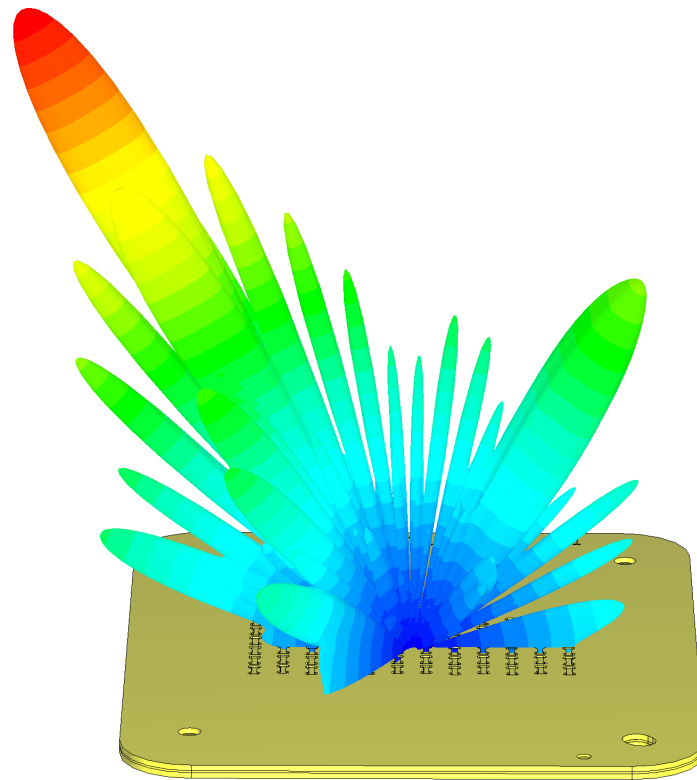




CHALMERS
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Design and Experimental Validation of a Passive Van Atta Retroreflector Array in Gap Waveguide Technology

For Wide-Angle RCS Enhancement in the 76–81 GHz Band

Master's thesis in Wireless, Photonics and Space engineering

ALEXANDER BOHLIN

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CHALMERS UNIVERSITY OF TECHNOLOGY
Gothenburg, Sweden 2026
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MASTER'S THESIS 2026

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Cover: Three-dimensional bistatic far-field scattering pattern of the Van Atta retroreflector when illuminated by a plane wave incident from 30° .

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Abstract

Automotive radar operating in the 77 GHz frequency band is increasingly implemented in modern cars. A challenge these radars face is the detection of vulnerable road users, such as pedestrians, whose radar cross section (RCS) can be low and highly variable. Compact passive retroreflectors, such as Van Atta arrays, can increase their radar visibility, but when implemented with patch antennas or substrate integrated waveguides (SIWs), these arrays suffer significant dielectric losses at mmWave frequencies.

Motivated by this limitation, this work uses Multilayer Waveguide (MLW) technology as the basis for a passive Van Atta array intended to enhance RCS over the 76–81 GHz frequency band. This is done by designing, modeling, manufacturing, and experimentally validating a prototype MLW Van Atta array. The prototype is designed for horizontal polarization and provides retrodirection in the azimuth plane. The Van Atta array consists of 12 slotted waveguide columns. Each column consists of six slots fed in series by a coaxial line in MLW. The columns are connected pairwise through transmission lines with equal electrical length. Several samples of the reflector are then manufactured and validated through an RCS measurement setup in an anechoic chamber.

All measured samples demonstrate a functional angular range of more than 90° across the 76–81 GHz frequency band, with an operational bandwidth of approximately 5 GHz. Within this range, the designed retroreflector provides an RCS improvement of approximately 10 dB to 20 dB compared with a metal plate of the same physical size. These results show that compact planar Van Atta retroreflectors with a thickness below 1 mm can be realized in MLW technology and provide wide-angle RCS enhancement. This makes the approach promising for increasing the radar visibility of vulnerable road users.

Keywords: multilayer waveguide (MLW), gap waveguide (GWG), Van Atta array, retrodirective array (RDA), passive retroreflector, millimeter wave (mmWave), radar cross section (RCS)

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Alexander Bohlin, Gothenburg, June 2026

List of Acronyms

Below is the list of acronyms used throughout this thesis, listed in alphabetical order:

ADAS	Advanced Driver Assistance Systems
dBsm	Decibel Square Meter
EBG	Electromagnetic Bandgap
FFT	Fast Fourier Transform
GWG	Gap Waveguide
IFFT	Inverse Fast Fourier Transform
ISAC	Integrated Sensing and Communication
L1–L3	MLW Layers 1–3
MLW	Multilayer Waveguide
PEC	Perfect Electric Conductor
PMC	Perfect Magnetic Conductor
RCS	Radar Cross Section
RDA	Retrodirective Array
SIW	Substrate Integrated Waveguide
TUT	Target Under Test
VNA	Vector Network Analyzer

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1

Introduction

This chapter presents the motivation and scope of the thesis. First, the background of modern automotive radar, vulnerable road users, and passive RCS enhancement is presented. Related work on passive retroreflectors and retrodirective arrays is then discussed, followed by the problem formulation and aim of the thesis. The chapter concludes with an outline of the remaining chapters.

1.1 Background and Motivation

Wireless systems are increasingly moving toward mmWave frequencies and beyond. This development is driven both by the limited availability of spectrum at lower frequencies and by the need for larger bandwidths. While 5G already includes mmWave bands, future 6G systems are expected to also exploit sub-THz frequency bands. These sub-THz bands are expected to support integrated sensing and communication (ISAC), which is one of the major new directions in 6G. In ISAC, the same signals and hardware are used both for wireless communication and for sensing the surrounding environment, for example the position and movement of users or other objects.

Automotive radar is another important example of the move toward mmWave frequency bands. Modern automotive radar sensors commonly use the 76–81 GHz band. The increased use of this band is closely connected to the expansion of advanced driver assistance systems (ADAS), which is driven by both market demand and safety regulations. For example, Intelligent Speed Assistance (ISA) and Advanced Emergency Braking Systems (AEBS) have been mandatory ADAS functions for all new cars sold in the EU since 2024 [1]. These systems support the development toward more automated driving while also reducing the risk of traffic accidents and improving the protection of vulnerable road users, including pedestrians and cyclists.

For these applications, automotive radar is very important, especially in adverse weather conditions such as fog or heavy rain where camera- and lidar-based systems may be severely degraded. Radar also faces its own challenges, especially in complex urban environments, where clutter, multipath, and interference from other vehicles pose real problems. These difficulties are further exacerbated for weak and hard-to-detect targets such as pedestrians and cyclists, which often have low and highly variable radar cross section (RCS) [2].

In these urban environments communication systems, automotive radar, and other applications will in the future need to operate together, sharing both spectrum and physical space. In addition to this they are also expected to become increasingly interconnected. With ISAC in 6G and higher levels of autonomous driving, these technologies will increasingly need to interact and support each other. This creates new possibilities for traffic monitoring, new safety systems, and improved situational awareness. For example, vehicles may receive information through the 6G network about objects or road users hidden behind a corner [3].

In both automotive radar and future 6G ISAC systems, it is of interest to make weak targets easier to detect by increasing their RCS. As mentioned above this is especially relevant for vulnerable road users, such as pedestrians and cyclists, which can be difficult for automotive radar to detect reliably due to their small and highly variable RCS. Similar challenges may also arise in future 6G ISAC systems operating in complex urban environments. One straightforward way of increasing the RCS of such targets is to use a passive reflector that scatters part of the incident signal back toward the sensor. How one possible urban scenario could look is shown in Figure 1.1.

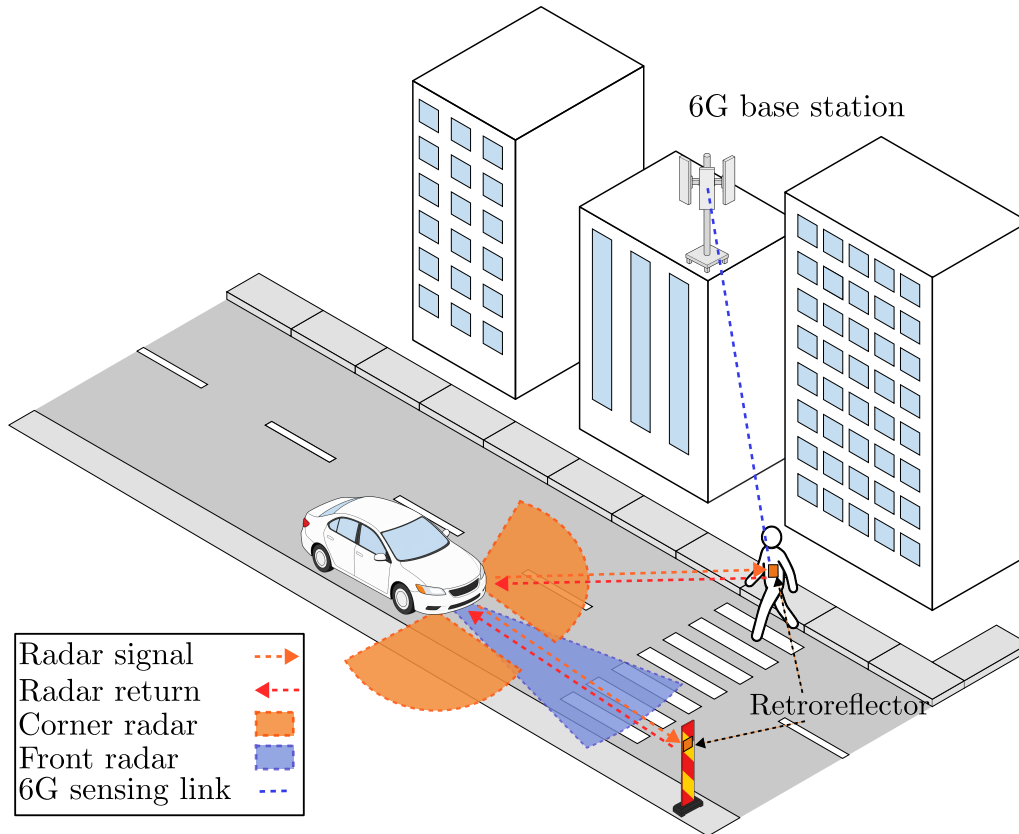


Figure 1.1: Example urban scenario with passive retroreflectors placed on a pedestrian and on roadside safety equipment. The reflectors strengthen the return seen by the radar. The illustration also includes a possible future 6G use case, where the pedestrian is passively observed by a 6G base station.

Such reflectors, either worn by vulnerable road users or integrated into roadside structures, could increase the detection range and give autonomous driving systems and ADAS more time to react and brake. They could also improve classification by providing stronger and more consistent radar returns. Existing passive retroreflectors and retrodirective arrays (RDA) provide different ways of achieving this but their suitability depends strongly on size, angular coverage and frequency band. The following section therefore reviews related work on passive RCS-enhancing structures, with particular focus on RDAs and their limitations at the automotive 77 GHz frequency band.

1.2 Related Work

The most basic radar reflectors used for increasing RCS are corner reflectors, which consist of perpendicular conducting plates that reflect incident waves back toward the source. Such reflectors have, for example, been used for decades in marine navigation to increase the RCS of objects with naturally low RCS, such as fiberglass sailboats [4]. Although they are fairly bulky, they have also been proposed for increasing the RCS of cyclists for automotive radar. In [5], a trihedral corner reflector is proposed to increase the rear RCS of a cyclist to approximately the same level as that of a sedan, around 14.7 dBsm. A dihedral corner reflector flattened to an angle of 135° instead of the usual 90° , combined with reflectarray surfaces, is also shown in [6] to increase the rear RCS of cyclists to 6 dBsm in the older 24 GHz automotive radar band. Other non-planar retroreflectors have also been proposed. One example is a metallic sphere reflector combined with a Luneburg lens array, which could potentially be made flexible for integration into, for example, clothing. In simulations, such an array was shown to provide retrodirective behavior from -65° to 65° in the 77 GHz frequency band [7].

Retrodirective arrays (RDAs) provide a more compact type of reflector. They may be passive or active, and their purpose is to reradiate the received energy back toward the direction of arrival without requiring prior knowledge of the source location [4]. Since they are antenna arrays, they can be implemented as thin planar structures, for example using patch antennas. A passive example is the Van Atta array. In this configuration, each antenna element is connected to the element mirrored about the array center, and the connecting transmission lines are designed to have the same electrical length. The reradiated field therefore obtains the opposite phase progression compared with the incident field. Planar Van Atta arrays are often implemented using patch antennas or substrate integrated waveguide (SIW) slot antennas [8, 9, 10]. Depending on the arrangement of the elements and interconnections, retroreflection can be obtained in either one or two dimensions.

Several works have implemented Van Atta arrays to increase RCS or achieve retrodirective behavior for other purposes. In [11], a one-dimensional Van Atta array with retrodirectivity in azimuth is presented, but an active phase-shifter component is included to allow the retrodirective plane in elevation to be changed for increasing the RCS of roadside structures such as road barriers. The work reports an RCS en-

hancement of 10 dB to 15 dB when placed on a roadside cylinder at X-band, which is not an automotive band. Another work implements a wideband Van Atta array in microstrip using Vivaldi antennas and reports an RCS improvement of 10 dB to 20 dB from 2.4 GHz to 29.7 GHz compared with a PEC plate of equal size [12].

While Van Atta arrays have been widely studied at lower frequencies and in the legacy automotive frequency band at 24 GHz, fewer implementations appear to have been reported at the 76–81 GHz band, where losses and fabrication constraints become more critical. One example of a Van Atta array operating for 77 GHz automotive radar is reported in [13], where a planar one-dimensional Van Atta array in microstrip patch technology is designed and manufactured. In that work, several such arrays are placed on a panel, and when mounted on a mannequin, an RCS improvement of 5.5 dB is measured.

1.3 Problem Formulation

As noted, there are challenges in building a compact planar RCS-enhancing structure operating in the automotive 77 GHz frequency band. Many traditional retroreflectors, such as spherical reflectors and corner reflectors, are inherently bulky. Patch and SIW antennas, which are widely used at lower frequencies, are based on dielectrics, which means that dispersion and losses limit their use in larger arrays at mmWave frequencies. At these frequencies, waveguide implementations become more attractive from a loss perspective. However, conventional waveguides are generally difficult to scale for large-scale manufacturing because of fabrication challenges such as soldered joints. This work therefore investigates whether Multilayer Waveguide (MLW) technology can be used to realize a planar Van Atta array with low loss, low dispersion, and practical manufacturability.

1.4 Aim

The aim of this thesis is to design, model, manufacture, and experimentally validate a planar Van Atta array in multilayer gap waveguide technology for increasing the effective RCS of vulnerable road users at the automotive 77 GHz frequency band. The work focuses on a prototype that is retrodirective in azimuth and operates for horizontal polarization. The prototype is evaluated through full-wave simulations, a simplified numerical model, manufacturing, and monostatic RCS measurements.

1.5 Outline of the Thesis

The thesis is structured as follows. Chapter 2 gives the theoretical basis for the work, covering automotive radar and RCS, antenna theory, MLW technology, and passive reflectors. Chapter 3 describes the design and methods used in the thesis, including the MLW coaxial line, the Van Atta reflector design, the numerical model, and the measurement setup. Chapter 4 presents the simulated and measured results, including the radiating elements, transmission networks, full-reflector simulations, measurement results, and results of the numerical model. The thesis concludes with Chapter 5, where the main findings are summarized and possible directions for future work are discussed.

2

Theory

This chapter gives the theory needed for the work. First, automotive radar and concepts related to radar cross section are presented. This is followed by the general antenna theory, as well as the theory and background of gap waveguide and Multi-layer Waveguide (MLW) technology. The final part of the chapter reviews passive retroreflectors, focusing mainly on the Van Atta array.

2.1 Automotive Radar and Radar Cross Section (RCS)

In this thesis, the reflector performance is primarily evaluated using radar cross section. Therefore, this section first gives the automotive radar context before introducing the basic RCS definitions and RCS in the context of measurements.

2.1.1 Automotive Radar

Radar sensors transmit radio waves and use the returned signal to estimate target range and radial velocity. In vehicles, radar is used together with cameras and lidar as part of ADAS. A key advantage of radar is that it is less affected by fog and rain, where camera and lidar performance can degrade, while it also provides a direct measurement of radial velocity. In recent years, automotive radar has moved from the legacy 24 GHz frequency band to the 77 GHz frequency band, where the wavelength is approximately 4 mm. This small wavelength enables compact antennas, while the available bandwidth at this band allows for high range resolution.

Because automotive radar operates in urban environments with short distances and many users sharing the spectrum, the output power of the radars is very limited. As a result, the range at which an automotive radar can detect a target is limited. For modern automotive radar sensors, reported maximum detection ranges are around 100 m to 300 m, depending on the sensor [14, 15, 16]. However, the actual detection range can be much lower depending on the situation and the target. This difference in detection range can be understood from typical RCS levels. Reported car RCS values are approximately 0 dBsm to 20 dBsm, depending on aspect angle [17], whereas pedestrian RCS values are much lower, approximately -20 dBsm to 0 dBsm [18]. Since a larger RCS gives a stronger radar return, cars are normally detectable at longer ranges than pedestrians.

2.1.2 RCS Definitions

Radar cross section describes the apparent scattering strength of a target as seen by a radar. For a specified incident direction and scattering direction, it is defined from the ratio between the scattered power density in the observation direction and the incident plane-wave power density at the target, with the usual $4\pi R^2$ scaling applied in the far field. The value of RCS therefore depends on the frequency, polarization, and aspect angle of the target. Mathematically, it is written as

$$\sigma = \lim_{R \rightarrow \infty} 4\pi R^2 \frac{|E_s|^2}{|E_i|^2} \quad (2.1)$$

where σ is the radar cross section, R is the distance from the target to the observation point, E_s is the scattered electric field, and E_i is the incident electric field at the target [19].

Depending on the relation between the incident and scattered directions, three RCS cases are commonly distinguished:

- **Monostatic RCS** (or **backscatter RCS**): scattering is observed in the direction back toward the radar.
- **Forward-scatter RCS**: scattering is observed in the forward direction of the incident wave.
- **Bistatic RCS**: scattering is observed in a direction that is neither the backscatter nor the forward-scatter direction.

When the term RCS is used without further context, it is usually understood to refer to the monostatic case. The forward- and backscatter directions may also be regarded as two special cases of the general bistatic case.

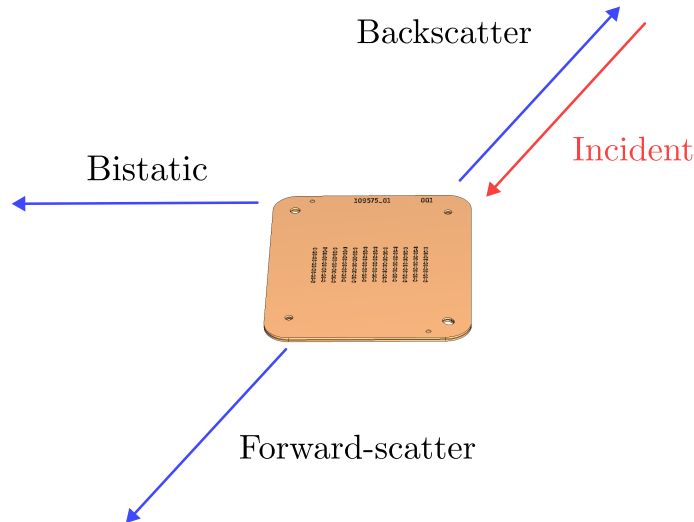


Figure 2.1: Definition of monostatic, bistatic, and forward-scatter RCS.

2.1.3 RCS Measurements

To measure the RCS of a target under test (TUT), the formal definition in Equation 2.1 is often not very useful. Instead, an equation relating the received power P_r

to the transmitted power P_t in a test setup is needed. In free space, this is described by the radar equation

$$P_r = \frac{P_t \mathbb{G}_0^2 \lambda^2 \sigma}{(4\pi)^3 R^4} \quad (2.2)$$

where P_t is the transmitted power, R is the distance from the radar to the TUT, $\mathbb{G}_0$ is the radar antenna gain, λ is the wavelength, and σ is the radar cross section of the TUT. Many of these variables are hard to determine with high precision in a test setup. In practice, the measurement is therefore often carried out by measuring a reference object with known RCS, σ_{ref} , such as a corner reflector or a metal sphere. If the measurement geometry is unchanged such that all other variables remain the same, the RCS of the TUT, σ_{TUT} , can then be calculated as

$$\sigma_{\text{TUT}}[\text{dBsm}] = \sigma_{\text{ref}}[\text{dBsm}] + P_{r,\text{TUT}}[\text{dB}] - P_{r,\text{ref}}[\text{dB}]. \quad (2.3)$$

2.2 Antenna Theory

In this section, the most important definitions and conventions used to describe antennas in this thesis are introduced. The notation, definitions, and general treatment in this section follow [20], unless stated otherwise.

2.2.1 Coordinate System

The spherical coordinate system used throughout this thesis is shown in Figure 2.2.

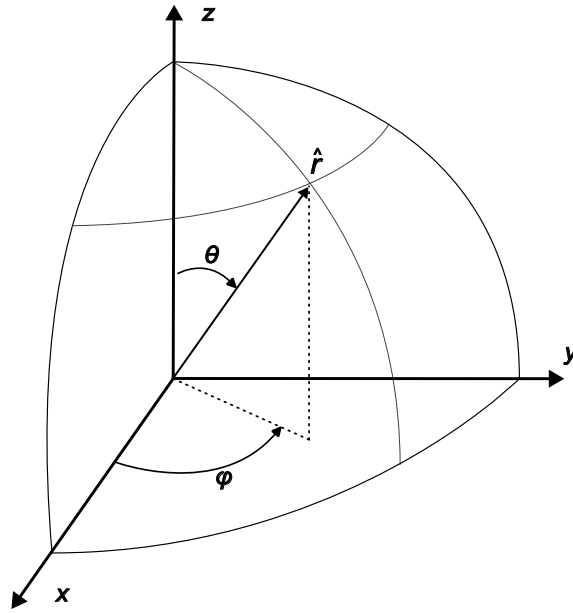


Figure 2.2: Spherical coordinate system.

2.2.2 Far-field Function

For an antenna with its spherical coordinate system centered at a point inside the antenna, the far-field electric field can be written as

$$\mathbf{E}(r, \theta, \varphi) = \frac{1}{r} e^{-jkr} \mathbf{G}(\theta, \varphi) \quad (2.4)$$

where r is the distance to the observation point in the far field, k is the wavenumber, and $\mathbf{G}(\theta, \varphi)$ is the far-field function. All angular dependence is contained in $\mathbf{G}(\theta, \varphi)$, while the radial dependence is contained in the factor e^{-jkr}/r [20].

The radiation intensity describes how much power is radiated in a given direction per unit solid angle. With the far-field function, this quantity is expressed as

$$U(\theta, \varphi) = \frac{1}{2\eta} |\mathbf{G}(\theta, \varphi)|^2 \quad (2.5)$$

where η is the wave impedance of free space [20].

2.2.3 Definitions of General Antenna Quantities

Total radiated power. The total power radiated by the antenna is calculated by integrating the radiation intensity over the far-field sphere,

$$P_{\text{rad}} = \int_0^{2\pi} \int_0^\pi U(\theta, \varphi) \sin \theta d\theta d\varphi \quad (2.6)$$

as defined in [20].

Directive gain. The directive gain is obtained by comparing the far-field function with that of an ideal isotropic radiator. Written in dBi, the directive gain becomes

$$(\mathbb{D}(\theta, \varphi))_{\text{dBi}} = 10 \log_{10} \left(\frac{|\mathbf{G}(\theta, \varphi)|^2}{|G_{\text{ISO}}|^2} \right) \text{ dBi} = 10 \log_{10} \left(\frac{4\pi |\mathbf{G}(\theta, \varphi)|^2}{P_{\text{rad}}} \right) \text{ dBi} \quad (2.7)$$

where G_{ISO} is the far-field function of an ideal isotropic radiator. An isotropic radiator is a theoretical antenna that radiates equally strongly in all directions, which means that its far-field function is independent of angle, that is $G_{\text{ISO}}(\theta, \varphi) = G_{\text{ISO}}$ [20].

Pointing direction and directivity. The pointing direction is the direction in which the directive gain is largest. For many antennas, this direction is the boresight direction, $(\theta, \varphi) = (0, 0)$. This largest value is called the antenna directivity, defined as

$$\mathbb{D}_0 = \max_{\theta, \varphi} \mathbb{D}(\theta, \varphi) \quad (2.8)$$

where $\mathbb{D}(\theta, \varphi)$ is the directive gain as defined in Equation 2.7 [20].

Realized gain. The previous definition uses the total radiated power to define directivity. If this is instead replaced by the power delivered to the antenna port, the corresponding quantity is the realized gain, often simply called the gain. This is

the gain quantity that can be measured experimentally. The realized gain is related to the directivity by

$$\mathbb{G}_0 = e_{\text{rad}} e_{\text{apol}} \mathbb{D}_0 \quad (2.9)$$

where e_{rad} accounts for ohmic losses and mismatch losses at the antenna input port, and e_{apol} accounts for polarization mismatch losses [20].

Effective aperture. When dealing with receiving antennas, or when analyzing the RCS of an antenna, another useful quantity is the effective aperture. This may correspond to a physical aperture, as in a horn antenna, or to a virtual aperture, for example over a planar array. For a plane wave incident from the pointing direction, the effective aperture relates the received port power, P_r , to the incident power density, W_r . The same quantity can be expressed using the realized gain as

$$A_e = \frac{P_r}{W_r} = \frac{\lambda^2}{4\pi} \mathbb{G}_0 \quad (2.10)$$

where $\mathbb{G}_0$ is the realized gain [20].

2.2.4 Polarization

The polarization of an electromagnetic wave is determined by how its electric-field vector varies in the plane transverse to the propagation direction. This transverse field can be represented using two orthogonal components. For a plane wave propagating in the positive z -direction, one possible representation is

$$\mathbf{E} = [E_x \hat{\mathbf{x}} + E_y \hat{\mathbf{y}}] e^{-jkz} \quad (2.11)$$

The polarization is then determined by the relative amplitudes and phases of E_x and E_y . More generally, any polarization can be described using two orthogonal basis vectors.

When designing an antenna, one polarization is usually chosen as the desired polarization and referred to as the co-polar polarization. The remaining orthogonal polarization is then treated as the cross-polar polarization. These two basis polarizations are represented by the complex unit vectors $\hat{\mathbf{c}}\mathbf{o}$ and $\hat{\mathbf{x}}\mathbf{p}$, which are both orthogonal to the propagation direction [20]. Using this basis, the electric field can be written as

$$\mathbf{E} = (E_{\text{co}} \hat{\mathbf{c}}\mathbf{o} + E_{\text{xp}} \hat{\mathbf{x}}\mathbf{p}) e^{-jkz} \quad (2.12)$$

A wave is linearly polarized when the electric field keeps the same orientation plane during propagation. In the component form used here, this occurs when the two orthogonal field components have no phase difference or a phase difference of π [20]. Horizontal and vertical polarization are two important linear polarization cases. For horizontal polarization, the electric field is orthogonal to both the propagation direction $\hat{\mathbf{z}}$ and the ground normal $\hat{\mathbf{n}}$. For vertical polarization, the electric field lies in the plane spanned by $\hat{\mathbf{z}}$ and $\hat{\mathbf{n}}$ [20]. These definitions do not require the wave to travel parallel to the ground. In this thesis, horizontal polarization means that the electric field lies in the xz -plane, while vertical polarization means that it lies in the yz -plane.

2.3 Waveguides and Gap Waveguide (GWG) Technology

This section introduces the waveguide technology used to realize the Van Atta array in this thesis. First, conventional rectangular hollow waveguides are discussed, with emphasis on their low-loss behavior and the manufacturing challenges caused by the need for galvanic contact. Gap waveguide technology is then introduced as a way to avoid this contact requirement. Finally, the multilayer waveguide (MLW) implementation used in this work is described, together with the electromagnetic bandgap (EBG) principle and relevant manufacturing approaches.

2.3.1 Traditional Waveguides

The rectangular hollow waveguide is one of the oldest microwave transmission lines. At lower frequencies, it has largely been replaced by planar implementations such as microstrip and stripline. However, it is still frequently used at mmWave frequencies and above because of its low loss. Since a hollow waveguide consists of a single conductor, it cannot support TEM modes, which require a voltage difference between two conductors. Instead, it supports TE and TM modes [21]. The electric field distribution of the fundamental TE_{10} mode is shown in Figure 2.3.

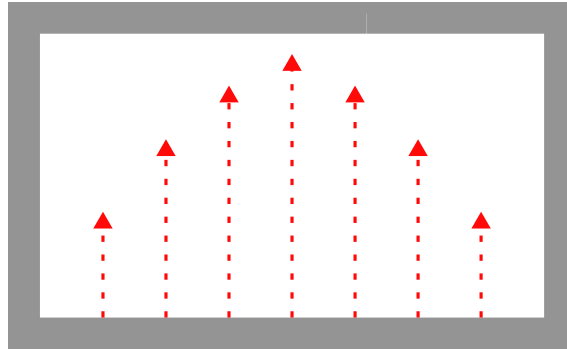


Figure 2.3: Electric field distribution of the TE_{10} mode in a rectangular hollow waveguide.

Because traditional hollow waveguides require good galvanic contact between all metal parts, they can be complicated to manufacture and assemble. The different parts often need to be joined by methods such as soldering or bonding. Any gap where the galvanic contact is broken can lead to field leakage and degraded performance. The severity of this problem depends on how the waveguide is split and assembled. In Figure 2.4b, a waveguide manufactured with a separate top plate is shown for the case where a small gap disrupts the current distribution on one side of the waveguide.

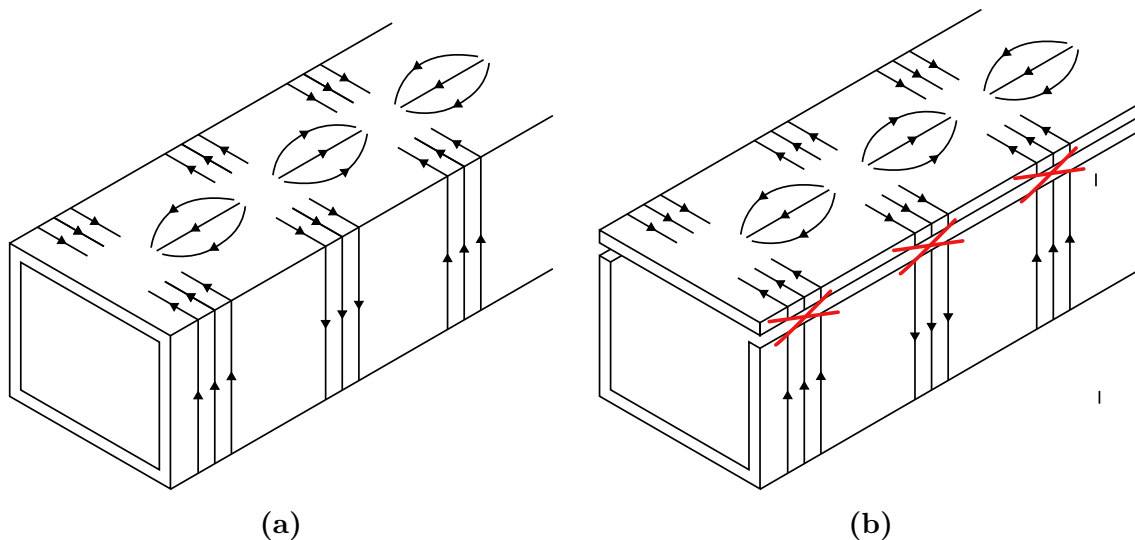


Figure 2.4: Current distributions for a rectangular hollow waveguide. (a) Ideal waveguide. (b) Waveguide with a small gap, resulting in a discontinuous current distribution.

This issue becomes especially important at high frequencies, such as mmWave frequencies, where the dimensions are small and the tolerances are tight. Advanced manufacturing techniques have therefore been developed to ensure good galvanic contact between the metal layers of a waveguide. One such method, which can be used at very high frequencies, is diffusion bonding of metal plates. This process uses high temperature, above 1000°C , and high pressure to join etched metal plates to create the waveguide [22]. While diffusion bonding can produce high-performance waveguide structures, it is a complex and expensive process. It is therefore more suitable for high-end scientific instruments and other specialized systems than for low-cost mass manufacturing.

This means that traditional hollow waveguides are not well suited for a Van Atta array intended for low-cost mass manufacturing, for example for pedestrian RCS enhancement. The galvanic contact issue can instead be avoided by using gap waveguide technology, which is introduced in the remainder of this section.

2.3.2 Gap Waveguide Technology

Gap Waveguide is based on research on hard and soft surfaces in electromagnetics. Kildal introduced these two concepts in 1990 [23]. They originate from acoustics, where a soft surface has zero acoustic pressure at the boundary, which essentially prevents wave propagation along it. In contrast, a hard surface forces the normal derivative of the pressure to be zero. Both types of surfaces are fully reflective, but for a hard surface the phase is preserved, which effectively doubles the pressure at the surface, whereas for a soft surface the phase is reversed, causing cancellation. The same distinction can be made for electromagnetic surfaces. In this context, soft surfaces inhibit waves from propagating along the surface, whereas hard surfaces support such propagation.

Many surfaces naturally behave as either hard or soft surfaces for a given propagation direction and field polarization. For example, consider an electromagnetic wave propagating parallel to a PEC plate, with the plate normal along z . For TE_z polarization, the electric field is parallel to the PEC surface. The PEC boundary condition forces this tangential electric field to vanish at the surface, so the plate behaves as a soft surface for the electric field. Specific structures have been developed to realize more general hard or soft surfaces. Examples include electric strips on a substrate [24], high-impedance mushroom structures [25], and metallic pins forming a bed of nails [26].

The metallic pin concept was later extended to gap waveguide (GWG) technology, where a ridge is positioned between two rows of periodic metallic pins below a metal plate. In a specific frequency interval, the periodic pin surface exhibits electromagnetic bandgap (EBG) behavior. Since the plate does not need to be galvanically connected to the pins, a small gap can still be present without causing leakage. This is because the EBG structure suppresses global parallel-plate modes and makes the surrounding surfaces exhibit very high surface impedance, so that they behave approximately as a perfect magnetic conductor (PMC). This gives the technology two important advantages. In contrast to microstrip and coplanar transmission lines, gap waveguides can be realized without a dielectric material. Compared with conventional metal waveguides, it also removes the need for electrical contact between all metal parts during assembly [27].

2.3.3 Multilayer Waveguide (MLW)

One implementation of GWG technology is the multilayer waveguide. It consists of a coaxial line running in an air-filled channel surrounded by a pin-like EBG structure. It is fabricated from several thin metal sheets that are chemically etched to form the EBG patterns and openings required for the structure [28]. One of the more recent MLW implementations, optimized for low cost and suitable for automotive applications, is built from three chemically etched thin copper layers. In this version, the EBG structure is etched into the top and bottom layers, while the opening for the stripline channel is etched into the middle layer [29, 30]. This is the MLW implementation used in this work, and its basic structure is shown in Figure 2.5, where the channel width w_{ch} and strip width w_{strip} are highlighted.

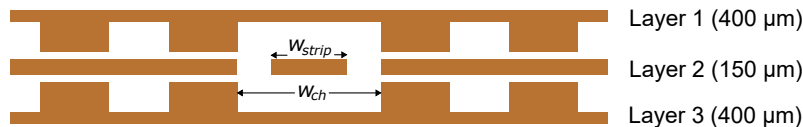


Figure 2.5: Basic geometry of the three-layer MLW implementation considered in this work. The EBG pattern is etched into the top and bottom layers, while the strip is etched into the middle layer. The channel width w_{ch} and strip width w_{strip} are indicated.

2.3.4 Electromagnetic Bandgap (EBG) Structures

As described above, EBG structures are periodic with a period comparable to the wavelength. Unlike free space, they do not support propagation at all frequencies. Instead, they support pass bands, where waves can propagate, and band gaps, where propagation is suppressed. Because the structure is periodic, the fields must satisfy Bloch-Floquet behavior. For a given wave vector $\mathbf{k}$, the periodic structure has a discrete set of eigenmodes with corresponding eigenfrequencies $\omega_n(\mathbf{k})$. A frequency interval where no such mode can propagate thus constitutes a band gap.

To avoid simulating the full infinite periodic structure, only a single unit cell of the EBG structure can instead be modeled with periodic boundary conditions. The wave vector $\mathbf{k}$ is then swept through the Brillouin zone along a high-symmetry path defined by the main symmetry points of the lattice. For each value of $\mathbf{k}$, an eigenvalue problem based on Maxwell's equations is solved to obtain the eigenfrequencies of the supported modes. Plotting these eigenfrequencies as a function of $\mathbf{k}$ gives the dispersion diagram, where each branch represents one mode of the periodic structure.

2.3.5 Manufacturing Techniques for GWG Technology

Traditionally, GWG structures were fabricated using computer numerical control (CNC) machining. At mmWave frequencies, this method becomes costly and time consuming, since GWG structures require tight tolerances and the periodic pin structures are small and intricate. This creates a need for very small CNC tools to realize the required dimensions. This has motivated research into more cost-effective manufacturing techniques for GWG structures. Some proposed techniques include:

- die-sink electrical discharge machining (EDM) [31],
- metallized 3D-printed polymer structures [32],
- metallized polymer injection-molded structures [33],
- dry-film photoresist micromachining [34],
- stacked chemically etched metal layers [30].

These approaches all aim to simplify the manufacturing of GWG structures and reduce fabrication cost. However, they have different advantages and limitations, and the most suitable method depends on the application, frequency band, required tolerances, and production volume. For the MLW coaxial line used in this work, the most relevant approach is chemically etched copper layers, since the structure is directly formed from several stacked metal layers.

2.4 Passive Reflectors

Passive reflectors are structures that scatter electromagnetic waves without using active components or external power. This section first introduces the distinction between antenna mode and structural mode scattering, which is important for understanding antenna-based reflectors. The section then focuses on the Van Atta

array, which is the retrodirective reflector studied in this thesis. Finally, thin metal plates and corner reflectors are described as reference reflectors for modeling and comparison.

2.4.1 Structural and Antenna Mode Scattering

Antenna scattering is commonly separated into two contributions: antenna mode scattering and structural mode scattering. Antenna mode scattering is the contribution associated with the receiving and reradiating behavior of the antenna. In this case, the incident field excites the antenna, the received signal propagates through the connected transmission line or waveguide, and the signal reflected by the termination is radiated back into space. The corresponding antenna mode RCS can be expressed using the antenna effective aperture as

$$\sigma_a = 4\pi |\Gamma|^2 \frac{A_e^2}{\lambda^2} \quad (2.13)$$

where σ_a is the antenna mode RCS, Γ is the reflection coefficient at the antenna termination, and A_e is the effective aperture, which is related to the realized gain through Equation 2.10. If the antenna is terminated by an ideal matched load, then $|\Gamma| = 0$, so the antenna mode RCS vanishes. The largest antenna mode RCS is obtained for an open-circuit or short-circuit termination, where $|\Gamma| = 1$.

By contrast, structural mode scattering corresponds to the scattering that would remain from the antenna body due to its geometry and material properties. The exact division between structural mode scattering and antenna mode scattering is somewhat arbitrary, since it depends on which currents and excitations are assigned to each contribution. A common convention, however, is to define the structural mode scattering as the scattering from the antenna when it is terminated with a matched load [35].

A purely theoretical prediction of antenna RCS is often difficult. In many cases, the antenna mode and structural mode contributions can be modeled fairly accurately independently. However, it is often very complicated to determine their phase relation [36].

2.4.2 Van Atta Array

Van Atta first introduced this type of retrodirective array (RDA) in 1959 [37]. To illustrate how a Van Atta array functions, consider the simple four-element linear Van Atta array in Figure 2.6 illuminated by a plane wave incident at θ_i .

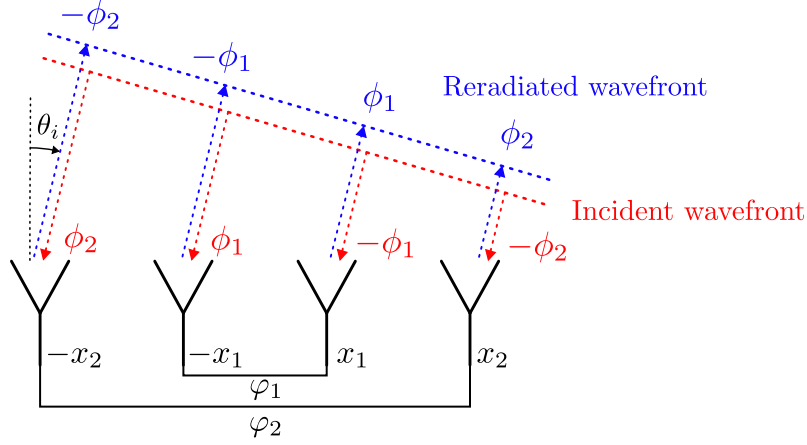


Figure 2.6: Operating principle of a linear Van Atta array. Symmetric element pairs are connected across the array center.

The incident phase at position x along the array is then given by

$$\phi_{\text{inc}}(x) = -kx \sin \theta_i, \quad (2.14)$$

where $k = 2\pi/\lambda$ is the wavenumber. In the four-element case shown in Figure 2.6, the element coordinates are $-x_2, -x_1, x_1, x_2$. Defining

$$\phi_1 = kx_1 \sin \theta_i, \quad \phi_2 = kx_2 \sin \theta_i, \quad (2.15)$$

the incident phases at the four element positions are therefore

$$\begin{aligned} \phi_{\text{inc}}(-x_2) &= \phi_2, \\ \phi_{\text{inc}}(-x_1) &= \phi_1, \\ \phi_{\text{inc}}(x_1) &= -\phi_1, \\ \phi_{\text{inc}}(x_2) &= -\phi_2. \end{aligned} \quad (2.16)$$

After the wave has propagated through the transmission lines, it is reradiated by the array elements with a new phase. Since the outer element pair and inner element pair may have different transmission line lengths, the reradiated phase distribution becomes

$$\begin{aligned} \phi_{\text{out}}(-x_2) &= \phi_{\text{inc}}(x_2) + \varphi_2 = -\phi_2 + \varphi_2, \\ \phi_{\text{out}}(-x_1) &= \phi_{\text{inc}}(x_1) + \varphi_1 = -\phi_1 + \varphi_1, \\ \phi_{\text{out}}(x_1) &= \phi_{\text{inc}}(-x_1) + \varphi_1 = \phi_1 + \varphi_1, \\ \phi_{\text{out}}(x_2) &= \phi_{\text{inc}}(-x_2) + \varphi_2 = \phi_2 + \varphi_2. \end{aligned} \quad (2.17)$$

When the two transmission lines have equal electrical length, so that φ_1 and φ_2 differ only by an integer multiple of 2π , the shared phase delay can be omitted, giving

$$\begin{aligned} \phi_{\text{out}}(-x_2) &\equiv -\phi_2, \\ \phi_{\text{out}}(-x_1) &\equiv -\phi_1, \\ \phi_{\text{out}}(x_1) &\equiv \phi_1, \\ \phi_{\text{out}}(x_2) &\equiv \phi_2. \end{aligned} \quad (2.18)$$

Thus, the phase gradient of the reradiated field is the reverse of the incident phase gradient in Equation 2.16. This means that the reradiated field sums coherently in the backscatter direction, so it propagates back toward the direction of incidence θ_i .

The phase delay of a transmission line is directly proportional to its physical length l through

$$\varphi(f) = \beta(f)l, \quad (2.19)$$

where $\beta(f)$ is the phase constant of the propagating mode. If two transmission lines have the same physical length, $l_1 = l_2 = l$, their electrical lengths are equal for all frequencies,

$$\varphi_1(f) = \varphi_2(f) = \beta(f)l. \quad (2.20)$$

In this ideal case, the bandwidth of the Van Atta principle itself is not limited by phase imbalance between the transmission lines. Instead, the bandwidth is limited only by the bandwidth of the antenna elements and the transmission lines.

If the transmission lines have different physical lengths, $l_1 \neq l_2$, they can only be made electrically equal at a single design frequency f_0 . This is done by choosing the lengths l_1 and l_2 such that

$$\varphi_1(f_0) - \varphi_2(f_0) = \beta(f_0)(l_1 - l_2) = 2\pi m, \quad m \in \mathbb{Z}. \quad (2.21)$$

Away from this frequency, the remaining phase error becomes

$$\Delta\varphi(f) = [\beta(f) - \beta(f_0)](l_1 - l_2), \quad f \neq f_0. \quad (2.22)$$

This phase error degrades the performance of the Van Atta array, since the antenna elements no longer add coherently. From Equation 2.22, it is clear that the phase error grows linearly with the physical length difference between l_1 and l_2 . It also depends directly on how $\beta(f)$ changes with frequency. For a nondispersive mode, $\beta(f)$ is approximately linear, so the phase error grows approximately linearly away from the design frequency. For a dispersive mode, $\beta(f)$ contains higher-order frequency dependence, which means that the phase error can grow quadratically, or even faster, away from the design frequency. This can significantly limit the bandwidth of the Van Atta array, especially if the difference in physical length is large.

The coherent addition of the reradiated field in the backscatter direction increases the RCS of the array. The achievable RCS enhancement depends primarily on the array size and the gain of the individual antenna elements. The theoretical maximum RCS for the antenna mode scattering of an ideal Van Atta array is given by

$$\sigma = \frac{n^2 \mathbb{G}_e^2 \lambda^2}{4\pi} \quad (2.23)$$

where $\mathbb{G}_e$ denotes the gain of each element and n denotes the total number of elements in the array [38]. Instead, using the effective aperture of the full array, the ideal antenna mode RCS can be written as

$$\sigma = \frac{4\pi A_{e,\text{array}}^2}{\lambda^2} \quad (2.24)$$

where $A_{e,\text{array}}$ is the effective aperture of the complete array.

2.4.3 Network-Based Van Atta Array Model

This section presents a network-based model for the antenna mode RCS of a Van Atta array. The model combines embedded antenna element far-field functions with an N -port network. The Van Atta array is modeled as a reciprocal N -element antenna array whose ports are interconnected by a passive N -port transmission network. By using an N -port formulation based on S-parameters, matching, insertion loss, phase, and isolation between the transmission lines are all taken into account.

For simplicity, only one cut with constant φ is considered, and only the co-polar component is included. The more general formulation follows analogously. Let the N antenna elements of the Van Atta array have the normalized embedded far-field functions

$$G_{\text{elem},n}(\theta) = 10^{\mathbb{G}_n(\theta)/20} e^{j\phi_n(\theta)}, \quad n = 1, \dots, N \quad (2.25)$$

where $\mathbb{G}_n(\theta)$ is the realized gain over angle of element n in the far field, expressed in dB, and $\phi_n(\theta)$ is the corresponding far-field phase.

The elements are located at the positions

$$x_n = \left(n - \frac{N+1}{2}\right) d, \quad n = 1, \dots, N \quad (2.26)$$

The corresponding far-field function of element n then becomes

$$G_n(\theta) = G_{\text{elem},n}(\theta) e^{jkx_n \sin \theta} \quad (2.27)$$

For a plane wave incident from direction θ_i with amplitude $|E_i| = 1$, the received wave at port n is

$$a_n^{\text{rx}}(\theta_i) = G_n^*(\theta_i) \quad (2.28)$$

The received port-wave vector is therefore

$$\mathbf{a}^{\text{rx}}(\theta_i) = \begin{bmatrix} a_1^{\text{rx}}(\theta_i) \\ \vdots \\ a_N^{\text{rx}}(\theta_i) \end{bmatrix} \quad (2.29)$$

The interconnecting N -port routing network then maps the received port-wave vector to the transmitted port-wave vector according to

$$\mathbf{b}^{\text{tx}}(\theta_i) = \mathbf{S} \mathbf{a}^{\text{rx}}(\theta_i) \quad (2.30)$$

where

$$\mathbf{S} \in \mathbb{C}^{N \times N} \quad (2.31)$$

is the S -parameter matrix of the N -port routing network.

The resulting scattered complex far-field function in a scattered direction θ_s is then the coherent sum over all transmitting elements,

$$F_s(\theta_s, \theta_i) = \sum_{n=1}^N b_n^{\text{tx}}(\theta_i) G_n(\theta_s) \quad (2.32)$$

From this, the scattered electric field at distance R is

$$E_s(R, \theta_s, \theta_i) = \frac{e^{-jkR}}{R} C F_s(\theta_s, \theta_i) \quad (2.33)$$

where C is a constant with units of meters introduced to recover the correct scaling and dimensionality, since $G_{\text{elem},n}$ is defined as a normalized dimensionless quantity. The bistatic RCS is defined as

$$\sigma(\theta_s, \theta_i) = 4\pi R^2 \frac{|E_s(R, \theta_s, \theta_i)|^2}{|E_i|^2} \quad (2.34)$$

Using $|E_i| = 1$ together with (2.33), this simplifies to

$$\sigma(\theta_s, \theta_i) = 4\pi C^2 |F_s(\theta_s, \theta_i)|^2 \quad (2.35)$$

For the isotropic reference case, the normalized scattered far-field quantity satisfies $|F_s| = 1$, while the corresponding RCS is $\lambda^2/(4\pi)$. Therefore, from (2.35), it follows that $4\pi C^2 = \lambda^2/(4\pi)$, giving $C = \lambda/(4\pi)$. This gives

$$\sigma(\theta_s, \theta_i) = \frac{\lambda^2}{4\pi} |F_s(\theta_s, \theta_i)|^2 \quad (2.36)$$

The monostatic RCS then follows by setting $\theta_s = \theta_i$,

$$\sigma_{\text{mono}}(\theta_i) = \sigma(\theta_i, \theta_i) \quad (2.37)$$

2.4.4 Reference Reflectors

This subsection presents the theory for the two reference reflectors used to model and compare with the Van Atta retroreflector.

2.4.4.1 Reflection from a Thin Metal Plate

For a thin electrically large PEC plate in the x - y plane, the bistatic co-polar RCS can be approximated as

$$\begin{aligned} \sigma_p(\theta_i, \varphi_i, \theta_s, \varphi_s) &\approx \frac{4\pi A^2}{\lambda^2} \cos^2 \theta_i \\ &\times \text{sinc}^2 \left[\frac{kL_x}{2} (\sin \theta_i \cos \varphi_i + \sin \theta_s \cos \varphi_s) \right] \\ &\times \text{sinc}^2 \left[\frac{kL_y}{2} (\sin \theta_i \sin \varphi_i + \sin \theta_s \sin \varphi_s) \right], \end{aligned} \quad (2.38)$$

where L_x and L_y are the side lengths of the plate, $A = L_x L_y$ is the plate area, and $k = 2\pi/\lambda$ is the wavenumber [39]. For the monostatic case, using the angular convention of Equation 2.38, the incident and scattered angles are equal, that is $\theta_s = \theta_i$ and $\varphi_s = \varphi_i$. This reduces the expression to

$$\begin{aligned} \sigma_p^{\text{mono}}(\theta_i, \varphi_i) &\approx \frac{4\pi A^2}{\lambda^2} \cos^2 \theta_i \\ &\times \text{sinc}^2[kL_x \sin \theta_i \cos \varphi_i] \text{sinc}^2[kL_y \sin \theta_i \sin \varphi_i]. \end{aligned} \quad (2.39)$$

2.4.4.2 Corner Reflectors

Corner reflectors are retroreflectors that reflect incident waves using three mutually perpendicular plates. For a given physical size, they can provide a relatively high RCS while remaining simple to manufacture. A common corner reflector used for radar testing is the triangular trihedral reflector, shown in Figure 2.7.

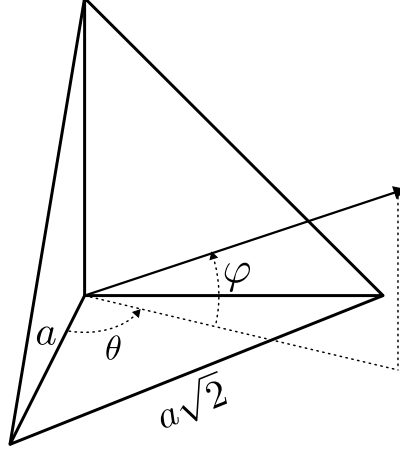


Figure 2.7: Geometry of a triangular trihedral corner reflector

If the corner reflector is much larger than a wavelength, an analytical expression for the monostatic RCS over viewing angle is presented in [40]

$$\sigma(\varphi, \theta) = \begin{cases} \frac{4\pi}{\lambda^2} a^4 \left(\frac{4c_1 c_2}{c_1 + c_2 + c_3} \right)^2, & \text{for } c_1 + c_2 \leq c_3 \\ \frac{4\pi}{\lambda^2} a^4 \left(c_1 + c_2 + c_3 - \frac{2}{c_1 + c_2 + c_3} \right)^2, & \text{for } c_1 + c_2 \geq c_3 \end{cases} \quad (2.40)$$

where c_1, c_2, c_3 are the ordered values of

$$\sin \varphi, \cos \varphi \sin \theta, \cos \varphi \cos \theta \quad (2.41)$$

such that $c_1 \leq c_2 \leq c_3$. In other words, the three angular expressions are first evaluated for the chosen direction, and then assigned to c_1, c_2 , and c_3 from smallest to largest. The effective area at boresight for a triangular trihedral corner reflector is given by

$$A_{\text{eff,tri}} = \frac{a^2}{\sqrt{3}}. \quad (2.42)$$

3

Design and Measurement Methods

This chapter describes the methods used to design, model, and validate the passive Van Atta array. The chapter first describes the design and verification of the MLW coaxial line. It then presents the radiating elements, the transmission network, and the final reflector assembly. Finally, the numerical reflector model and the measurement setup used for validation are described.

3.1 Simulations

All electromagnetic simulations needed to design and validate the Van Atta array were performed in CST Studio Suite 2024. The CST eigenmode solver was used to simulate the dispersion of the EBG structures. For all other simulations, the frequency-domain solver was used.

To design and tune the subcomponents of the Van Atta array, such as the slotted waveguide columns and transmission networks, the structures were excited using waveguide ports at the interfaces where the subcomponents are connected. The simulated S-parameters and far-field patterns were then used to evaluate matching, coupling, phase delay, and radiation performance. To extract the monostatic and bistatic RCS of the complete Van Atta array, the plane wave excitation in CST was used instead. For all simulations of the MLW Van Atta array a nominal air gap of $5\mu\text{m}$ was used between the layers. The material used for the simulations was annealed copper.

3.2 MLW Coaxial Line Design

The Van Atta array is implemented using an MLW coaxial line. The first step of the design was therefore to choose and verify the basic parameters of the line. Three important parameters are the strip width, the channel width, as outlined in Figure 2.5, and the dimensions of the periodic pin structure. The aim when choosing these dimensions is to guide the TEM mode in the channel without also exciting a TE_{10} -like mode, while ensuring that the pin structure creates a stopband over the frequency band of interest.

The EBG structure was first verified using the unit-cell simulation approach described in subsection 2.3.4. The dispersion diagram in Figure 3.1 was obtained for a square-symmetric pin structure with dimensions adapted from [30]. The simulated stopband extends from approximately 39.6 GHz to 176.4 GHz, even at a gap of $g = 0.01$ mm.

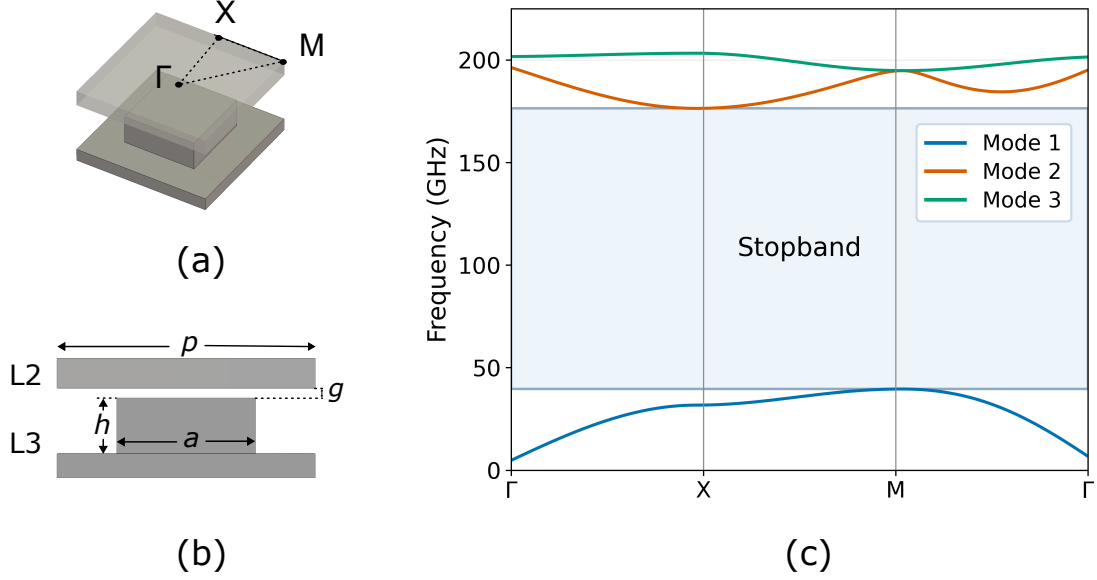


Figure 3.1: Dispersion analysis for the square-symmetric EBG pin structure. (a) Simulated unit cell with the high-symmetry path in the Brillouin zone. (b) Cross section of the unit cell with the main geometrical parameters, where $p = 1.3$ mm, $a = 0.7$ mm, $h = 0.28$ mm, and $g = 0.01$ mm. (c) Dispersion diagram along the Γ -X-M- Γ path, showing a stopband from 39.6 GHz to 176.4 GHz.

The values for the strip and channel widths, $w_{\text{strip}} = 0.75 \text{ mm}$ and $w_{\text{ch}} = 1.45 \text{ mm}$, were adapted from [30]. To verify that these values are suitable, the structure in Figure 2.5 was simulated using the same unit-cell approach, but with the $\mathbf{k}$ -vector swept only in the waveguide direction over one pin period p . Figure 3.2 shows the propagating modes of this structure. Across the band of interest, the desired TEM mode is the only propagating mode. This mode is almost perfectly nondispersive, which is important since dispersion in the transmission network would introduce additional frequency-dependent phase errors. The cutoff frequency of the unwanted TE_{10} -like mode is 90.2 GHz, placing it well above the band of interest. At much lower frequencies outside the EBG stopband, several modes can propagate both in the waveguide channel and in the surrounding EBG structure. However, these modes are far below the operating band and are therefore not relevant. Overall, this shows that the selected MLW structure is suitable for the 77 GHz automotive radar band.

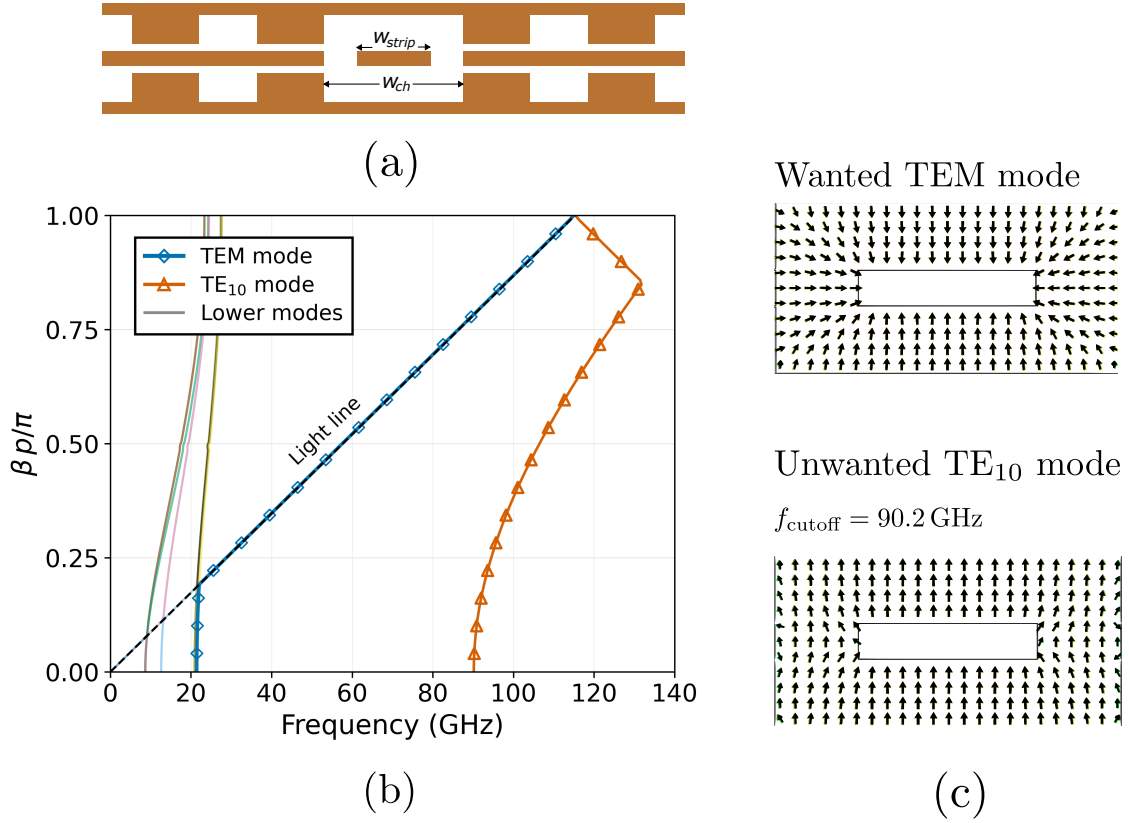


Figure 3.2: MLW channel dispersion in the waveguide direction for $w_{\text{strip}} = 0.75 \text{ mm}$, $w_{\text{ch}} = 1.45 \text{ mm}$, $p = 1.3 \text{ mm}$, and a $5 \mu\text{m}$ gap between layers. (a) Simulated unit cell. (b) Dispersion diagram with the desired TEM mode, the unwanted TE_{10} -like mode, and lower-order modes. The cutoff frequency of the TE_{10} -like mode is approximately 90.2 GHz. (c) Simulated field distributions for the desired TEM mode and the unwanted TE_{10} -like mode.

There are also mechanical considerations for the MLW design. One of these is that, for long transmission lines in MLW, the coaxial line needs mechanical support to avoid sagging. To achieve this, pairs of thin inductive support stubs were placed between the stripline and the surrounding metal, as shown in Figure 3.3. These support stubs negatively affect the matching of the line, so the distance between the stubs in each pair, sup_d , and the distance between pairs, sup_p , were tuned locally for each transmission line section to minimize their effect on the matching.

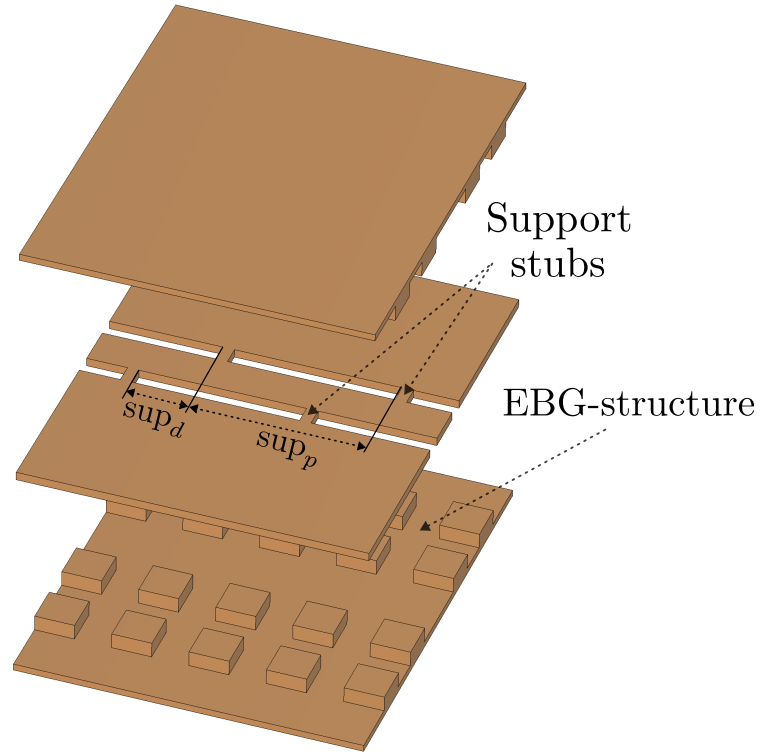


Figure 3.3: Exploded view of a section of MLW line showing the support stubs.

3.3 Van Atta Array Design

This section describes the design of the different subcomponents that make up the passive Van Atta array and how they are combined. This includes the radiating elements, the transmission network, and the final assembly.

3.3.1 Slotted Waveguide Column Design

The choice of radiating element was made to achieve a compact array with an element spacing as close to $\lambda/2$ as possible. To achieve this, a series-fed slotted waveguide column was chosen. In the feeding direction, conventional slots are too long to allow an element spacing of $\lambda/2$, and a dogbone-shaped slot design was therefore used instead. The slots are excited by a meandering coaxial line running in the layer beneath them. Below each slot, the coaxial line is offset to the right for one half of the slot and to the left for the other half. Since the design is intended to provide retroreflection only in azimuth, a relatively narrow elevation beamwidth can be used to increase the antenna gain. However, making this beamwidth too narrow would make the array more sensitive to orientation, which can be problematic if the reflector is not mounted perfectly vertically or if a vehicle approaches from a slope. Because of this trade-off, a slotted column with six slots was chosen, giving a beamwidth of approximately 16° in elevation. The slot orientation was chosen to produce horizontal polarization. The resulting geometry is shown in Figure 3.4.

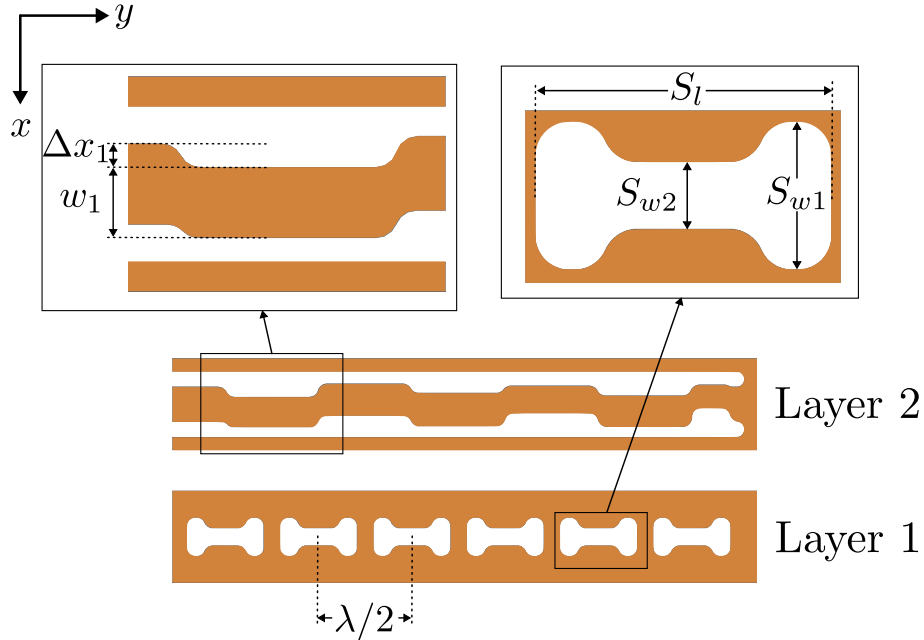


Figure 3.4: Proposed slot column structure. Layer 1 has been offset in x to show the coaxial line running beneath feeding the slots. All slots are identical, but the meandering varies with $0.10 \text{ mm} \leq \Delta x_i \leq 0.17 \text{ mm}$ and $0.50 \text{ mm} \leq w_i \leq 0.75 \text{ mm}$. The remaining dimensions are $S_l = 1.7 \text{ mm}$, $S_{w1} = 0.85 \text{ mm}$, and $w_2 = 0.39 \text{ mm}$.

The main challenge in designing the series-fed slot array is to achieve controlled slot

excitation in order to minimize beam squint in the elevation plane for all frequencies. To achieve this, the offset of each meander section, Δx , and the width of that section, w , were tuned. The dimensions of the dogbone slots were also tuned according to the parameters shown in Figure 3.4. To evaluate and optimize this, the slot column was simulated by exciting it with a waveguide port and extracting the far-field pattern in azimuth and elevation. The optimization targets were input matching, directivity, sidelobe level, and a main beam pointing angle of 0° at 76 GHz, 78.5 GHz, and 81 GHz. For this optimization, CST's covariance matrix adaptation evolution strategy (CMA-ES) algorithm was run for 200 iterations with the goals and parameters described above. The finalized element performance is presented in section 4.1.

To create the array, the slotted columns were grouped into the two six-element banks shown in Figure 3.5, with every other element fed from opposite sides of the array to leave more space for routing. The banks themselves are not mirror symmetric about the center point. Instead, each column is identical to its connected counterpart in the other antenna bank, for example element 1 with element 12, element 2 with element 11, and so on.

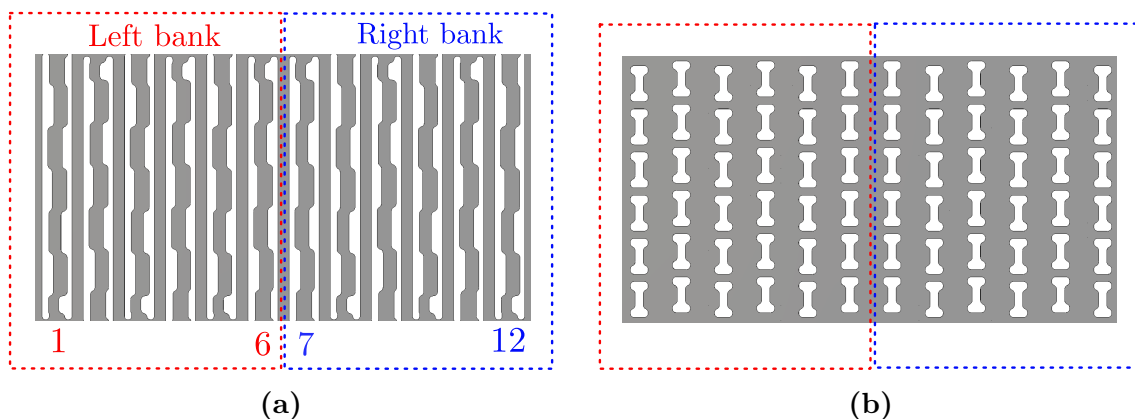


Figure 3.5: Van Atta array radiating element layout. (a) L2, meandered stripline. (b) L1, slot aperture.

3.3.2 Transmission Network Design

To connect the slot columns in the array, one transmission network was designed on each side of the radiating elements. For the Van Atta array design, the key requirement is equal phase delay through all transmission lines in the networks. However, since the lines have different physical lengths, this condition can only be satisfied at a single frequency. Here, the lines were designed to have equal phase delay at the center frequency, 78.5 GHz. This leads to phase mismatch at the band-edge frequencies, since the electrical lengths of the lines vary with frequency. Larger physical length differences between the transmission lines lead to larger phase mismatch. For this reason, the transmission lines were kept as short as possible in the design.

Each transmission network was simulated individually in CST to verify input matching, low coupling between the individual transmission lines, and the phase delay of

each connected element pair. The two networks were first tuned so that the lines within each network had matched phase delay at 78.5 GHz. A common length extension was then added to one of the networks so that all six connected element pairs had the same phase delay at 78.5 GHz.

In Figure 3.6, the array with the two transmission networks is shown rotated by 90° to better fit the page. The six lines are outlined, and the increase in physical length can be seen clearly. The line between elements 1–12 is the longest, at approximately 32.8 mm, while the line between elements 6–7 is the shortest, at approximately 7.4 mm.

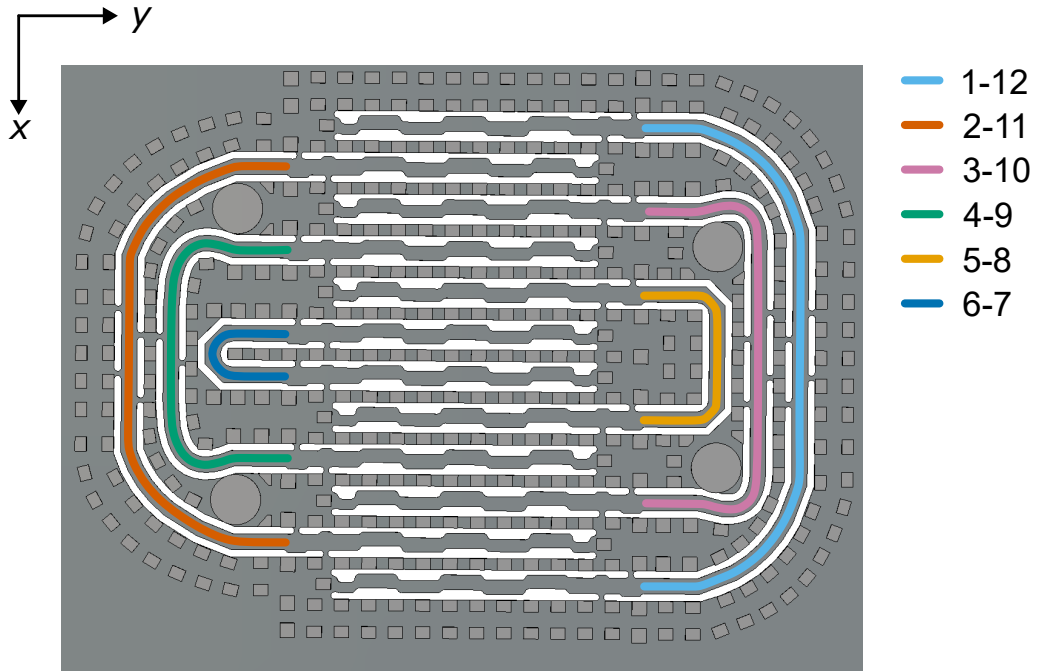


Figure 3.6: Final transmission line layout of the complete Van Atta array showing layer 2. The colored lines indicate the six connected element pairs, starting with the outermost pair 1–12 and ending with the innermost pair 6–7, with the intermediate pairs 2–11, 3–10, 4–9, and 5–8.

3.3.3 Full Retroreflector Assembly and Manufacturing

In Figure 3.7, the final reflector is shown together with mechanical features such as alignment holes for keeping the layers aligned during manufacturing. As described above, layers 1 and 3 are $400\text{ }\mu\text{m}$ thick, while layer 2 is $150\text{ }\mu\text{m}$ thick. This gives the reflector a total thickness of only 0.95 mm. The EBG structure, coaxial lines, and other features are formed by etching the copper layers. After etching, the three layers are stacked and joined using laser welding. In the reflector simulations, a nominal air gap of $5\text{ }\mu\text{m}$ was included between the layers. However, as described in section 2.3, a wide range of air gaps can be tolerated before any major impact on performance is observed. This makes the manufacturing requirements less strict.

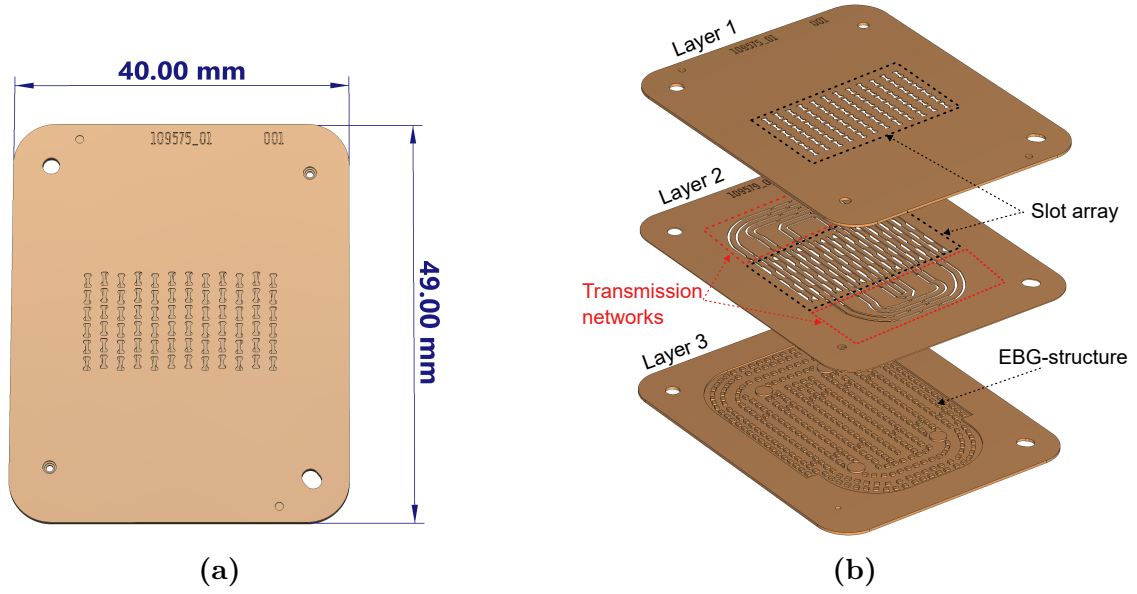


Figure 3.7: Final reflector design. (a) CAD model of the assembled reflector. (b) Exploded view showing the individual layers.

Figure 3.8 shows the manufactured reflector. The laser-welding marks can be seen on the reflector, together with scratches and stains from manufacturing and handling.

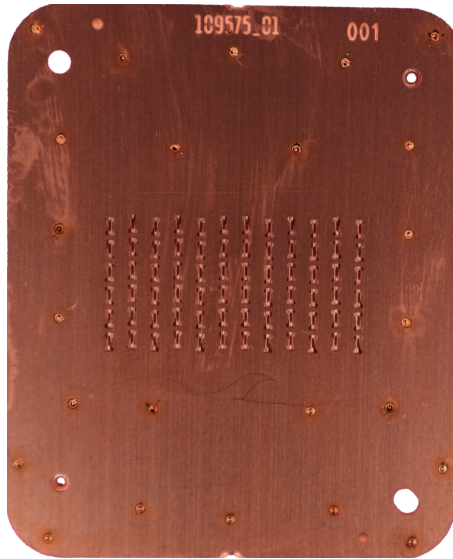


Figure 3.8: Fabricated Van Atta array reflector.

3.4 Numerical Van Atta Array Model

The network-based Van Atta array model introduced in subsection 2.4.3 was implemented numerically to quickly evaluate how design changes affect the scattering response of the array. The model uses separately simulated embedded far-field functions from the slot columns and the S-parameter matrix of the transmission network

as input, and calculates the monostatic and bistatic RCS. This provides a faster alternative to repeated full-wave simulations of the complete Van Atta array in CST, which become increasingly time-consuming as the array size grows. For example, if a slotted waveguide column is modified, only the embedded far-field patterns of that element need to be updated. Similarly, if the transmission network is modified, only the corresponding S-parameter matrix needs to be re-simulated. These separate simulations can be performed using waveguide ports and typically require significantly shorter simulation time.

3.4.1 Python Implementation

The numerical model was implemented in Python. The simulated data from CST were exported as S-parameters in a .s12p file and embedded far-field functions of the slotted columns. The far-field patterns were exported at the center and edge frequencies, 76 GHz, 78.5 GHz, and 81 GHz. These data were then imported into Python and used as input to the numerical model.

The calculation of F_s from Equation 2.32 was implemented in Python as a matrix multiplication in NumPy. In the implementation, the incident angle θ_i is treated as a scalar, while the scattered angle is sampled at M discrete angles, such that

$$\boldsymbol{\theta}_s = [\theta_{s,1} \quad \cdots \quad \theta_{s,M}]^T. \quad (3.1)$$

The scattered field over the sampled angular space can then be written as

$$\mathbf{F}_s(\boldsymbol{\theta}_s, \theta_i) = \mathbf{G}(\boldsymbol{\theta}_s) \mathbf{S} \mathbf{g}^*(\theta_i), \quad (3.2)$$

with

$$\mathbf{G}(\boldsymbol{\theta}_s) = \begin{bmatrix} G_1(\theta_{s,1}) & \cdots & G_N(\theta_{s,1}) \\ \vdots & \ddots & \vdots \\ G_1(\theta_{s,M}) & \cdots & G_N(\theta_{s,M}) \end{bmatrix}, \quad (3.3)$$

and

$$\mathbf{g}(\theta_i) = \begin{bmatrix} G_1(\theta_i) \\ \vdots \\ G_N(\theta_i) \end{bmatrix}. \quad (3.4)$$

To obtain a monostatic sweep over K incident angles, the model is evaluated K times with

$$\boldsymbol{\theta}_s = [\theta_i], \quad (3.5)$$

such that the scattered angle calculated is only equal to the incident angle in each evaluation.

3.4.2 Structural Mode Approximation

The numerical model only captures the antenna mode scattering of the reflector. In practice, the structural mode scattering also contributes to the total reflector RCS, especially for angles close to normal incidence. To account for this, a simple approximation is to model the reflector as a thin, large metal plate and use Equation 2.38 with the physical dimensions of the full reflector, including non-radiating areas, $L_x = 40$ mm and $L_y = 49$ mm. Let the corresponding scattered field be denoted by E_{plate} . The total scattered field can then be written as

$$E_{\text{tot}} = E_{\text{plate}}e^{j\phi_{\text{plate}}} + E_s \quad (3.6)$$

where ϕ_{plate} is an unknown phase offset between the structural mode and antenna mode contributions, and E_s is the antenna mode scattered field defined in Equation 2.33. The total RCS is then calculated from E_{tot} using the definition in Equation 2.34.

3.5 Measurements and Validation

In this section, the physical measurement setup is first presented. This is followed by an overview of how the measured data were processed to extract the RCS of the measured reflector.

3.5.1 Measurement Setup

To measure the monostatic RCS of the Van Atta array, an anechoic chamber normally used for antenna measurements was adapted. The target under test (TUT), in this case the reflector, was placed on a camera gimbal that could be rotated by hand to a specific angle, allowing the monostatic RCS to be swept over azimuth. The rotating structure was placed behind an absorber to reduce unwanted reflections from the mounting structure. A standard gain horn with 20 dBi gain was connected to a VNA through a frequency extender. The horn was mounted on a robot arm, which was used to control its position and rotation so that it was centered on the reflector.

Before the RCS measurements, a standard short-open-load calibration was performed at the output of the extender. The open termination was realized using a short together with a $\lambda/4$ shim. This calibration was used to correct the measured S_{11} levels before the corner reflector and TUT measurements were performed.

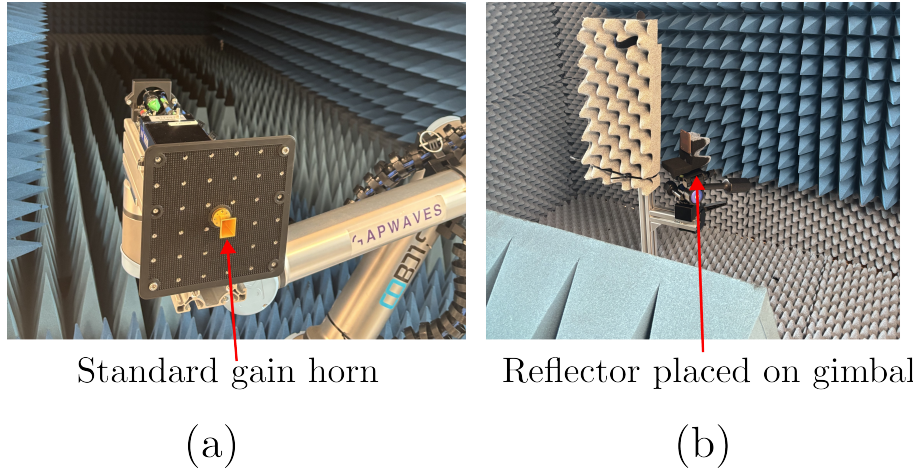


Figure 3.9: Measurement setup inside the anechoic chamber. (a) Standard gain horn and robot arm. (b) Reflector on a gimbal behind an absorber wall.

To calibrate the setup, a trihedral corner reflector with a known RCS of 10 dBsm at 77.5 GHz was measured first. The Van Atta reflector was then mounted on the gimbal for the measurement. The same nominal position was used for both the corner reflector and the TUT. The horn polarization was also aligned with that of the reflector.

The gimbal was then manually rotated in approximately 4.5° steps, with an S_{11} frequency sweep recorded by the VNA at each position. Measurements were taken over the full azimuth interval from -90° to 90° . Since the angular sweep was performed manually, the angle setting introduces some uncertainty in the measured RCS pattern. In total, this was repeated for three manufactured reflectors.

The frequency sweep covered 70 GHz to 87 GHz with a frequency step of 25 MHz. This corresponds to a bandwidth of 17 GHz, giving a range resolution of approximately

$$\Delta R = \frac{c}{2B} \approx 0.9 \text{ cm.} \quad (3.7)$$

The frequency step gives a range period of

$$R_{\text{period}} = \frac{c}{2\Delta f} \approx 6 \text{ m.} \quad (3.8)$$

However, since only the positive-time values from the transformation to the time domain can be used, the usable unambiguous range is half of this value,

$$R_{\text{max}} = \frac{c}{4\Delta f} \approx 3 \text{ m.} \quad (3.9)$$

The measurement was performed at a range of approximately 1 m, which is well within the usable unambiguous range. This range was selected as a compromise between far-field condition and available dynamic range. Considering only the radiating aperture areas of the reflector and horn, which are approximately 116 mm^2 and 291 mm^2 , respectively, the far-field condition is fulfilled at approximately 8 cm

and 20 cm. However, if the full reflector size, including the non-radiating parts, is considered, with dimensions 40 mm by 49 mm, the far-field condition is fulfilled only at approximately 2.1 m. The dynamic range of the measurement setup did not allow measurements at this longer distance, which is why the measurement was performed at 1 m instead.

The measurement was therefore not fully in the far-field region when the complete reflector geometry is considered. This may have a small effect on the result, especially on the exact ripple close to boresight, where structural mode scattering is important. However, the distance was considered sufficient for validating the reflector response, especially away from boresight, where the antenna mode scattering dominates.

Residual reflections from the horn, mounting structure, and chamber were reduced by placing absorber material in front of the rotating structure. Remaining unwanted reflections were further suppressed by the time-gating procedure described in the next subsection.

3.5.2 Data Processing

As mentioned above, the measurement setup was limited to one-port measurements of S_{11} , since only a single antenna and port could be used. A simplified representation of the one-port measurement geometry is shown in Figure 3.10. The antenna reference plane is denoted by $z = 0$, and the TUT is located at range R . At the antenna interface, a_1 denotes the incident port wave delivered to the antenna, while b_1 denotes the reflected port wave measured at the same interface. The measured reflection coefficient is therefore

$$S_{11} = \frac{b_1}{a_1}. \quad (3.10)$$

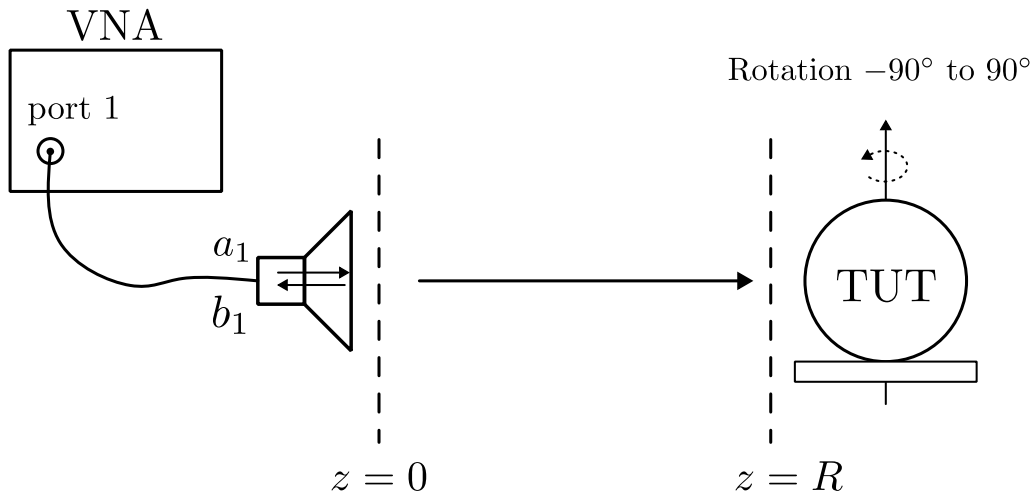


Figure 3.10: Simplified one-port measurement geometry used for the S_{11} time-gating framework. The outgoing port wave is denoted by a_1 , while the received wave is denoted by b_1 .

The measured S_{11} power level is proportional to the received-to-transmitted power ratio

$$|S_{11}|^2 = \frac{|b_1|^2}{|a_1|^2} \propto \frac{P_r}{P_t}. \quad (3.11)$$

Therefore, the gated S_{11} power level can be used together with a known reference object to calibrate the RCS of the TUT using Equation 2.2 and Equation 2.3. However, the raw S_{11} measurement cannot be used directly because it is dominated by the first reflections from the cable connections and the horn antenna. Because of this, the measured S_{11} first has to be transformed to the time domain, so that the reflection from the TUT can be time gated. The signal processing chain used to isolate the reflection from the TUT is summarized in Figure 3.11. Each processing step shown in the figure is described in more detail below.

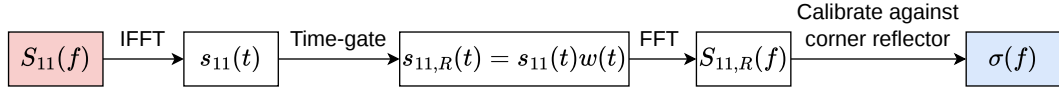


Figure 3.11: Signal processing chain for extracting the reflector RCS.

The measured $S_{11}(f)$ was transformed to the time domain using an inverse fast Fourier transform (IFFT),

$$s_{11}(t) = \text{IFFT} \{S_{11}(f)\}. \quad (3.12)$$

The time response is related to the physical measurement distance by

$$R = \frac{ct}{2}, \quad (3.13)$$

where c is the speed of light, and the factor of two accounts for the two-way propagation path. The reflector response at range R can then be separated in the time domain by applying a window function $w(t)$ around the corresponding delay. In this work, a Kaiser window was used. After applying the time gate, the response becomes

$$s_{11,R}(t) = s_{11}(t)w(t). \quad (3.14)$$

A fast Fourier transform (FFT) was then applied to return the gated response to the frequency domain,

$$S_{11,R}(f) = \text{FFT} \{s_{11,R}(t)\}. \quad (3.15)$$

An example of the time-gating procedure applied to a measured S_{11} response is shown in Figure 3.12. The raw frequency-domain response is first transformed to the time domain, where the target reflection is isolated using the time gate, before being transformed back to the frequency domain.

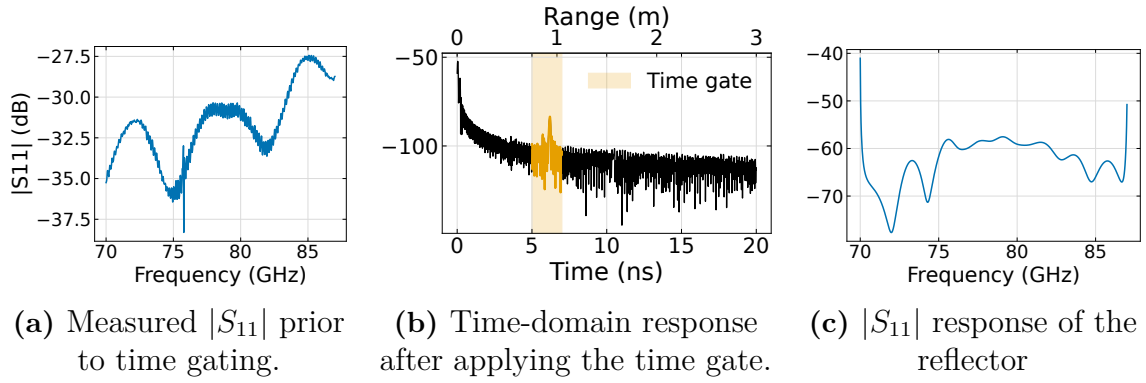


Figure 3.12: Processing of the measured S_{11} response using time gating.

This procedure was first applied to the trihedral corner reflector to obtain a reference response. The TUT measurement was processed in the same way. By comparing the gated S_{11} power level of the TUT with the gated S_{11} power level of the reference corner reflector, the absolute RCS of the TUT can be determined by

$$\sigma_{\text{TUT}}[\text{dBsm}] = \sigma_{\text{ref}}[\text{dBsm}] + S_{11,\text{TUT}}[\text{dB}] - S_{11,\text{ref}}[\text{dB}]. \quad (3.16)$$

4

Results and Validation

4.1 Slotted Waveguide Column

The optimized slotted waveguide column is evaluated in this section. Figure 4.1 presents the simulated azimuth and elevation radiation patterns of the final design. Across the frequency band, the maximum realized gain lies between 12.8 dBi and 13.4 dBi. From Equation 2.23, this corresponds to a maximum theoretical antenna mode RCS between -11.9 dBsm and -11.2 dBsm for a 12-element Van Atta array. In the elevation plane, the main beam stays close to broadside, with the pointing angle varying from -0.7° to -1.5° across the band. The 3 dB beamwidth is 82° in azimuth and 16° in elevation, and the elevation sidelobe level remains below -10 dB at all simulated frequencies.

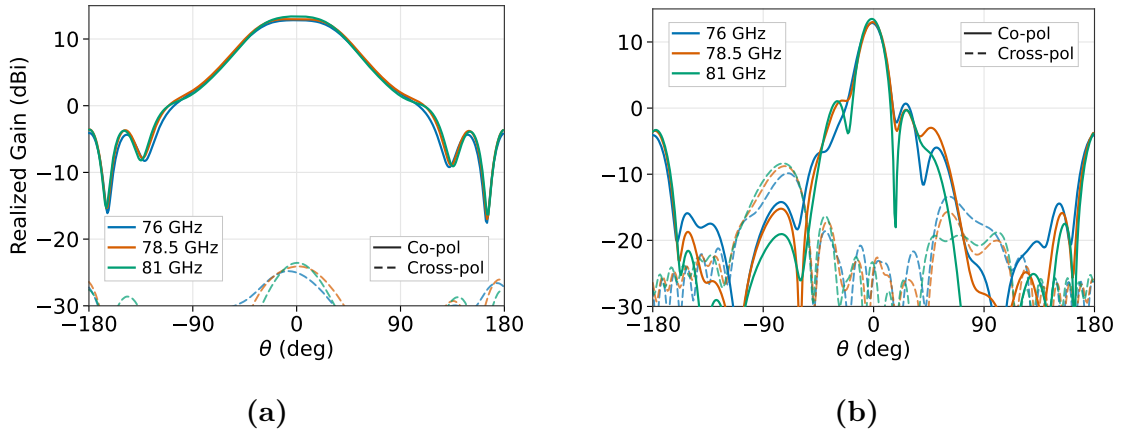


Figure 4.1: Simulated realized gain of the final slot column design for co-polar and cross-polar radiation. (a) Azimuth cut. (b) Elevation cut.

The nominal simulated input matching of the six-slot column is shown in Figure 4.2. Across the frequency band of interest, the nominal matching remains better than -15 dB. The simulation also shows a radiation efficiency that is better than 96 % over frequency.

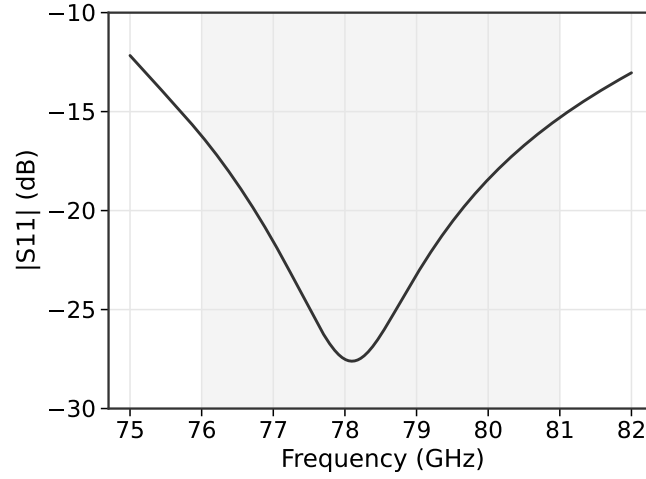


Figure 4.2: Simulated input matching, $|S_{11}|$, of the six-slot column.

4.2 Transmission Networks

The two transmission networks were simulated independently to evaluate input matching, phase delay, and insertion loss. Figure 4.3 presents the phase difference and insertion loss for the six transmission lines. Since the design target was equal phase at the center frequency of 78.5 GHz, the phase difference is smallest there. At this frequency, the largest phase difference is 10.8° . At the edge frequencies of 76 GHz and 81 GHz, as expected, the transmission lines with the largest difference in physical length also show the largest phase difference. These are the transmission lines between elements 1–12 and 6–7, for which the worst-case phase difference at the band edges is 76° . As expected, the insertion loss also increases with transmission line length, and even for the longest transmission line it remains very low at approximately -0.3 dB in band.

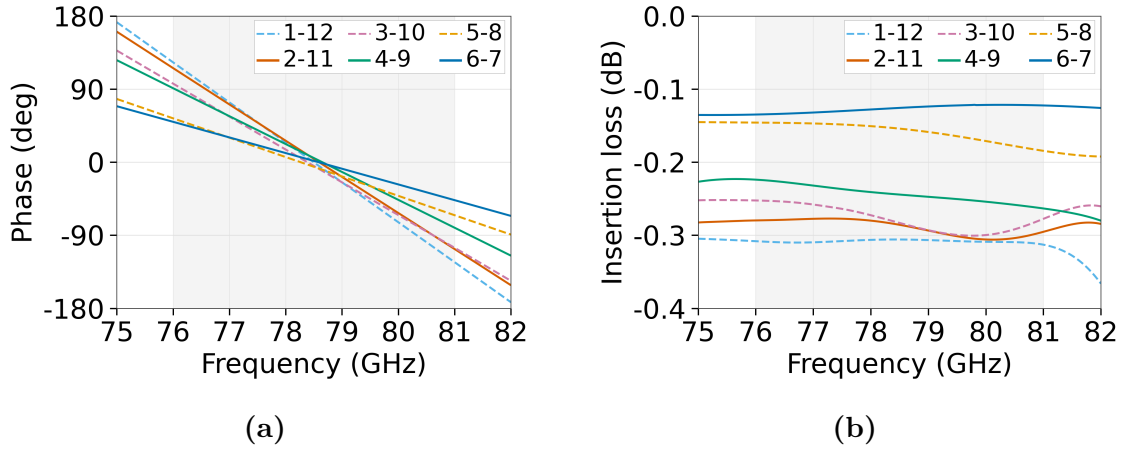


Figure 4.3: Simulated transmission network performance over frequency. (a) Phase delay. (b) Insertion loss.

Isolation between neighboring transmission lines and the input matching were also simulated. The results showed that the input match was better than -19 dB for all ports, and that the coupling between all combinations of neighboring transmission lines was below -50 dB.

4.3 Reflector Simulations

While the monostatic RCS is the most important aspect of a retroreflector, since it shows how much energy is reflected back toward a radar sensor, it is also useful to study the bistatic scattering of the Van Atta array to see how the energy is distributed for different incident angles. In Figure 4.4, simulated bistatic patterns for four different incident angles are presented. It is immediately clear that most of the energy is scattered in the specular direction, $\theta_s = -\theta_i$, for all incident angles. This is expected, since the metal area around the slots is large and the structural mode scattering dominates. There is also a clear retroreflection peak around the

incident direction $\theta_s = \theta_i$, which shows that the retroreflection principle is working. The retroreflection peaks are caused by the antenna mode scattering, with the final level affected by interference with structural mode scattering. This can produce stronger peaks in some cases, but split peaks and lower retroreflection in others. The interference pattern also changes with both observation angle and frequency. For example, at $\theta_i = 15^\circ$ the interference is destructive at 81 GHz, while at $\theta_i = 30^\circ$ it is destructive at 76 GHz.

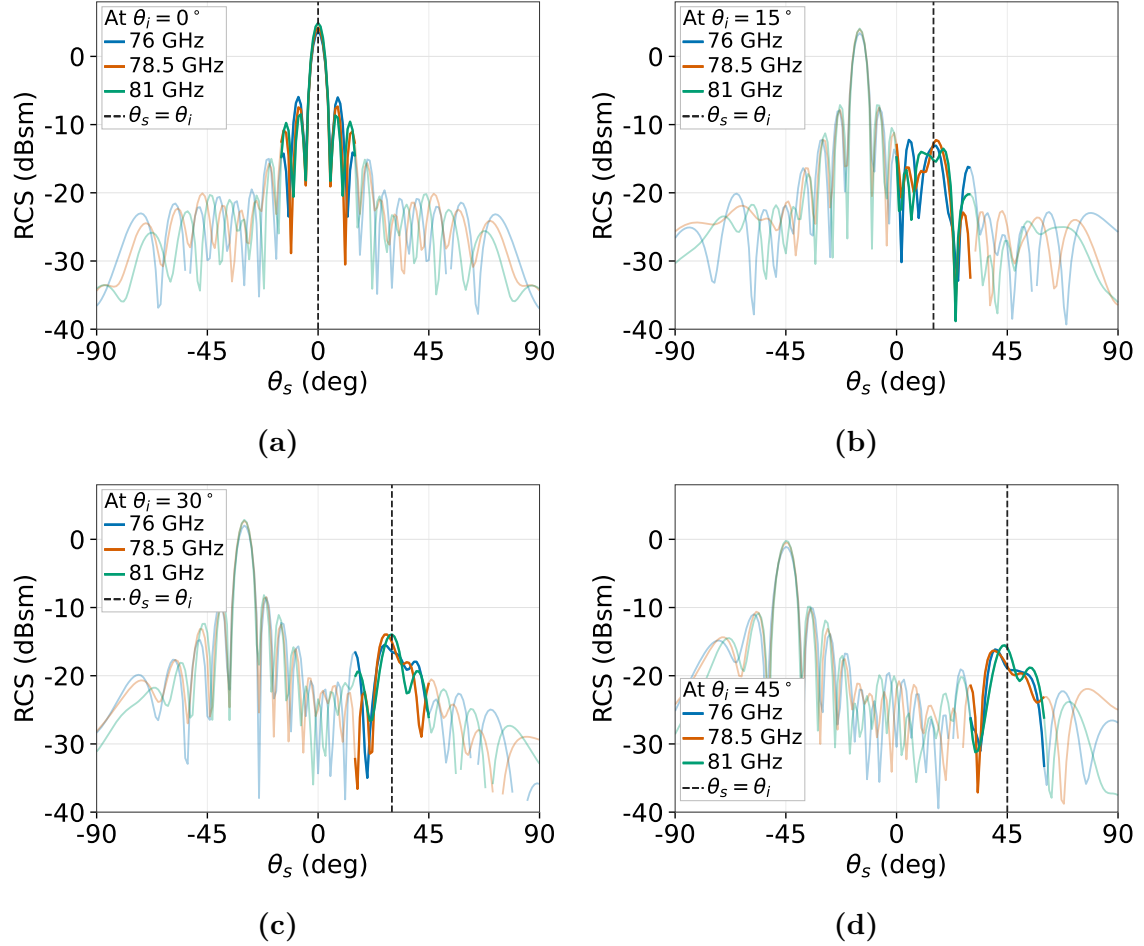


Figure 4.4: Simulated bistatic RCS patterns for different incident angles in the azimuth plane. (a) $\theta_i = 0^\circ$. (b) $\theta_i = 15^\circ$. (c) $\theta_i = 30^\circ$. (d) $\theta_i = 45^\circ$.

In Figure 4.5, the simulated monostatic RCS pattern is shown for azimuth and elevation cuts, and for both HH and VV polarization. As expected, strong retroreflection is only observed in the azimuth cut for HH polarization, which is the case considered in the design. In the other cases, the response is mainly due to structural mode scattering.

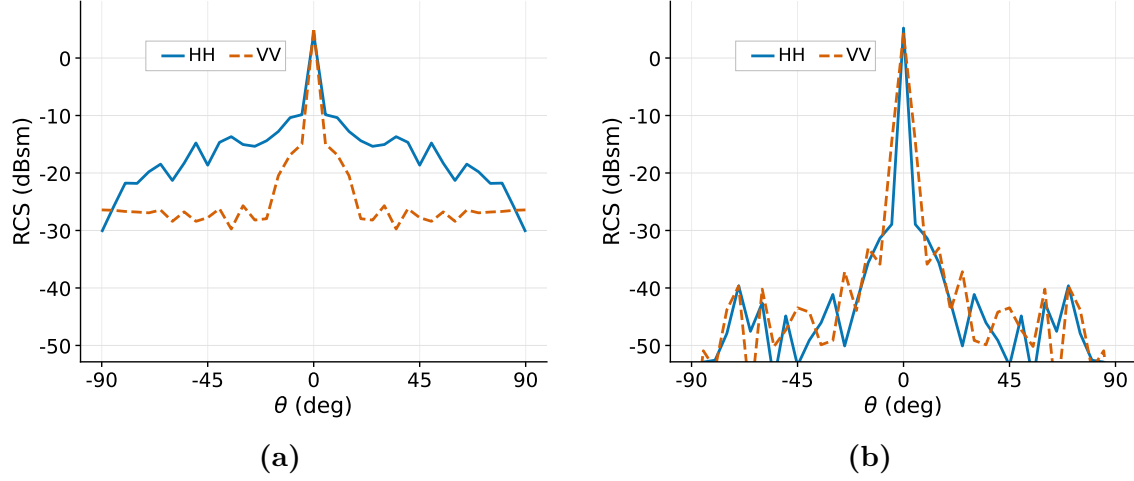


Figure 4.5: Monostatic RCS simulation at 78.5 GHz for HH and VV polarization. (a) Azimuth cut, $\varphi = 0^\circ$. (b) Elevation cut, $\varphi = 90^\circ$.

4.4 Reflector Measurements

In this section, the measured reflector results are presented. The results are compared with reference reflectors, and the functional angular range is introduced as a metric for quantifying how well the reflectors perform.

4.4.1 Measured Monostatic RCS

In Figure 4.6, all three reflectors are compared with the simulated values at three frequencies in the band. The measured responses of the three reflectors are very similar, which indicates that the design can be manufactured repeatably and that the fabrication tolerances are manageable. A larger number of samples would be needed to fully confirm this, but the results provide a good first indication. The figure also shows that the retrodirective effect is present at all three frequencies, with retroreflection over approximately -45° to 45° , even though the phase mismatch reaches about 60° at the band-edge frequencies.

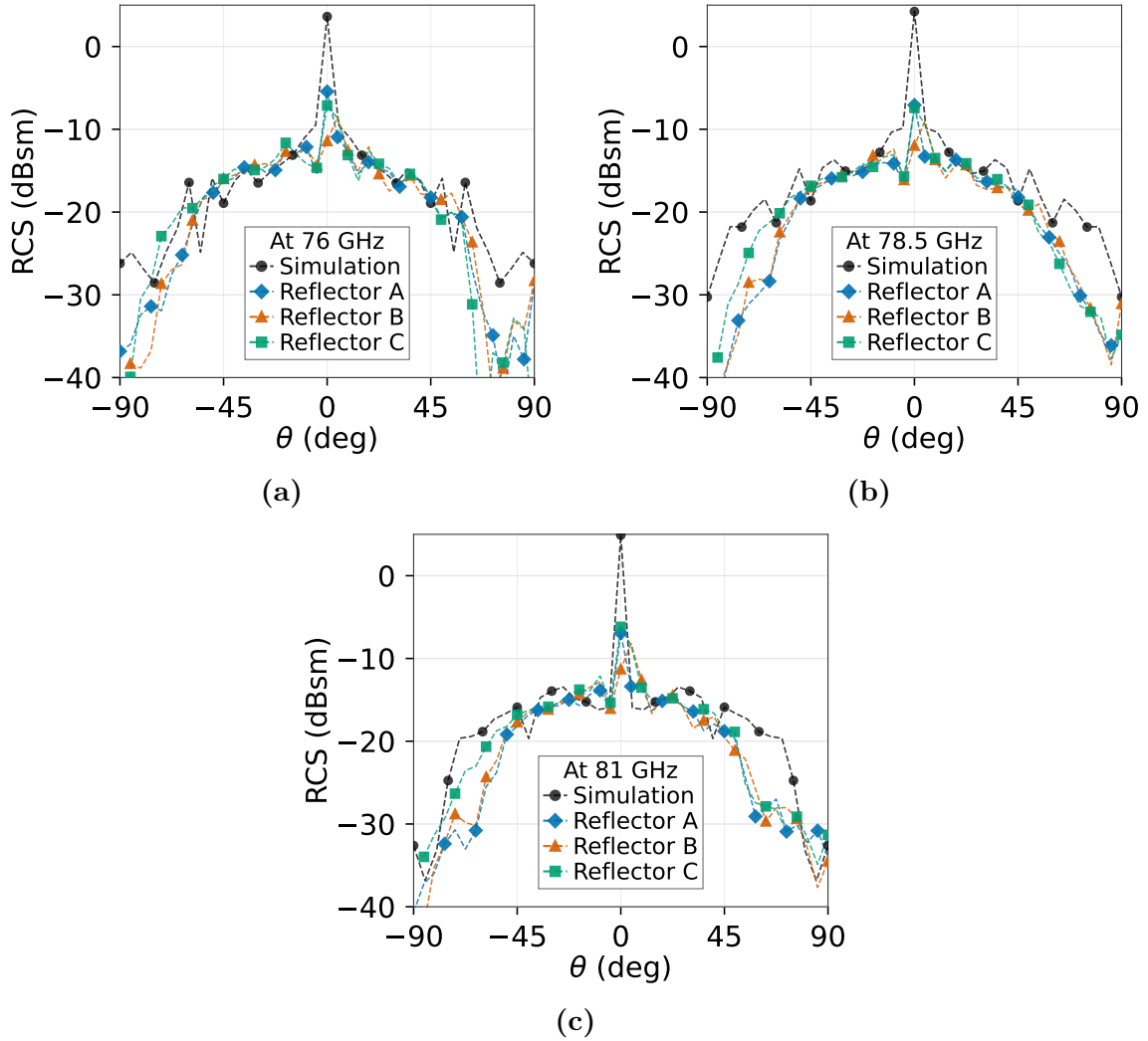


Figure 4.6: Comparison between measured and simulated monostatic RCS for the three manufactured reflectors. (a) 76 GHz. (b) 78.5 GHz. (c) 81 GHz.

4.4.2 Comparison to Simulation and Reference Reflectors

To compare the measured result with the simulated result in more detail, reflector A is compared with the simulated RCS at 78.5 GHz in Figure 4.7. It is also compared with an ideal simulated PEC plate of the same size and with the upper limit for

antenna mode scattering given by Equation 2.24. As expected, the structural mode scattering dominates at boresight, and the RCS therefore exceeds the maximum antenna mode value.

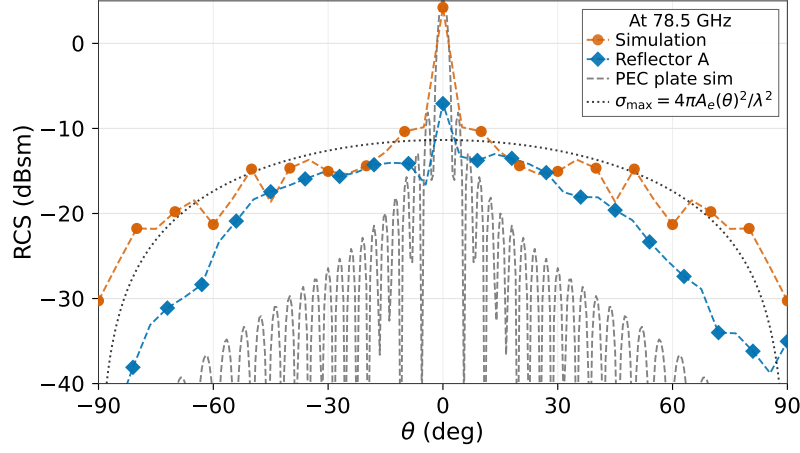


Figure 4.7: Measured monostatic RCS of reflector A compared with the simulated reflector RCS, the simulated PEC plate RCS, and the maximum antenna mode RCS based on the effective aperture of the full array at 78.5 GHz.

The measured reflector has an RCS approximately 10 dB lower than the simulated reflector at boresight. This is expected, since the manufactured reflector has several small holes and artifacts from laser welding on its surface. Because the measured angles have some uncertainty, the measurement point closest to boresight may not correspond to exact boresight. Since the RCS level falls off very quickly close to boresight, this angular uncertainty has a large effect in this region.

The ripple is also much less pronounced than in the simulation, and the measured level is more stable. This could indicate that the structural mode scattering is weaker for the manufactured reflector, which would reduce the interference with the antenna mode scattering. Another difference between the simulation and measurement is that the measured RCS falls off more quickly with angle. For angles larger than 60° , there is a large difference between measurement and simulation. This could be because the reflector was mounted on an absorber, which may have dampened reflections from the edges. It could also be caused by the reflector not being rotated exactly around its center. Another possible explanation is that the edge diffractions of the reflector are much weaker than in simulations. In general, the measurement result is more expected than the simulation result, since the retrodirective range is limited by the beamwidth of the antenna elements. The beamwidth of the designed slotted waveguide column corresponds more closely to a retrodirective range of approximately -45° to 45° than to the approximately -80° to 80° range shown in the simulation.

In the main retroreflection range from -45° to 45° , the simulated level agrees fairly well with the measurement. The ripple is not reproduced exactly, but the overall

level is similar. In this range, the reflector also gives a significant improvement of approximately 10 dB to 20 dB compared with the PEC plate.

To also compare the manufactured reflector with a reference reflector, the measured response is compared with ideal triangular trihedral corner reflectors modeled with Equation 2.40. This is not a completely direct comparison, since a trihedral corner reflector is a three-dimensional structure, while the Van Atta array is planar. This makes the choice of physical size for the comparison ambiguous.

The comparison therefore uses two trihedral corner reflector cases. The first has an effective area, defined using Equation 2.42, equal to the physical area of the complete designed retroreflector, $A_{\text{eff}} = 40 \text{ mm} \times 49 \text{ mm} = 1960 \text{ mm}^2$. The second has an effective area equal to the effective aperture of the Van Atta array, $A_{\text{eff}} = 292 \text{ mm}^2$. The comparison is shown in Figure 4.8 for reflector A at 78.5 GHz.

The larger corner reflector has a much higher RCS level inside approximately -45° to 45° , but its RCS drops off much more quickly with angle. The smaller corner reflector has a higher RCS close to boresight than the Van Atta reflector, but it also falls off more quickly with angle.

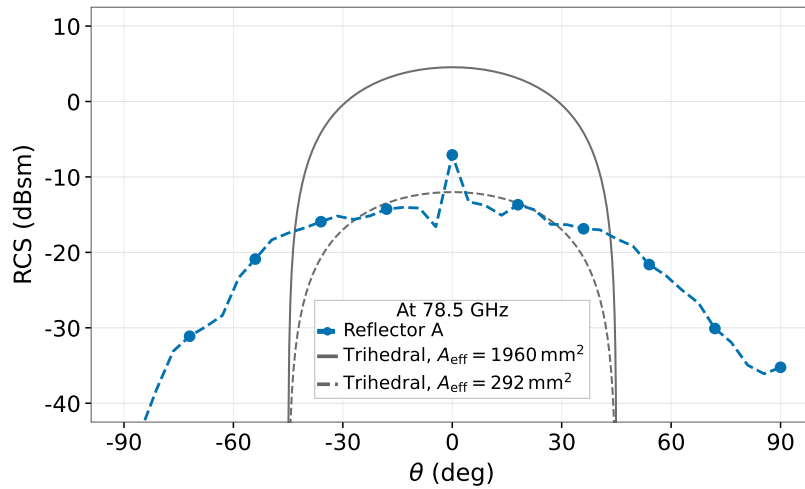


Figure 4.8: Measured monostatic RCS of reflector A compared with two simulated ideal trihedral corner reflectors with different effective apertures at 78.5 GHz.

4.4.3 Functional Angular Range and Mean RCS

To quantify the measured reflector performance, a metric is needed for the angular range over which the retroreflection remains effective. A possible choice would be the 6 dB beamwidth of the monostatic RCS pattern. However, because the boresight response is strongly dominated by structural mode scattering, a conventional beamwidth definition based on the global maximum is not meaningful. Instead, the functional angular range is defined here as the angular interval where the RCS remains within 6 dB of the maximum value outside the boresight region, here defined

as $|\theta| > 5^\circ$.

The resulting functional angular ranges at the lower edge, center frequency, and upper edge of the automotive radar band are shown in Table 4.1. For all three manufactured reflectors, the usable angular range exceeds 90° over the full 76 GHz to 81 GHz band. As a reference, the corner reflector in Figure 4.8 gives a functional angular range of approximately 50° . The slightly lower usable angular range at the band edges is expected, since the phase mismatch in the transmission networks increases away from the center frequency. This introduces both loss and beam squint, which becomes especially important far from boresight, where the phase gradient from the incident wave is larger across the antenna array.

Table 4.1: Measured functional angular range of the reflectors.

	76 GHz	78.5 GHz	81 GHz
Reflector A	97.7°	102.3°	97.5°
Reflector B	94.1°	93.1°	92.4°
Reflector C	95.6°	105.1°	92.9°
Average	95.8°	100.2°	94.3°

To evaluate the complete measured band, and not only the center and edge frequencies, the same functional angular range was extracted over frequency. The outer 2.5 GHz at each end of the measured sweep was excluded to avoid artifacts from the time-gating procedure. The resulting functional angular range is shown in Figure 4.9. All three reflectors maintain a functional angular range above 90° over the full 76 GHz to 81 GHz band. The figure also shows that, if a usable angular range above 90° is used as the limiting criterion, the angular bandwidth is larger than the designed 5 GHz band for several of the reflectors. The usable angular range is also significantly larger than for the ideal trihedral corner reflector modeled using Equation 2.40. Outside the designed frequency band of the Van Atta retroreflector, the functional angular range becomes less stable and shows sharper variations as individual angular points move above or below the 6 dB threshold.

To also evaluate the RCS level over the band, the RCS was averaged linearly over the fixed angular range -45° to 45° , excluding the boresight region. Figure 4.9 also includes this average RCS result. Within the 76 GHz–81 GHz frequency band, the average level remains relatively stable, with only small variations across the band. Outside the design band, the level drops more clearly, which indicates that the average RCS level, rather than the functional angular range, is the limiting factor for the effective bandwidth of the reflector.

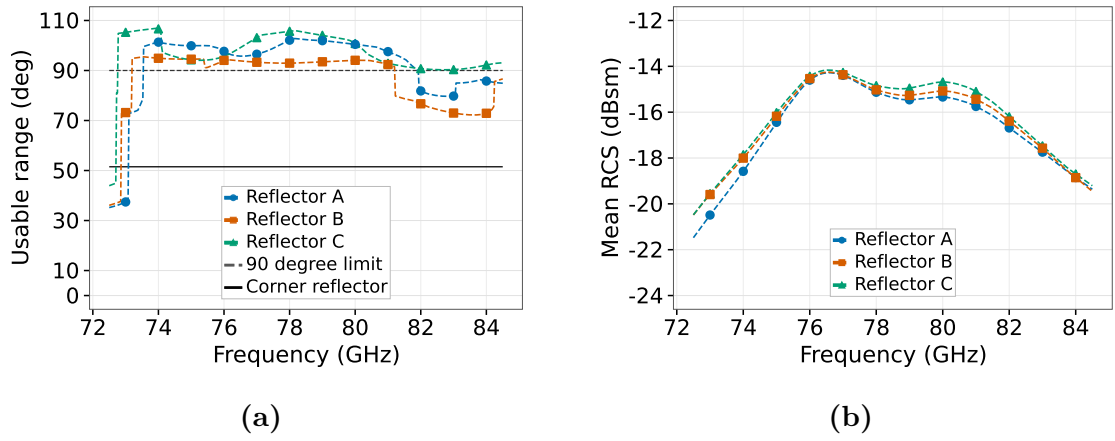


Figure 4.9: Measured reflector performance as a function of frequency. (a) Functional angular range over frequency, using the 6 dB criterion defined above, compared with an ideal trihedral corner reflector. (b) Average RCS over the fixed angular range -45° to 45° , excluding the boresight region.

4.4.4 Full Monostatic RCS Frequency Dependence

To show the monostatic RCS over the measured frequency range, the results are shown as heatmaps in Figure 4.10. Outside the designed band, the RCS is less stable and deeper nulls appear. Far outside the designed band, the RCS also drops more quickly with angle. The PEC plate reference in Figure 4.10d shows much deeper nulls, and outside boresight its RCS is much weaker than that of the reflectors.

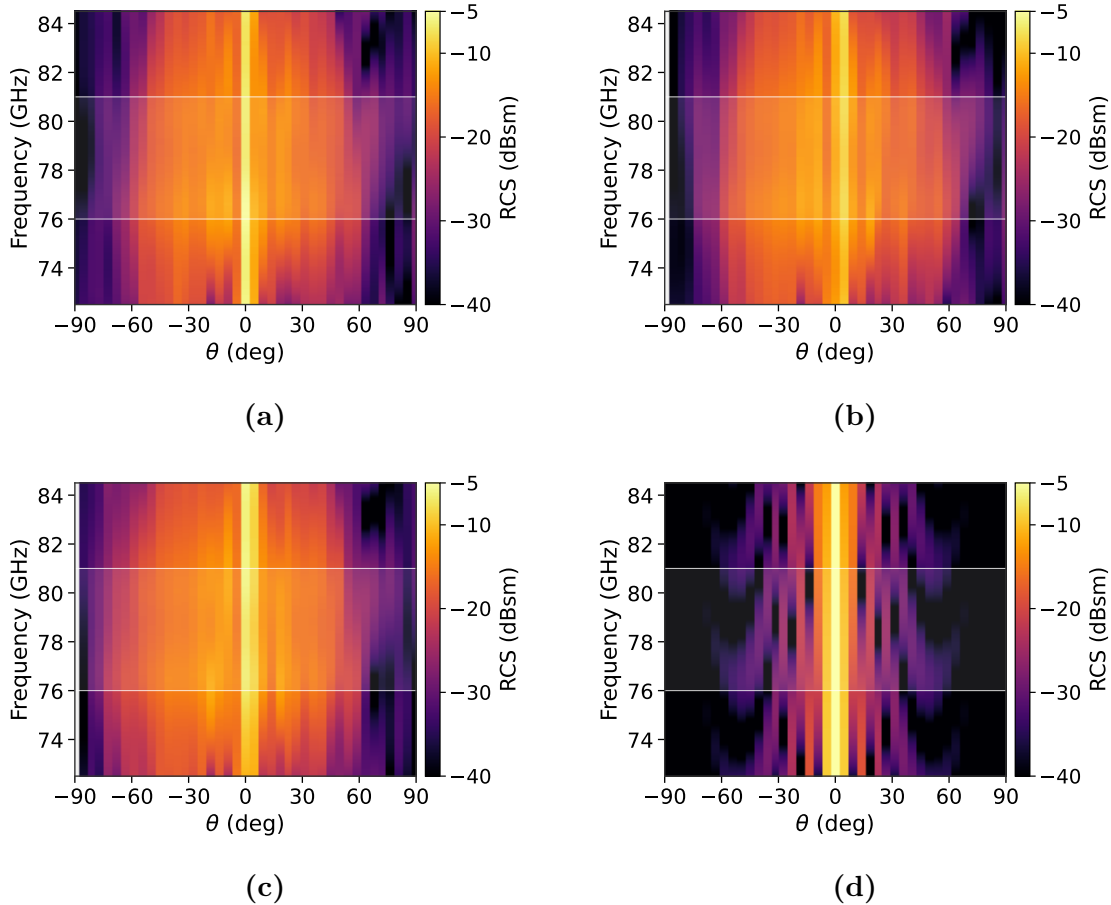


Figure 4.10: Monostatic RCS over angle and frequency. (a) Measured reflector A. (b) Measured reflector B. (c) Measured reflector C. (d) Simulated PEC plate reference.

4.5 Validation of Numerical Van Atta Model

The main advantage of the implemented numerical model is the reduced computation time, but another useful feature is that the antenna mode and structural mode contributions can be evaluated separately. However, when the two contributions are combined, the phase relationship between them is difficult to determine. The numerical model can be used to explore how this phase relationship affects the total RCS. To illustrate this, the bistatic pattern in Figure 4.11 is shown for an incident wave at 30° and a single frequency. The result shows that the main retroreflective

lobe level is strongly affected by the phase. For a phase difference of 0° , the two contributions add destructively, whereas for a phase difference of 180° , they add constructively.

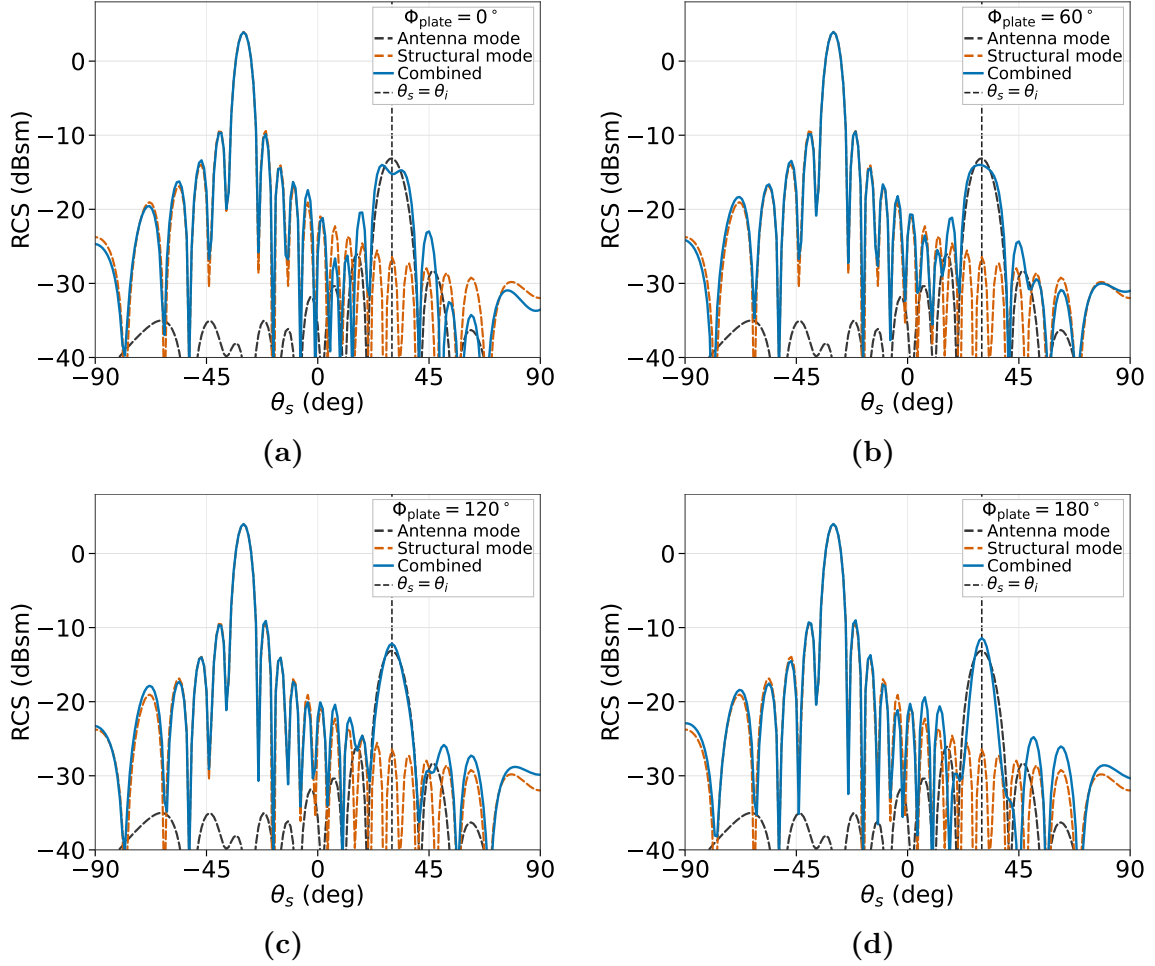


Figure 4.11: Bistatic RCS pattern calculated using the numerical model for an incident wave from 30° and different relative phases between structural mode and antenna mode scattering. (a) $\Phi_{\text{plate}} = 0^\circ$. (b) $\Phi_{\text{plate}} = 60^\circ$. (c) $\Phi_{\text{plate}} = 120^\circ$. (d) $\Phi_{\text{plate}} = 180^\circ$.

The Python model is mainly validated by comparing it with the CST full-wave results. For the case $\theta_i = 30^\circ$, the comparison is given in Figure 4.12. Overall, the bistatic patterns show good agreement. As noted above, however, the relative phase between antenna mode and structural mode scattering is not known. As a result, the numerical model does not capture the interference correctly at every frequency. In this example, the chosen phase offset gives the correct level at 81 GHz, but does not capture the exact effect at 76 GHz and 78.5 GHz.

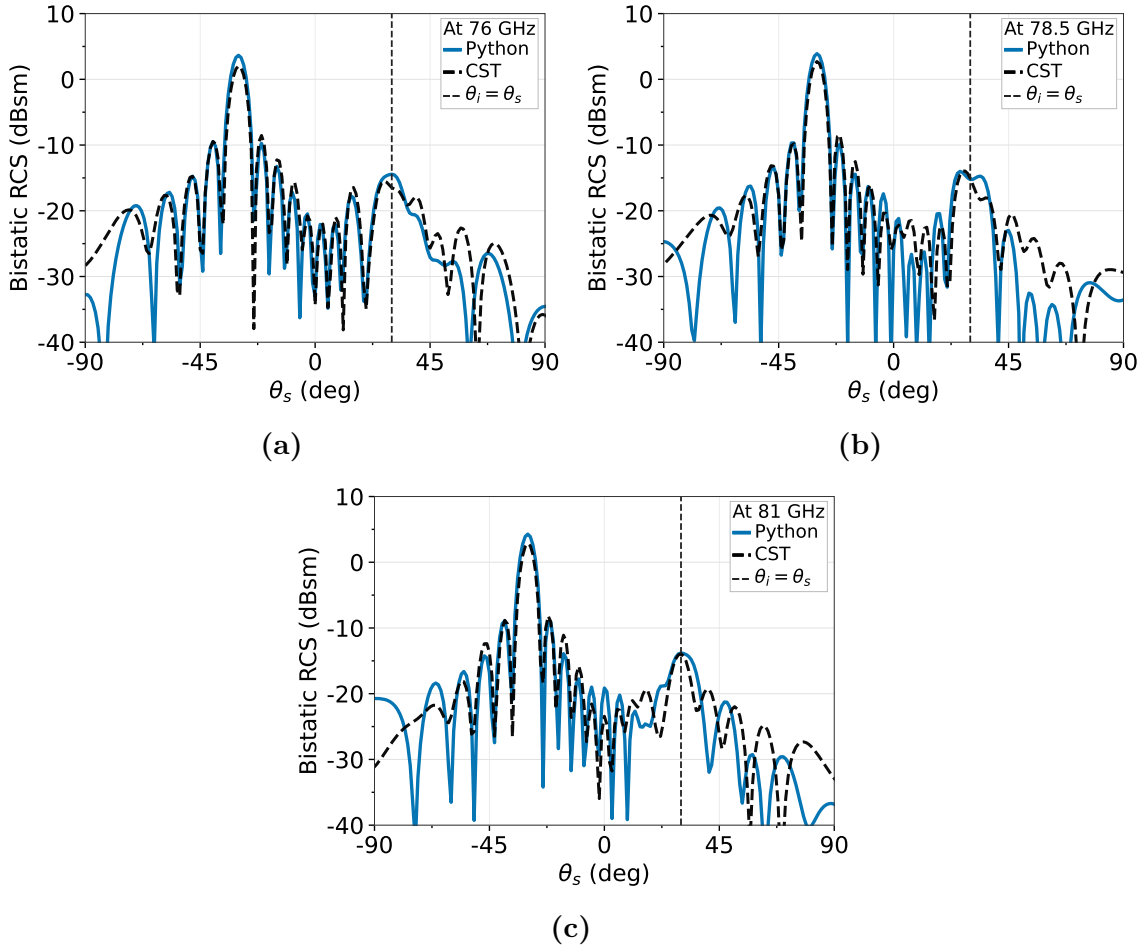


Figure 4.12: Bistatic RCS comparison between the numerical model and CST simulation for the reflector at $\theta_i = 30^\circ$ incidence. (a) 76 GHz. (b) 78.5 GHz. (c) 81 GHz.

Since the numerical model can simulate the antenna mode scattering independently, the measured response of reflector A can be compared directly with the modeled antenna mode, as shown in Figure 4.13. Outside boresight, the measured reflector is very close to the antenna mode scattering predicted by the numerical model. At 78.5 GHz and 81 GHz, the measured RCS is a few dB lower in the usable angular range, but the overall shape is very similar. This further supports the interpretation that the measured response outside boresight is mainly dominated by antenna mode scattering, with a weaker structural mode contribution than in the full-wave simulation.

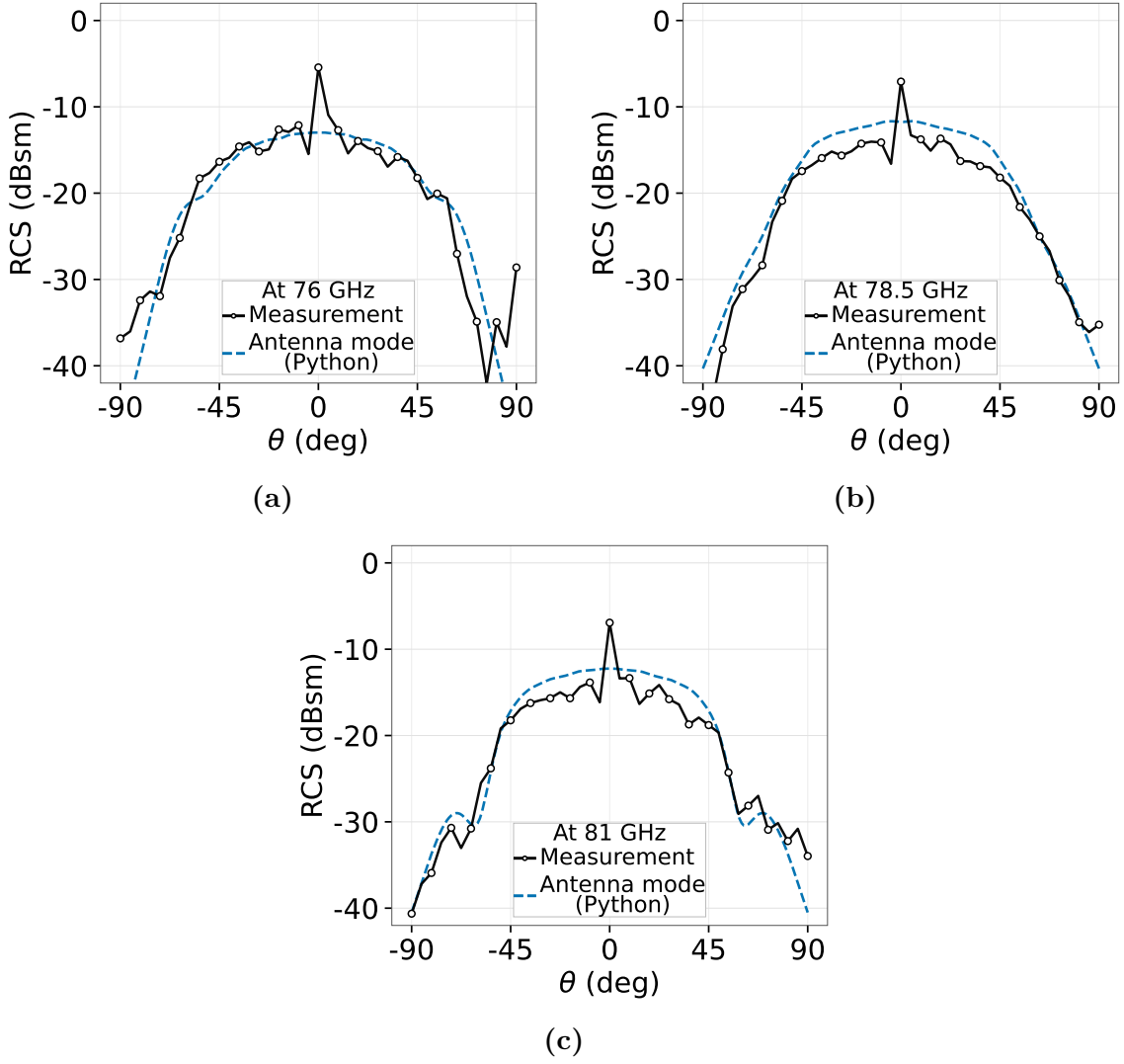


Figure 4.13: Measured monostatic RCS of reflector A compared with the antenna mode RCS predicted by the numerical model. (a) 76 GHz. (b) 78.5 GHz. (c) 81 GHz.

To compare the computational time required by the numerical model and full-wave simulation, a monostatic angular sweep from -90° to 90° with 5° steps was considered. A full-wave simulation of the complete reflector requires one simulation for each angle, giving 37 simulation points in total. This was compared with the numerical model workflow, where the slot array is simulated once with 12 waveguide ports, the two transmission networks are simulated separately with six ports each, and the angular sweep is then calculated using the Python implementation.

All simulation times were obtained on the same workstation, which used two Intel Xeon Silver 4215R CPUs at 3.20 GHz and 512 GB of RAM. The reported times should therefore be interpreted as a relative comparison between the two workflows. In general, the speedup from using the numerical model is expected to become even larger on slower hardware, since the full-wave scattering simulations are the dominant part of the computation time.

Table 4.2 shows the time for the two methods. Even if all three subcomponents are re-simulated, using the numerical model reduces the computation time by more than 120 h. In addition, the Python implementation can be evaluated using a much finer angular spacing without any meaningful increase in computation time. In the numerical model, the angular resolution is only limited by the angular sampling of the simulated slot array, which in this case was 1° .

Table 4.2: Comparison of computation time between an angular sweep using full-wave scattering simulations and the numerical model.

Simulation step	Computation time
Full-wave scattering simulation, 37 angles	124 h 19 min
Slot array, 12 waveguide ports	2 h 30 min
Transmission network 1, six ports	52 min
Transmission network 2, six ports	35 min
Python implementation, 37 angles	< 1 min
Total numerical model workflow	3 h 58 min
Time reduction	> 120 h

Overall, the numerical model provides a useful complement to full-wave simulations. It greatly reduces the computation time and gives good agreement with full-wave simulations for the main antenna mode response. The comparison with measurement also shows that the antenna mode predicted by the model gives a good estimate of the measured total response outside boresight. This makes the model especially useful for quickly evaluating the effect of design changes and for estimating the functional angular range of the reflector. In addition, it allows antenna mode and structural mode scattering to be studied separately.

4.6 Discussion and Comparison with Previous Work

While many Van Atta arrays have been reported at X-band and below, considerably fewer implementations have been reported at mmWave frequencies. Some of the reported passive mmWave Van Atta arrays are summarized in Table 4.3 and compared with this work. A direct comparison is difficult, since different works report their results using different metrics. For example, the angular response is not defined in the same way in all references. In addition, few works report bandwidth at these frequencies. When the result is reported only at a single frequency, the bandwidth is listed as single-frequency in the table. The comparison should also be interpreted with some caution because not all designs have the same purpose. Some are mainly intended for RCS enhancement, while others also prioritize properties such as bistatic selectivity, polarization rotation, or circular polarization.

4. Results and Validation

Table 4.3: Comparison of reported passive mmWave Van Atta arrays

Category	[41]	[10]	[42]	[43]	[44]	This work
Center frequency	78.5 GHz	30 GHz	25.15 GHz	79 GHz	60 GHz	78.5 GHz
Bandwidth	Single-frequency	Single-frequency	24.9–25.5 GHz (2.4%)	Single-frequency	Single-frequency	76–81 GHz (6.4%)
Retrodirective scanning ability	1-D	1-D	1-D	1-D	1-D	1-D
Measured monostatic retrodirective range	80° stable response	120° 7 dB above same-size plate	60° –3 dB	34° –3 dB	144° –10 dB	94° functional angular range
Maximum RCS level	–23.5 dBsm reported maximum	N.C	–18 dBsm estimated maximum	–31 dBsm estimated maximum	–26 dBsm reported maximum	–11 dBsm outside boresight
Antenna type	Yagi-Uda	SIW slot	SIW diamond slot	SIW slot	Circular patch	MLW slot
Antenna elements	8	8	32	16	16	12
Physical size, $L \times W \times H$ in λ_c	$4.85 \times 3.39 \times 0.027$	$5.20 \times 5.38 \times \text{N.R.}$	$12.1 \times 11.2 \times 0.09$	$7.90 \times 15.80 \times \text{N.R.}$	$8.80 \times 3.00 \times \text{N.R.}$	$12.83 \times 10.47 \times 0.25$
Polarization	Single linear	Single linear	Circular	Polarization rotation	Polarization rotation	Single linear
Transmission network	Microstrip line	SIW transmission line	SIW transmission line	SIW transmission line	LTCC microstrip line	MLW coaxial line
Equal physical transmission line lengths	No	Yes	No	No	No	No

N.R.: not reported. N.C.: not comparable. Estimated values were extracted from published plots. Outside boresight values are used to exclude the structural mode dominated response at boresight when applicable. The reported RCS level in [10] is not included in the comparison because it appears inconsistent with the physical size of the reported Van Atta array.

From the comparison in Table 4.3, it can be seen that most reported mmWave Van Atta arrays, including this work, use transmission lines with different physical lengths and are only retrodirective in one dimension. This contrasts with reported Van Atta arrays at lower frequencies, such as [8], where the implementation is often based on equal-length transmission lines and two-dimensional retrodirectivity. The difference is likely a practical layout constraint. Making all transmission lines the same physical length puts strong demands on the layout, and at mmWave frequencies the design becomes especially sensitive to additional loss, bends, coupling, and manufacturing tolerances. The same challenge becomes even more severe when the Van Atta array is also designed for retrodirectivity in two dimensions.

The comparison also shows that most passive mmWave implementations are either very narrowband or only reported at a single frequency. This is likely connected to the use of transmission lines with large physical length differences and dispersive behavior, which causes the phase error between connected element pairs to increase away from the design frequency. In contrast, this work is experimentally validated over the full 76 GHz–81 GHz frequency band. One explanation for this higher bandwidth is that the MLW coaxial line used in this work has very low dispersion and a phase velocity close to that of free-space propagation.

It is also clear from the comparison that this work achieves a higher absolute RCS

level than the other reported implementations, even when the boresight-dominated response is excluded. However, this comparison must be interpreted carefully, since many of the reported arrays are smaller in physical size. In addition, absolute RCS is strongly affected by structural mode scattering and by how the RCS level is extracted from the reported results.

The comparison also indicates one of the main limitations of dielectric-based transmission networks in the 77 GHz automotive radar band. For example, [41] reports that adding a fourth element pair gives only a small additional RCS increase, which is attributed to loss in the transmission lines. This suggests that the solution proposed in [41] would be difficult to scale to a much larger array.

Overall, the comparison shows that the proposed Van Atta array achieves a relatively large bandwidth and a good functional angular range compared with the reported passive mmWave implementations. The comparison also highlights the advantage of using an MLW coaxial line with low loss and low dispersion. While Table 4.3 should not be interpreted as an absolute ranking, it supports that MLW is a promising technology for realizing passive Van Atta arrays in the 77 GHz automotive radar band.

5

Conclusion and Future Work

5.1 Conclusion

This thesis has investigated the design, modeling, measurement, and experimental validation of a planar passive Van Atta retroreflector in multilayer waveguide (MLW) technology. The purpose of the reflector is to increase the radar cross section (RCS) of vulnerable road users in the 77 GHz automotive radar band.

This was done by designing a prototype Van Atta array with retrodirectivity in azimuth for horizontal polarization. The prototype was also modeled numerically to enable faster iterative design and to separate the antenna mode and structural mode scattering contributions. In addition, a method for measuring and validating the retroreflector using a single-port monostatic RCS measurement in an anechoic chamber was presented.

The proposed numerical model reduces the computation time by more than 120 h compared with a full angular sweep in CST. In addition, the modeled antenna mode contribution agrees well with the measured total RCS outside boresight. This makes the model useful both for fast design iteration and for interpreting the measured results by separately studying the antenna mode and structural mode scattering.

All three measured reflectors show a functional angular range of more than 90° over the full 76 GHz–81 GHz frequency band. In this range, they provide an improvement of approximately 10 dB to 20 dB compared with a metal plate of the same physical size. The simulations and measurements agree well in the functional angular range, although the full-wave simulations overestimate the RCS outside this range, likely due to differences in structural mode scattering. The achieved bandwidth of 5 GHz indicates that the low dispersion of MLW allows relatively large physical length differences in the transmission lines before the effective bandwidth of the Van Atta array is significantly degraded. Together with the high radiation efficiency, low transmission network loss, and compact form factor with a thickness below 1 mm, these results validate that MLW is a suitable technology for building a wide-angle Van Atta retroreflector in the 76 GHz–81 GHz frequency band.

The achieved absolute RCS of approximately -17 dBsm to -11 dBsm shows the level obtained for this first prototype design. Further improvement is most likely needed before the concept can meaningfully increase the detectability of vulnerable

road users in a practical real-world scenario. Some of the remaining challenges are discussed in the next section.

5.2 Future Work

One important direction for future work is to investigate how the reflector can be made retrodirective in both azimuth and elevation. This is especially important for pedestrians, since they may bend or move in ways that make it difficult to keep the reflector perfectly vertical. This is less critical for roadside structures or bicycles, where the reflector orientation can be more controlled. One possible approach is to use an MLW design with more layers, for example six layers, so that one layer can be used exclusively for the transmission lines between elements. This would also allow the structure to be even more compact.

Although most automotive radars today mainly use horizontal polarization, it would also be interesting to investigate whether the design can be extended to operate for vertical polarization as well. This could be especially relevant for future 6G applications, where both polarizations may be used.

Another approach that could be investigated is the integration of nonlinear components into the design. In addition to increasing the RCS, such modulation could create a recognizable Doppler-like or frequency-shifted signature for the Van Atta array, which could significantly improve its detectability for radars. These components could be passive devices, such as diodes, or, in a more advanced implementation, active components operating at very low power.

Bibliography

- [1] European Parliament and the Council of the European Union, “Regulation (eu) 2019/2144 on type-approval requirements for motor vehicles and their trailers, and systems, components and separate technical units intended for such vehicles, as regards their general safety and the protection of vehicle occupants and vulnerable road users,” 2019, official Journal of the European Union, L 325, 16 December 2019, pp. 1–40.
- [2] S. Tebaldini, M. Manzoni, D. Tagliaferri, M. Rizzi, A. V. Monti-Guarnieri, C. M. Prati, U. Spagnolini, M. Nicoli, I. Russo, and C. Mazzucco, “Sensing the Urban Environment by Automotive SAR Imaging: Potentials and Challenges,” *Remote Sensing*, vol. 14, no. 15, p. 3602, Jan. 2022.
- [3] R. Liu, M. Hua, K. Guan, X. Wang, L. Zhang, T. Mao, D. Zhang, Q. Wu, and A. Jamalipour, “6G Enabled Advanced Transportation Systems,” *IEEE Transactions on Intelligent Transportation Systems*, vol. 25, no. 9, pp. 10 564–10 580, Sep. 2024.
- [4] V. Fusco and N. Buchanan, “Developments in retrodirective array technology,” *IET Microwaves, Antennas & Propagation*, vol. 7, no. 2, pp. 131–140, 2013, _eprint: <https://ietresearch.onlinelibrary.wiley.com/doi/pdf/10.1049/iet-map.2012.0565>.
- [5] A. Nielsen and B. Kahler, “Radar Reflector Design For Improved Detection Of Cyclists,” in *NAECON 2025 - IEEE National Aerospace and Electronics Conference*, Jul. 2025, pp. 1–4, iSSN: 2379-2027.
- [6] C. Charlo, S. Méric, and R. Gillard, “RCS-enhancement for improving the detectability of bikes in road safety applications,” in *2020 17th European Radar Conference (EuRAD)*, Jan. 2021, pp. 222–225.
- [7] J.-H. Wang, J. Wang, L. Zhou, Z. Zhan, Q.-M. Cai, X. Cao, and Y. Zhu, “Research on Road Safety-Based Wide-Angle RCS Enhancement Flexible Surface,” in *2025 Photonics & Electromagnetics Research Symposium - Spring (PIERS-Spring)*, May 2025, pp. 1–6, iSSN: 2831-5804.
- [8] X.-F. Li, Y.-L. Ban, Q. Sun, Y.-X. Che, J. Hu, and Z. Nie, “A Planar 2-D 8×8 Van Atta Retrodirective Array for RCS Enhancement in a Wide Angular Range,” *IEEE Transactions on Antennas and Propagation*, vol. 70, no. 12, pp. 12 364–12 369, Dec. 2022.
- [9] M. Qu, F. Liu, Z. Li, J. Su, Y. Liu, and L. Y. Yang, “Design of Low-Profile Polarization-Reconfigurable Van Atta Retrodirective Array,” *IEEE Microwave and Wireless Technology Letters*, vol. 35, no. 12, pp. 1961–1964, Dec. 2025.

- [10] A. A. M. Ali, H. B. El-Shaarawy, and H. Aubert, "Millimeter-Wave Substrate Integrated Waveguide Passive Van Atta Reflector Array," *IEEE Transactions on Antennas and Propagation*, vol. 61, no. 3, pp. 1465–1470, Mar. 2013.
- [11] Y. Chen, T. Wang, Q. Liu, H. Wang, and C. Jin, "Orientation Controllable RCS Enhancement Electromagnetic Surface to Improve the Road Barriers Detectability for Autonomous Driving Radar," *Sensors*, vol. 25, no. 13, Jun. 2025.
- [12] D. He, S. Li, L. Chen, L. Deng, and Y. Shu, "A Wide-Angle and Ultrawideband Van Atta Array for Monostatic RCS Enhancement," *IEEE Antennas and Wireless Propagation Letters*, vol. 24, no. 1, pp. 68–72, Jan. 2025.
- [13] P. Hallbjörner and S. Cheng, "Improvement in 77-GHz Radar Cross Section of Road Work Jacket and Side Screen by Use of Planar Flexible Retrodirective Reflectors," *IEEE Antennas and Wireless Propagation Letters*, vol. 12, pp. 1085–1088, 2013.
- [14] smartmicro, "UMRR-96 Type 153 Automotive Radar," Datasheet, 2020, accessed: 2026-05-05. [Online]. Available: https://www.smartmicro.com/fileadmin/media/Downloads/Automotive_Radar/Sensor_Data_Sheets_76-81GHz/UMRR_Automotive_Type_153_Data_Sheet.pdf
- [15] Aptiv, "Gen 7 Radar Family," Product page, 2025, accessed: 2026-05-05. [Online]. Available: <https://www.aptiv.com/en/solutions/intelligent-perception/radars/gen-7-radar-family>
- [16] Magna International, "IAA 2022 Media Kit: ICON Radar," Media kit, 2022, accessed: 2026-05-05. [Online]. Available: https://www.magna.com/docs/default-source/2022-iaa/media-kit_destinction-experience_magna_iaa22.pdf
- [17] S. B. J. Gowdu, A. Schwind, R. Stephan, and M. A. Hein, "Monostatic RCS Measurements of Representative Road Traffic Objects in the 76 ... 81 GHz Frequency Band," in *2020 IEEE Radar Conference (RadarConf20)*, Sep. 2020, pp. 1–6, iSSN: 2375-5318.
- [18] European Commission. Joint Research Centre. Institute for the Protection and the Security of the Citizen., *Radar cross section measurements of pedestrian dummies and humans in the 24/77 GHz frequency bands: establishment of a reference library of RCS signatures of pedestrian dummies in the automotive radar bands*. LU: Publications Office, 2013.
- [19] "IEEE Std 1502-2020 (Revision of IEEE Std 1502-2007) IEEE Recommended Practice for Radar Cross-Section Test Procedures," 2020.
- [20] P.-S. Kildal, *Foundations of antenna engineering: a unified approach for line-of-sight and multipath*. Pixbo: Kildal Antenn, 2015.
- [21] D. M. Pozar and D. M. Pozar, *Microwave Engineering*. Chichester, UNITED STATES: Wiley, 2012.
- [22] D. Kim, J. Hirokawa, M. Ando, J. Takeuchi, and A. Hirata, "64-Element and 32-Element Slot Array Antennas Using Double-Layer Hollow-Waveguide Corporate-Feed in the 120 GHz Band," *IEEE Transactions on Antennas and Propagation*, vol. 62, no. 3, pp. 1507–1512, Mar. 2014.
- [23] P.-S. Kildal, "Artificially soft and hard surfaces in electromagnetics," *IEEE Transactions on Antennas and Propagation*, vol. 38, no. 10, pp. 1537–1544, Oct. 1990.

-
- [24] E. Lier, “Analysis of soft and hard strip-loaded horns using a circular cylindrical model,” *IEEE Transactions on Antennas and Propagation*, vol. 38, no. 6, pp. 783–793, Jun. 1990.
 - [25] D. Sievenpiper, L. Zhang, R. Broas, N. Alexopolous, and E. Yablonovitch, “High-impedance electromagnetic surfaces with a forbidden frequency band,” *IEEE Transactions on Microwave Theory and Techniques*, vol. 47, no. 11, pp. 2059–2074, Nov. 1999.
 - [26] M. G. Silveirinha, C. A. Fernandes, and J. R. Costa, “Electromagnetic Characterization of Textured Surfaces Formed by Metallic Pins,” *IEEE Transactions on Antennas and Propagation*, vol. 56, no. 2, pp. 405–415, Feb. 2008.
 - [27] P.-S. Kildal, A. Zaman, E. Rajo-Iglesias, E. Alfonso, and A. Valero-Nogueira, “Design and experimental verification of ridge gap waveguide in bed of nails for parallel-plate mode suppression,” *IET Microwaves, Antennas & Propagation*, vol. 5, no. 3, pp. 262–270, Feb. 2011.
 - [28] A. Vosoogh, H. Zirath, and Z. S. He, “Novel Air-Filled Waveguide Transmission Line Based on Multilayer Thin Metal Plates,” *IEEE Transactions on Terahertz Science and Technology*, vol. 9, no. 3, pp. 282–290, May 2019.
 - [29] A. Vosoogh, A. Haddadi, and C. Bencivenni, “Novel Low-loss Coaxial Slot Array Based on Gap Waveguide Technology for E-band Automotive Radar Applications,” in *2023 17th European Conference on Antennas and Propagation (EuCAP)*, Mar. 2023, pp. 1–5.
 - [30] J.-L. A. Lijarcio, A. Vosoogh, C. Bencivenni, and A. U. Zaman, “Low-Cost Center-Fed Slot Array Based on Gap Waveguide MLW Coaxial Line Technology for E-Band Automotive Radar,” *IEEE Transactions on Antennas and Propagation*, vol. 72, no. 7, pp. 5674–5681, Jul. 2024.
 - [31] A. Vosoogh, P.-S. Kildal, and V. Vassilev, “A multi-layer gap waveguide array antenna suitable for manufactured by die-sink EDM,” in *2016 10th European Conference on Antennas and Propagation (EuCAP)*, Apr. 2016, pp. 1–4.
 - [32] A. Tamayo-Domínguez, J.-M. Fernández-González, and M. Sierra-Pérez, “Groove Gap Waveguide in 3-D Printed Technology for Low Loss, Weight, and Cost Distribution Networks,” *IEEE Transactions on Microwave Theory and Techniques*, vol. 65, no. 11, pp. 4138–4147, Nov. 2017.
 - [33] S. Farjana, M. Ghaderi, A. U. Zaman, S. Rahiminejad, T. Eriksson, J. Hansson, Y. Li, T. Emanuelsson, S. Haasl, P. Lundgren, and P. Enoksson, “Realizing a 140 GHz Gap Waveguide-Based Array Antenna by Low-Cost Injection Molding and Micromachining,” *Journal of Infrared, Millimeter, and Terahertz Waves*, vol. 42, no. 8, pp. 893–914, Aug. 2021.
 - [34] S. Farjana, M. Ghaderi, A. U. Zaman, S. Rahiminejad, P. Lundgren, and P. Enoksson, “Low-Loss Gap Waveguide Transmission Line and Transitions at 220–320 GHz Using Dry Film Micromachining,” *IEEE Transactions on Components, Packaging and Manufacturing Technology*, vol. 11, no. 11, pp. 2012–2021, Nov. 2021.
 - [35] E. F. Knott, J. F. Shaeffer, and M. T. Tuley, *Radar Cross Section*. Institution of Engineering and Technology, Jan. 2004.
 - [36] Y. Liu, Y. Jia, and S. Gong, *Antenna Radar Cross Section: Theory and Design*. Singapore: Springer Nature, 2025.

- [37] L. C. V. Atta, “Electromagnetic reflector,” US Patent US2908002A, Oct. 6, 1959, u.S. Patent.
- [38] S. Drabowitch, A. Papiernik, H. D. Griffiths, J. Encinas, and B. L. Smith, *Modern Antennas*. Boston, MA: Springer US, 2005.
- [39] C. G. Moschovitis, H. Anastassiou, and P. V. Frangos, “Scattering of Electromagnetic Waves from a Rectangular Plate Using an Extended Stationary Phase Method Based on Fresnel Functions (SPM-F),” *Progress In Electromagnetics Research*, vol. 107, pp. 63–99, 2010.
- [40] A. W. Doerry, “Reflectors for SAR Performance Testing – second edition,” Sandia National Laboratories (SNL-NM), Albuquerque, NM (United States), Tech. Rep. SAND–2014-0882, Feb. 2014.
- [41] S. Ghasemi, L. Guo, S. Harisha, and A. Eid, “mmWave Retroreflective Road Markers for Automotive Radar Vision,” in *2024 IEEE International Conference on RFID (RFID)*, Jun. 2024, pp. 89–94, iSSN: 2573-7635.
- [42] Z.-W. Miao, Z.-C. Hao, and Q. Yuan, “A Passive Circularly Polarized Van Atta Reflector For Vehicle Radar Applications,” *IEEE Antennas and Wireless Propagation Letters*, vol. 16, pp. 2254–2257, 2017.
- [43] H. Zhou, W. Hong, L. Tian, X. Jiang, X.-c. Zhu, M. Jiang, L. Cheng, and J.-x. Zhuang, “A Retrodirective Antenna Array With Polarization Rotation Property,” *IEEE Transactions on Antennas and Propagation*, vol. 62, no. 8, pp. 4081–4088, Aug. 2014.
- [44] K. Trzebiatowski, M. Ihle, B. Sikorski, L. Kulas, and K. Nyka, “Millimeter Wave Retrodirective Van Atta Arrays in LTCC Technology,” in *2024 18th European Conference on Antennas and Propagation (EuCAP)*, Mar. 2024, pp. 1–5.

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