demonstration,” *IEEE Transactions on Antennas and Propagation*, vol. 62, no. 4, pp. 1817–1829, Apr. 2014.
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- [P9] **A. Al-Bassam**, S. Otto, and C. Caloz, “Dual-channel radiation using a single double-asymmetric periodic leaky-wave antenna,” in *IEEE International Symposium on Antennas and Propagation (APSURSI)*, Fajardo, PR, USA, Jun. 2016.
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- [P11] W. Alshrafi, **A. Al-Bassam**, C. Culotta-Lopez, and D. Heberling, “Printed yagi-helix antennas,” in *Loughborough Antennas & Propagation Conference (LAPC)*, Loughborough, UK, Nov. 2017.

- [P12] **A. Al-Bassam**, D. Heberling, and C. Caloz, “Multi-plane frequency scanning using a leaky-wave antenna array with double-end series feeding network,” in *IEEE International Symposium on Antennas and Propagation (APSURSI)*, Boston, MA, USA, Jul. 2018.
- [P13] **A. Al-Bassam**, D. Heberling, and C. Caloz, “Two-dimensional periodic leaky-wave antenna for oblique frequency scanning,” in *12th European Conference on Antennas and Propagation (EuCAP)*, London, UK, Apr. 2018.
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- [P17] **A. Al-Bassam**, D. Heberling, and C. Caloz, “Full-space coverage leaky-wave antenna based on successive space harmonics,” in *Photonics & Electromagnetics Research Symposium (PIERS)-Spring (PIERS-Spring)*, Rome, Italy, Jun. 2019.
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- [P23] A. C. Granich, R. Moch, **A. Al-Bassam**, and D. Heberling, “Radiation pattern measurements using an active radar module,” in *Antenna Measurement Techniques Association Symposium (AMTA)*, Denver, CO, USA, Oct. 2022.
- [P24] **A. Al-Bassam**, D. Heberling, and C. Caloz, “Exceptional point perspective of periodic leaky-wave antennas,” in *IEEE International Symposium on Antennas and Propagation (APSURSI)*, Portland, OR, USA, Jul. 2023.

- [P25] G. François, **A. Al-Bassam**, T. M. Eckroth, and D. Heberling, “All-dielectric artificial magnetic conductor at millimeter-wave frequencies,” in *IEEE International Symposium on Antennas and Propagation (APSURSI)*, Portland, OR, USA, Jul. 2023.

Curriculum Vitae

Personal Details

Name	Amar Al-Bassam
Place of birth	Baghdad, Iraq

Education

2014	Master of science (M. Sc.) in electrical engineering and information technology at University of Duisburg-Essen.
2011	Bachelor of science (B. Sc.) in electrical and electronic engineering at University of Duisburg-Essen.

Work Experience

10/2015 – today	Research assistant at Institute of High Frequency Technology (IHF), RWTH Aachen University.
04/2015 – 09/2015	Research assistant at Chair of High Frequency Technology (HFT), University of Duisburg-Essen.
10/2014 – 06/2015	Research assistant, IMST GmbH, Department Antenna and EM Modeling.
05/2013 – 10/2013	Research Intern, PolyGrames Research Center, École Polytechnique de Montréal, Canada.

Awards & Grants

2019	PIERS Best Student Paper Award – 1 st Place, Rome, Italy.
2015	Innovation prize 2014 of the Best Master Thesis in Engineering at University of Duisburg-Essen sponsored by Sparkasse am Niederrhein.
2011 – 2012	“Deutschlandstipendium” scholarship, partly sponsored by Federal Ministry of Education and Research.