Advances in Beam-Steerable and Low-Scattering Antennas for Communications and Sensing

Mikko K. Leino
Advances in Beam-Steerable and Low-Scattering Antennas for Communications and Sensing

Mikko K. Leino

A doctoral dissertation completed for the degree of Doctor of Science (Technology) to be defended, with the permission of the Aalto University School of Electrical Engineering, at a public examination held via a remote connection on 23 April 2021 at 11.

Aalto University
School of Electrical Engineering
Department of Electronics and Nanoengineering
Abstract

Communication networks are being developed to handle the increased wireless data traffic that the existing systems cannot handle. Fifth-generation mobile networks (5G) are adopting millimeter-wave (mm-wave) frequencies for higher spectral efficiency and wider spectral bands in order to increase network capacity. Additionally, the number of network antennas will increase exponentially, since the propagation at mm-waves suffers from intrinsic atmospheric attenuation. Furthermore, the mm-wave antennas are required to be highly efficient, and beam-steering capabilities are also necessary to focus capacity where it is needed.

The first part of this thesis discusses phased array designs for the mm-wave base-station applications and tools for the analysis and optimization of the antennas. The presented antennas operate in Kₘ and E-bands, and they combine low-loss, waveguide-based power division networks and antenna elements with phase shifters that are integrated on a printed circuit board (PCB). The resulting proposals are antennas with high efficiency, where the majority of the losses are caused by the used phase shifters. The performances of both antennas (e.g., their beam-steering capabilities) have been validated with measurements. Furthermore, the antenna diagnostic results based on the holography data are presented, along with optimization methods that allow performance enhancement in terms of higher antenna gain and lower side lobes.

Because antenna development can be a time-consuming and costly process, utilizing the same antenna in multiple different scenarios is desirable. The second part of the thesis explains how the previously presented Kₘ-band antennas, which were initially developed for communications, can be used in imaging applications. A frequency-diverse imaging method is explained, in which a computational algorithm is used to reconstruct the image from the observation data. A theoretical evaluation and experimental test results are presented. The proposed method has been used successfully to reconstruct an image of the scene locating a conducting sphere, and future research paths are discussed.

Modern radar applications may require co-locating multiple antennas together, especially if the area or the volume reserved for the antennas is limited. Therefore, electrically invisible or transparent antennas that do not affect the performance of the co-existing antennas are required. The third and final part of the thesis focuses on this topic and describes the design steps required to realize a low-scattering antenna, i.e., an inductively loaded, chopped dipole that is transparent at a higher frequency than where it operates. It is experimentally demonstrated how the radar cross section of the designed antenna is reduced at the higher frequency 15 dB, while the radiation efficiency of the dipole decreases 0.4 dB at its operation frequency due to the inductive loading.

Keywords beam steering, phased arrays, phase shifters, sensing, imaging, low scattering, loaded antennas

ISSN (printed) 1799-4934 ISSN (pdf) 1799-4942
Location of publisher Helsinki Location of printing Helsinki Year 2021
Pages 145
Preface

In his last message, Lord Baden-Powell famously urged us scouts to leave this world a little better behind than how we found it. The message of this quote has been one of my inspirations to do research. I sincerely believe that by expanding our understanding of the surrounding world, the world will eventually become a better place for us all. The aim of the doctoral dissertation is to create new knowledge and with this thesis finished, I hope that I have contributed even a bit towards what B-P asked us to do.

The research presented in this thesis was carried out in the Department of Electronics and Nanoengineering at Aalto University School of Electrical Engineering during 2017-2021. First, I would like to thank my supervisor Prof. Ville Viikari for offering me the opportunity to work in the world of antennas and for guiding me to become a better researcher. Second, my biggest thanks to my advisor Dr. Juha Ala-Laurinaho who always found time to calmly listen my rambling questions and whose deep knowledge and support helped me to overcome many problems along the way.

I am grateful to all my co-authors for their valuable contributions that helped me to achieve the results presented in this thesis. Additionally, I would like to thank the industrial collaborators Nokia Bell Labs and Saab AB for funding the research projects. Furthermore, I am grateful to the Walter Ahlström Foundation, the Jenny and Antti Wihuri Foundation, and the Nokia Foundation for the personal financial support.

I am also thankful to Prof. Olivier Lafond and Prof. Claire Migliaccio for finding time to read through my thesis and for providing me highly valued constructive feedback during the preliminary examination process. It helped me to improve the thesis. Merci beaucoup!

This thesis was finalized in a situation where I was forced to work alone from home, and although I enjoyed the company of our cats, the new situation emphasized me the importance of our research group and my co-workers, both old and new. I have greatly enjoyed our time together at work and during the free time. Special thanks to Dr. Joni Kurvinen for your friendship and being a great roommate, Mr. Jan Bergman for the unparalleled speed at providing different codes I requested, Mr. Jaakko Haarla for providing the best memes especially...
during the past year, Dr. Jari-Matti Hannula for guiding us “youngsters”, Mr. Sabin Karki for your kindness and company during various trips, Mr. Riku Kormilainen for the discussions where we both could let the steam out, Mr. Henri Kähkönen for all the discussions from mm-waves to trekking and Gundams, Mr. Rasmus Luomaneni for sharing your expertise regardless the topic, and Dr. Resti Montoya Moreno for the siestas, questions, and football. I would also like to thank Dr. Jari Holopainen and Dr. Anu Lehtovuori for their valuable insights and for motivating me to learn more by sharing their passion for teaching. In addition, I want to express my gratitude for Prof. Ari Sihvola for offering me opportunities as a teaching assistant and in the Doctoral Programme Committee as well as for making the world of electromagnetics so much more fascinating and enjoyable. Furthermore, the service and technical personnel of the university have shared their expertise with me and helped me multiple times. Special thanks to Mr. Eino Kahra and Mr. Matti Vaaja for your help in manufacturing and measuring the prototypes.

There are many friends that have showed me there is life outside the research and supported me along the path towards the completed thesis. Special thanks to everyone in Viinimafia, to Veikko Sompa and his driving students, and to Vapaateekkarit. In addition, I am grateful to the Rauha family: thank you Ismo, Jenni, and Tilda for your support and for reminding me what is important in life.

Finally, this thesis could not have been done without the help of those closest to me. I want to thank my family: my parents Tuija and Jyrki, and my brother Tuomo. I am thankful for all the support and encouragement you have provided me. Last and definitely not least, my overflowing love and gratitude for Riina. It's a cliche, but very true in this case, that this thesis couldn’t have been done without your endless support and caring for me. Thank you for being there when I needed it the most.

Espoo, March 23, 2021,

Mikko K. Leino
<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.2.2 28-GHz antenna with integrated electronics</td>
<td>33</td>
</tr>
<tr>
<td>3.2.3 Discussion on antenna performance</td>
<td>36</td>
</tr>
<tr>
<td>3.3 Phased array diagnostics</td>
<td>37</td>
</tr>
<tr>
<td>3.4 Phased array performance optimization</td>
<td>39</td>
</tr>
<tr>
<td>4. Imaging with Millimeter-Wave Phased Array</td>
<td>43</td>
</tr>
<tr>
<td>4.1 Millimeter-wave imaging with phased arrays</td>
<td>43</td>
</tr>
<tr>
<td>4.2 Frequency-diverse imaging algorithm</td>
<td>45</td>
</tr>
<tr>
<td>4.3 Experimental results in $K_a$-band</td>
<td>47</td>
</tr>
<tr>
<td>5. Electrically Invisible Antenna</td>
<td>51</td>
</tr>
<tr>
<td>5.1 Low-scattering antennas</td>
<td>51</td>
</tr>
<tr>
<td>5.2 Design of a low-scattering antenna</td>
<td>52</td>
</tr>
<tr>
<td>5.2.1 Minimizing the structural scattering</td>
<td>52</td>
</tr>
<tr>
<td>5.2.2 Load optimization</td>
<td>54</td>
</tr>
<tr>
<td>5.3 Chopped dipole with inductive loading</td>
<td>55</td>
</tr>
<tr>
<td>5.3.1 Radiation performance</td>
<td>56</td>
</tr>
<tr>
<td>5.3.2 Scattering cross section</td>
<td>58</td>
</tr>
<tr>
<td>6. Summary of Publications</td>
<td>61</td>
</tr>
<tr>
<td>7. Conclusions</td>
<td>63</td>
</tr>
<tr>
<td>References</td>
<td>65</td>
</tr>
<tr>
<td>Publications</td>
<td>73</td>
</tr>
</tbody>
</table>
List of Publications

This thesis consists of an overview and of the following publications which are referred to in the text by their Roman numerals.


Author’s Contribution

Publication I: “E-Band Beam-Steerable and Scalable Phased Antenna Array for 5G Access Point”

The idea for the design is based on collaborative discussion between Prof. Viikari and Dr. Ala-Laurinaho. Dr. Islam had the leading role in the design work, preparing the content for the paper, and writing the manuscript. The author designed the transitions and the antenna elements and was involved with the phase shifter design. Mr. Luomaniemi carried out the measurements and was involved with the analysis of the results. Dr. Song designed the power division network. Dr. Ala-Laurinaho and Dr. Valkonen instructed the work. Prof. Viikari supervised the project.

Publication II: “Waveguide-Based Phased Array with Integrated Element-Specific Electronics for 28 GHz”

The author had the leading role in preparing the content for the paper and writing the manuscript. He designed the transitions, the PCB integration of the phase shifters, and the control software for beamforming. He was also involved with the measurements and the analysis of the results. Dr. Montoya Moreno designed the E-plane and H-plane power dividers and the matching steps for the E-plane divider. In addition, he studied the beam-steering capabilities of the antenna array. Dr. Ala-Laurinaho and Dr. Valkonen instructed the work. Prof. Viikari supervised the project.

Publication III: “Diagnostics of the Phased Array for E-band Using Holography Data”

This paper extends the work conducted in [I]. The author had the main responsibility for the paper and he performed the holography analysis. Dr. Islam assisted in the analysis. Dr. Ala-Laurinaho contributed to the validation of the analysis results and instructed the work. Prof. Viikari supervised the work.
Publication IV: “Beam Optimization for 28 GHz Phased Array Utilizing Measurement Data”

This is an extension of the work done in [II]. The author wrote the manuscript and the algorithm for the gain optimization method based on the concept by Dr. Ala-Laurinaho. The author also performed the measurements and was responsible for the analysis of the results. Mr. Bergman was involved with the measurements and derived the algorithm for the Taylor optimization method with instructions from the author and Dr. Ala-Laurinaho. Prof. Viikari supervised the work.

Publication V: “Millimeter-wave Imaging Method based on Frequency-Diverse Subarrays”

The author had the leading role in preparing the content for the paper and writing the manuscript. He performed the simulations and the analysis of the results with assistance from Dr. Ala-Laurinaho. All co-authors were involved with the derivation of the image-reconstruction algorithm and writing the paper.

Publication VI: “Millimeter-wave Frequency-Diverse Imaging with Phased Array Intended for Communications”

The author wrote the manuscript and was responsible for the analysis of the results. Mr. Bergman assisted the author in the analysis, and he also performed the measurements and wrote the algorithm for the image reconstruction with guidance from the author and Dr. Ala-Laurinaho. Prof. Viikari supervised the work.


This paper is the result of collaborative work. In addition to writing the manuscript, the author designed the antenna and derived the design principles for the low-scattering dipole structure. He studied the antenna performance through the simulations and performed the measurements and the analysis of the results with assistance from Dr. Ala-Laurinaho. Dr. Ala-Laurinaho, alongside Dr. Höök and Dr. Svensson, instructed the work, and Prof. Viikari supervised the project.
List of Abbreviations and Symbols

Abbreviations

2D Two-dimensional
3D Three-dimensional
5G Fifth generation
6G Sixth generation
AESA Active Electronically Scanned Array
AUT Antenna Under Test
CST Computer Simulations Technology
dc Direct Current
EM Electromagnetic
GPS Global Positioning System
HB High Band
IoT Internet of Things
LB Low Band
LTE Long-Term Evolution
MIMO Multiple-Input Multiple-Output
mm-wave Millimeter-wave
MoM Method of Moments
MVG Microwave Vision Group
NF Near Field
NR New Radio
PCB Printed Circuit Board
PEC Perfect Electric Conductor
RCS Radar Cross Section
RF Radio Frequency
Rx Receiver
List of Abbreviations and Symbols

SAR  Synthetic Aperture Radar
SIW  Substrate Integrated Waveguide
S-parameters Scattering parameters
SSR  Secondary Surveillance Radar
Tx   Transmitter
TwIST Two-step Iterative Shrinkage/Thresholding Algorithm
VNA  Vector Network Analyzer

Symbols

$\alpha$  desired phase shift
$\beta_i$  phase shift of phase state $i$
$\varepsilon_r$  relative permittivity
$\delta_{cr}$  cross-range resolution
$\delta_r$  range resolution
$\Gamma$  reflection coefficient
$\lambda$  wavelength
$\sigma$  radar cross section
$\theta_{Rx}$  observation angle
$\theta_{Tx}$  illumination angle

$a$  radius of a sphere
$A$  field data matrix
$A_{eff}$  effective aperture of an antenna
$A_i$  projected amplitude of phase state $i$
$a_i$  vector presentation of phase state $i$
$a_{\iota}$  amplitude of phase state $i$
$B$  imaging system bandwidth
$c$  speed of light
$d$  distance to an imaging plane
distance to an imaging plane
$d_{1,2}$  probe length inside a waveguide
$D$  directivity
$D_{ant}$  antenna dimension
$E_m$  electric field of receiver $m$
$E_n$  electric field of transmitter $n$
$f$  frequency
$f_k$  observed frequency sample
$G$  minimizer vector of iteration algorithm
List of Abbreviations and Symbols

\[ G_{Rx} \] receiver antenna gain
\[ G_{Tx} \] transmitter antenna gain
\[ h_1, h_2 \] probe distance from closed end of a waveguide
\[ I \] identity matrix
\[ L_{Rx} \] receiver aperture size
\[ L_{Tx} \] transmitter aperture size
\[ O \] observation data vector
\[ P_{Rx} \] received power
\[ P_{Tx} \] transmitted power
\[ r_{ff} \] far-field distance
\[ r_i \] distance to a voxel \( i \)
\[ r_{Rx} \] distance from receiver to a target
\[ r_{Tx} \] distance from transmitter to a target
\[ S \] scattering matrix
\[ S_{DD} \] coupling between ports on a chopped wire
\[ S_{DF} \] coupling from ports on a chopped wire to Floquet ports
\[ S_{DF} \] coupling from Floquet ports to ports on a chopped wire
\[ S_{FF} \] coupling between Floquet ports
\[ S_{ini} \] initial simulation model
\[ S_L \] load matrix
\[ S_{sys} \] two-port system model
\[ T \] target data vector
\[ T_t \] target data vector at iteration step \( t \)
\[ \hat{u} \] unit vector
\[ v \] iteration constant
\[ X \] reactance
\[ y_{mn} \] observation from transmitter \( n \) to receiver \( m \)
\[ \tan \delta \] dielectric loss tangent
\[ Z_0 \] characteristic impedance
1. Introduction

Wireless communications have evolved significantly within recent years. Mobile communication devices have become part of our everyday lives; whereas a couple of decades ago they were mainly used for calls and messages, they now offer many other applications, e.g., instant access to the Internet, video streaming services, and positioning systems, such as GPS. The world around us has become more connected with applications, such as the Internet of Things (IoT), and this has led to a continuous increase in mobile data traffic with no end in sight [1]. Fifth-generation mobile networks (5G) are developed for this purpose, and 5G NR (New Radio) radio access technology is currently being deployed and standardized.

Communication systems are forced to advance continuously so that they can meet the requirements of the envisioned future systems [2]–[4]. Current mobile networks are based on the Long-Term Evolution (LTE) standard, but some of the upcoming services, like autonomous vehicles and stable wireless data transfer at several Gbit/s, are beyond the capacity of existing networks. One way that 5G expands the capacity is by adding more spectral resources. This is done in 5G NR by allocating new frequency bands in sub-6 GHz range and in millimeter-wave (mm-wave) frequencies.

The lengths of the electromagnetic (EM) waves are in the scale of millimeters in the radio frequency range from 30 to 300 GHz. The frequencies belonging to the band are thus called mm-wave frequencies. Traditionally, the mm-wave frequencies have been used in radio astronomy, remote sensing, and radar applications. However, the research both in the academia and in the industry for the mm-wave region has increased exponentially in recent years, primarily because of the interest shown by the telecommunications industry. As previously asserted, the mm-wave range offers more frequency resources than the currently used frequency range. In addition, the absolute bandwidth of the data transmission is increased, which allows for higher capacity and thus faster data transmission.

The environment has greater effect for radio-wave propagation at the higher frequencies, and, in particular, the atmospheric absorption increases [5]. To compensate for the higher propagation loss, mm-wave antennas normally have
Introduction

a narrow beam and high directivity so that the radiated power can be directed as energy-efficiently as possible in the desired direction. The antenna beam-steering is an essential feature in the applications that require the antenna beam to track a moving object, whether the object is an end-user of a mobile network or a target to be detected by a radar.

The next generation 6G networks even beyond 5G are already envisioned; one such vision sees localization and sensing becoming a more integral part of future networks, where they naturally coexist with communications [6]. Bringing the two applications closer together is worth researching. Mm-waves allow imaging due to their small wavelength, and security and military sectors have a long history with the frequency band. Although the antennas are often designed with a particular application in mind, using them in multiple operation scenarios is desirable, since antenna development can be both expensive and time consuming. The number of beam-steerable mm-wave antennas surrounding us will increase exponentially when the 5G networks are constructed [7], and the possibility to use the communication antennas in imaging applications should be considered.

In addition, the antennas operating at lower frequencies can interfere with the beamforming of a high-band (HB) antenna and thus decrease the HB antenna performance. In radar applications in particular, this problem is already quite significant, for example, with L-band secondary surveillance radars that locate next to the primary antennas operating in X-band. However, the problem may also arise in the mobile networks with the spread of 5G. For this reason, the low-band (LB) antenna solutions that do not interfere with the operation of the HB antennas should be developed.

1.1 Objectives of this work

This work has three main objectives: first, to design antennas that are suitable for mm-wave base stations and to develop tools to analyze them. Second, to investigate the suitability of said antennas for sensing applications. Third, to develop electrically invisible antennas that can co-exist with the high-band counterparts. Electrical invisibility in the scope of this thesis denotes “low-scattering”, and these antennas can find their applications in radar and communication systems. These objectives can be summarized by the following research questions:

- How to design and analyze beam-steerable mm-wave antennas for communications that simultaneously have high efficiency and contain integrated electronics?

- Is it possible to use the frequency-diverse, non-conformal beams of the mm-wave phased array intended for communications in sensing applications?

- How to design a LB antenna that can co-exist in a shared aperture with a HB antenna without affecting the performance of the HB antenna?
This thesis aims to answer these questions, and, in doing so, it strives to provide design tools and insight for the antennas in future communication and sensing systems.

1.2 Main scientific merits

The main scientific merits of this thesis are as follows:

1. The development of waveguide-based, low-loss antenna array and phase shifter integration concept: antennas combining waveguide-based power division networks and antenna elements with PCB-mounted, element-specific phase shifters while demonstrating low transmission losses [I]–[II].


3. Performance optimization techniques for phased arrays by utilizing element-specific measurement data [IV].

4. The development and validation of a frequency-diverse imaging method for phased arrays that are intended for communication platforms [V]–[VI].

5. Concept and design principles for the invisible antenna: an antenna that efficiently radiates at its nominal frequency and simultaneously has low scattering at higher frequencies. The realization of said antenna for a secondary surveillance radar application [VII].

1.3 Contents and organization of the thesis

This thesis consists of an overview and seven publications. The overview provides the background for the thesis and presents the key findings of the publications. More detailed descriptions of the proposed antennas and methods can be found in the publications at the end.

First, Chapter 2 presents the background information regarding the main concepts in the scope of the thesis, and it also discusses the tools of antenna design. The research results are summarized in the following three chapters. Each chapter starts with an introduction to the discussed topic to give a more comprehensive overview of the subject at hand.

Chapter 3 presents the low-loss phased arrays for 5G applications and tools to characterize and analyze them. Chapter 4 discusses the frequency-diverse imaging method for phased arrays and presents the experimental verification of the method with antennas intended for communication. Chapter 5 then describes the design principles of the electrically invisible antenna concept.
and the measured performance of this antenna. Finally, Chapter 6 presents summarizations of each publication, and Chapter 7 concludes the thesis with some proposal for future work.
2. Background and Research Methods

This chapter discusses the basic information and concepts that are most central to the scope of this thesis. First, brief introductions to the wireless systems and antennas are presented in Sections 2.1 and 2.2. Section 2.3 then discusses the scattering concepts related to the thesis. Finally, Section 2.4 explains the research methods of the work through the tools utilized in the antenna design and analysis.

2.1 Radio spectrum

The basis of communication and imaging systems is the electromagnetic waves that are constructed from the time-varying electric and magnetic fields propagating at the speed of light $c$. The waves can be categorized through their frequencies $f$ or wavelengths $\lambda$.

The radio spectrum consists of frequencies from 3 Hz to 300 GHz, and it is divided into different regions and bands [8]. For example, the mm-wave frequencies cover the range 30–300 GHz. The boundaries between bands are arbitrary and are often assigned without any physical reasoning. They are simply a matter of convention; they exist mainly to make the frequency allocation easier and to allow the collection of similar applications under the same band in order to avoid accidental interference between different systems.

Table 2.1 lists the frequency ranges, the corresponding band designations, and the studied applications that are discussed in the context of this thesis. The majority of the research has been conducted in regard to mm-waves and the potential 5G bands. The low-scattering studies have been done at lower frequencies, since their intended applications are in civil aviation radars that operate at those frequencies.
Table 2.1. Frequency ranges and bands considered in this thesis.

<table>
<thead>
<tr>
<th>Band</th>
<th>Frequency (GHz)</th>
<th>Studied application</th>
<th>Publication</th>
</tr>
</thead>
<tbody>
<tr>
<td>L</td>
<td>1–2</td>
<td>Sensing</td>
<td>[VII]</td>
</tr>
<tr>
<td>X</td>
<td>8–12</td>
<td>Sensing</td>
<td>[VII]</td>
</tr>
<tr>
<td>K_a</td>
<td>24–32</td>
<td>Sensing</td>
<td>[VI]</td>
</tr>
<tr>
<td></td>
<td>26–30</td>
<td>Communication</td>
<td>[II], [IV]</td>
</tr>
<tr>
<td></td>
<td>36–44</td>
<td>Sensing</td>
<td>[V]</td>
</tr>
<tr>
<td>E</td>
<td>71–86</td>
<td>Communication</td>
<td>[I], [III]</td>
</tr>
</tbody>
</table>

2.2 Antenna fundamentals

In wireless systems, transmitters (Tx) are used to send the encoded data in EM waves to the free-space media, and these waves are then captured and sensed with receivers (Rx). The antenna is the part of the wireless system that is designed to radiate or to receive these EM waves [9]. In short, antennas act as interfaces between a guided media (e.g., a transmission line) and an open media (i.e., the free space). They are crucial components in a communication system; antennas largely determine how much energy is transferred between the Tx and Rx and how much energy is collected from unwanted, possibly interfering sources or transmitted in unintended directions. This section describes the most relevant antenna parameters for this thesis and the basics of the antenna arrays. In addition, transmission lines are briefly discussed, since they are often an integral part of the antenna.

2.2.1 Antenna parameters

The electrical operation of an antenna is often described using its matching, efficiency, gain, and polarization. Like any other electrical component, antennas also have an impedance, typically observed from their feeding points. For an antenna to radiate, it must accept power from a source or generator. If the impedances between the antenna and the generator do not match, part of the input power is reflected back to the source. Matching efficiency describes how well the power available from the source is transmitted to the antenna. Generally, a matching level is considered acceptable when the reflection coefficient is below –10 dB in the required band.

The radiation efficiency of an antenna describes how well the antenna transforms the accepted power into the EM radiation. For example, part of the power is dissipated in the antenna structure by dielectric and resistive losses. While radiation efficiency is a good parameter for antenna characterization, it does not take into account the matching. The total efficiency is the product of both the matching and the radiation efficiency, and it is often a useful figure of merit, since it is easy to measure in practice.
Antenna directivity specifies the antenna’s ability to focus energy in a certain direction. The directional dependency of the directivity is described by the three-dimensional (3D) radiation pattern of an antenna. An isotropic radiator would distribute the power equally in angular space, but it is impossible to realize in practice. Furthermore, isotropic radiators would not be practical at mm-waves, since the increased absorption loss requires more efficient concentration of the power so that the communication range can be extended. Therefore, the mm-wave antennas have typically focused radiation patterns and narrow, pencil-like main beams.

Although directivity is an important parameter of an antenna, it does not take into account the losses in the antenna structure. A realized gain is a product of the directivity and the total efficiency, and realized gain values are typically given relative to the isotropic radiator in decibels (dBi). The realized gain should not be confused with the term “antenna gain,” which does not take into account the matching losses but only the radiation efficiency.

The antenna directivity $D(\theta, \phi)$ can be related to the effective size of the antenna aperture $A_{\text{eff}}$ as [8]:

$$D(\theta, \phi) = \frac{4\pi}{\lambda^2} A_{\text{eff}}(\theta, \phi). \quad (2.1)$$

With aperture antennas, such as those in [I] and [II], the effective aperture size is typically the same or close to the actual physical size, and the physical size can hence give a good estimate for the directivity and, assuming low losses, for the gain. Furthermore, the equation above highlights how a larger aperture can increase the gain at the operation frequency and why many of the mm-wave antenna realizations generally have large apertures compared to the wavelength.

Antenna polarization defines that of the transmitted (or received) wave. Antennas presented in this thesis are linearly polarized, meaning that the electric field follows a single line transverse to the propagation direction. Linearly polarized antennas are also often described by their E- and H-planes. These are “the plane containing the electric/magnetic-field vector and the direction of maximum radiation” [10]. Aperture antennas typically have easily distinctive E- and H-planes, and these planes can be used to describe the direction of the beam-steering if this function is applicable to the discussed antenna.

Furthermore, antennas typically operate in each other’s far-fields, and the radiation properties and patterns are associated with radiating fields that are observed in the far-field. However, close to the antenna, the reactive parts of the fields are present, although they decrease rapidly when the distance from the antenna increases. The typical boundary between the near- and far-field is at the distance [10]

$$r_{\text{ff}} = \frac{2D_{\text{ant}}^2}{\lambda}, \quad (2.2)$$

where $D_{\text{ant}}$ is the maximum measure of the antenna. This limit holds especially well for mm-wave antennas that are normally electrically large compared to the wavelength. However, the equation also shows that the far-field distance can be
quite large for the mm-waves, which can make the antenna measurements in far-field challenging. This is discussed in more depth in Section 2.4.3.

2.2.2 Antenna arrays

One straightforward way to increase the size of the antenna aperture and the gain is by utilizing multiple antennas together in an array constellation. Typical antenna arrays are constructed using identical antenna elements that are evenly distributed either on a line (linear array) or on a plane (planar array). The radiation pattern of such an array depends on the pattern of an antenna element and how the elements are distributed and fed in regard to each other; namely, the array factor.

The array factor describes how the radiating waves from isotropic point sources are combined depending on their corresponding locations and initial weights, i.e., amplitudes and phases. Based on the source locations and weights, the EM fields add together either constructively or destructively in different directions in the far-field, forming the total pattern. Normally, the elements are in fixed locations, and the radiation pattern is adjusted by changing the excitation signal phases to different elements. Such antennas are phased arrays, and they allow scanning of the free space with their main beam.

In phased arrays, the main beam is steered by applying a progressive phase shift to every element. The spacing between the elements is normally chosen to be $\lambda/2$ because with larger spacing, the grating lobes can emerge in the visible range depending on the main beam steering angle. The radiation pattern of the phased array can be also shaped by adjusting the excitation amplitude of each element. The benefit of this operation is that amplitude control can be used to achieve lower side lobes at the cost of a wider main beam. With uniform excitation, the side-lobe level for square (or linear) array is $-13.3$ dB. By applying a Taylor distribution to the excitation, the side-lobe level can be, for example, as low as $-25$ dB [8]. A more detailed description and analysis of the phased arrays can be found, for instance, in [11].

2.2.3 Transmission lines

The feed signal must be transferred to the antenna element from the source. Feeding networks are integral parts of antenna arrays, since the fed signal must be distributed to the array elements. The transmission losses in the feeding networks are due to the limited conductivity of the metals and the losses in the dielectrics. The chosen transmission line structure typically relies on the chosen antenna structure—the two go normally hand in hand so that they can be implemented together easily.

Planar transmission lines on printed circuit boards (PCBs) are alluring: the PCB technology allows for cost-effective manufacturing and the facile integration of electrical components. However, in planar transmission structures, such as
microstrip lines or co-planar waveguides, the losses increase with the frequency and can be rather high at mm-waves. In these structures, a large portion of the propagating wave is inside the dielectric, where the losses tend to increase with frequency [12]. In addition, these structures are at least partly open, and, at higher frequencies, the losses will increase due to the radiation from the transmission lines. Similarly, the power dissipation in dielectrics is a problem in coaxial cables as well, even though the wave itself is confined to the transmission structure, thus preventing radiation losses.

Air-filled waveguides can support high power levels and offer low losses in mm-waves compared to most other transmission structures. However, they are bulky and do not facilitate facile component integration. Additionally, the manufacturing costs are typically higher than those of PCB structures, although the cost difference may be decreasing due to the emergence of 3D-printing technology advancements. Nevertheless, in terms of the most efficient energy transfer in mm-waves, the air-filled waveguides continue to outperform other solutions.

2.3 Scattering and radar cross section

Sensing with radio waves has a long history. Radar, acronym for radio detection and ranging, is one of the well-known examples, with early implementations dating back to World War II and British Chain Home radar system [13]. Radars are prime examples of active imaging/sensing systems: Tx is used to illuminate the target space, and Rx observes the reflections from the possible targets. Active imaging is often mandatory to achieve sufficiently high dynamic range in the sensing system. Whereas the radars are typically used to sense the far-field, the near-field domain has its applications in medical imaging and security screening systems [14]–[16]. Furthermore, long ranges in sensing are generally easier to achieve in practical realizations with lower frequencies than with higher ones like mm-waves, similarly as with the communication systems.

Resolution is an important figure of merit when describing a sensing system. Resolution can be evaluated, for example, with the help of the point-spread function [17]. Furthermore, image reconstruction generally requires some form of post-processing, since the information about the target is collected through the received signal. This can be done, for instance, by recording the propagation time of the signal or by analyzing the change in the signal phase and amplitude. This thesis and the imaging method presented in Chapter 4 focus on calculating the image computationally from various combinations of the transmitter and receiver radiation patterns.

Scattering explains how the target object affects the propagation of the incident wave, that is, how much the incident wave interacts with the object. The larger the scattering, the better the visibility of the object to a radar. In contrast, an object with strong scattering in close vicinity of an antenna can disturb its
Background and Research Methods

operation. In the following sections, the concepts of antenna scattering and radar cross section are discussed, providing important background for consideration of the low-scattering antennas that are discussed in Chapter 4.

2.3.1 Antenna scattering

Scattering from an antenna has been studied extensively [18]–[20], and it can be analyzed with the help of two different scattering modes: structural and antenna-mode scattering [18], [20]. The structural scattering is independent of the load in the antenna port, and it describes how the incident EM wave scatters from the surface of the antenna, even when the antenna is matched to its load.

In contrast, antenna-mode scattering depends on the radiation properties of the antenna. In essence, if there is a mismatch between the antenna load and the signal that was induced by the incident wave, part of the energy is not accepted into the antenna port and re-radiates into free space. While wide-band impedance matching can reduce the antenna-mode scattering, the structural scattering will still remain. It can thus be more desirable to reduce the structural scattering, since it can also affect how the incident wave couples to the antenna structure.

2.3.2 Radar cross section

In this thesis, the scattering is analyzed through a radar cross section (RCS), a far-field property for a target object that explains how the illuminated structure scatters energy towards the receiver. The RCS depends on the incident angle of the illumination and on the observation angle. If the Tx and Rx angles are the same, the situation and the RCS are said to be monostatic, and only the backscattering from an object matters. If the Rx observes the scattering from a different angle than the Tx, the situation and the RCS are described as bistatic. Fig. 2.1 further illustrates the difference between the two cases.

![Figure 2.1. Radar setup and scattering evaluation in different situations: (a) monostatic case and (b) bistatic case.](image-url)
The RCS can be calculated by reorganizing the radar equation for the bistatic case [21]:

$$\sigma(\theta_{Tx}, \theta_{Rx}) = \frac{P_{Rx}(4\pi)^3r_{TX}^2r_{RX}^2}{P_{Tx}G_{TX}G_{RX}\lambda^2},$$

(2.3)

where $\theta_{Tx}$ and $\theta_{Rx}$ are the illumination and observation angles, respectively. $P_{Tx}$ and $P_{Rx}$ are the transmitted and received powers, while $G_{TX}$ and $G_{RX}$ are the antenna gains in the target direction for the transmitter and receiver, respectively. $r_{TX}$ and $r_{RX}$ describe the distances between the target and the transmitting and receiving antennas, respectively. Although the above equation is for bistatic RCSs, the monostatic result can be obtained by setting $\theta_{Tx} = \theta_{Rx}$ and $r_{TX} = r_{RX}$.

It should be again highlighted that the RCS is only observed from a single direction at a time. The structure scatters the incident wave in the other directions as well, and even if the scattering towards an observed direction is low, the scattering over the whole space can be significant. The total scattering can be evaluated by simulations in a straightforward manner: a total RCS gives the scattering cross section value with which the received power is calculated over the whole solid angle. Measuring this value for all the transmission and receiving direction pairs would be quite impractical even though reciprocity often applies. However, the simulations can give a good initial guess of which (bistatic) angles are worth of measuring.

A sphere has the same monostatic RCS response from every illumination angle. It is hence a suitable object to describe different scattering regions. Fig. 2.2 shows the simulated monostatic RCS for a sphere made from perfect electrical conductor (PEC). The sphere with radius $a$ has a physical cross section of $\pi a^2$ intercepting the incident wave, and the RCS is normalized to this value. Fig. 2.2 also highlights the wavelength, i.e., the frequency dependency of the RCS.

![Figure 2.2. Scattering regions explained with a monostatic RCS from a conducting sphere. The monostatic RCS is obtained through simulations by illuminating a PEC sphere with a plane wave.](image-url)
Three distinctive regions can be identified [21]: the Rayleigh region on the far left of the spectrum, where the scattering is lowest but depends strongly on the wavelength; the Mie or resonant region in the middle, where the scattering has periodical variation; and the optical region on the far right of the spectrum, where the scattering begins to be independent on the frequency, and the RCS approaches the physical size of the intercepting area.

Fig. 2.2 indicates that the scattering cross section of the sphere is much smaller than its physical cross section in the Rayleigh region, that is, when the sphere is small compared to the wavelength. The same result applies for most other shapes as well. Therefore, objects can be made fairly weak scatterers by ensuring that they are in the Rayleigh region. However, with antennas, this can be problematic, because the antenna size is typically such that the antenna appears in Mie region at its operation frequency. This problem is further discussed in Chapter 5.

### 2.4 Antenna design and analysis

Antennas, like other EM structures, are typically first studied and designed with simulations. These provide the necessary information for prototyping and manufacturing, after which the realized antennas are measured to verify their performance in reality. This has also been the design process in this thesis and the following describes the used tools more closely.

#### 2.4.1 Simulation tools

The EM simulations in this thesis have been conducted utilizing a commercial software, CST Studio Suite [22]. The software has several 3D full-wave solvers (e.g., the frequency domain, the time domain, and the integral equation solvers) that have been used to model the different antennas presented in this thesis. Additionally, CST also allows the combining of circuit simulations with EM simulations, which has been particularly useful while investigating component integration into the antennas.

#### 2.4.2 Antenna measurements

Each measurement setup presented in this thesis is constructed around a vector network analyzer (VNA). The VNA enables the measuring of the frequency responses of the antennas under test (AUTs) that are connected to the VNA ports. This is done through scattering parameters (S-parameters). A single port can be used to measure the reflection from an antenna, and multiple ports are required to measure the radiation properties of the AUT. For example, the near-field and far-field measurement systems use the additional ports for the measurement probes that characterize field values of the AUT. The near-field
and far-field measurements are discussed in more depth in the following. In addition, the two-port measurements have been also used in [VI] and [VII], since both works required bistatic sensing setups. A more detailed description of the respective measurement setups can be found in the corresponding chapters.

Measurement data generally require processing: the data may be compared to the results of the known reference antenna to find out the actual values of, for example, gain and efficiency. Furthermore, the signal can be presented in time domain or frequency domain, namely as a function of time or frequency, respectively. Both presentations contain the same information, but different properties can be observed from them. As previously explained, the VNA measures the AUT response in the frequency domain. The frequency domain view allows, for example, the characterization of the AUT bandwidth and matching. In contrast, the time domain view allows the identification of the structures in the measurement setup that are not part of the AUT but that still affect the measured results. In the time-gating process, the signal is filtered in the time domain so that the effect of the error sources are omitted from the final processed results. Time-gating has been particularly important in [VII].

2.4.3 Near-field measurements and holography

In the near-field measurements, the probe collects the field values in the AUT near-field. Electrically large mm-wave antennas are generally characterized by near-field measurements because their far-field distances are often large, thus leading to impractically large measurement ranges, while the high atmospheric loss can also reduce the accuracy of the measurement system. Near-field measurements are also considered better for antennas with adjustable beams and can provide lots of data for the subsequent AUT diagnostics. The far-field values are calculated computationally from the near-field data, typically by a measurement software. In this thesis, a planar near-field scanner, NSI-5 × 5, and its software, NSI-2000, have been used to characterize mm-wave antennas in [I]–[IV]. Similarly, MVG’s StarLab 6 GHz antenna measurement system has been used in [VII] to measure the total efficiencies of different dipole antennas.

The collected near-field data can be also used to evaluate fields on the aperture plane of the AUT. Whereas the far-field transformation computationally propagates the measured fields away from the AUT, the back propagation moves the fields towards the antenna [23]. The fields at the aperture plane are called holograms, and they have been used to diagnose antennas in [II] and [III].

2.4.4 Far-field measurements

In the far-field measurements, the observation of the AUT occurs in the far-field: the measuring probe is in the far-field of the AUT and vice versa. Measurement ranges are typically in anechoic chambers, and chambers are generally shielded to prevent interference due to external signals. In this thesis, the far-field
measurements in the anechoic chamber of Aalto Electronics-ICT have been used in the RCS evaluation of the different dipole antennas discussed in [VII]. The RCS is a far-field property and cannot be acquired through a similar near-field transformation to the far-field as the radiation patterns.
3. Low-Loss Phased Arrays for 5G Base-Station Applications

This chapter presents antennas that are designed for 5G base stations. These phased array designs aim to reduce transmission losses often present at mm-wave frequencies. The antennas are discussed in Publications [I] and [II], and they are part of the main contribution of this thesis. Section 3.1 gives an overview of the mm-wave base-station antenna designs and relates the presented antennas to the state-of-the-art technology. Section 3.2 discusses in detail the antenna designs and their respective performances.

Phased arrays are complex antenna systems that often require antenna performance diagnosis and optimization. Publications [III] and [IV] support the main contribution by presenting methods for diagnosing and calibrating the aforementioned antennas, and the methods have provided important information that should be utilized to improve the current designs. Furthermore, these approaches are also applicable to other phased arrays. Sections 3.3 and 3.4 describe the methods.

3.1 Millimeter-wave antennas for base stations

Electrical beam-steering is one of the main requirements for 5G mm-wave base-station antennas. Reconfigurable antenna beams are essential to realize envisioned deployment schemes and to focus capacity where it is needed [24]. Similarly, a high directivity is required to maximize the range and to compensate for the increased path loss at mm-wave frequencies [5]. The small sizes of 5G network cells and the subsequent increase in the number of the base-station antennas dictates its own conditions: the antennas should preferably support mass production for lower manufacturing costs, while they should simultaneously have a low profile for imperceptible installation. Additionally, antenna efficiency significantly affects the energy consumption of mobile communications and should therefore be high.

Many different antenna architectures have been proposed as potential solutions for mm-wave beam-steering. These include, leaky wave antennas [25], reconfigurable reflect- and transmitarrays [26]–[28], and passive beam-forming
structures, such as a Rotman lens with switchable feeds [29], [30]. However, the research for 5G mm-waves has mainly focused on two distinctive antenna types: integrated lens antennas and phased arrays.

Integrated lens antennas operate similarly to the optical lenses; the shape of the lens is used to focus the passing EM waves through refraction. While the directivity of the lens antenna can be high, its disadvantage is its bulkiness. Furthermore, electrical beam-steering with a lens antenna requires a focal feed array and a beam-switching network, which often leads to high losses. Nevertheless, several promising solutions for base-station applications have been presented for E-band and K_a-band [31]–[34].

In a phased array, the phase of each element is controlled independently using a phase shifter. Printed circuit boards offer a suitable platform for the component integration and mass production. Moreover, PCB-based phased arrays are attractive solutions, and multiple realizations have been presented for mm-wave communications [35]–[38]. While each of these solutions indicate very good performance (e.g., a wide beam-scanning region and a high directivity), the transmission losses in the power division networks are high, primarily due to their planar realizations.

Losses can be reduced by introducing more efficient transmission structures, and good results have been achieved utilizing gap waveguides [39], [40] and substrate integrated waveguides (SIW) [41], [42]. However, these structures might become complex and challenging to realize, thus waveguide-based solutions are attractive in their simplicity. Based on the work published in [43], [44], one commercial example of the low-loss, waveguide-based antenna array for mm-wave communications is Nokia MetroHopper. The antenna supports inexpensive manufacturing through plastic injection molding, a mass production method suitable for mm-waves, as shown in [45] and [46]. However, the MetroHopper antenna is a fully passive structure, and it is not electrically beam-steerable, but it has inspired part of the research performed in this dissertation.

Antennas presented in [I] and [II] are four-by-four phased array prototypes that are designed for E-band (71–86 GHz) and lower K_a-band (26–30 GHz), respectively. Air-filled rectangular waveguides are used as the main building blocks of the proposed antennas, since they are commonly known to exhibit low losses at mm-waves. However, since the majority of the electronic components used in antenna systems are PCB-based, integrating these components with the waveguide structures is necessary. Furthermore, the presented antenna structures are scalable in the number of antenna elements, for example, up to 64.

One feature missing from the discussed antennas that is included in the designs that have been published afterwards [47]–[49] is dual-polarized elements. Dual-polarization is needed in, for example, polarization diversity and MIMO [50], [51]. Nevertheless, the antennas in [I] and [II] have demonstrated one of the lowest insertion losses when considering the antenna systems including the power division network.
3.2 Waveguide-based phased arrays with integrated phase shifters

In the following, the design principles of the waveguide-based phased array antennas are first described, and then the performances of both antennas are presented separately in Sections 3.2.1 and 3.2.2. Finally, Section 3.2.3 discusses advantages and future improvements for these antennas in detail.

Fig. 3.1 presents the overview of the proposed low-loss antenna structure, which combines waveguide-based feeding network and antenna elements with PCB-integrated components. Four main building blocks can be identified: a power division network, waveguide-to-PCB transitions, PCB with embedded phase shifters, and pyramidal horn antenna elements. The operation principle is simple: the power division network distributes the input signal equally to 16 outputs, one for each antenna element. The input signal is then transferred from the waveguide environment to the PCB via transition. Element-specific phase shifters control the phase of each element independently, and after the phase shifter is another transition back to the waveguide environment. Pyramidal horns are used as antenna elements to radiate the energy.

The feeding network is constructed from one primary and four secondary power dividers, depicted in Fig. 3.2 (a) and (b), respectively. The primary network divides the input signal in the E-plane to four branches, and the following secondary networks split the signal to a further four outputs, this time in the H-plane. The power divider is designed to divide the power equally between the output ports. However, the phase delay is different between some output ports due to the non-symmetric structure of the power divider.

The phase shifters operate in a microstrip environment on a multilayer PCB immediately after the power division network. The PCBs in both antennas have a similar architecture. The RF layer on top is made from a low-loss RF substrate, RO5880 ($\varepsilon_r = 2.20$, $\tan\delta = 0.0009$). Layers below are made from FR-4 and are reserved for dc routing where the control lines for the phase shifters can be implemented.
The side-view of the phase shifter integration including the transition structures between the PCB and the waveguides is illustrated in Fig. 3.3. The waveguides are tapered in the H-plane to fit the phase shifters in-between the elements. The transitions utilize microstrip lines as probes, which is a commonly used technique [52], [53]. The performance is optimized to provide maximum transmission by adjusting the probe lengths inside the waveguides ($d_1–d_2$) and the probe distances from the closed ends of the waveguides ($h_1–h_2$).
Fig. 3.3 also depicts how the cavities are processed in the PCB around the waveguides. Only the RF substrate remains inside the waveguide with the microstrip probe. The edges of the single cavity are metallized so that the surrounding waveguide remains coherent, and with the via fences on the PCB, they ensure that the incoming wave does not leak through the PCB to the adjacent elements.

### 3.2.1 E-band antenna with time-delay lines

Chronologically, the antenna presented in [I] was designed and manufactured first. It is designed to operate at E-band (71–76 and 81–86 GHz), a potential 5G band. Fig. 3.4(a) presents the manufactured antenna with element numbering and beam-steering planes. Since there were no commercial phase shifters available at the time of manufacturing, the beam-steering is realized with the microstrip-based, true time-delay lines. Several PCB assemblies have been manufactured for different beam-steering angles, with each PCB having adjusted delay lines for the required element phasing. Fig. 3.4(b) depicts the PCB realization for the broadside beam. The aforementioned figures show how the array elements are slightly offset in the azimuth plane. The reason for this is to reduce the grating lobes in azimuth steering, which are due to the large element spacing. The grating lobes and the reason behind the large element spacing are discussed in more detail in section 3.2.3.

The transitions must be properly designed for minimal insertion loss. Fig. 3.5 illustrates the S-parameter results of the simulation model that incorporates the both transitions (waveguide-to-PCB and PCB-to-waveguide) and the phase shifter. The reflection coefficient is well below $-10\,\text{dB}$ for most of the lower E-band (71–76 GHz), and the transmission losses are less than 2 dB for the whole band. The phase shifter in the simulation model provides a $45^\circ$ phase delay at 72.5 GHz.

![Figure 3.4](image)

**Figure 3.4.** E-band antenna with time-delay lines: (a) element numbering and steering directions. Adapted from [III]. (b) Time-delay lines for the broadside beam [III] © 2018 EuMA.
The far-field properties of the antenna at 72.5 GHz have been measured using a near-field scanner. The simulated and measured results show good agreement, even though the absolute values differ marginally. The measured gain is 15.2 dBi, while the simulated gain is 16.9 dBi. Potential error sources, such as uncertainties in measurements and material parameters, can explain this difference. Furthermore, the simulations do not consider the surface roughness, which is known to have an effect at mm-waves [54]–[56].

The measured beam-steering capabilities of the antenna at 72.5 GHz are illustrated in Fig. 3.6 for the elevation and azimuth angles. The antenna has approximately 5 dB scan loss at 40° in both planes. The element offset in the azimuth plane appears to work as intended, since the azimuth steering is grating-lobe free. However, the situation for elevation steering is quite the opposite: the grating lobe is visible already with the 30° beam. Moreover, for the 40° beam-steering angle, the grating lobe is higher than the beam in the desired direction. Future designs should address this problem, and one possible solution is discussed in Section 3.2.3.
3.2.2 28-GHz antenna with integrated electronics

The design from [I] was scaled for the lower operation band 26–30 GHz. However, some changes were made based on the experience gained from the earlier E-band design. For example, the feeding network in [II] was improved upon that presented in [I] to support facile manufacturing with milling. The antenna in [II] follows the same steering planes and element numbering as the one in [I], but it is realized without the element offset in the azimuth plane, as shown in Fig. 3.7.

The antenna presented in [II] is implemented with electronically controlled 5-bit phase shifters, namely TGP2100 [57]. A major difference is hence that the PCB realization and the phase shifter integration presented in [II] are more complex than that in [I]. The dc routing is realized inside the FR-4 section of the PCB, and Fig. 3.8 illustrates the stack-up of the manufactured PCB. Fig. 3.9 demonstrates the phase shifter integration from the top. The phase shifter is connected directly to the RF ground plane from its bottom, in a cavity that is
specifically milled for this purpose. The control line pads and the RF signal pads on the top of the PCB are connected with bond wires to the corresponding pads on the phase shifter.

Fig. 3.10 illustrates the simulated S-parameter results of the two transitions placed back to back. The transmission losses due to the transitions are small—less than 0.4 dB over the majority of the observed band. The matching is also better than −10 dB. However, the passive phase shifters are known to be lossy components, as is evident in Fig. 3.11(a), which shows the simulated transmission coefficient of the entity, including both transitions and the phase shifter. The insertion loss depends on the chosen phase shifter state (phase shift), and the variation has been highlighted. Similar results for the reflection coefficient are shown in Fig. 3.11(b). Matching with the phase shifter is rather poor due to a design flaw in the integration: the bond wires require larger pads (flares) at the end of the microstrip lines to compensate for the inductance caused by the wires. This is easy to fix in the next revision of the PCB.

![Figure 3.9. Top view of the PCB-integrated phase shifter in the 28-GHz phased array. Reprinted from [II].](image1)

![Figure 3.10. Simulated S-parameters of the transition structures for the 28-GHz phased array. The phase shifter is not included in the evaluation. Reprinted from [II].](image2)
Fig. 3.11. Simulated results of the transition structure with the phase shifter for the 28-GHz phased array: (a) transmission coefficient and (b) reflection coefficient. Reprinted from [II].

Fig. 3.12 (a) and (b) demonstrate the measured beam-steering performance of the phased array in the azimuth and elevation planes, respectively. The maximum gain towards the broadside is 11.3 dBi after a simple phase calibration. The scan loss is around 3 dB until a $30^\circ$ steering angle in both planes, and, while the azimuth steering works rather well, the elevation plane is plagued with early rising grating lobes. According to the array theory, the grating lobes should appear in the visible region when the steering angle exceeds $20^\circ$, and the performance should be similar in both H- and E-planes. However, the visible difference in the beam-steering performance is due to a stronger mutual coupling between the elements in E-plane. Because of this stronger inter-element coupling, the phase shifters do not provide the intended amplitude and phase in elevation steering, thus making the beam patterns not fully coherent in E-plane.

Figure 3.12. Beam-steering performance of the 28-GHz phased array: (a) azimuth plane and (b) elevation plane. Reprinted from [II].
3.2.3 Discussion on antenna performance

The main advantages of these antennas are the low losses in the power division and in the antenna elements. The phased arrays provide high realized gains and adequate beam-steering capabilities in 2D for their respective operation frequencies. The measured efficiency for the E-band antenna is quite high, indicating that the losses in the whole antenna system are less than 4 dB. In contrast, the majority of the losses in the 28-GHz array are due to the phase shifters, but they could be replaced with, for example, active vector modulators. The antenna structure itself has very low losses—less than 1 dB. In the loss comparison, the waveguide antennas are clearly superior to the planar realizations on the PCB. The presented antennas also support manufacturing by plastic injection molding to some extent. This could lower the manufacturing costs in mass production.

The main trade-off in using the waveguide-based feeding network and antenna elements is that the element spacing increases above $\lambda/2$. The reason for this is the size of the waveguide in the H-plane: the cut-off frequency dictates this dimension, and, for the air-filled waveguides, it is larger than $\lambda/2$ at the operation bandwidth of the antenna. In the E-plane, the tightened spacing could be otherwise implemented, but there must be room reserved for the phase shifter integration.

The consequence of the large spacing is a reduced beam-steering range as the grating lobes appear earlier to the visible region when the antenna beam is steered. The spacing could be tightened utilizing waveguides with dielectric filling, which allows smaller waveguide dimensions. This has been demonstrated by the authors in [58] for the E-band, but this implementation will increase the losses inside the waveguides as more power is dissipated in the dielectrics. Ridged waveguides are also known to have a lower cut-off frequency compared to that of the rectangular waveguide of the same width and might thus enable tighter spacing [59]. However, the feeding network and the transition structure between the waveguide and PCB might be more challenging to design and realize.

It should be also noted, that the both E-band and 28-GHz antennas are capable of wide-band performance; the feeding networks, transitions, and antenna elements have been designed to support this feature. However, the radiation results are only shown for point frequencies because the element phasings are optimized for fixed frequencies as well. Especially with E-band antenna, the phase shifters were initially designed to demonstrate the beam-steering capabilities at 73.5 GHz. Due to a manufacturing error, the center frequency performance is observed at slightly lower frequency of 72.5 GHz, and hence the beam-steering results are presented at that frequency. With 28-GHz antenna, the phasing is optimized for that frequency, which is the center frequency of the observed band. When the antenna pattern is observed at other frequencies, the main beam steers away from the wanted direction and does not remain
fully coherent. Additionally, low cross-polarization levels have been observed in both antennas, as expected with linearly polarized horn antennas. However, the cross-pol performance naturally decreases in beam-steering, especially with diagonal angles.

Furthermore, there are small design and manufacturing flaws in both antennas that should be addressed in the later designs. The diagnosis of the E-band antenna described in [III] shows that the power division network is not working as intended over the whole E-band. For the antenna presented in [II], the integration of the phase shifter has shortcomings, as described above. However, the error sources are known, and the problems can be fixed with relative ease in the future versions of the proposed antennas. Discrepancies between the simulations and measurements have already been studied with several revised simulation models, and those have indicated the possible areas that require improvement. Nevertheless, further improvements could be still made, for example by introducing surface roughness to the simulation models.

A final comment on the performance of the proposed waveguide-based antennas is related to the utilized electronics. When the presented antennas were first proposed, most of the commercial electronic chips for 5G frequencies had only a single channel. As a result, every array element has been designed to include their own component. However, the industry has since focused on multi-channel chips [60]. Therefore, some re-designing should be undertaken if those components are still to be used. Even despite this, the antennas in [I] and [II] have proven to be potential solutions for realizing mm-wave antenna arrays.

### 3.3 Phased array diagnostics

Holography data are very useful for characterizing the single elements in phased arrays. They are especially convenient for locating the faulty elements in the arrays, as described by the authors in [61]. For example, in [II], the holography data has been evaluated to locate the defective antenna element in the lower left corner of the 28-GHz phased array.

When the E-band antenna presented in [I] was first measured, its performance at some frequencies was not what the simulations suggested. This inspired research on the E-band antenna diagnostics and resulted in [III]. Individual element performance is evaluated with each PCB assembly that generates the varying steering angle. Common factors between different PCBs (14 in total), possible frequency relations, and the proper operations of the single fixed phase shifters have been observed.

As explained in Section 2.4.3, the field results at the antenna aperture can be achieved from the measured near-field data through back propagation. Fig. 3.13 shows the field intensity results, that is, normalized amplitudes at different frequencies when the array beam is directed to broadside. Normalization against the reference horn antenna is done for each frequency separately. Individual
elements have been visually identified and localized. It is already clear from this small sample that the power flow through the elements depends on the observed frequency.

The holography data are used to calculate the weighted average amplitudes and phases of each element in every PCB. Weighting is achieved by applying a two-dimensional, tapered cosine window (Tukey window) to the complex holography data points inside the evaluated element. The average amplitude data are used to determine whether the element is working at the measured frequency for that particular PCB, the threshold being $-10$ dB compared to the reference horn antenna amplitude. Similarly, the average phase data are used to verify that the data windowing is done correctly by calculating the radiation pattern through the array factor with the obtained average phase results. The array factor calculations are compared to the measured results that are in good agreement. Simultaneously, the average phase results indicate that the phase shifters work as intended.

Fig. 3.14 illustrates the individual element working rate over the frequency. Results show a clear frequency dependence between the elements at the same position in the secondary power divider, namely between the elements 4, 8, 12, and 16. It has been concluded that the errors in manufacturing and design
are the likely causes for the observed performance and are why the secondary power divider does not work as intended over the whole E-band. No particular problems in phase shifter performance have been detected.

### 3.4 Phased array performance optimization

Phased arrays are susceptible for beam-pointing errors [62], and quite often the calibration is applied to correct the phase error in order to increase the gain [63], [64]. However, sometimes an imbalance of the phase shifter losses can also be the contributing factor for lower gain [65]. In [II], a phase calibration has been conducted for the presented antenna, whose realized gain was improved by 0.6 dB as the result of the calibration. This inspired research for more effective beam optimization for the 28-GHz antenna, since it was seen as a potential way to improve the phased-array performance. The work resulted in [IV]. Two optimization methods presented in [IV] have different objectives: the first method calculates the maximum realized gain towards a wanted steering angle, while the second aims to decrease the side-lobe level. Experimental studies to maximize the gain have been presented before [66], but as with [IV], the methods are often rather application specific. Each element in the 28-GHz phased array has been individually measured, and the acquired amplitude and phase results are utilized in the presented optimization schemes.

Fig. 3.15(a) shows the measured amplitudes and phases for element 6 at 28 GHz. Each result corresponds to one out of 32 phase shifter states. There is a large amplitude variation between the states due to the phase shifter integration design, which is also reflected by the colors in the plot. The maximum gain is achieved when a single element in the array provides the highest possible
amplitude with constructive phasing in the desired direction. The element phasing through beam-steering dictates what phase shift value is desired in the element. The phase states are projected to this phase shift, implying how large amplitude contributions they add to the given phasing. This is mathematically expressed for every phase state $i$ (indexing from 1 to 32) in

$$A_i(\alpha) = \hat{u}(\alpha) \cdot a_i(\beta_i) = a_i(\sin(\alpha)\sin(\beta_i) + \cos(\alpha)\cos(\beta_i)), \quad (3.1)$$

where $A_i$ is the projected amplitude and $\alpha$ is the desired phase shift from the element. $\hat{u}$ is a unit vector towards the desired phase shift on a unit circle, and $a_i$ is a vector that implies the amplitude $a_i$ corresponding the phasing $\beta_i$ that is provided by the phase state $i$.

Fig. 3.15(b) illustrates an example of the projection when the optimization is done towards $\alpha = 45^\circ$. The state giving the highest amplitude projection is used for that phase shift. In addition, the required element-wise phase shift $\alpha$ changes when the phase of the reference element is varied, because the phasing in the array is progressive. There are hence multiple phase state configurations for a single steering angle, and the one giving the maximum gain is chosen. Fig. 3.16 presents the flow chart describing the maximum gain optimization method in detail.

Large amplitude variation also allows for the use of amplitude tapering schemes to reduce the side-lobe levels. In the algorithm developed to reduce the side-lobes, the elements are set to follow a Taylor distribution while providing the best gain towards the desired steering angle. Results of this optimization method, as well as the maximum gain method, are presented in Fig. 3.17 for different azimuth steering angles. Both optimization methods improve the performance of the phased array in the ways that they are intended to. By choosing the correct phase state configurations, the gain can be increased, or the side-lobes can be decreased. However, the optimization mainly works in the azimuth plane. The elevation corresponds to the E-plane, and due to the stronger inter-element coupling there, the individually measured element amplitudes and phases used in optimization are not the same that occur when the beam is steered in that plane.
**Figure 3.15.** Measured amplitude and phase responses of the single element in the 28-GHz phased array. (a) Phase shifter states described in unit circle. Colors in the plot are used to visualize the obtained amplitude levels. (b) Example of the maximum gain optimization and how the states are projected toward the required steering angle. In this example, radiation is optimized for $45^\circ$. Red color indicates the chosen state, blue is reserved for the others [IV] © 2020 EurAAP.

**Figure 3.16.** Block diagram describing the maximum gain optimization method [IV] © 2020 EurAAP.
Figure 3.17. Comparison of the optimization methods to the performance of the 28 GHz phased array in several directions: (a) azimuth pattern in broadside, (b) elevation pattern in broadside, (c) azimuth pattern in 10°, (d) azimuth pattern in −10°, (e) azimuth pattern in 20°, and (f) azimuth pattern in −20° [IV] © 2020 EurAAP.
4. Imaging with Millimeter-Wave Phased Array

The mm-wave phased arrays developed for modern communication systems [37], [38], as well as the ones presented in [I] and [II], have similar architectures and operation principles as state-of-the-art radar systems [67]–[69]. Furthermore, the communication and radar antennas can potentially share the spectrum, and several joint operation schemes have already been proposed [70], [71]. In addition, the mm-wave frequencies allow imaging with sufficient spatial resolution due to their small wavelengths. Therefore, using communication antennas in sensing is a viable option, and this opens up an interesting research path that is investigated in the following.

Computational imaging schemes, such as using the frequency diversity of the non-conformal beams, offer benefits over the traditional phased array sensing applications, mainly due to their imaging speed, flexibility, and accuracy [72]. This chapter presents a frequency-diverse imaging method for phased arrays, which is tested with the antenna originally discussed in [II]. This topic is summarized in Publications [V] and [VI], which are part of the major contribution of the thesis. As explained in the previous chapter, the phased array used in the imaging was initially developed for a 5G base station. This chapter explains how the modern phased arrays intended for communications can also be utilized in sensing applications.

Section 4.1 gives a brief overview of the imaging method and the related previous work. Section 4.2 describes the mathematical foundation of the image reconstruction algorithm, and Section 4.3 presents the reconstructed images based on the actual measurements.

4.1 Millimeter-wave imaging with phased arrays

The traditional way of sensing with antennas that have steerable beams is to use the generated pencil beams to scan the target space [73]. However, this can be a rather time-consuming way to reconstruct an image. Each pixel of the target space has to be scanned and measured separately, and even with the electrical beam steering, this can be a relatively slow and arduous task.
A frequency-diverse imaging method offers an alternative way of imaging and can potentially enable a real-time image reconstruction. The method is based on illuminating the whole target space with non-focused beams and measuring the responses with frequency sweeps. Less beams need to be measured than with the traditional phased-array scanning, hence allowing faster imaging. Frequency-diverse systems typically require wide operation bandwidth to include the required information for the image reconstruction. Pioneering work on the wide-band mm-wave imaging has been reported for example in [74], [75].

Frequency-diverse imaging has previously been studied and conducted with frequency-sensitive metasurfaces, notably by Smith et al. [76]–[80]. The option to use the imaging method with phased arrays is also discussed, for example, in [76], but [V] and [VI] present the first advances regarding the topic. Publication [V] is a simulation study of the frequency-diverse imaging method in the 36–44 GHz range: the aim is to test the proposed method for phased arrays using an ideal antenna. Publication [VI] shows the first results of the proposed imaging method when it is tested in practice. Studied frequency range in [VI] is 24–32 GHz.

Fig. 4.1 illustrates the concept of the frequency-diverse imaging method for the phased arrays: the method is driven by the idea that a large phased array can be divided into subarrays, and each subarray has its unique, non-focused radiation pattern. These subarrays will then sequentially illuminate the target space, and the other subarrays will sample the reflected wavefront in parallel. The image is then computationally reconstructed from the subarray observations; an image-reconstruction algorithm has been created for this purpose. The frequency-diverse patterns are required to bring enough diversity to the observations, thus allowing faster illumination of the target space and also enabling better spatial resolution.

Additionally, the frequency-diverse imaging method has other benefits as well. The state-of-the-art radars are active electronically scanned arrays, (AESAs), and they have element-wise transceivers in place of the phase shifters when
compared to the regular phased arrays [81], [82]. The transceivers are one of the most expensive components in AESAs, and reducing the number of them could allow great savings in manufacturing costs. With the proposed imaging method and the subarray division related to it, this becomes possible; each subarray would require only one transceiver that would serve several elements.

Finally, MIMO radars are worth mentioning regarding the proposed imaging method. MIMO radars, such as [83], create spatial diversity for the system through different waveforms, each of which is addressed to the distinctive probing signal. The orthogonal waveforms thus increase the spatial resolution of the MIMO radar. In a sense, the frequency-diverse imaging system works quite similarly, but the spatial resolution improvement is acquired through the beam-diversity of the utilized imaging antennas. This eases the requirements for the transceivers because there is no need to record and separate different waveforms from each other.

4.2 Frequency-diverse imaging algorithm

Probably the best-known computational imaging application in the radio spectrum is a synthetic aperture radar (SAR) [84]. Like the basic SAR algorithm, the frequency-diverse imaging method relies on solving the image by using multiple different echoes from the target. SAR image reconstruction has been studied extensively [85]–[87], but there are also many other disciplines demonstrating advanced ways to solve target locations through clever signal processing. For example, the authors in [88] describe a sophisticated algorithm for efficient brain source localization that is used in medical imaging.

However, in the simplest approach, the image can be reconstructed through a linear inverse problem. This approach has been taken in [V] and [VI] since the idea is to discern whether the frequency-diverse imaging method is applicable with phased arrays. The first step of the image reconstruction is to formulate a signal model for the problem.

Equation (4.1) describes a signal model for one Tx-Rx pair, referring to the situation illustrated in Fig. 4.1. The signal model works under the assumption, that the Tx and Rx are placed closely together, so the situation can be assumed as monostatic. One observation $y_{mn}$ for the Tx-Rx pair includes reflections from all $N$ voxels that lie in the target space.

$$y_{mn}(f_k) = \sum_{i=1}^{N} (E_n(\theta_i, \phi_i, f_k) \cdot E_m(\theta_i, \phi_i, f_k)) \Gamma_i \frac{e^{-j4\pi f_k r_i}}{(4\pi r_i)^2}. \quad (4.1)$$

The target space is sampled at several frequencies for every Tx-Rx pair. All these observations are collected to the observation vector $O$, which has a length of $M$ after the number of Tx-Rx pairs and frequency samples. The vector can be further expressed as a product of the field data matrix $A$ and the target data vector $T$, as shown in (4.2). $A$ and $T$ are constructed for every voxel in the target
space (identified with index $i$), thus their size is affected by the mesh applied to the imaged volume.

$$\mathbf{O}(y_{mn}(f_k)) = \mathbf{A}(P_{mn}(\theta_i, \phi_i, f_k), r_i)\mathbf{T}(\Gamma_i).$$

(4.2)

$\mathbf{A}$ also contains the electric field values projected towards the target space from each Tx-Rx pair at every frequency sample there is an observation from. These are obtained with simulations or measurements before the imaging. A single element of the field data matrix $\mathbf{A}$ can be postulated as

$$A_{m,n,i,k} = P_{mn}(\theta_i, \phi_i, f_k) e^{\frac{j4\pi f_k r_i}{c}}. (4.3)$$

where

$$P_{mn}(\theta_i, \phi_i, f_k) = E_n(\theta_i, \phi_i, f_k) \cdot E_m(\theta_i, \phi_i, f_k).$$

(4.4)

$\mathbf{A}$ is hence arranged as $M \times N$-matrix, where $N$ indicates the total number of voxels, and $M$ is the total number of observations/measurement modes, that is, a product of different pattern pairs and frequency samples.

The actual unknowns in the sensing are the reflection coefficients of each voxel $\Gamma_i$. These are collected in the target data vector $\mathbf{T}$, and in order to solve them, (4.2) is rearranged to

$$\mathbf{T} = \mathbf{A}^+ \mathbf{O}. (4.5)$$

The aforementioned equation is the linear inverse problem that is required to be solved, and the image is then reconstructed from the obtained $\mathbf{T}$. This has been shown to work with ideal simulations in [V]. The simulation model in [V] has been constructed around an ideal array with 8 subarrays, and the observed environment is also ideal, the effect of the noise is hence left out of the study. However, the solution is not as straightforward in practice. Measurement uncertainties and noise are present in the real-life observations, and they disturb the image reconstruction enough that a more complex, iterative approach must be taken.

The two-step iterative shrinkage/thresholding algorithm (TwIST) [89] has shown to be beneficial to microwave imaging in the K-band (18–27 GHz) [77, 90, 91]. Since the algorithm is open-source material, its iterative algorithm, along with the monotonicity requirement for the convergence, have been used as inspiration to improve the image quality.

At its core, the utilized algorithm is a least-squares iterative method that is used to minimize the following optimization function

$$f(\mathbf{T}) = \frac{1}{2} || \mathbf{O} - \mathbf{A} \mathbf{T} ||^2. (4.6)$$

In more sophisticated algorithms such as TwIST, there is typically a regularizer at the end of (4.6) for further noise reduction. However, it is left out in the current algorithm that aims for simplicity.
The iterative steps are as follows; first a minimizer \( G \) is calculated for the initial value of the target data vector \( T \) as

\[
G = A^T (O - AT) \tag{4.7}
\]

Minimizer is then used to calculate the next iteration of the \( T_{t+1} \):

\[
T_{t+1} = T_t + \frac{G}{v} \tag{4.8}
\]

where the constant \( v \) must be chosen so that

\[
f(T_{t+1}) < f(T_t) \tag{4.9}
\]

The constant \( v \) is real and it is initialized as 1. This algorithm has been able to produce results for real-life scenarios quite effectively even in this simplified form, as shown in the following section.

### 4.3 Experimental results in \( K_a \)-band

The frequency-diverse imaging method for phased arrays has been tested in practice using two antennas, those discussed in [II]. The results of this test are presented in [VI]. As explained in Section 4.1, the suggested imaging concept involves a division of the larger array to the subarrays, which are driven separately. However, the experimental imaging setup does not include subarrays: those are virtually modeled with the two measurement antennas with varying beam-patterns. One of the antennas is used as Tx, the other as Rx.

Fig. 4.2 shows both the illustration and the actual photograph of the imaging system setup that has been used to verify the performance of the method. For this practical setup, the signal model presented in (4.1) must be updated to take into account that the two imaging antennas are spatially separated. The effect of this update is subsequently seen in the construction of the field data matrix \( A \).

A metallic sphere is used as the imaged object, and its location is varied in the target space using a sled that is normally reserved for NF-scanner probes. This allows for the precise positioning of the object. The antenna beams are controlled with a computer, and the same computer is used to reconstruct the images from the samples recorded by a VNA. Over the frequency band from 24 GHz to 32 GHz, 161 frequency samples are obtained with 36 different pattern pairs, thus totaling 5796 observations. The frequency sweep is done for each pattern pair separately before switching the antenna pattern configuration. The drifting of the VNA is assumed to be small between different pattern pairs.

The different patterns are obtained by using different phase shifter configurations, that is, different phase states in the phased array elements as discussed in the previous chapter and in [II]. Both Tx and Rx antenna have each 6 unique patterns, thus the total combination of possible pattern pairs is 36. The patterns
have been chosen so that together they illuminate the whole target plane, with emphasis that individual patterns provide nulls and peaks in different locations at the same frequency. The different beam patterns have been manually and tested with simulations for this first practical study. This is done to if arbitrarily chosen patterns can provide the required information for the image reconstruction.

The field data matrix $A$ is constructed from the different pattern pairs that have been measured together, while also taking into account the possible mutual coupling between the Tx and Rx antennas. However, the measured phase of the formed convolution pattern is not considered, only the phase due to the pixel location is. The measurement system does not allow calibration and might be a cause for phase uncertainty that has been found to affect the image reconstruction.

However, the setup did not allow moving the object in $z$ direction, and the analysis is hence conducted for the 2D $xy$-plane. It should be also pointed out that the imaging range is quite modest at $\approx 0.5$ m. The range in the test setup has been limited primarily by the dynamic range of the measurement equipment.

The range and cross-range resolution estimations, $\delta_r$ and $\delta_{cr}$, respectively, are calculated according to (4.10) and (4.11) [92]. Although the range component is not evaluated, the wide-band imaging antennas suggest that the range resolution of 20 mm could be achieved with 8 GHz band. In addition, the simulation studies conducted in [V] have already proposed that the wide band could reduce the error in range location. The cross-range resolution for the system is suggested to be 70 mm according to (4.11) at 28 GHz. The distance to the imaging plane $d$ is 410 mm, and the transmitting and receiving antennas have the same square aperture of $31.5 \times 31.5$ mm$^2$, hence $L_{\text{Tx}} = L_{\text{Rx}} = 31.5$ mm.

$$\delta_r = \frac{c}{2B}.$$  

(4.10)
\[ \delta_{cr} = \frac{\lambda d}{L_{\text{Tx}} + L_{\text{Rx}}} \]  \hspace{1cm} (4.11)

Fig. 4.3 shows the reconstructed images of the sphere alongside the actual locations of the object. The sphere is correctly located in several different places inside the target space. The current algorithm has shown to work quite well, although the size of the object appears slightly larger than it actually is. The 3-dB width of the imaged object averages around 60 mm against the actual size of 30 mm. However, the computed size is more in line with the resolution estimation and hence seems to be appropriate. It is likely that even a point target would appear close to the resolution estimation in both range and cross-range directions.

Ultimately, the study in [VI] has shown that the suggested frequency-diverse imaging method works in practice with phased arrays. The finding also paves the way for several future studies: the imaging algorithm can certainly be improved, and the image reconstruction offers great possibility to test other image-enhancing algorithms that already exist, for example, for SAR [93]–[95].
Multiple targets, dielectric objects, and resolution targets could be used to further characterize the imaging system or they could act as test objects when the system is improved.

Another important research question relates to the optimization of the antenna beams, namely how to increase the diversity of the observations through different beam patterns. The diversity can also be achieved with antennas having multiple polarizations. Existing two-polarized phased arrays intended for communications, such as [47]–[49], could be used in the imaging system to test this proposal. Finally, neural networks could also allow more accurate and efficient imaging, as shown by the authors in [96]. Since the optimization problem is extremely complex, the machine learning approach might prove to be the most beneficial one. The machine learning techniques could be utilized to determine the most useful beam patterns for the image reconstruction or even to replace the reconstruction algorithm altogether. Neural networks could be taught to locate the objects straight from the measurement data.
5. Electrically Invisible Antenna

The volume or area reserved for antennas can be limited in some applications. This can be overcome by co-locating antennas into a shared volume, but this often leads to problems in antenna performance. In mobile devices, this is a common problem that has recently been investigated in great depth [97]–[99].

This chapter presents design principles to realize an electrically invisible antenna and the verified performance of such antenna configuration. Electrical invisibility can be defined in multiple ways, and, in this thesis, it is described as low scattering. The design steps and the antenna are presented in Publication [VII], and they are part of the main contribution of this thesis.

First, Section 5.1 presents an overview of the low-scattering antennas. In Section 5.2, the design considerations for an electrically transparent antenna are discussed. Finally, Section 5.3 describes the antenna performance and the invisibility of the proposed antenna.

5.1 Low-scattering antennas

Low-scattering antennas have been widely researched, and there are two straightforward ways to reduce the antenna scattering. The first method is to modify the antenna structure in order to lower scattering. This is demonstrated by the authors in [100] for an octagonal patch antenna. This method often requires the subtraction of the metal areas in the reshaped antenna.

The second method to affect scattering is achieved by loading [101]–[104]. This is likely the most common way to make the antennas appear transparent, and several examples can be found in the literature [105]–[108]. These solutions aim to become minimum-scattering antennas, which are ideal antennas that achieve invisibility with reactive loading in the antenna port [109]. A special case where the invisibility is achieved with an open port is called a canonical minimum-scattering antenna [110].

The most recent development in low-scattering antennas has seen metamaterial cloaks and other frequency-selective realizations [111]–[114]. However, these studies include whole antenna systems, and their design principles are
not applicable for configurations where each component is arranged individually. Furthermore, these realizations can be highly complex to manufacture.

One characteristic of the aforementioned antennas is that they are typically invisible only at their operation frequencies. If the antennas support invisibility at other frequencies, they must shift between the invisible and the radiating state. The presented antennas cannot simultaneously be in both.

Many applications could benefit from the antennas capable of dual-band performance, namely those that are invisible in one band while radiating in the other. The antenna presented in [115] has addressed this design challenge. Nevertheless, the investigation is rather scarce and leaves room for improvement.

Publication [VII] has proposed an improved solution. It introduces clear steps to design an invisible dipole antenna element and presents analysis on how invisibility affects the dipole performance at its nominal frequency. The antenna presented in [VII] has been designed to operate in L-band (1–2 GHz), while the scattering is minimized in X-band (8–12 GHz). One application for this type of dipole is in secondary surveillance radars (SSRs) that are used in aircrafts in particular. However, the design principles of the invisible antenna are presented to be applicable in a more general situation, thus making them more valuable for the antenna community.

5.2 Design of a low-scattering antenna

Compared to the mobile devices in which the design paradigm is often to make sure that the high-band mm-wave antennas do not affect the performance of the low-band sub-6 GHz antennas, the approach presented in [VII] is the opposite. The objective is to ensure that the operation of the high-band (HB) antenna is not disturbed by the low-band (LB) antenna.

The design principles proposed in [VII] are simple: first, the LB antenna is chopped into pieces that minimize the structural scattering at the HB, thus ensuring that the chopped parts appear in the Rayleigh scattering region. Second, the chopped pieces are connected together using optimal loads that minimize the scattering from the formed entity. Effectively, the antenna appears uniform at the LB but still discrete at the HB. The following sections explain in detail how these principles are realized in practice to design the low-scattering dipole presented in [VII].

5.2.1 Minimizing the structural scattering

To give a more general perspective, the structural scattering is first evaluated with simulations for an ideal wire that is made from the PEC for simplicity. Fig. 5.1(a) shows the scattering at 10 GHz as a function of the wire length. The wire is illuminated with a linearly polarized plane wave whose electric field is aligned with the wire. The scattering is evaluated with the total RCS, and the RCS is
Figure 5.1. (a) The normalized total RCS of a PEC wire as a function of the wire length \(l\) at 10 GHz. (b) Total RCS of two PEC wires as a function of the gap width \(d\) between the wires at 10 GHz. The radius of each wire in (a) and (b) is 0.5 mm. Taken from [VII].

normalized to the line length and radius (mm\(^2\)/(mm × mm)) in order to ensure that the RCS values of different lengths are comparable with each other. This way, for example, the resonance effects are visible in the evaluation. The results indicate that the single wire length should be less than 10 mm. Below that limit, the wire is assumed to be in the Rayleigh scattering region at 10 GHz.

The next step is to identify the suitable gap between the adjacent wire parts. Two wires of equal length are placed one after another, and the gap width between them is varied. There is coupling between the wires when they are placed close together, and this affects the electrical length of the single wire part, thus resulting in more scattering than what the sum of two independent wires would otherwise suggest. The results for different wire lengths as a function of the gap width are presented in Fig. 5.1(b). It has been further concluded that the wire length should be 8 mm (0.267\(λ\) at 10 GHz) at most, and the gap width should be at least 2 mm (0.067\(λ\) at 10 GHz) to ensure low structural scattering. These are good compromises for practical realizations: increasing the gap width has little effect beyond the 2-mm point, and the wire cannot be cut to indefinitely small pieces. Moreover, shorter wire lengths lead to an increased number of loads in the final structure, assuming that the original size of the dipole remains the same. This translates to a more complex antenna structure and a potential decrease in the antenna performance. In addition, the larger wire length leaves more room for adjustment. It is later shown that the load to connect the wires should be inductive. When the dipole is inductively loaded, it becomes electrically longer at 1 GHz, shifting the resonance down in frequency. This can be easily compensated by shortening the wire parts. The compensation could be achieved by simply reducing the number of chopped parts, but the adjustment through wire length is more straightforward and enables continuous tuning of the resonance frequency.
5.2.2 Load optimization

The second-most important step in designing the low-scattering dipole is to choose the proper load between the wires to minimize the scattering. The scattered fields and the load impedances connecting the adjacent wire parts should be related, which is not a simple formulation. Harrington and Mautz have described the relation through the theory of characteristic modes [116], and they have presented how the optimization can be achieved for maximum scattering [117]. Furthermore, Hirasawa has demonstrated in [118] how passive loading can be used to reduce the radar cross section of a wire, and he has conducted the analysis with the method of moments (MoM). However, the solutions, while sophisticated, are already quite complex for simple, ideal structures, and they do not consider the reduction of the structural scattering. More realistic antenna structures can be difficult to optimize with the proposed methods.

The EM simulation model presented in Fig. 5.2(a) has been created in this thesis for more straightforward approach. The antenna structure with N-ports is placed in the middle of a unit cell that has two Floquet ports on opposite sides of the bounding box. The other sides of the unit cell have periodic boundary conditions. The Floquet ports are used to analyze the power transfer through the structure, while the ports on the wire are used to load the antenna structure. Since there are no ohmic losses, maximizing the transmitted power past the wire and minimizing the scattering from the wire are identical goals. A relationship can be established between the transferred power and the load impedances, and this allows the transformation of the initial EM problem into the circuit problem. The circuit model of the optimization problem with the loads connected to the ports on the wire is illustrated in Fig. 5.2(b).

The S-parameter presentation of the initial simulation model $S_{ini}$ relates all...
the ports to each other, and it can be written with sub-matrices as follows:

\[
S_{\text{ini}} = \begin{bmatrix}
S_{FF} & S_{FD} \\
S_{DF} & S_{DD}
\end{bmatrix}.
\] (5.1)

Here \( S_{FF} \) and \( S_{DD} \) are the coupling terms between the Floquet ports and the ports on the wire, respectively. Similarly, \( S_{FD} \) describes the coupling from the Floquet ports to the ports on the wire, and \( S_{DF} \) describes the inverse relationship.

The loads in different positions on the wire are expressed in the matrix \( S_L \). The same load, with reactance \( X \), is assumed in every port to simplify the optimization, but the following formulation is relevant for more complex systems as well. The \( N \times N \) load matrix is written as follows:

\[
S_L = \begin{bmatrix}
\frac{jX-Z_0}{jX+Z_0} & 0 & \cdots & 0 \\
0 & \ddots & \ddots & 0 \\
0 & \cdots & \frac{jX-Z_0}{jX+Z_0}
\end{bmatrix}.
\] (5.2)

The earlier sub-matrices and load matrix are then used to reduce the initial \((N+2)\)-port matrix \( S_{\text{ini}} \) to a two-port system model \( S_{\text{sys}} \) with the following equation [119]:

\[
S_{\text{sys}} = S_{FF} + S_{FD}S_L(I - S_{DD}S_L)^{-1}S_{DF}.
\] (5.3)

The elements in \( S_{\text{sys}} \) are functions of load reactance \( X \). The minimum-scattering requirement is then achieved by maximizing the transmission coefficient \( S_{\text{sys,21}} \) with the proper load termination. The sign of \( X \) decides whether the load should be inductive or capacitive.

The use of the presented S-parameter model in the antenna design is a novel approach, but the same formulation also has benefits in the antenna characterization. The authors in [120] describe how a measurement setup in polarimetric RCS measurements can be similarly divided to sub-matrices; with different known loads in the AUT input, its characteristics are found utilizing the same S-parameter formulation.

### 5.3 Chopped dipole with inductive loading

The design methods presented in Section 5.2 have been used in [VII] to realize practical models of low-scattering dipoles. The manufactured antennas have been etched on a PCB, and their performance is compared to a half-wave dipole whose nominal frequency is at 1 GHz. The differences in radiation performance at 1 GHz and the scattering at 10 GHz have been recorded and analyzed.

The manufactured prototypes are shown in Fig. 5.3. The chopped dipole parts are placed 2 mm apart, and they are connected with the load inductances. The optimal reactance value is 367 \( \Omega \) at 10 GHz for the chopped copper line on a 0.25-mm thick RO4350B substrate \((\varepsilon_r = 3.48, \tan\delta = 0.0037)\). The width of the line is
Electrically Invisible Antenna

Figure 5.3. Manufactured prototypes with meandered lines as load inductances: (a) a 2-meander inductance and (b) a 4-meander inductance. Taken from [VII].

1 mm. The positive reactance value translates to the inductance of 5.8 nH, which can be realized on a PCB with a narrow, meandered line. Two meander-line implementations are presented in [VII], and they have been named 2-meander and 4-meander inductances after the number of turns in the inductance line. The lengths of the inductance lines are 9.8 mm and 12.2 mm for the 2- and 4-meander lines, respectively, when the width of the line is 90 μm. There is likely more parasitic capacitance in the 4-meander realization, hence the longer length.

Length of a single wire part is 4.75 mm in the 2-meander model and 5.0 mm in the 4-meander model. As explained above, the inductive loading of a half-wave dipole shifts the resonance down in frequency from the nominal one. This must hence be compensated for by shortening the length of a single wire to maintain the radiation at the original frequency; in the case of [VII], at 1 GHz.

5.3.1 Radiation performance

The antenna performance at 1 GHz is evaluated by examining the impedance matching and the total efficiency. The inductive loading affects the input impedance of the dipole. The simulated impedances for the 2- and 4-meander models at 1 GHz are 35.7 Ω and 38.5 Ω, respectively. Fig. 5.4(a) illustrates the simulated and measured reflection coefficients of the meander-line dipoles and the reference half-wave dipole. The results indicate that the antennas have
resonance at 1 GHz, and the matching for meander-line dipoles is better than −8 dB. The reference dipole is better matched, and this can be explained by the commercial Johnson 1720BL15B0050E balun used in the feeds. When a dipole is connected to an unbalanced feed, such as a coaxial line, a balun is required to make a transition from the unbalanced lines to the balanced lines that the dipole is constructed from. Most commercial baluns are designed with half-wave dipoles in mind, and matching to these can hence be expected to be better than to those that have an adjusted input impedance. This is evident in the results shown in Fig. 5.4(a).

The effect on radiation efficiency due to the antenna loading is particularly interesting and is the most suitable figure of merit to benchmark the meander-line dipoles against the normal half-wave dipole. The radiation efficiency is embedded in the total efficiency, which can be easily measured. Fig. 5.4(b) illustrates the total efficiency of the reference and the designed structures and how they compare with the simulations. The agreement between the simulated and measured results is excellent, and the total efficiency of the meander-line models is around −1.6 dB. The decrease in comparison to the reference half-wave dipole is 1 dB. Part of the decrease is explained by the poorer matching, and the rest is due to the lower radiation efficiency. According to the simulations, the deterioration of radiation efficiency is 0.6 dB for the loaded dipoles. It has also been noted that the feeding affects the radiation efficiency; if the antenna elements only are considered then the decrease is 0.4 dB. It has been speculated that the feed affects the current distribution in the dipole, subsequently decreasing the radiation efficiency. The reduced efficiency is the only clear weakness of the low-scattering dipole. In the following, it is shown that this trade-off is more than acceptable as the improvement in antenna invisibility is so remarkable.

Figure 5.4. Comparison between measured and simulated results for a chopped dipole: (a) reflection coefficient and (b) total efficiency. Taken from [VII].
Figure 5.5. Measurement setup for the RCS inside the anechoic chamber. The AUT is in the middle. Illumination from 0°. (a) Top view illustration showing the measured bistatic angles and (b) photograph of the illustrated setup. Taken from [VII].

5.3.2 Scattering cross section

The invisibility of the low-scattering antennas and the reference dipole are evaluated through RCSs. The feeds of the dipoles are represented by the 50-Ω loads as the interest is in the RCSs of the antenna elements. Fig. 5.5(a) depicts the top view of the measurement setup with the measurement antennas placed to record the bistatic RCS response at 315°. Fig. 5.5(b) is the photograph of the same setup. The Tx antenna illuminates the AUT placed in the middle of an anechoic chamber. The Rx antenna is moved around the AUT, and it records the scattered signal. This has been conducted with 22.5° steps. The measurement at 0° gives the (quasi-)monostatic RCS result, while the other angles give bistatic responses.

The free-space results have been measured before and after every AUT measurement, and their average (base-line result) has been removed from the recorded Rx signal. The signal has also been time-gated to ensure that the observation only considered the scattered power. The scattering from the AUT has been identified in the time domain signal while considering the travel-time of the measurement signal and the distances between the Tx and Rx and the
AUT. The applied time-gate window is not constant for every measured angle, and it must be varied to consider the changing measurement setup and the Rx location with respect to the Tx antenna. Especially with the forward bistatic angles from 90° to 270°, the line-of-sight path between the Tx and Rx antenna is evident, causing a strong component in the measured signal. The RCS values for each angle are calculated using the bistatic radar equation, and the results at 10 GHz are collected in Fig. 5.6(a) alongside the simulated ones.

The simulations and measurements are in excellent agreement and the meander-line dipoles show more than 10 dB lower RCS values at every measured point. The maximum RCS values over 4–18 GHz are shown in Fig. 5.6(b) to present the wide-band performance. The scattering is reduced at best by 15 dB, and the 10-dB RCS-reduction bandwidth is greater than 3 GHz. Measured and simulated values show good agreement, though the measured values deviate slightly from the simulated ones as they are shifted down in frequency. One reason for this might be that the manufactured inductance is marginally larger than the simulated one, and hence the shift is present. Fortunately, this can be easily corrected by, for example, shortening the inductance line to return the minima right at 10 GHz.

The design methods described in [VII] have proven to be useful when the co-located antenna configurations are considered. The method has been verified with inductively loaded, chopped dipoles that show great improvement in the antenna invisibility, with the trade-off being a minor decrease in the radiation efficiency.
6. Summary of Publications

I. E-Band Beam-Steerable and Scalable Phased Antenna Array for 5G Access Point

This paper introduces a novel, low-loss, waveguide-based phased array suitable for 5G access points. It demonstrates how the PCB-based phase shifters should be integrated with the waveguide-based feeding networks and antenna elements. Due to the lack of suitable electronic phase shifters at E-band, the phase shift required in beam-steering is realized with discrete, microstrip-based, true time-delay lines. The antenna performance has been validated with measurements, and the beam-steering has been demonstrated in the E-, H-, and diagonal planes.

II. Waveguide-Based Phased Array with Integrated Element-Specific Electronics for 28 GHz

The waveguide-based phased array is implemented for 28 GHz in this publication. The feeding network has been improved from that presented in [I], and the beam-steering is realized with electronically controlled phase shifters. Therefore, a more complicated PCB structure for integrating and controlling the phase shifters is presented. The antenna performance has been validated with measurements, and the results show good agreement with the simulations. It has been shown that the antenna structure and transmission lines themselves have low losses, the majority of the insertion loss being due to the phase shifters.

III. Diagnostics of the Phased Array for E-band Using Holography Data

The antenna presented in [I] has been further characterized in this work. The paper describes a method where the holography data from the near-field measurements are utilized to accurately diagnose element-specific phases and amplitudes in phased arrays. The diagnostic evaluation indicates that the phase shifters of the antenna work as intended. However, the feeding network is frequency sensitive and is thus the primary reason for the decreased performance between 75–82 GHz.
IV. Beam Optimization for 28 GHz Phased Array

The antenna presented in [II] has been optimized in this work to enhance the antenna performance. The measured amplitude and phase data of each element are used in two optimization schemes: one aims to achieve maximum gain towards the desired steering angle by choosing the phase shifter states that project the highest element amplitudes there. The purpose of the other method is to achieve lower side lobes by choosing the phase states that produce amplitude tapering in the array according to Taylor distribution. Both methods have resulted in clear improvements in antenna performance for their intended uses.

V. Millimeter-wave Imaging Method based on Frequency-Diverse Subarrays

An imaging method aimed at mm-wave phased arrays is introduced in this paper. The method is based on frequency-diverse subarrays, and the simulation study of the method has been conducted at 36–44 GHz. The performance and accuracy of the imaging method is evaluated, and the results show that the method is a viable option for phased array image reconstruction at mm-waves.

VI. Millimeter-wave Frequency-Diverse Imaging with Phased Array Intended for Communications

This paper extends the work conducted in [V]. The imaging method has been further developed and tested in practice using the antennas presented in [II]. A metallic sphere has been successfully located in the sensed target space. It has been shown that the phased arrays intended for 5G communication applications could also be used for imaging purposes.

VII. Low-Scattering Chopped Dipole for Secondary Surveillance Radar

A novel, chopped-dipole antenna with inductive loading and reduced scattering is presented in this publication. The antenna radiates well in L-band (1–2 GHz), while having a reduced RCS in X-band (8–12 GHz). The straightforward steps to design such an antenna are introduced in this paper, with especial focus on how the dipole structure and periodical load should be optimized for the lower scattering.
7. Conclusions

This thesis focuses on advancing the beam-steerable mm-wave antennas and low-scattering antennas for communication and sensing applications. The main scientific merits of the work are as follows: the development of the beam-steerable phased arrays with low-loss feeding structures, their characterization and use in imaging systems, and the design of low-scattering antennas that are electrically invisible to the radiation of an antenna operating at a higher frequency. The thesis is divided into three main parts, each of which presents the developments regarding the discussed topic.

The first part of the thesis presents the waveguide-based base-station antennas for the upcoming 5G systems that operate at mm-waves. The new proposed antenna designs have been summarized in Publications [I] and [II]. The antennas have demonstrated low insertion losses in their power division networks and suitable beam-steering ranges, which exhibit limitations due to large inter-element spacing.

Publications [III] and [IV] have introduced diagnostic and optimization methods for the waveguide-based phased arrays. These studies provide important knowledge about the antennas and highlight possible ways to improve their designs and optimize their performance. Although the diagnostics and optimization methods have been developed for these specific antennas, they are applicable to the other phased arrays as well.

The second part of the thesis discusses the use of phased arrays in frequency-diverse imaging schemes. The main intention is to describe how antennas that were initially developed for communications can be used for sensing. The work in Publications [V] and [VI] has resulted in the first experimental demonstration of a frequency-diverse phased array imaging setup: the imaging system has shown how a simple target can be found in an observed space by computationally calculating the image from the samples that are created with different Tx-Rx pattern pairs.

The third and final part of the thesis focuses on developing an electrically invisible antenna, namely an antenna that operates at a lower frequency while maintaining low-scattering from high-frequency waves. The work has been summarized in Publication [VII]. Design principles have been described for a
low-scattering dipole, which demonstrates a reduced RCS over a wide band. The dipole is chopped into uniform pieces that are connected to adjacent ones with loads that minimize the scattering from the incident wave. The realized dipole has its application in an aviation radar system as an SSR antenna.

Whilst the presented designs and methods are already appropriate solutions, they offer plenty of room for further studies. The waveguide-based antennas could be re-designed to support multichannel chips, which is the direction in which the industry seems to be heading. Moreover, the fabrication of antennas through plastic-injection molding offers great opportunities to research how the mass-production of mm-wave antennas could be undertaken without sacrificing accuracy in manufacturing. A further important issue to study is the cooling for the active electronics that will be integrated in the future systems.

The study of frequency-diverse imaging for phased arrays has narrowed the gap between mm-wave communications and sensing antennas. However, the imaging method is in its early stages, and there are many ways for improvement. Much can currently be achieved by advancing the image-reconstruction algorithm and utilizing the existing image-enhancing formulas. In addition, the radiation patterns used for imaging require optimization, meaning that lots of work regarding the research topic is yet to be studied.

Finally, the question remains of how the low-scattering antennas will actually affect the performance of the high-band antenna, especially when the beam of the high-band antenna is steered. Moreover, realizing these invisible antennas for mm-waves should provide an interesting path for further study.

Antenna design, especially in regard to mm-waves, will continue to evolve greatly over the following years as requirements become clearer and mm-wave networks for 5G are constructed and standardized. Challenges are likely to arise that require talented engineers and scientists to solve them. Besides, the most progressive studies are already looking beyond 5G and laying the groundwork for 6G networks. Research such as that conducted in this thesis will continue to be relevant for years to come.
References


Development of modern communication networks is a major driving force in current antenna research. Envisioned antennas for future communication and sensing systems have many different requirements like electrical beam steering and electrical transparency, that is, low scattering.

This thesis discusses the advances in the field of communication and sensing antennas by presenting new antenna realizations for beam-steerable phased arrays and for low-scattering loaded dipoles. Diagnostic and optimization tools support the contribution along with imaging studies that close the gap of the joint use of antennas in communications and sensing.