Henna Paaso

DIRECTION OF ARRIVAL ESTIMATION ALGORITHMS FOR LEAKY-WAVE ANTENNAS AND ANTENNA ARRAYS
HENNIA PAASO

DIRECTION OF ARRIVAL ESTIMATION ALGORITHMS FOR LEAKY-WAVE ANTENNAS AND ANTENNA ARRAYS

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Abstract
The focus of this thesis is to study direction of arrival (DoA) estimation algorithms for reconfigurable leaky-wave antennas and advanced antenna arrays. Directional antennas can greatly improve the spectrum reuse, interference avoidance, and object and people localization. DoA estimation algorithms have also been shown to be useful for applications such as positioning for user tracking and location-based services in wireless local area networks (WLANs).

The main goal is to develop novel DoA estimation algorithms for both advanced antenna arrays and composite right/left-handed (CRLH) leaky-wave antennas (LWAs). The thesis introduces novel modifications to existing DoA estimation algorithms and shows how these can be modified for real-time DoA estimation using both antenna types. Three modified DoA estimation algorithms for CRLH-LWAs are presented: 1) modified multiple signal classification (MUSIC), 2) power pattern cross-correlation (PPCC), and 3) adjacent power pattern ratio (APPR). Additionally, the APPR algorithm is also applied to advanced antenna arrays.

The thesis also presents improvements to the modified MUSIC and APPR algorithms. The complexity of the algorithms is reduced by selecting a smaller number of received signals from different directions. The results show that the selection of the radiation patterns is very important and that the proposed algorithms can successfully estimate the DoA, even in a real-world environment. Based on the results, this thesis provides a good starting point for future research of DoA estimation algorithms to enhance the performance of future-generation wireless networks and the accuracy of localization.

Keywords: adjacent power pattern ratio (APPR) algorithm, composite right/left-handed leaky-wave antenna (CRLH-LWA), direction of arrival (DoA) estimation algorithms, directional antennas, multiple signal classification (MUSIC) algorithm, power pattern cross-correlation (PPCC) algorithm
Paaso, Henna, Suunnanestimointialgoritmit vuotoaaltoantenneille ja antenniryhmille.
Oulun yliopiston tutkijakoulu; Oulun yliopisto, Tieto- ja sähköteknikan tiedekunta

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Oulun yliopisto, PL 8000, 90014 Oulun yliopisto

Tiivistelmä

Tässä väitöskirjassa tutkitaan suunnanestimointialgoritmeja uudelleen konfiguroituville vuotoaaltoantenneille (LWA, leaky wave antenna) ja kehittyneille antenniryhmille. Suuntaavilla antennilla voidaan parantaa huomattavasti spektrin uudelleen käyttöä ja esineiden ja ihmisten sijaintipaikkannusta sekä pienentää häiriöitä. Suunnanestimointialgoritmit ovat myös osoittautuneet hyödylliseksi esimerkiksi seuranta- ja sijaintipaikannuspalvelusovelluksille langattomissa lähi-verkoissa.

Työn päätavoite on kehittää uusia suunnanestimointialgoritmeja sekä kehittyneille antenniryhmille että vuotoaaltoantenneille (composite right/left-handed (CRLH) LWA). Työssä osoitetaan, miten olemassa olevia suunnanestimointialgoritmeja voidaan muokata uudella tavalla, jotta ne soveltuisivat molemmissa antennityypeissä realiaikaiseen suunnanestimointiin. Vuotoaaltoantennille on kehitetty kolme erilaista suunnanestimointialgoritmia: 1) muunneltu MUSIC-(multiple signal classification), 2) säteilykyvyn tehojen ristikorrelaatio- (PPCC, power pattern cross correlation) ja 3) vierrekkäisten säteilykyvyn tehosuhtealentalgoritmi (APPR, adjacent power pattern ratio). APPR-algoritmina on myös sovellettu kehittyneelle antenniryhmälle.

Työssä esitetään myös paranneksia muunnelluille MUSIC- ja APPR-algoritmeille. Algoritminen kompleksisuutta voidaan piententää valitsemalla vähemmän vastaanotettuja signaaleja. Tulokset osoittavat, että signaalien valinta on hyvin tärkeää ja ehdotetut algoritmit estimoivat onnistuneesti saapuvan signaalin suunan todellisuudessa mitausympäristössä. Yhteenvetona voidaan sanoa, että tämä väitöstyö on hyvä lähtökohta suunnanestimointialgoritmitutkimukselle, jonka tavoitteena on parantaa tulevien sukupolvien langattomien verkkojen suorituskykyä ja paikannuksen tarkkuutta.

Asiasanat: APPR-algoritmi, MUSIC-algoritmi, PPCC-algoritmi, suunnanestimointialgoritmit, suuntaavat antennit, vuotoaaltoantenni (CRLH-LWA)
Preface

The research work for this thesis was conducted at the VTT Technical Research Centre of Finland Ltd in 2012–2018 and during my one-year research visit at Drexel University. The principal supervisor of this thesis was Professor Aarne Mämmelä, and Professor Jari Iinatti was the other supervisor. The research work was conducted as part of the READS and FUNERA projects, cofunded by Tekes, in 2012–2014 and 2015–2017, and as part of VTT’s internal AWARENESS project. I would like to thank VTT, and particularly my former and present managers, Mr. Kyösti Rautiola, Dr. Jussi Paakkari, Mr. Pekka Rantakari, and Dr. Tauno Viitala-Heikkilä, for providing me with the opportunity to work in various international and national projects and to acquire knowledge of a variety of topics, and ultimately to write and publish this doctoral dissertation. I would also like to thank the EU-funded DENSE project which provided me with the opportunity to apply my DoA estimation research to radars. My thesis has also been supported by scholarships from Jenny and Antti Wihuri Foundation (three times), Ulla Tuominen Foundation, Tauno Tönning, Kaute Foundation, Emil Aaltonen Foundation, and Riitta and Jorma J. Takanen Foundation. They are gratefully acknowledged.

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Valkama, Professor Aarne Mämmelä, Mr. Matti Somersalo, Mr. Mikko Hiivala, Mr. Mika Hoppari, and Mr. Juha Korpi.

I am grateful to my friends outside of work. In particular, I would like to thank my basketball friends, who have played an important role in helping me forget this thesis work altogether every once in a while. I am also grateful to my parents, my brothers, my parents-in-law, and my other close relatives, for their care and support throughout my life. Finally, I would like to express my deepest gratitude to my husband Tuomas for his love, understanding, and support throughout all the years we have shared together, our children, Torsti and Sohvi, for being there and helping me forget work-related matters.

Oulu, September 18, 2018 Henna Paaso
### Abbreviations

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<tr>
<td>AAS</td>
<td>Adaptive array systems</td>
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<tr>
<td>ACI</td>
<td>Adjacent channel interference</td>
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<td>ADD</td>
<td>Addition</td>
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<td>A/D</td>
<td>Analog-to-digital</td>
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<td>AN</td>
<td>Access node</td>
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<td>APPR</td>
<td>Adjacent power pattern ratio</td>
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<tr>
<td>BER</td>
<td>Bit error rate</td>
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<td>BM</td>
<td>Butler matrix</td>
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<td>BS</td>
<td>Base station</td>
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<td>CRLH</td>
<td>Composite right/left-handed</td>
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<td>D/A</td>
<td>Digital-to-analog</td>
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<td>DBF</td>
<td>Digital beamforming</td>
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<tr>
<td>DC</td>
<td>Direct current</td>
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<tr>
<td>DIV</td>
<td>Division</td>
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<td>DML</td>
<td>Deterministic maximum likelihood</td>
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<td>DoA</td>
<td>Direction of arrival</td>
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<tr>
<td>EKF</td>
<td>Extended Kalman filter</td>
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<td>ESPAR</td>
<td>Electronically steerable parasitic array radiator</td>
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<td>ESPRIT</td>
<td>Estimation of signal parameters via rotational invariance techniques</td>
</tr>
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<td>EVD</td>
<td>Eigenvalue decomposition</td>
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<tr>
<td>FFT</td>
<td>Fast Fourier transform</td>
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<tr>
<td>FPGA</td>
<td>Field-programmable gate array</td>
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<td>HFSS</td>
<td>High-frequency structure simulator</td>
</tr>
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<td>HetNet</td>
<td>Heterogenous network</td>
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<td>HPBF</td>
<td>Half-power beamwidth</td>
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<td>HSPA</td>
<td>High-speed packet access</td>
</tr>
<tr>
<td>ICI</td>
<td>Intercarrier interference</td>
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<tr>
<td>IFFT</td>
<td>Inverse fast Fourier transform</td>
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<td>ILA</td>
<td>Integrated lens antenna</td>
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<tr>
<td>IoT</td>
<td>Internet of Things</td>
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<tr>
<td>ISI</td>
<td>Intersymbol interference</td>
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<td>Abbreviation</td>
<td>Description</td>
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<tr>
<td>LH</td>
<td>Left-handed</td>
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<tr>
<td>LMS</td>
<td>Least-mean square</td>
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<tr>
<td>LoS</td>
<td>Line-of-sight</td>
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<td>LTE</td>
<td>Long-Term Evolution</td>
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<tr>
<td>LTE-A</td>
<td>LTE-Advanced</td>
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<tr>
<td>LUT</td>
<td>Look-up table</td>
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<tr>
<td>LWA</td>
<td>Leaky-wave antenna</td>
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<tr>
<td>MIMO</td>
<td>Multiple-input multiple-output</td>
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<tr>
<td>ML</td>
<td>Maximum likelihood</td>
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<td>MMSE</td>
<td>Minimum mean-square error</td>
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<tr>
<td>mm-wave</td>
<td>Milli meter wave</td>
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<tr>
<td>MS</td>
<td>Mobile station</td>
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<tr>
<td>MUL</td>
<td>Multiplication</td>
</tr>
<tr>
<td>MUSIC</td>
<td>Multiple signal classification</td>
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<tr>
<td>MV</td>
<td>Minimum variance</td>
</tr>
<tr>
<td>MVDR</td>
<td>Minimum variance distortionless response</td>
</tr>
<tr>
<td>NLoS</td>
<td>Non-line-of-sight</td>
</tr>
<tr>
<td>OFDM</td>
<td>Orthogonal frequency division multiplexing</td>
</tr>
<tr>
<td>O&amp;M</td>
<td>Operation and maintenance</td>
</tr>
<tr>
<td>PD</td>
<td>Power detector</td>
</tr>
<tr>
<td>PoP</td>
<td>Point of presence</td>
</tr>
<tr>
<td>PPCC</td>
<td>Power pattern cross-correlation</td>
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<tr>
<td>P2MP</td>
<td>Point-to-multipoint</td>
</tr>
<tr>
<td>P2P</td>
<td>Point-to-point</td>
</tr>
<tr>
<td>QoS</td>
<td>Quality of service</td>
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<tr>
<td>RARE</td>
<td>Rank reduction estimation</td>
</tr>
<tr>
<td>RF</td>
<td>Radio frequency</td>
</tr>
<tr>
<td>RH</td>
<td>Right-handed</td>
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<tr>
<td>RiMAX</td>
<td>Parameter estimation algorithm</td>
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<tr>
<td>RLS</td>
<td>Recursive least squares</td>
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<tr>
<td>RMSE</td>
<td>Root-mean-square error</td>
</tr>
<tr>
<td>RSSI</td>
<td>Received signal strength indicator</td>
</tr>
<tr>
<td>RX</td>
<td>Receiver</td>
</tr>
<tr>
<td>SAGE</td>
<td>Space-alternating generalized expectation-maximization</td>
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<tr>
<td>SBS</td>
<td>Switched-beam system</td>
</tr>
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<td>Abbreviation</td>
<td>Description</td>
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<tr>
<td>SDP</td>
<td>Semi-definite program</td>
</tr>
<tr>
<td>SINR</td>
<td>Signal-to-interference and noise ratio</td>
</tr>
<tr>
<td>SISO</td>
<td>Single-input single-output</td>
</tr>
<tr>
<td>SLC</td>
<td>Sidelobe canceller</td>
</tr>
<tr>
<td>SMI</td>
<td>Sample matrix inversion</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal-to-noise ratio</td>
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<tr>
<td>SON</td>
<td>Self-organizing network</td>
</tr>
<tr>
<td>TL</td>
<td>Transmission line</td>
</tr>
<tr>
<td>TX</td>
<td>Transmitter</td>
</tr>
<tr>
<td>UDN</td>
<td>Ultra-dense network</td>
</tr>
<tr>
<td>UKF</td>
<td>Unscented Kalman filter</td>
</tr>
<tr>
<td>ULA</td>
<td>Uniform linear array</td>
</tr>
<tr>
<td>UN</td>
<td>User node</td>
</tr>
<tr>
<td>WARP</td>
<td>Wireless open-access research platform</td>
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<tr>
<td>WiFi</td>
<td>Wireless fidelity</td>
</tr>
<tr>
<td>WiGig</td>
<td>Wireless Gigabit Alliance</td>
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<tr>
<td>WiMAX</td>
<td>World Interoperability for Microwave Access</td>
</tr>
<tr>
<td>WLAN</td>
<td>Wireless local area network</td>
</tr>
<tr>
<td>WPAN</td>
<td>Wireless personal area network</td>
</tr>
<tr>
<td>3GPP</td>
<td>Third Generation Partnership Project</td>
</tr>
<tr>
<td>1D</td>
<td>One-dimensional</td>
</tr>
<tr>
<td>2D</td>
<td>Two-dimensional</td>
</tr>
<tr>
<td>3D</td>
<td>Three-dimensional</td>
</tr>
<tr>
<td>5G</td>
<td>Fifth generation</td>
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<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
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<tbody>
<tr>
<td>a</td>
<td>Steering vector</td>
</tr>
<tr>
<td>A</td>
<td>Steering matrix</td>
</tr>
<tr>
<td>c</td>
<td>Speed of light</td>
</tr>
<tr>
<td>$C_L$</td>
<td>Series capacitor</td>
</tr>
<tr>
<td>$C_R$</td>
<td>Shunt capacitor</td>
</tr>
<tr>
<td>d</td>
<td>Distance between antenna elements</td>
</tr>
<tr>
<td>D</td>
<td>Number of signals</td>
</tr>
<tr>
<td>$D_m$</td>
<td>Measured radiation pattern</td>
</tr>
<tr>
<td>$E{\bullet}$</td>
<td>Statistical expectation</td>
</tr>
<tr>
<td>f</td>
<td>Wave frequency</td>
</tr>
</tbody>
</table>
\( F(\theta, \phi) \) Array factor
\( G_m \) Antenna gain of the radiation pattern
\( H \) Hermitian transpose
\( I \) Unit matrix
\( I_n \) Amplitude function
\( J \) Length of the specific range of \( \theta \)
\( J_0 \) Number of the DoA estimation points
\( k \) Time instant
\( k_0 \) Free space wavenumber
\( L \) Number of incident signals
\( L_s \) Shunt inductor
\( L_R \) Series inductor
\( m \) \( m^{th} \) antenna element
\( M_A \) Number of antenna elements
\( M \) Number of selected radiation beampatterns
\( n \) Noise vector
\( N \) Number of unit cells
\( N_s \) Number of snapshots
\( p \) Cell size
\( P \) Received power of the antenna
\( P_M \) MUSIC pseudospectrum
\( P_{\text{norm}} \) Normalized received power of the antenna
\( r \) Reactance value
\( \mathbf{R}_{ss} \) Received signal matrix
\( \mathbf{R}_{xx} \) Covariance matrix
\( s(k) \) Signal waveform
\( T \) Transpose
\( U \) Number of beams
\( V_S \) DC bias voltage
\( V_{SH} \) DC bias voltage
\( w \) Complex weight coefficient
\( \mathbf{w} \) Weight vector
\( x_m(k) \) Received signal
\( \mathbf{x}(k) \) Received signal vector
\( y(k) \) \hspace{1cm} \text{Output signal of the beamformer}

\( \alpha \) \hspace{1cm} \text{Leakage factor or attenuation factor}

\( \alpha_c \) \hspace{1cm} \text{Progressive phase shift}

\( \beta \) \hspace{1cm} \text{Propagation constant or phase constant}

\( \theta \) \hspace{1cm} \text{Elevation angle}

\( \theta_0 \) \hspace{1cm} \text{Direction of arrival of the received signal}

\( \lambda \) \hspace{1cm} \text{Wavelength}

\( \lambda_g \) \hspace{1cm} \text{Guided wavelength}

\( \lambda_m \) \hspace{1cm} \text{Eigenvalue}

\( \Gamma \) \hspace{1cm} \text{Cross-correlation coefficient}

\( \gamma \) \hspace{1cm} \text{Complex propagation constant}

\( \phi \) \hspace{1cm} \text{Azimuth angle}

\( \sigma_N^2 \) \hspace{1cm} \text{Common variance}
List of original publications

This thesis is based on the following papers, which are referred to in the text by their Roman numerals (I–V):


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1 Introduction

1.1 Background and motivation of the thesis

The focus of this thesis is to study advanced directional antenna techniques for reconfigurable leaky-wave antennas (LWAs) and antenna arrays. In recent years, these antennas have attracted significant attention as wireless data communications are expected to continue their exponential growth [1–6]. Current networks and systems cannot support such a massive mobile data increase, and thus a new research area, i.e., research on advanced directional antenna systems such as reconfigurable antennas, has emerged with the aim to solve this challenge. Advanced directional antennas have been shown to provide additional gains in both single-user [7–9], and multi-user [10] multiple-input multiple-output (MIMO) systems. They are also a significant element in the design of enhanced small-cell deployments. Wireless system performance can be heavily restricted by interference from nearby users in a dense wireless network. In such environment, antennas are one of the final tools a designer can use before transmitting signals into the wireless channel. Direction of arrival (DoA) estimation methods have played a significant role in practical implementations of antennas [11–13]. The demand for DoA estimation arises from the need of positioning and tracking of signal sources in both civilian and military applications [14].

The significance of DoA estimation and directional antenna communications will be further highlighted in future ultra-dense networks (UDN). For example, inter-site distances can range from a few meters in indoor deployments to up to around 50 m in outdoor deployments [15]. Hence, the probability of line-of-sight (LoS) conditions increases approximately to 70% when the expected distance between access nodes (ANs) is 50 m [16, 17]. As a result, the importance of spatial direction and DoA estimation is clearly highlighted compared with traditional radio networks.

Furthermore, advanced antennas can enable extending the transmission range, increasing data throughput, enhancing spectrum reuse, and substantially decreasing co-channel interference, which is one critical problem in future heterogeneous wireless communications systems, where classical macrocell networks and more advanced small-cell systems coexist. These advantages are studied in both academia and industry alike, e.g., by all the major network and device vendors (such as Samsung and Nokia), and the body of work is gradually accumulating into new standards.
Adaptive antennas can be divided into two classes: 1) phased arrays and 2) reconfigurable antennas. Traditional phased array systems use an array of many elements to control the beam direction and radiation pattern shape [18, 19], which can lead to very large antennas. This issue affects antennas operating, for example, in wireless fidelity (WiFi) frequency bands. However, conventional phased array systems could also be practical in millimeter-wave mobile communications systems, in situations where, for example, the 60 GHz band can be used because the distance between antenna elements is shorter. As such, large antenna arrays are predicted to be come a potentially disruptive technology in future wireless communications systems. Although antenna arrays can be considered one of the key technologies for fifth generation (5G) cellular networks. The problem with large arrays is the complexity of the algorithms since each antenna element requires its own weight coefficient. One solution to the problem is to use partially adaptive algorithms instead of fully adaptive beamforming [83]. One goal of this thesis is to study DoA estimation algorithms and beamforming for advanced antenna arrays.

In contrast to multi-element antenna arrays, CRLH reconfigurable LWAs [20] do not require multi-element antenna arrays or feeding networks [21]. In practical scenarios, LWAs have also many other benefits: low manufacturing costs, low direct current (DC) power consumption, full-space beam scanning, and it requires significantly less printed circuit board space. Considering these benefits, especially compactness and beamsteering, CRLH-LWAs also offer significant potential benefits for DoA estimation systems. This thesis focuses on the use of reconfigurable antennas in new generation small-cell radio systems.

Several DoA estimation algorithms for traditional antenna arrays have been discussed in previous literature [22–26]. However, these conventional algorithms cannot be directly applied to DoA estimation for CRLH-LWAs because of the inherent difference in design and operation of LWAs compared with traditional antenna arrays. The reason for this difference is that an LWA has only a single observation available at each sampling instance, unlike in the case of multi-element antenna arrays, where signals can be received from the different elements of the antenna array. For example, the traditional multiple-signal classification (MUSIC) algorithm forms a spatial correlation matrix of the signal samples received from the different elements of an antenna array [22]. In addition, the performance of this algorithm has a significant effect on a variety of array characteristics, such as the number of elements, array geometry, and the mutual coupling between the elements [27, 28]. These problems may be mitigated or avoided by using a CRLH-LWA.
1.2 Objectives and scope

The purpose of this thesis is to use adaptive antenna techniques in new generation (5G) small-cell networks. The main objective is to design DoA estimation algorithms for CRLH-LWAs and advanced antenna arrays. CRLH-LWAs have a different design compared with conventional antenna arrays. As a result, conventional DoA estimation algorithms for antenna arrays cannot be directly applied to reconfigurable antennas. The thesis presents that the spatial capabilities of the conventional antenna array can be virtually generated with a radio frequency (RF) one- or two-port antenna to receive $M$ arriving signals, each obtained with different radiation patterns from $M$ different directions $\theta_m$ measured from one or two antenna ports. The system model is illustrated in Figure 1. In this thesis, all presented DoA estimation algorithms are based on evaluating signals received from different radiation patterns, as presented in Figure 1.

The first research question of this thesis is how can DoA estimation algorithms be modeled for CRLH-LWAs. The thesis presents solutions to this question and introduces

![Fig. 1. System model where $\theta_1 < 0$ and $\theta_2 > 0$.](image)
three DoA estimation algorithms for a two-port CRLH-LWA: 1) modified MUSIC, 2) LWA-based power pattern cross-correlation (PPCC), and 3) LWA-based adjacent power pattern ratio (APPR). The performances of these three algorithms are qualitatively compared between each other, as well as with a low-complexity power detector (PD). The proposed MUSIC algorithm is introduced in two conference articles (Papers III – IV), which discuss the performance of DoA estimation algorithms in an anechoic chamber at Drexel University. In contrast to these studies, and other DoA estimation techniques discussed in literature [29–34], Papers I and II explore the DoA estimation capabilities of a two-port CRLH-LWA in a real-world multipath indoor environment and the use of both antenna ports at the same time. Consequently, the papers also examine utilizing the beam symmetry characteristic of LWAs by steering bidirectional beams. As a result, the overall signal acquisition and estimation time was halved and the length of the periodical training sequence was truncated. The related measurement results are introduced in Paper I. The paper also presents an extensive comparison and analysis of the DoA estimation performance achievable with an LWA and comparable antenna arrays. The results demonstrate that the DoA estimation of the receiving wave can be successfully performed by using a two-port CRLH-LWA with measured channels, even in the presence of severe multipath interference.

The thesis also presents improvements for the modified MUSIC and LWA-based APPR algorithms. The complexity of calculation is reduced by selecting a smaller number of received signals from different directions. It should be noted that the configuration selection is very important. The selection of different directions gives different results. The modified MUSIC and LWA-based APPR algorithms are tested in a different channel environment, and it is also examined how the algorithms perform if a smaller number of received signals is chosen. The results of these experiments are presented in Paper II.

The second research question is how can the complexity of DoA estimation algorithms and beamforming of the advanced antenna arrays be reduced. In future wireless 5G systems, it also possible to use a large number of antenna elements due to the use of high frequencies. One problem with large antenna arrays is the complexity of the algorithms, since each antenna element requires its own weight coefficient. One solution to this problem is to use partially adaptive algorithms where beamforming weight coefficients of antenna elements are calculated only for a fraction of the array elements. This thesis presents the DoA estimation algorithm for an advanced 4x4 antenna array, which has only four weight coefficients and one output port. The antenna
array is developed by an antenna expert of VTT. The size of the antenna can be easily expanded. Thus, these same algorithms can also be used for bigger antenna arrays. In this thesis, the APPR algorithm is applied to an advanced 2.4 GHz 4x4 antenna array and the algorithm is tested by carrying out several measurements in real-world environments. The measurement results are presented in Paper V. In short, the thesis also considers a feasible DoA estimation algorithm for advanced antenna arrays.

To sum up, the dissertation has two main objectives: 1) to design feasible DoA estimation algorithms for CRLH-LWAs (Papers I−IV) and 2) to design DoA estimation algorithms for advanced antenna-arrays (Paper V) that can be used in future 5G small-cell networks. In this thesis, both the studied adaptive antennas, i.e. the CRLH-LWA and smart antenna array, are designed for the 2.4 GHz WiFi frequency band, and thus the research focuses on this band. However, the proposed DoA estimation algorithms can also be applied to higher frequency bands. The thesis is based on five original papers, which are summarized in Chapter 3.

1.3 Author’s contribution

The focus of this thesis is to study advanced directional antenna techniques for reconfigurable antennas and advanced antenna arrays in future wireless communications systems. The goal is to develop DoA estimation algorithms that are able to estimate the direction of the arriving signal for reconfigurable antennas and advanced antenna arrays. The research methods include analysis, simulations, and measurements.

This dissertation is based on five peer-reviewed original papers. The contents of the papers can be divided into two main categories. Novel DoA estimation algorithms for CRLH-LWAs and related performance measurements are discussed in Papers I–IV. A DoA estimation algorithm for advanced 4x4 antenna arrays is presented in Paper V.

The author had the main responsibility for writing Papers I–III and V. The measurements carried out at Drexel University, discussed in Papers I–II, were obtained together with Aki Hakkarainen from Tampere University of Technology, Nikhil Gulati, and Damiano Patron from Drexel University. In addition, Damiano Patron developed the CRLH-LWAs described in Papers I–IV, while the author had the main responsibility for developing the DoA estimation algorithms for the LWA and analyzing the performance of the algorithms. The theoretical model for the algorithms was developed jointly by the author and Aki Hakkarainen in Paper II. The Cramer-Rao bound analysis discussed in Paper I was provided by Janis Werner and Aki Hakkarainen.
The author’s contribution to Paper V is as follows: The author had the main responsibility for developing the DoA estimation algorithms for the advanced antenna array and analyzing the performance of the algorithms. Matti Somersalo developed advanced antenna array and Mika Hoppari modeled the software-defined radio system described in Paper V. In the case of all the original papers, the co-authors also provided valuable comments and criticism. The original papers are summarized in Chapter 3.

1.4 Outline of the thesis

The dissertation is organized as follows: The first chapter outlines the research problem and objectives and provides some background information regarding the topic. Chapter 2 provides a review of the relevant literature on the use of advanced antennas in future generation small-cell networks. A summary of the original papers is presented in Chapter 3. Chapter 4 discusses the main findings and limitations of the thesis, future work, and potential applications of the results. Finally, Chapter 5 presents a summary and conclusion of the thesis.
2 Advanced antennas in future generation small-cell networks

Future wireless systems and services are dependent on the design of reconfigurable radio architectures that can adapt to the rapidly varying wireless environment. Furthermore, several studies have shown that mobile traffic volumes continue their exponential growth [1–6]. Existing cellular radio access networks are based on macrocells that have a range of up to tens of kilometers. However, the existing networks cannot fulfill the requirements of future mobile broadband services, unless a capacity boost is achieved at specific locations such as hotspots, cell edges, or indoor locations [35]. Single, static, and omnidirectional antennas are employed in current small-cell base stations (BSs), which include pico cells, femto cells, and ultra-dense small cells [36, 37]. Small cell networks can have cell sizes ranging from tens of meters to hundreds of meters [37]. The use of multiple antenna methods to mitigate interference from nearby macro BSs and other co-located small-cell BSs in co-channel deployments has recently attracted significant attention [37]. Furthermore, there is growing pressure to employ advanced antenna techniques to maximize spectral efficiency to meet the demand for increased data rates [37, 38].

Advanced antenna techniques, such as MIMO techniques, antenna diversity, spatial multiplexing, and beamforming, are significant elements of wireless standards such as IEEE 802.11n WiFi, IEEE 802.16 World Interoperability for Microwave Access (WiMAX) and Third Generation Partnership Project (3GPP) High Speed Packet Data Access (HSPA) and Long-Term of Evolution (LTE)/LTE-advance (LTE-A) [37]. In future wireless 5G networks, the dominant theme is multilayered heterogeneity and densification [6]. Small-cell BSs will also play a significant role, and advanced antenna techniques help to fulfill the foreseen demands.

It is envisioned that future 5G networks will have a very high spatial density of access nodes [39–43]. Consequently, the probability of LoS conditions will significantly increase because it is likely that user nodes (UNs) are within the range of several ANs [16, 17, 44]. Achieving LoS condition is beneficial not only for communications purposes but also in terms of estimating the direction of the received signal. In 5G networks, ANs are expected to be equipped with adaptive antenna solutions, such as antenna arrays. DoA information can be used to achieve higher accuracy for a device
localization. Improvements in the localization accuracy also creates opportunities for various future location-based applications, such as autonomous vehicles, intelligent traffic systems, and proactive radio resource management [45].

Conventional BSs serve a fixed direction or a BS sector and suffer from azimuthal interference [37]. On the other hand, small-cell BSs must adapt the direction of the antenna and can suffer from random azimuthal and vertical interference. These factors need to be considered when selecting the antenna scheme. In a small-cell deployment, system performance can be improved using adaptive directional antenna systems to steer the main beam of the antenna towards the desired users and the place nulls towards the sources of interference. In such systems, the achievable system performance enhancement highly depends on how well the DoA of the desired source signal and the interference signals is known. Hence, DoA estimation plays a key role in small-cell wireless communications systems equipped with adaptive directional antennas.

The challenges and requirements related to small-cells and the considerations in using advanced directional antennas can be summarized as follows: Directional adaptive antenna techniques should be low in cost, power consumption, and complexity. Small-cell BSs are energy efficient because they are located in close proximity to the user equipment. Thus the required transmission power is very low when compared with macrocell BSs [37]. Antenna arrays must be designed efficiently with small antenna sizes, close antenna spacing, and mitigation techniques for potential increased mutual coupling effects.

Heterogeneous networks have two layers of wireless networks: a macrocell layer and a small-cell layer. Their key challenge is cross-tier interference between the small-cell and macrocell networks [38], particularly when the same frequency band uses both the small-cell network and the surrounding macrocell network. Another difficult problem is interference between co-located small cells, i.e. the co-tier interference. These issues can be resolved or mitigated by using multi-element antenna techniques and processing in the spatial domain [46]. Beamforming techniques can also be used to reduced cross-tier and co-tier interferences which are critical challenges in small-cell networks.

Frequency selection plays an important role particularly in small-cell deployments because of the physical features, deployment costs, and available spectrum at the given frequency band. Attenuation of radio signals increases as the function of the frequency. For that reason, high frequencies require higher gains compared with lower frequencies. Moreover, the antenna size is comparable with the wavelength of the given frequency enabling a higher antenna gain in a smaller size when the frequency increases.
Furthermore, if antenna gain increases, the directivity of the antenna increases and the main beam of the antenna narrows. Narrow antenna beams must be carefully aligned at both ends of the transceiver, and they are vulnerable to antenna misalignment and outages. Recently, the use of higher frequencies has been seen as a very attractive alternative for high-speed and low-cost small-cell backhaul because higher frequencies have more available spectrum with light licensing. In addition, narrow beams and high attenuation increase the frequency reusability and decrease the risk of interfering nearby devices. ‘Millimeter-wave frequencies’ refers to the 60 GHz and 70/80 GHz frequency bands in the context of small-cell backhaul. Capacities are improving and BS distances decreasing. Therefore, these frequency ranges offer a very potential alternative for a number of reasons. Usually, the 60 GHz and 70/80 GHz frequency bands are unlicensed or lightly licensed, which means that the licensing costs also remain reasonable. Transmission speed can be increased to a few gigabits per second with very modest transmission methods because of the wide bandwidth. Furthermore, the higher frequency bands also enable a higher antenna gain, smaller antenna design and lower production costs. [47, 48]

Microwave frequencies in the range 6–60 GHz have been employed to wireless backhaul links for many decades [48]. These frequencies also require narrow beams to survive path loss as millimeter-wave frequencies. Wider beams require less accurate antenna alignment. Millimeter-wave frequencies are suitable for densely located BSs, whereas microwave frequencies are more suitable for long-distance rooftop-to-rooftop connections because the less spectrum is available and the path loss is lower. Millimeter-wave frequencies are suitable for densely located BSs, whereas microwave frequencies are more suitable for long-distance rooftop-to-rooftop connections because the less spectrum is available and the path loss is lower. Figure 2 illustrates an example of a wireless small cell backhaul deployment. Small-cells are placed on lamp posts at street level and they are closely located. Thus, they are suitable for the 60 GHz because the range is short and interference is low. In the bottom left-hand corner of the figure, trees block the LoS, and thus the sub-6 GHz technology is used to cope with the NLoS. However, recent studies [5, 35, 49] have shown that it is possible to plan NLoS backhaul deployments in bands above 20 GHz that can provide a high network performance. [47, 48]

Backhaul is among the most important research topics in the small-cell community. This is an area of intense innovation, driven by the many new operators in the field and an increased presence of established backhaul vendors. However, much work is still needed to ensure that backhaul does not become a cost and performance bottleneck in small-cell deployments. Small-cell BSs are most likely densely located in urban areas with a high traffic demand, i.e. in areas where typical BS installations are fixed, for
example, to a lamp post or building wall. The dense network of high-speed small cells creates new challenges regarding backhaul. [47]

The number of the small cells may become large, but the total cost of ownership should be considerably lower in comparison with more traditional BS deployments. Consequently the backhaul solution used in 5G mobile broadband systems needs to be more cost-efficient, and more easy to scale and install compared with traditional backhaul solutions. If the 5G network enable ubiquitous ultra-high data rate and low latency mobile services, backhaul transport must provide native support for this by enabling ultra-high capacity data transfer between the access point and core network. In small-cell networks, costs such as installation, maintenance, and device costs should be as low as possible. Consequently, an ideal solution would be a plug and play solution that would enable fixing a low cost backhaul device is located in a desired location with minimum configuration and network design effort. Furthermore, in a wireless small-cell the backhaul device should automatically adapt to the changing environment without human involvement. [37, 47, 50, 51]
Beamforming techniques for millimeter-waves have been defined in standards such as IEEE 802.15.3c (TG3c) for indoor wireless personal area networks (WPAN), IEEE 802.11 ac/ad, IEEE 802.11, and Wireless Gigabit Alliance (WiGig) for wireless local area network (WLAN). However, although beamforming methods are commonly employed to indoor scenarios, they do not easily extend to outdoor scenarios with longer distances, outdoor propagation, and other environmental factors such as wind. Small-cell units are typically mounted to outdoor structures such as walls and poles. Poles can move and vibrate due to strong winds and gusts, which may result in an unacceptable outage if beam alignment is not frequently performed. [52]

Massive MIMO is a promising technology with proven and remarkable advantages. However, its wider deployment is difficult, mainly due to the highly increased hardware complexity of front-end RF circuits. In the case of a conventional large antenna array with independently fed antenna elements, the total cost and the hardware burden increase linearly with the number of RF chains and the number of RF power amplifiers, in particular. The requirement for highly linear characteristics and reduced the power dissipation regarding the power amplifiers is another significant problem. Indeed, the increased RF front-end complexity and cost are among the main factors restricting the number of elements in massive MIMO systems, and thus the overall potential performance of cellular networks. In addition the higher number of antenna elements consumes more circuit power, although it also improves spectrum efficiency. Thus, there is a trade-off between energy efficiency and spectrum efficiency. [53, 54]

This thesis focuses on studying DoA estimation techniques for advanced antennas. The following sections provide a literature review concerning directional adaptive antennas and DoA estimation techniques for antenna arrays and reconfigurable antennas.

2.1 Directional adaptive antennas

Directional adaptive antennas commonly refer to antennas steering a narrow beam in a single specific direction. The direction of the beam can be adaptively adjusted according to its environment. Transatlantic wireless communications was first demonstrated in 1901 by Marconi and et al. [55]. Marconi was also first to use the term ‘antenna’ for large emission facilities. in 1983, IEEE specified ‘antenna’ as means for radiating or receiving radio waves [56]. Marconi used an early antenna array to improve the link gain of the trans-Atlantic transmission of Morse codes. The antenna array consisted of
Adaptive antennas can be generally categorized based on two different approaches of beamforming: switched-beam systems (SBS) and adaptive array systems (AAS). An SBS uses a fixed beamforming network that produces predefined beams, whereas adaptive array systems can create a beam to any direction. Probably the most applied solution in SBSs is Butler matrix (BM) \cite{58}. SBSs use a switching network for selecting the most appropriate beam to receive a signal from a particular mobile station (MS). Figure 3(a) shows that it may not be possible to steer the maximum of the chosen beam to the desired direction because each beam typically serves more than one MS. On the other hand, an AAS system can steer the beam separately for each user, as is shown in Figure 3(b). However, the implementation of an AAS is much more complex compared with a SBS.

The simplest link alignment method is an exhaustive method where all possible beam pairs between the transmitter and receiver are measured. The number of the required beam searches is $2^U$, if the nodes have $U$ beams. Thus, the tracking estimation time increases exponentially with the number of beam directions. \cite{47}

Directional adaptive antennas enable the use of multi-level beam searching. This means that a wide beam is used to estimate the initial direction of the radio link (rough
estimation) and the beamwidth is narrowed once the approximate direction is found (precise DoA estimation). In the IEEE 802.11ad [59] and IEEE 802.15.3c [60] standards, beamforming is exploited following a similar principle. Two-level beamforming training is described in [61] and in IEEE 802.11ad. It includes sector-level search and refinement stages. In [62], Wang proposed a three-level training procedure which was adopted with slight changes in IEEE 802.15.3c. The proposed method covers 1) quasi-omni-level, 2) sector-level, and 3) beam-level training [63]. Both of the above standards apply the two-level training system, which reduces the number of required beam searches compared with the exhaustive method. However, if the beams become narrower, the number of beam searches increases. IEEE 802.15.3c ignores the first training level, i.e. the quasi-omni-level training. In the sector-level search, both ends of the transceiver transmit training symbols in different directions by using wide sectors and exchange measurement data to identify the estimated direction of the connection. After that, the devices start to transmit narrower beams in the estimated directions and exchange data to select the best possible beam pair.

Figure 4 illustrates a classification of the different antenna types. The literature of directional antennas, such as conventional antenna array elements, metamaterial leaky-wave antennas (LWA), Alford loop antennas, integrated lens antennas (ILA), and electronically steerable parasitic array radiator (ESPAR) antennas. Directional antennas can be categorized into two main classes: 1) phased arrays and 2) reconfigurable antennas. Phased array antennas have multiple of radiating elements that are excited by different phases but the reconfigurable antenna has only one reconfigurable element. Conventional phased array systems use an array of many elements to control the beam direction and radiation pattern shape [18, 19], which can lead to a very large antenna size. This issue affects antennas operating, for example, in WiFi frequency bands. However, traditional phased array systems could also be practical in millimeter-wave mobile communications systems, for example, when the 60 GHz band can used due to the small distance between the antenna elements. As such, large antenna arrays are expected to become a potentially disruptive technology in future wireless communications systems. The problem with large arrays is the complexity of the algorithms, since each antenna element requires its own weight coefficient. One solution to the problem is to use partially adaptive algorithms instead of fully adaptive beamforming [83]. In contrast to large antenna arrays, reconfigurable CRLH-LWAs [20] do not require multi-element antenna arrays or feeding networks [21].
Leaky-wave antennas have been studied for a long time [64–66]. An LWA has a radiating transmission line (TL) structure and either a uniform or a periodic configuration. LWAs have been widely studied because they offer a flat profile, low weight and cost, relatively high gain, and ease of fabrication. However, the structure of LWAs is forward-wave propagation, which means that conventional LWAs have limited its scanning capability. Veselago introduced the concept of metamaterials in 1967 [67]. He proposed that materials with simultaneous negative permittivity and permeability are physically permissible and have a negative index of refraction [67]. Veselago presented only right-handed (RH) metamaterials. Thirty years later, Smith and et al. developed the first left-handed (LH) metamaterials with an artificial and effectively homogeneous structure [68]. The initial metamaterial structure was not practical for engineering purposes. The main disadvantages were the high losses and narrow bandwidth. Around the same time in 2002, three research groups introduced the first approach to metamaterials that involved using transmission lines [69–71]. Metamaterial transmission lines are non-resonant in nature, and thus these structures have low losses and a broad bandwidth. This thesis studies in more detail DoA estimation algorithms for advanced antenna arrays.
and reconfigurable CRLH-LWAs in new generation small-cell radio systems. Next the principles of adaptive antenna arrays, reconfigurable antennas, and DoA estimation algorithms are discussed in more detail.

### 2.2 Beamforming and DoA estimation with antenna arrays

Adaptive antenna arrays must solve two main estimation problems: the DoA estimation and the calculation of antenna element weight coefficients to enable beamsteering and beamforming [19]. Adaptive antenna arrays can automatically control the directionality of the radiation pattern by changing the weight coefficients of each branch of the array in accordance with the changing signal environment. If the DoA is not known, it is estimated by using DoA algorithms such as MUSIC [22], and estimation of signal parameters via rotational invariance techniques (ESPRIT) [26]. Next, adaptive beamforming methods adapt the beam by adjusting the weight coefficient of each antenna element so that the main signal power is steered towards the desired user and nulled towards interfering users [72]. First, this section discusses antenna arrays and beamforming in general. Then, it presents a literature review of DoA estimation algorithms for antenna arrays.

#### 2.2.1 Beamforming

Early antenna arrays were introduced over 100 years ago [73]. The research on adaptive antenna arrays started in 1959 [74–76]. It begun with the invention of the intermediate frequency sidelobe canceller (SLC) in 1959 by Howells [74]. Abblebaum [76] introduced the concept of a fully adaptive array in 1966. Abblebaum’s algorithm was based on the maximization of the signal-to-noise ratio (SNR) at the output of the antenna array. Applebaum’s research was a generalization of a coherent SLC algorithm. Later in the same decade, Widrow presented a very similar method, using the least mean square (LMS) algorithm [72]. The difference between the two techniques is that the Howells-Applebaum algorithm requires previous knowledge of the beam steering vector whereas the LMS algorithm requires knowledge of the received signal [57]. Array processing has been studied extensively, and there are good review papers available on the topic, for example, in [12, 77–79]. Array processing has also been discussed in books on antenna arrays [11, 19, 80–82].
Conventional antenna arrays consist of multiple closely located antenna elements. Beamforming is an antenna array that uses phased arrays. The basic idea behind beamforming is that the different channels must be correlated. In the beamforming approach, antenna elements are spaced closely enough so that the Nyquist criterion for spatial sampling is fulfilled. Consequently, transmitted or received signals add constructively in some directions and add destructively in other directions. Furthermore, these directions can be adjusted by modifying the array weights. This is referred to as spatial filtering. A conventional beamformer steers the beam towards the desired signals. The maximum antenna gain is proportional to the number of antenna elements, $M_A$. Thus, antenna gain and the accuracy of directivity can be enhanced by increasing the number of antenna elements.

There are many possible antenna arrays [83, 84]. The radiating elements can be arranged in different ways along a line or on a plane. Linear, circular and planar arrays have been used in various systems. As shown in Figure 5, the uniform linear array (ULA) typically consists of $M_A$ identical radiators which are equally spaced at distance $d$ [83]. The beam can be steered to the desired direction by changing the progressive phase shift $\alpha_z$ electronically. This can be done by simple RF phase shifters. The progressive phase shift can be written as

$$\alpha_z = \frac{2\pi}{\lambda} d \sin \theta_0$$  \hspace{1cm} (1)

where $\lambda$ is the wavelength, $\lambda = c/f$ where $c$ is the speed of light and $f$ is frequency of the wave, and $\theta_0$ is the direction of arrival of the received signal. The progressive phase shift can also be called the progressive phase factor.

The radiation pattern of the array can be analyzed by means of an array factor. The array factor represents the far-field radiation pattern of an array of isotropically radiating elements. The ideal three-dimensional (3D) array factor of a phased array is given by

$$F(\theta, \phi) = \sum_{m=1}^{M_A} A_m e^{j(m-1)(k_0 d \sin \theta \sin \phi - \alpha_z)} = \sum_{m=1}^{M_A} A_m e^{j(m-1)\psi_{M_A}} = \sum_{m=1}^{M_A} a_m(\theta, \phi)$$  \hspace{1cm} (2)

where $k_0 = 2\pi/\lambda$ is the free space wavenumber, $\psi_{M_A} = k_0 d \sin \theta \sin \phi - \alpha_z$, and the angle $\theta$ is the elevation angle. The angle $\theta$ defines the $zy$-plane. The two-dimensional (2D) array factor is a special case of 3D, so that $\phi$ is $90^\circ$. The steering vector of the antenna array consists of the elements of the array factors $a_m(\theta, \phi)$ and can be defined as

$$a(\theta, \phi) = [a_1(\theta, \phi), a_2(\theta, \phi), ..., a_{M_A}(\theta, \phi)]^T$$  \hspace{1cm} (3)

where $T$ is the transpose operation.
If distance $d$ between the antenna elements is too large compared with wavelength $\lambda$, a second main lobe can appear in the radiation pattern [84]. The extra main lobes are referred to as grating lobes. To prevent such lobes from emerging, the distance between antenna elements should follow

$$\frac{d}{\lambda} < \frac{1}{1 + |\cos \theta|}.$$  

(4)

The antenna element spacing criterion depends on the desired direction of the main lobe. If the lobe is scanned close to the endfire ($0^\circ$ or $180^\circ$), the spacing must be one half a wavelength to prevent grating lobes from appearing. Placing antenna elements to the array always involves minor errors. Especially in high frequencies, the accuracy of the placing is very important. The wavelength at 60 GHz is 5 mm. An error of 0.1 wavelengths corresponds to a 0.5 mm error in placing. Thus, the distance between the elements has to be very precise to avoid grating lobes.

Beamformers were first developed for narrowband signals that can be sufficiently characterized by a single frequency [85]. The general narrowband receiver beamformer structure is illustrated in Figure 6. In this type of a beamformer, signals from each antenna element are multiplied using complex weights and added to the array output. Signal $x(k)$ received by the antenna array as a single plane wave and can be mathematically expressed as

$$x(k) = As(k) + n(k)$$  

(5)

where $s(k)$ is the $D\times1$ signal sources vector where $D$ is the number of the transmitted signals, including interfering transmitters. $A$ is the $M_A \times D$ steering matrix, and $n(k)$ is
the $M_A \times 1$ noise vector. The steering matrix, signal source vector, and noise vector can be presented, respectively, as

$$A = [a(\theta_1), a(\theta_2), \ldots, a(\theta_D)], \quad (6)$$

$$s(k) = [s_1(k), s_2(k), \ldots, s_{M_A}(k)]^T, \quad (7)$$

and

$$n(k) = [n_1(k), n_2(k), \ldots, n_{M_A}(k)]^T. \quad (8)$$

Thus, the output of the beamformer becomes

$$y(k) = \sum_{m=1}^{M_A} w_m x_m(k) = w^H x(k) \quad (9)$$

where $x_m(k)$ is the received signal by the $m^{th}$ antenna element at the time instant and $w_m$ is the complex weight corresponding to the $m^{th}$ antenna element. The weight vector and the received signal can be represented as

$$w = [1, \exp[jk_0d\sin(\theta)], \ldots, \exp[j(M_A - 1)k_0d\sin(\theta)]]^T \quad (10)$$

and

$$x = [x_1(k), x_2(k), \ldots, x_{M_A}(k)]^T. \quad (11)$$
The purpose of the steering and modification of the radiation pattern is to improve the reception of the desired signal and simultaneously to mitigate the effect of interfering signals through the selection of complex weights $w_m$ [83].

In adaptive beamforming, the maximum number of interferers that can be suppressed is upper-bounded by the number of antenna elements minus one ($M_A - 1$) [86, 87]. Thus, if the size $M_A$ of the adaptive arrays is increased, better interference mitigation and better resolution can be obtained. However, this may also result in challenges related to computational burden and convergence speed. Furthermore, the high-resolution capability resulting from using large arrays also makes the array more sensitive to various imperfections.

Adaptive beamforming can be designed to be fully adaptive or only partially adaptive [83]. In fully adaptive beamforming, all available degrees of freedom are utilized. This refers to the number of unconstrained or “free” weights that can be used to form a beam. Thus, each element or beam is individually adaptively controlled to achieve the maximum control of the radiation pattern. Fully adaptive beamforming offers the following three benefits: 1) all of the array’s degrees of freedom are available to the beamformer, 2) the maximum aperture gain can be retained because the entire array is employed, and 3) the maximum spatial resolution can be retained because the entire array aperture is used. Fully adaptive beamforming may be difficult to implement in situations where the number of array elements would be large [83], such as in massive antenna arrays. However, a number of strategies can be used to reduce the complexity. For example, only a fraction of the array elements can be judiciously selected for the adaptive control, which means that the adaptivity is implemented at the element level [88]. Alternatively, if all of the array elements can be grouped into subarrays, beamforming can be carried out separately for each subarray, and the outputs of the subarray beamforming can then be adaptively controlled [89]. Finally, the entire array can be used to form beams which can then be steered in an adaptive way [90].

The adaptive techniques require some sort of reference signals in their adaptive optimization process. The reference signal usually means explicit information about the signals of interest. An explicit reference can be divided into two parts, i.e. spatial reference and temporal reference. The spatial reference usually refers to the DoA information of the desired signal. The temporal reference signal may be a pilot signal that is correlated with the desired signal. The used reference signal depends, in particular, where adaptive beamforming is to be implemented. [83]
The beamforming techniques for orthogonal frequency division multiplexing (OFDM) systems can be divided into pre-fast Fourier transform (FFT) and post-FFT techniques at the receiver and post-inverse FFT (IFFT) and pre-IFFT techniques at the transmitter [92]. Pre-FFT beamforming is also called symbol-wise beamforming and post-FFT beamforming is called subcarrier-wise beamforming at the receiver [91]. In [92], the performance and complexity of the pre-FFT and post-FFT schemes are investigated. The results show that pre-FFT techniques can offer reliable performance in cases where there are no strong multipath components and no strong interference signals in the channel. If there are strong multipath components and interferences, post-FFT beamforming is required. Consequently, it is important to verify this observation by a target channel model in use. In general, time domain processing has a low complexity because only one FFT is required [92]. In the frequency domain, spatial signal processing of individual subcarriers is believed to provide the optimum performance, but with a much higher complexity. The post- and pre-FFT sample matrix inversion (SMI) beamforming techniques are also compared with each other in [93].

### 2.2.2 DoA estimation

In general, a beamformer requires advance knowledge of the DoA of the desired signal so that it can steer the main beam to the desired direction. In practice, this information is not precisely known. DoA estimation algorithms play a significant role in the practical implementation of antenna arrays [12]. DoA estimation algorithms can be divided into acquisition and tracking algorithms. This thesis focuses on existing acquisition methods for , which are mainly discussed in [12, 80]. The acquisition methods can be categorised to data aided and blind systems. In data-aided systems, the data \( s_0 \) is known. In blind systems, the data is unknown.

Several DoA estimation algorithms are based on the antenna array covariance matrix. Therefore, it is next explained how the covariance matrix and its eigenvalue decomposition (EVD) can be defined. The signal received by the array is defined in (5); it should be noted that the signal is noise-corrupted. Typically the noise is uncorrelated and data signals received by the different elements are correlated when they are transmitted from the same source. This property can be used to extract the required DoA information. The DoA information can be specified by measuring the spatial correlation matrix of the received data, a.k.a. the covariance matrix. The covariance
matrix for the $N_s$ number of snapshots can be defined as

$$\hat{\mathbf{R}}_{xx} = \frac{1}{N_s} \sum_{k=1}^{N_s} \mathbf{x}(k)\mathbf{x}^H(k)$$

(12)

where $\mathbf{H}$ denotes the Hermitian operator. The exact covariance matrix of $\hat{\mathbf{R}}_{xx}$ is not known and it is difficult to specify due to the limited number of received data samples. Therefore, an estimate of the covariance matrix must be used. The covariance matrix can be decomposed into two orthogonal spaces: signal and noise, as expressed by

$$\hat{\mathbf{R}}_{xx} = \mathbf{E}\{\mathbf{s}(k)\mathbf{A} + \mathbf{n}(k)\}\mathbf{E}^H\{\mathbf{s}(k)\mathbf{A} + \mathbf{n}(k)\}$$

(13)

$$= \mathbf{A}\mathbf{E}\{\mathbf{s}(k)\mathbf{s}^H(k)\}\mathbf{A}^H + \mathbf{E}\{\mathbf{n}(k)\mathbf{n}^H(k)\}$$

$$= \hat{\mathbf{R}}_{ss}\mathbf{A}\mathbf{A}^H + \sigma_N^2\mathbf{I}_{M_A}$$

where $\mathbf{E}\{\cdot\}$ denotes the statistical expectation, $\hat{\mathbf{R}}_{ss}$ is the $M_A \times M_A$ signal covariance matrix, $\sigma_N^2\mathbf{I}_{M_A}$ is the noise covariance matrix, $\sigma_N^2$ is the common variance of the noise, and $\mathbf{I}_{M_A}$ is the identity matrix of $M_A \times M_A$. The diagonal elements of the signal covariance matrix represent the source power and the off-diagonal elements represent the source correlations. The covariance matrix of noise is diagonal because noise in each channel is highly uncorrelated. The eigenvalues of $\hat{\mathbf{R}}_{xx}$ are $[\lambda_1, ..., \lambda_L]$ where $L$ is the number of the incident signals. These are sorted according to size, as expressed by

$$\lambda_1 \geq \lambda_2 \geq ... \geq \lambda_{M_A} > 0$$

(14)

where $D$ largest eigenvalues correspond to the signal while $M_A - D$ correspond to noise. It should be be noted that if any vector orthogonal to $\mathbf{A}$ is an eigenvector of $\hat{\mathbf{R}}_{xx}$ with the eigenvalue $\sigma_N^2$ (there are $M_A - D$ linearly independent of such vectors), then the remaining eigenvalues are all larger than $\sigma_N^2$, and the eigenvalue and vector pairs can be divided into noise eigenvectors and signal eigenvectors. Thus, the covariance matrix can be represented as

$$\hat{\mathbf{R}}_{xx} = \hat{\mathbf{E}}_s\hat{\mathbf{A}}\mathbf{E}_s^H + \hat{\mathbf{E}}_n\hat{\mathbf{A}}_n\mathbf{E}_n^H$$

(15)

where $\hat{\mathbf{A}}_n = \sigma_N^2\mathbf{I}_{M_A}$. Thus, all noise eigenvectors are orthogonal to $\mathbf{A}$, the columns of $\hat{\mathbf{E}}_s$ has to cover the range space of $\mathbf{A}$ whereas those of $\hat{\mathbf{E}}_n$ cover its orthogonal complement. The covariance matrix has a major effect on the performance of DoA estimation algorithms because many of those algorithms use the spectral matrix. Using this EVD, the total number of signals $D$ must be known or estimated. [14, 78]
DoA estimation techniques can be classified into two categories: spectral-based and parametric methods [78], as illustrated in Figure 7. Spectral-based methods create a spectrum-like function of the parameters of interest, for instance the DoA. DoA estimates are the highest peaks of the function. On the other hand, parametric methods require a simultaneous search for all parameters of interest. Maximum likelihood (ML) methods are parametric techniques. These methods can be divided into two approaches: stochastic ML and deterministic ML. ML methods are more computationally complex than other DoA estimators. However, the estimates are often more accurate, particularly in low SNR conditions [14]. [78]

Spectral-based methods can be divided into conventional beamforming techniques and subspace methods. Conventional methods refer to beamforming and null steering techniques such as delay-and-sum method, Bartlett, and minimum variance distortionless response (MVDR). The conventional DoA estimation methods direct beams in all possible directions and search for peaks in the output power. These types of algorithms require a large number of antenna elements to obtain a high resolution. The Bartlett [94] algorithm applies a classical spectral Fourier analysis to spatial analysis. It maximizes the power of the beamformer output signal for a given input signal. Another conventional algorithm is the MVDR algorithm, also known as the Capon minimum variance (MV) algorithm [95]. It maintains gain along the desired direction according to the MV criterion subject to a constraint on the array response while it simultaneously minimizes
the output power in other directions. The estimation accuracy of the algorithms depends on the number of the snapshots and the length of the array.

Subspace methods, such as MUSIC, rank reduction estimation (RARE) and ESPRIT, employ the structure of the received data. Thus, the resolution significantly better compared with previous methods. Subspace-based approaches can be divided into element-space and beam-space methods [83]. Element-space beamforming means that the data symbols are directly weighted with the array weights to form a beam. In beam-space beamforming, the data signal is processed by a multi-beam beamformer to form a suite of orthogonal beams, which are then are processed in beam-space. The subspace DoA estimation methods provide a high resolution. These algorithms are more accurate and they are not limited to the physical size of aperture [80].

The MUSIC algorithm is one of the most common high-resolution DoA estimation algorithms for antenna arrays. Thus, the main focus of this thesis is to apply the high-resolution MUSIC algorithm to reconfigurable LW As, as discussed in section 1.2. Next, the conventional MUSIC algorithm is presented in more detail. The algorithm uses the estimated covariance matrix to define the DoAs. The main purpose of the MUSIC algorithm is to find the $D$ directions that can minimize the projection of the steering vector for the entire noise subspace. The spectral decomposition can be presented in equations (15) and (12), where it is assumed that $\mathbf{AR}_n\mathbf{A}^H$ has full rank and the diagonal matrix consists of $M_A$ largest eigenvalues. Then, the eigenvectors in $\hat{\mathbf{E}}_n$ are orthogonal to $\mathbf{A}$. This implies that

\[ \hat{\mathbf{E}}_n\mathbf{a}(\theta) = 0, \theta \in \{\theta_1, \ldots, \theta_{M_A}\}. \]  

(16)

Finally, the MUSIC pseudospectrum can be defined as [22]

\[ P_M(\theta) = \frac{\mathbf{a}^H(\theta)\mathbf{a}(\theta)}{\mathbf{a}^H(\theta)\hat{\mathbf{E}}_n\hat{\mathbf{E}}_n^H\mathbf{a}(\theta)}. \]  

(17)

The DoAs of the signals received by the array are the $D$ largest peaks in the MUSIC pseudospectrum.

The criterion used for optimizing the weights is typically the minimum mean-square error (MMSE), maximum signal-to-interference and noise ratio (SINR), or MV [83]. In addition, adaptive algorithms are also commonly used for updating the weights. These algorithms include LMS, direct sample covariance matrix inversion, and recursive least squares (RLS).
2.3 DoA estimation with reconfigurable antennas

Reconfigurable antennas have a single-antenna geometry, and their electrical and radiation characteristics can be adaptively modified in real time. Reconfigurable antennas have four reconfiguration properties: the antenna can change the frequency of operation, radiation pattern, polarization behavior, or a combination of any of these properties [96]. It depends on the use of the antenna which reconfigurable properties an antenna designer should select. Reconfigurable antennas consist of only a single radiation element, and they do not require expensive phase shifters. Antenna arrays, on the other hand, have several antenna elements for receiving incident signals and use phase shifters. The antenna array DoA estimation algorithms cannot be directly applied to reconfigurable antennas due to the inherent difference in their design and operation. There are different kinds of reconfigurable antennas, as illustrated in Figure 4. This thesis focuses on microstrip antennas, and specifically on CRLH-LWAs. These antennas have several advantages, as explained above in Section 1.1. There are only a few previous articles that discuss LWA-based DoA estimation algorithms in the literature. In addition, ESPAR-based DoA estimation algorithms are also applied to LWAs in this thesis.

The ESPAR antenna is a single-port array antenna and has only one central active monopole element that is surrounded by reactively controlled parasitic elements [97]. Radiation patterns of the ESPAR antenna can be controlled by changing the reactance values \( \{r^{(1)}, r^{(2)}, \ldots, r^{(M)}\} \). The single RF chain connects to the active element, which is then fed to a low-noise amplifier with load impedance. Parasitic elements are attached to variable reactors, which maintain the reactance loads of the parasitic elements. In the literature, different types of DoA estimation algorithms for ESPAR antennas have been introduced. A hand-held microwave DoA finder was proposed in [98]. The basic idea of this method is that the DoA finder receives signals from 12 different directions. It selects as the DoA estimate the pattern direction that achieves the highest antenna output antenna gain. The resolution of the algorithm is only 30°. Higher-resolution algorithms for ESPAR antennas are achieved by using antenna output antenna gain and premeasured directional radiation patterns of the antenna; see [99, 100]. The PPCC algorithm was introduced in [99] and the APPR algorithm was first proposed in [100]. The MUSIC [34], unitary MUSIC [31], and ESPRIT [101] algorithms for ESPAR antennas are discussed in [31, 34, 101–103]. These types of subspace algorithms require the evaluation of the covariance matrix on the basis of measurements and assume a large
number of snapshots for data samples. The MUSIC algorithm for ESPAR antennas can be developed by transmitting the same data the number of times corresponding with the number of directional radiation patterns used. However, this data transmission scheme decreases the transmission rate. It is still satisfactory in applications such as terminal position location, or when DoA estimation is required from time to time for tasks such as forming a node position location table in an ad-hoc network where the nodes do not frequently relocate [101]. In this thesis, the PPCC and APPR algorithms are applied to LWAs, and thus these methods are introduced next in more details.

The principle behind the PPCC algorithm is that \( M \) power patterns correspond with \( M \) reactance vectors. The correlation coefficient between the calculated output power of the antenna for each corresponding reactance vector and the antenna power pattern set is the highest at the signal DoA angle. The PPCC algorithm measures the cross-correlation coefficient between the measured powers and the measured radiation patterns for \( M \) directions. The key steps of the algorithm are as follows [99]:

1. Select \( M \) different sets of reactances \( \{r^{(1)}, r^{(2)}, \ldots, r^{(M)}\} \). Calculate the power of the antenna for each set. The power pattern value of the antenna at angle \( \theta \), which corresponds the \( m \)th set, is presented as \( D_m(\theta) \). The first step is performed only once.
2. Next the corresponding output power of the antenna for each set of reactances \( \{P_1, P_2, \ldots, P_M\} \) is calculated.
3. Finally, the correlation coefficient \( \Gamma(\theta) \) between the calculated power patterns and the measured output powers of the antenna is measured when \( \theta \) is between 0° and 360°. The correlation coefficient \( \Gamma(\theta) \) can be presented as

\[
\Gamma(\theta) = \frac{\sum_{m=1}^{M} D_m(\theta) P_m}{\sqrt{\sum_{m=1}^{M} D_m^2(\theta)}} \sqrt{\sum_{m=1}^{M} P_m^2}.
\]  

(18)

4. The DoA estimate is the highest value of the function \( \Gamma \).

The basic idea behind the simple APPR algorithm is that it compares adjacent pattern power ratios. The steps of this algorithm are as follows [100]:

1. First, the powers of the received signal \( P_m(\theta) \) are measured as the ESPAR antenna changes over the set of \( M \) reactance vectors \( r^m \). Thus the powers are measured from \( M \) different directions.
2. Then the radiation pattern with the maximum signal value is selected.
3. Adjacent pattern power ratios are measured with respect to the maximum signal power pattern, and the DoA estimation range is narrowed. The two APPRs for the
measured radiation patterns can be defined as

\[ \Gamma_{m+1}(\theta) = \frac{D_m(\theta)}{D_{m+1}(\theta)} \]  

\[ \Gamma_{m-1}(\theta) = \frac{D_m(\theta)}{D_{m-1}(\theta)} \]  

These ratios for the chosen \( \theta \) range are measured and stored into a look-up table (LUT) beforehand to reduce run-time computations. Thus, the DoA is estimated over the \( \theta \) range.

4. Then, the APPR is calculated for the received power. It can be expressed as

\[ \hat{\Gamma}_{m+1}(\theta_0) = \frac{P_m(\theta_0)}{P_{m+1}(\theta_0)} \]  

\[ \hat{\Gamma}_{m-1}(\theta_0) = \frac{P_m(\theta_0)}{P_{m-1}(\theta_0)} \]  

5. Finally, the DoA \( \theta_0 \) is estimated by comparing these ratios on the predefined ratios from the LUT.

In this thesis, different types of DoA estimation algorithms were developed for reconfigurable CRLH-LWAs. Next, the principle behind the CRLH-LWAs and LWA-based DoA estimation algorithms is discussed in more detail. A microstrip implementation of a CRLH-TL was introduced in [104]. CRLH-TLs support both RH and LH propagation. The reconfigurable CRLH-LWA can be implemented as a two-port radiating element with tunable radiation properties. The layout of the antenna is cascaded to achieve a periodic structure from port 1 to port 2, and it is made by a series of \( N \) metamaterial unit cells [20, 105]. The LWA is based on a traveling wave that progressively leaks out power while it propagates along the metamaterial waveguide. In other words, when a signal is received at one of the input ports, the traveling wave leaks out energy as it progressively travels towards the second port. This energy leakage defines the directivity of the radiated main beam and is a function of the propagation constant \( \beta \) along the structure. Thus, the radiation properties can be defined by the complex propagation constant \( \gamma = \alpha + j\beta \), where \( \alpha \) is an attenuation constant, which can also be called a leakage factor [104]. In literature [106, 107], the propagation constant \( \beta \) can also be called a phase constant.

The unit cell is designed so that the propagation constant sweeps within the radiated region of the dispersion curve \( |\beta| < k_0 \) [20]. Each unit cell is populated with two pairs: a
series capacitor \( C_L \) and a shunt inductor \( L_L \) to provide the left-handedness and series inductor \( L_R \) and a shunt capacitor \( C_R \) to provide the right-handedness, as illustrated in Figure 8. The CRLH behavior is determined by designing unit cell with proper series capacitance and shunt inductive component through a microstrip stub. The propagation constant along the waveguide structure can be electronically modulated from left-hand \((\beta < 0)\) to right-hand \((\beta > 0)\) regions through the two DC voltages \( V_S \) and \( V_{SH} \). As a result, these variations of the wave propagation characteristic become responsible for steering the beam from broadside \( \theta_1 = \theta_2 = 0^\circ \), to the backward direction (to the left, \( \beta < 0 \)) when \( \theta_1 < 0^\circ \) (using port 1), and to the forward direction (to the right, \( \beta > 0 \)) when \( \theta_2 > 0^\circ \) (using port 2), as illustrated in Figure 9 [20]. The relationship between the propagation constant \( \beta \) and the beam steering angle \( \theta \) can be expressed as [20]

\[
\theta_1 = -\theta_2 = \arcsin\left(\frac{\beta}{k_0}\right)
\]  

where the propagation constant \( \beta = \frac{1}{p} \left( \frac{\omega \sqrt{L_R C_R} - \frac{1}{\omega \sqrt{L_L C_L}}}{\omega \sqrt{L_R C_R} + \frac{1}{\omega \sqrt{L_L C_L}}} \right) \) can be varied through the varactor diodes that can vary \( C_R \) and \( C_L \). Thus, the CRLH-LWA can steer the main beam from broadside to backward and forward angles [20]. This type of an antenna can be used for beamforming in MIMO systems. LWAs have an effectively homogenous structure, which means that the average cell size \( p \) is much smaller than the guided wavelength \( \lambda_g \). Hence, the average cell size should be smaller than \( p < \lambda_g/4 \) [20].
Fig. 9. Sketch of a two-port CRLH-LWA and example of beam steering capabilities from broadside $0^\circ$ to backward $\theta_1$ and forward $\theta_2$ angles.

There are only a limited number of papers introducing DoA estimation algorithms for LWAs. An energy-based DoA estimation system is evaluated at a fixed frequency employing a CRLH-LWA in [29]. In [30], the DoA of the wideband pulsed signal is measured from the estimated cross-correlation coefficients using the spectra of the received signals with a predetermined incident angle-dependent spectra. A DoA estimation method for sectorized antennas, such as an LWA, are proposed in [108]. The algorithm assumes a noncooperative transmitter. Similar work on DoA estimation as in this thesis, has also been independently reported in [32, 33]. These papers discuss the use of the MUSIC algorithm for CRLH-LWA. However, in [32], only experimental results are presented and it is not explained how the MUSIC algorithm can be generated using LWA. Furthermore, only experimental studies of the use of the MUSIC algorithm in an anechoic chamber are reported, and the DoA estimation is performed using a single port antenna [32, 33].

2.4 Comparison of conventional antenna arrays and CRLH-LWAs

This section provides a comparison of conventional antenna arrays and LWAs. These antennas have inherent differences in terms of their design and operation. An LWA
Fig. 10. Array factor: (a) conventional uniform linear antenna array, (b) LWA. © [2013] IEEE

has only a single observation available at each sampling instance. In conventional antenna arrays, signals are received from different elements of the antenna array. Therefore, conventional DoA estimation algorithms cannot be directly applied to the DoA estimation and beamforming in CRLH-LWAs. However, the array factor of the LWA can be formed almost in the same way as for a conventional uniform linear array, as illustrated in Figure 10 [109]. The amplitude function has a constant value $I_n = I_0$ in a conventional uniform array. On the other hand, it is an exponential function $I_n = I_0 \exp\left(-\alpha(n-1)d\right)$, with a leakage factor $\alpha$ and a cell size $p$ in the case of the LWA [109]. The cell size $p$ can be considered the same as the distance between antenna elements $d$. In addition, each individual LWA unit cell can be considered a single antenna array element. Thus, the array factor can be written as [109]

$$AF(\theta) = \sum_{n=1}^{N} I_n \exp(j(n-1)k_0d\sin(\theta) + j\xi_n)$$

$$= \sum_{n=1}^{N} a_{\text{mod},n}(\theta)$$

(24)

where $\xi_n = (n-1)k_0d\sin(\theta_0)$. The beam angle $\theta_0$ is defined by using equation (23).

As previously noted, the radiation methods of the antennas also differ. Radiation in a traveling wave antenna occurs because the RF signal leaks out energy while it travels through the antenna waveguide [20]. In antenna arrays, the total field is the vector sum
of the fields radiated by the individual elements, which is related to the mutual coupling between the antennas [19]. This is one of the major distinctions between these two types of beamsteering antennas. The functional principles of phased arrays (phase shifters, mutual coupling, etc.) cannot be mixed with the behavior of LWAs.

The optimal distance between the elements, \( d \), is \( 0.5\lambda \) for traditional antenna arrays. However, the average cell size \( p \) should be smaller than \( p < \lambda_g/4 \) [20]. Thus, the cells can locate more densely in LWAs compared with antenna array elements. If the distance between antenna elements is less than \( 0.5\lambda \), mutual coupling effects can emerge. However, in practice, mutual coupling can be compensated very effectively (see, e.g. [110]), at the expense of more complicated array calibration and increased processing complexity.

Regarding the comparison between LWAs and phased arrays: the first benefit of LWAs is wider range in beamsteering compared to phased arrays. A planar phased array hardly achieves \( \pm 40^\circ \) [111, 112], while an LWA can easily direct the beam up to \( \pm 60^\circ \). This larger spatial range is a key advantage in terms of DoA estimation. The second advantage is the cost. A single LWA costs much less in terms of substrate area and the number of lumped components compared to a phased array, which has multiple antenna elements and phase shifters. Each antenna element is also connected to a radio frequency converter and an analog-to-digital converter. The third benefit is the small size of the antenna. An LWA requires a smaller printed circuit board area than a phased-array at the same frequency. Furthermore, LWAs have a low DC power consumption. Taking these advantages into consideration, especially beamsteering and compactness, CRLH-LWAs offer significant potential benefits for DoA estimation systems.
3 Summary of the original papers

This chapter summarizes the contents and key results of the five original publications with respect to the prevailing gaps in the literature review. It describes the contributions of the original Papers I to V in detail and addresses the research question and assumption presented in section 1.2, along with references to the literature discussed in Chapter 2. The contents of the original papers of this thesis cover DoA estimation algorithms for LWAs and advanced antenna arrays. As noted in Chapter 2, only a limited number of previous papers have proposed DoA estimation algorithms for CRLH-LWAs [29–34].

This chapter is divided into two sections: section 3.1 covers the topic of DoA estimation algorithms for CRLH-LWAs, and section 3.2 discusses DoA estimation algorithms for advanced antenna arrays. Section 3.1 provides a summary of Papers I to IV. The papers present three different types of novel DoA estimation algorithms for CRLH-LWAs: Modified MUSIC, LWA-based APPR, and LWA-based PPCC. The APPR algorithm is also applied to advanced antenna arrays in Paper V. The summary of Paper V is presented in section 3.2.

The starting point for the thesis work was Paper III, which discusses how the conventional MUSIC algorithm can be modified for CRLH-LWAs. Similar work concerning the application of the MUSIC DoA estimation algorithm to CRLH-LWAs has also been independently reported in [32, 33]. Papers III and IV demonstrate the use of MUSIC algorithms using an LWA in an anechoic chamber. Papers I and II describe the performance of novel DoA estimation algorithms in compact CRLH-LWAs in indoor environment. Paper V discusses the performance of the APPR DoA estimation algorithm in advanced antenna arrays in an indoor environment. Papers I and II are journal articles. Papers III to V have been published in conference proceedings. Table 1 shows how the studies concerning the use of DoA estimation algorithms for CRLH-LWAs and advanced antenna arrays are related to the papers.

3.1 DoA estimation algorithms for CRLH-LWAs

In Papers I to IV, the proposed algorithms were designed to perform DoA estimation using Drexel University’s CRLH-LWAs. The antenna has been designed using full-wave electromagnetic simulator (HFSS) [113] and adjusted to operate within the 2.4 GHz 802.11 WiFi band. The LWA used in Papers I–III consists of 12 metamaterial CRLH
Table 1. Relation of the original papers to the DoA estimation algorithm studies.

<table>
<thead>
<tr>
<th>DoA estimation algorithms</th>
<th>Antenna type</th>
<th>Paper</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>LWAs</td>
<td>Antenna array</td>
</tr>
<tr>
<td>Modified MUSIC</td>
<td>x</td>
<td>x</td>
</tr>
<tr>
<td>APPR</td>
<td>x</td>
<td>x</td>
</tr>
<tr>
<td>PPCC</td>
<td>x</td>
<td>x</td>
</tr>
</tbody>
</table>

unit cells, and the dimensions of the milled prototype are 156 mm in length and 38 mm in height. In the LWA used in Paper IV, the number of the unit cells is 11 and the dimensions are 140 mm in length and 40 mm in height. The cells are populated with two varactor diodes in series and one in a shunt configuration. By adapting the two DC bias voltages ($V_S$ and $V_{SH}$) across the varactors, the antenna can direct the beam from broadside to backward and forward directions, as illustrated in Figure 9. The LWA is a single antenna entity. This is a very significant difference compared to multi-element antenna arrays, where signals can be received from different RF branches and finally combined after phase shifting.

The single unit cell has been optimized in the left-hand region ($\beta < 0$) to achieve the maximum beam coverage when using each port. In other words, when Port 1 is used, the beam can be oriented at angles between $0^\circ$ and $-60^\circ$ but when switching to Port 2, the beam covers the symmetrical quadrant between $0^\circ$ and $+60^\circ$, as illustrated in Figure 9. The LWA has a 1.6 mm thick Rogers 4360 substrate, and it is characterized by a relatively high dielectric constant $\varepsilon_r = 6.15$ and low loss tangent = 0.003 to decrease the antenna form factor and increase the radiation efficiency.

The RF ports of the Drexel University’s LWA can be used either one at a time or both at the same time. If both ports are used simultaneously, the antenna generates two symmetrical beams with respect to the broadside $\theta = 0^\circ$ direction. In other words, two signals can be received, namely $y_1(k)$ and $y_M(k)$, at a time from two symmetrical directions. Consequently, the beam symmetry characteristics of LWAs can be utilized as an additional benefit. In Papers III and IV and in [32, 33], the LWA uses only one antenna port at a time. The both antenna ports are used simultaneously in Papers I and II. The papers show that the symmetry feature can significantly reduce the processing time during the DoA estimation.

By taking advantage of the LWA’s compact dimensions and pattern reconfigurability, three different DoA estimation algorithms were developed: Modified MUSIC (Papers I to IV), LWA-based APPR (Paper II), and LWA-based PPCC (Paper I). The DoA estimation
algorithms for traditional antenna arrays cannot be directly used for DoA estimation using CRLH-LWAs because of the inherent differences in design and operation of LWAs versus conventional antenna arrays. An LWA has only a single observation available at each sampling instance, whereas traditional antenna arrays have access to multiple elements and signals can be received through the different elements of the antenna array [114]. Papers I to IV show how the spatial capabilities of the conventional antenna arrays can be virtually generated with the CRLH-LWA.

The spatial capabilities of the conventional antenna arrays can be virtually created with the CRLH-LWA to receive signals from $M$ different directions. In Papers I to IV, all proposed DoA estimation algorithms for LWAs apply the same principle, as illustrated in the flowchart in Figure 11. The first step is to measure the transmitted signals $M$ times through $M$ different radiation patterns. These $M$ received signals are stored and they all are used to estimate the DoA. The LWA-based MUSIC algorithm requires that the transmitted signal is the same in all directions. In Papers I and III, the number of the used directions $M$ is the same as the number of the LWA unit cells $N$.

Now the spatial diversity has been created for the LWA. Next, a steering vector and a correlation matrix can be modeled for the LWA to estimate the DoA by using

![Flowchart of the DoA estimation algorithms for CRLH-LWAs.](image)

Fig. 11. Flowchart of the DoA estimation algorithms for CRLH-LWAs. © [2013] IEEE
the MUSIC algorithm. The array factor has been formulated in equation (24) for the LWA. In the system, the LWA has \( N \) unit cells and the transmitted signal \( u(k) \) has been measured from \( M \) different radiation patterns. Hence, each of these received signals can be presented as \( x_m \) and vector \( x \) can be expressed as (Paper III)

\[
x = a_{\text{mod}}u(k) + n
\]

where \( x = [x_1(k) \ldots x_M(k)]^T \), \( n = [n_1(k) \ldots n_M(k)]^T \), and \( a_{\text{mod}} \) is the steering vector of the LWA. It can be represented as

\[
a_{\text{mod}} = \begin{bmatrix} a_{\text{mod},1}(\theta_1) \\ a_{\text{mod},2}(\theta_2) \\ \vdots \\ a_{\text{mod},M}(\theta_M) \end{bmatrix} = A_{\text{LWA}} a(\theta_0)
\]

where \( A_{\text{LWA}} \) is

\[
A_{\text{LWA}} = \begin{bmatrix} a_{\text{LWA}}^T(\theta_1) \\ a_{\text{LWA}}^T(\theta_2) \\ \vdots \\ a_{\text{LWA}}^T(\theta_M) \end{bmatrix},
\]

\[
a_{\text{LWA}}(\theta_m) = \begin{bmatrix} I_1 \\ I_2 \exp(jk_0d\sin(\theta_m)) \\ \vdots \\ I_N \exp(j(N-1)k_0d\sin(\theta_m)) \end{bmatrix},
\]

and

\[
a(\theta_0) = \begin{bmatrix} 1 \\ \exp(jk_0d\sin(\theta_0)) \\ \vdots \\ \exp(j(N-1)k_0d\sin(\theta_0)) \end{bmatrix}.
\]

Once the received signals \( x_m(k) \) are collected using the \( M \) different radiation patterns, the covariance matrix \( \hat{R}_{xx} \) can be estimated in the same way as for antenna arrays; see the equations (12) and (15). In (15), the noise subspace matrix \( \hat{E}_n \) is defined by using the EVD. Then, the MUSIC pseudospectrum can be defined for the LWA

\[
P_{\text{MUSIC}}(\theta) = \frac{a_{\text{LWA}}^H(\theta)a_{\text{LWA}}(\theta)}{a_{\text{LWA}}^H(\theta)\hat{E}_n\hat{E}_n^H a_{\text{LWA}}(\theta)}.
\]
In Papers III and IV, the performance of the modified MUSIC algorithm is demonstrated using an LWA in an anechoic chamber, as illustrated in Figure 12. In the measurements, the signal was transmitted from 12 different directions ($\theta = -60^\circ$ to $\theta = +50^\circ$ in steps of $10^\circ$) by changing the bias voltages $V_S$ and $V_{SH}$. The measurements were made from four different DoA angles: -15°, -25°, -35°, and -45°. The difference between Papers III and IV is that Paper IV introduces a CRLH-LWA with reduced dimensions and improved beam symmetry, which were achieved through a new design of the metamaterial unit cell. In Paper III, the length of the LWA is 15.6 cm, while in Paper IV, it is 14.0 cm. The new CRLH-LWA has also been applied to DoA estimation performed using the modified MUSIC algorithm. The estimated DoAs are in a very good agreement with the expected angles in both Papers III and IV, as seen in Figure 13 and Table 2. It should be noted that the performance of the DoA estimation has slightly reduced in the improved design of the LWA (Paper IV) compared with in the older design (Paper III).

In Papers I and II, the MUSIC algorithms were tested in a real-world environment and two novel DoA estimation algorithms are introduced for LWAs: LWA-based PPCC (Paper I) and APPR (Paper II). The algorithms proposed in these papers use two antenna ports at the same time. Unlike in the previous conference papers (Papers III, and IV)
Table 2. Relation of the original papers to the DoA estimation algorithm studies.

<table>
<thead>
<tr>
<th>Expected DoA</th>
<th>Paper III</th>
<th>Paper IV</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>DoA</td>
<td>Error</td>
</tr>
<tr>
<td>-15°</td>
<td>-17°</td>
<td>1.6 %</td>
</tr>
<tr>
<td>-25°</td>
<td>-25°</td>
<td>0.0 %</td>
</tr>
<tr>
<td>-35°</td>
<td>-35°</td>
<td>0.0 %</td>
</tr>
<tr>
<td>-45°</td>
<td>-47°</td>
<td>1.7 %</td>
</tr>
<tr>
<td>Average</td>
<td></td>
<td>0.83 %</td>
</tr>
</tbody>
</table>

or in any other DoA estimation techniques previously described in literature [32, 33], Papers I and II utilize the beam symmetry characteristic of LWAs by using both antenna ports at the same time and by steering bi-directional beams. As a result, the overall signal acquisition and estimation times are halved, and the length of the periodical training sequence can be truncated. In addition, the algorithms are also suitable for use in MIMO systems.

The experiments described in Papers I and II were carried out in the premises of Drexel University. A wireless open-access research platform (WARP) and a field-programmable gate array (FPGA)-based software defined radio platform were used in the experiments. The indoor setup was a closed lobby with stairs and glass walls. In
the measurements, both LoS and NLoS components were expected due to the severe multipath interference between the transmitters (TXs) and receivers (RXs). Figure 14a illustrates the layout of the measurements. In the RX nodes, the reconfigurable two-port CRLH-LWAs were used, and the TXs were equipped with two standard omnidirectional antennas, as illustrated in Figure 14b. The measurements were carried out using the open WiFi frequency range of 2.452 GHz to 2.472 GHz, as the LWAs had been calibrated for this range. Due to the various WiFi access points that are in close proximity all WiFi traffic received was affected by co-channel interference, making the measurement environment very challenging.

In Paper I, the modified MUSIC and PPCC algorithms for LWAs are qualitatively compared with each other, as well as with a low-complexity PD. The PPCC algorithm was first introduced for ESPAR antennas in [99]. Paper I discusses how to apply the PPCC algorithm to LWAs. The PPCC algorithm calculates the cross-correlation coefficient between the measured powers $P_m$ and the measured radiation patterns $D_m(\theta)$, as described in section 2.3. The received powers can be calculated as

$$P_m = \frac{1}{N_x} \sum_{k=1}^{N_x} |y_m(k)|^2. \quad (31)$$
Then, the measured radiation pattern of each direction $m$ is normalized
\[
\mathcal{D}_m(\theta) = \frac{D_m(\theta)}{\sqrt{\sum_{l=1}^{M} D_l^2(\theta)}}. \tag{32}
\]
After that, the normalized received powers are measured, presented as
\[
\overline{P}_m = \frac{P_m}{\sqrt{\sum_{l=1}^{M} P_l^2}}. \tag{33}
\]
Finally, the cross-correlation coefficient is expressed as
\[
\Gamma(\theta) = \sum_{m=1}^{M} P_m \mathcal{D}_m(\theta) \tag{34}
\]
and the angle that maximizes (34) is the estimated DoA.

PPCC algorithms generate more accurate DoA estimates compared with the PD. However, the cost of the cross-correlation computation increases with the resolution of $\theta$. Thus, the complexity of the PPCC algorithm is higher than that of the PD. Compared with the MUSIC algorithm, the PPCC algorithm has a much lower complexity because the PPCC does not rely on eigenvalue decomposition or on computationally intensive matrix multiplications. Paper I introduces the measurement results of DoA estimation in a multipath indoor environment, conducted in a large lobby area at Drexel University, as illustrated in Figure 14a. The results show that the DoA estimation of the received signal can be successfully performed by using a two-port CRLH-LWA with measured channels, even in the presence of severe multipath interference, as seen in Figure 15.

Paper II introduces the APPR algorithm for LWAs and compares it with the modified MUSIC algorithms and the PD. The APPR algorithm was first introduced for ESPAR antennas in [100]. Paper II shows how to apply the APPR algorithm to LWAs and how configuring the antenna radiation patterns for signal observations can impact the performance of the MUSIC and APPR algorithms, which was not discussed in [100]. In addition, the paper presents a complexity analysis of the algorithms.

First, the received powers $P_m$ are calculated from $M$ different directions and then normalized as $P_m^{\text{norm}} = P_m / G_m$, where $G_m$ is the measured antenna gain of set $m$. After that, the radiation pattern which has the maximum received power is chosen. Then, the adjacent pattern power ratio between adjacent patterns and the selected pattern is measured. The APPR is first calculated on the basis of the radiation patterns measured in the anechoic chamber, and these patterns are also normalized as $D_m^{\text{norm}} = D_m / G_m$. 

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Fig. 15. DoA estimation results for the best case, TX1-RX6, (top) and for the worst case, TX1-RX2, (bottom): (a) and (c) modified MUSIC pseudospectrum, (b) and (d) PD and PPCC power spectrums. © [2017] IEEE

The APPR can be expressed as

\[
\Gamma_m^{+} (\theta) = \frac{D_m^{\text{norm}} (\theta)}{D_{m+1}^{\text{norm}} (\theta)}
\]

\[
\Gamma_m^{-} (\theta) = \frac{D_m^{\text{norm}} (\theta)}{D_{m-1}^{\text{norm}} (\theta)}
\]

These ratios for the selected \( \theta \) range can be calculated and stored into an LUT beforehand to decrease run-time computations. The selection of the radiation pattern configuration influences the length of the \( \theta \) range. In Figure 16a, the search areas are illustrated with a dotted line.

Second, the APPR is calculated based on the received powers, and can be introduced as

\[
\hat{\Gamma}_m^{+} (\theta) = \frac{P_m^{\text{norm}} (\theta)}{P_{m+1}^{\text{norm}} (\theta)}
\]

\[
\hat{\Gamma}_m^{-} (\theta) = \frac{P_m^{\text{norm}} (\theta)}{P_{m-1}^{\text{norm}} (\theta)}
\]
Finally, the APPRs are compared. The choice of the $\Gamma_{m+1/m-1}$ is denoted as follows: If $p_{m+1}^{\text{norm}}(\theta_0) > p_{m-1}^{\text{norm}}(\theta_0)$, $\Gamma_{m+1}(\theta)$ and $\hat{\Gamma}_{m+1}(\theta)$ are compared. If $p_{m+1}^{\text{norm}}(\theta_0) < p_{m-1}^{\text{norm}}(\theta_0)$, $\Gamma_{m-1}(\theta)$ and $\hat{\Gamma}_{m-1}(\theta)$ are compared. As a result, the estimated DoA can be presented as

$$\hat{\theta} = \begin{cases} 
\arg \min |\Gamma_{m-1}(\theta) - \hat{\Gamma}_{m-1}(\theta_0)| & \text{if } p_{m+1}^{\text{norm}}(\theta_0) < p_{m-1}^{\text{norm}}(\theta_0) \\
\arg \min |\Gamma_{m+1}(\theta) - \hat{\Gamma}_{m+1}(\theta_0)| & \text{if } p_{m+1}^{\text{norm}}(\theta_0) > p_{m-1}^{\text{norm}}(\theta_0)
\end{cases}.$$  

(39)

The performance of the MUSIC and APPR estimation algorithms in the case of LWAs has been verified and validated. Paper II presents several simulations carried out to numerically evaluate the performance and to conduct experimental measurements. The effect of the number of radiation patterns on the accuracy of the estimated DoAs was also researched. Different radiation pattern options were also considered for estimating the DoA. From Figure 16b, it can be noted that if the number of radiation patterns is five, the radiation patterns cover very well the estimation area. From Figures 16a and 16c, it can be seen that the radiation patterns can be too near to (Figure 16a) or too far from (Figure 16c) each other. Table 3 provides the radiation pattern solutions used in Cases 1 to 3. The first column presents Cases 1 to 3, which correspond to the different choices as the number of radiation patterns and their main beam directions.

The results of Paper II show that the radiation pattern configuration should be carefully selected for the APPR algorithm. For example, if five radiation patterns were chosen instead of 11 radiation patterns, the overall signal storing time would be halved and the power of the signal could be measured fewer times. However, if fewer radiation patterns were used, the DoA estimation range would broaden because the gap between...
Table 3. Selection of radiation patterns in DoA estimation cases.

<table>
<thead>
<tr>
<th>Main beam directions (°)</th>
<th>-47</th>
<th>-39</th>
<th>-28</th>
<th>-18</th>
<th>-8</th>
<th>0</th>
<th>8</th>
<th>18</th>
<th>28</th>
<th>39</th>
<th>47</th>
</tr>
</thead>
<tbody>
<tr>
<td>Case 1</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td>x</td>
</tr>
<tr>
<td>Case 2</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td>x</td>
<td></td>
<td>x</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Case 3</td>
<td></td>
<td>x</td>
<td>x</td>
<td>x</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 4. DoA estimation results with PD, APPR, and MUSIC algorithms.

<table>
<thead>
<tr>
<th>Cases</th>
<th>Total RMSE (°)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>PD</td>
</tr>
<tr>
<td>Case 1</td>
<td>13.9</td>
</tr>
<tr>
<td>Case 2</td>
<td>15.0</td>
</tr>
<tr>
<td>Case 3</td>
<td>14.1</td>
</tr>
</tbody>
</table>

The adjacent radiation patterns increases. In addition, the amount of data to be processed offline and online would be reduced if fewer radiation patterns were chosen.

The simulation results concerning the AWGN channel, presented in Paper II, show that the best root mean square error (RMSE) results for the APPR algorithms were achieved in Cases 2 and 3, and Case 1 gave the best results for the MUSIC algorithm. Thus, these three cases were also analyzed in a real-world multipath environment. The summary of the results for the LWA-based APPR, modified MUSIC and PD is presented in Table 4. Based on these experimental results, it seems that the APPR algorithm works better in a real-world multipath environment than the MUSIC algorithm. In the measurements, the received signals experienced both LoS and NLoS components due to the severe multipath interference between the TX and RX nodes in an indoor environment. The correlated multipath signals have an effect that the eigenvalue decomposition of the signal covariance cannot be split into the signal and noise subspaces, thus the RMSE increases significantly when compared with the AWGN channel simulations. It can be concluded that the performance gap between these two algorithms increases when the number of radiation patterns reduces, as can be seen from Table 4. The table also shows that the performance of the APPR algorithm is almost the same in all cases. The difference of the RMSE is only $1.4°$ between Case 1 and Case 3. Thus the overall signal storing time can be decreased by selecting fewer radiation patterns. However, the results of Paper II also show that the adjacent radiation patterns cannot be too far away from each other. In sum, the proposed DoA estimation algorithms for CRLH-LWAs can
Paper II also introduces a computational complexity analysis of the modified MUSIC and LWA-based APPR algorithms. The complexity of these DoA estimation algorithms is considered in terms of basic operations \[115\], i.e., additions and subtractions, multiplications, and divisions, and which are referred to by ADD, MUL, and DIV, respectively. In the APPR algorithm case, part of the computation can be done offline and stored into the LUT, in which case the calculations do not have to be performed in real time. The analysis is presented in Table 5. There, \( J_0 \) denotes the number of the DoA estimation points. In the measurements, it is equal to 109 (-54\(^\circ\) to 54\(^\circ\) with 1\(^\circ\) resolution). Furthermore, \( J \) is the length of the specific range of \( \theta \) for the APPR algorithm, which depends on the selection of the radiation pattern configuration. Choosing \( M = 11 \) results in \( J = 15 \), whereas for \( M = 5 \) results in \( J = 31 \). It can be clearly seen that the APPR algorithm has a much lower computational complexity compared with the MUSIC algorithm.

### 3.2 DoA estimation algorithms for advanced antenna arrays

In Paper V, the proposed APPR algorithm is applied to a smart advanced 2.4 GHz antenna array, as illustrated in Figure 17. The antenna array design is simplified to reduce the hardware complexity. The proposed antenna array has 4\( \times \)4 antenna elements, which beamform the radiation pattern using only four weight coefficients. The antenna has one output port. This type of an antenna can be easily expanded. Thus, the proposed algorithm can also be applied to larger antenna arrays. The beam of the antenna can be directed between -45\(^\circ\) and 45\(^\circ\) with 13 fixed beam directions, as shown in Figure 17.

Paper II discusses how the selection of the antenna radiation patterns affects the performance of the MUSIC and APPR algorithms developed for LWAs. In Paper V, the APPR algorithm is applied to an advanced 2.4 GHz antenna array. Only five

<table>
<thead>
<tr>
<th>Algorithm</th>
<th>MUSIC</th>
<th>APPR</th>
</tr>
</thead>
<tbody>
<tr>
<td>MUL</td>
<td>( M^2(N_s + 2J_0) )</td>
<td>( 2MN_s )</td>
</tr>
<tr>
<td>DIV</td>
<td>( J_0 )</td>
<td>2</td>
</tr>
<tr>
<td>ADD</td>
<td>( M^2(N_s - 1 + 2J_0) - 2J_0M )</td>
<td>( J + M(N_s - 1) )</td>
</tr>
<tr>
<td>LUT</td>
<td>-</td>
<td>((MJ_0)MUL \ (2MJ + MJ_0)DIV)</td>
</tr>
</tbody>
</table>

Table 5. The computational complexity analysis. © [2017] Hindawi
Fig. 17. Smart WLAN antenna and its measured radiation patterns. © [2018] IEEE

Fig. 18. The radiation pattern configuration for the APPR algorithm.

radiation patterns are selected because the patterns cover very well the estimation range -50° to 50°. The selected main beam directions are -36°, -17°, 1°, 19°, and 38°, which are not too far from or too near to each other, as illustrated in Figure 18. The selection of fewer radiation patterns has many advantages. For example, the overall signal storing time reduces and the power of the signal needs to be calculated fewer times. In addition, the amount of data that needs to be processed off-line and the on-line processing time decrease. A disadvantage is that the DoA estimation range increases
when fever radiation patterns are selected due to the increased gap between the adjacent radiation patterns.

The APPR estimation algorithm for the proposed antenna array can be calculated in the same way as in Paper II for LWAs and as shown in formulas (35)-(38). The performance of the APPR algorithm developed for the advanced antenna array was estimated by carrying out experimental measurements in a multipath indoor environment. The indoor environment was a large lecture room in the premises of the VTT Technical Research Centre of Finland Ltd, Oulu. The measured signals experience both LoS and NLoS conditions because of the multipath between TXs and RXs. Two TX nodes were equipped with a standard omnidirectional antenna, which transmitted real-time LTE-downlink signals. Six RX nodes were equipped with a directional smart antenna array. The broadside directions $\theta = 0^\circ$ of the RX antennas are illustrated with arrows in Figure 19. Only one TX-RX pair was active at a time during the measurements.

Figure 19 illustrates the layout and results of the measurements. These results show that the estimated DoAs with the proposed APPR algorithm are in good agreement with the predicted angles. The total RMSE for all the measurement cases was 5.6° for the APPR algorithm and 7.8° for the PD. The results indicate out that the APPR algorithm is more accurate compared with the PD. To sum up, the combination of the APPR DoA estimation algorithm and the advanced antenna array can provide fairly accurate DoA estimation in a real-world measurement environment.
Additionally, Paper V presents a real-time demonstration environment. The RF signals are difficult to visualize because they are invisible. The purpose of the demonstration was to visualize how the radiation patterns of a smart antenna changes during DOA estimation measurements, as illustrated in Figure 20. The radiation pattern changes could also be presented in real time during APPR DoA estimation measurement. The 3D radiation patterns were modeled by using Matlab, and the demonstration was built using universal software radio peripheral platforms and the LTE application framework by National Instruments [116].
4 Discussion

4.1 Main findings

Wireless data demands are exponentially increasing in future wireless radio networks. New technology solutions are needed to fulfill the high performance requirements. The use of advanced directional antenna systems in smart wireless devices has increased to improve data throughput and to achieve a high quality of service. These types of antennas can adaptively adjust the radiation pattern towards a desired wireless user and according to the changing environment. They can steer the main beam towards the desired user, which improves the channel capacity and link robustness and significantly reduces the interference from adjacent wireless networks. The importance of the directional antenna techniques will further increase in future ultra-dense small-cell networks, where LoS conditions are increased. This thesis investigates advanced directional antenna algorithms as a solution to these future requirements.

The thesis particularly focuses on the development of DoA estimation algorithms for reconfigurable CRLH-LWAs and advanced antenna arrays. It was studied how a high-resolution MUSIC algorithm should be modified for LWAs. Due to the inherent differences in the design of antenna arrays and the CRLH-LWA, the proposed LWA-based DoA estimation algorithms use \( M \) received signals, each obtained using a different radiation pattern and measured from two antenna ports. Then, it is possible to virtually generate the spatial capabilities of the conventional antenna array using the CRLH-LWA, and a correlation matrix can be formed using the \( M \) received signals. Three DoA estimation algorithms were developed for LWAs: 1) modified MUSIC, 2) LWA-based PPCC, and 3) LWA-based APPR. The performances of these three algorithms were qualitatively compared with each other, as well as with a low-complexity PD.

The thesis also showed how the complexity of the algorithms can be reduced by selecting a smaller number of received signals. It was noted that the configuration selection is very important because different direction selections provide different results. The comparison of the computational complexity of the MUSIC, APPR, and PPCC algorithms was shown in Table 6 with 11 radiation patterns and \( N_S = 10 \). The complexity of these DoA estimation algorithms was evaluated in terms of basic operations, i.e., ADD, MUL, and DIV. Part of the calculations can be measured offline and stored into the LUT in the case of the APPR algorithm. Thus, they do not need to be calculated in real
Table 6. The computational complexity analysis. © [2017] Hindawi

<table>
<thead>
<tr>
<th>Algorithm</th>
<th>MUSIC</th>
<th>APPR</th>
<th>PPCC</th>
</tr>
</thead>
<tbody>
<tr>
<td>MUL</td>
<td>28919</td>
<td>220</td>
<td>396</td>
</tr>
<tr>
<td>DIV</td>
<td>109</td>
<td>2</td>
<td>330</td>
</tr>
<tr>
<td>ADD</td>
<td>25069</td>
<td>114</td>
<td>165</td>
</tr>
<tr>
<td>LUT</td>
<td>-</td>
<td>1199 MUL 1529 DIV</td>
<td>-</td>
</tr>
<tr>
<td>Square root</td>
<td>-</td>
<td>-</td>
<td>16</td>
</tr>
</tbody>
</table>

Conventional phased array systems use all antenna elements of the array to adjust the direction of the beam and the shape of the radiation pattern [18, 19]. If the number of elements is huge, the complexity of the algorithm also increases because each antenna element requires its own weight coefficient. Partially adaptive algorithms are one solution to the problem [83]. They can calculate beamforming weight coefficients only for a fraction of the array elements. In Paper V, the APPR DoA estimation algorithm was applied to a smart antenna array with a reduced complexity in the calculation antenna beam steering. The complexity of the APPR algorithm and storing time were decreased by selecting only five out of 13 radiation patterns for the DoA estimation. The results show that the required level of accuracy can be reached when the proposed antenna is used for APPR DoA estimation. A Matlab 3D radiation pattern was also modeled to display the changes of the radiation pattern a real-time demonstration during the DoA estimation measurement.

4.2 Limitations and future work

There are some limitations in the studies discussed here. Many of these could also be topics for future work. Therefore, recommendations for future work are also discussed in this section.

The APPR and PPCC DoA estimation algorithms are power-based methods that select the most powerful signal. Thus, the performance of these algorithms is heavily dependent on the power of the received signal and, they cannot estimate the DoA of the
desired received signal very well if there are stronger signals interfering with the desired signal. The studies discussed in this thesis show that if the power level of the received signal is low, it is difficult to estimate the DoA. These problems could be solved if both a known reference signal and a received power can be used to estimate the DoA.

In addition, the developed algorithms can only estimate the DoA for the most powerful received signal. They cannot estimate multiple DoAs at the same time. In this thesis, it was assumed for the developed algorithms that only one desired signal is received at a time. In further studies, the proposed algorithms could be modified so that they can estimate DoAs for more than one received signal at a time.

Furthermore, the LWA-based MUSIC algorithms requires that all signals received from the different directions are the same. This means that LWA-based MUSIC algorithms is not suitable for applications that do not support transmitted signal repetition. In addition, the proposed data transmission scheme decreases the transmission rate. In the thesis, attempts were made to minimize the length of the periodical training sequence by selecting fewer radiation patterns and using two receiving ports at the same time. The algorithm is suitable for applications such as terminal position location, or when DoA estimation is required from time to time tasks such as forming a node position location table in an ad-hoc network where the nodes do not often relocate [101]. Furthermore, multipath propagation caused some problems which can be solved by using a narrower beam and wider bandwidth to improve resolution in space and time, respectively.

Throughout the thesis, the proposed DoA estimation algorithms considered that the antenna steers only on the azimuth plane. In future research, DoA estimation algorithms could be adjusted so that they could estimate DoAs at both azimuth and vertical angles. Such algorithms are also needed in future wireless communications networks. Furthermore, the author is also planning to study DoA estimation algorithms for autonomous radars.

4.3 Applications of the results

It has been estimated that in the next generation networks that DoA will play a key role in estimating the user’s location accurately. Localization has many applications, for example in emergency services, position-based billing, hostage rescue missions, etc. Adaptive directional antennas enable DoA estimation that can serve a basis for localization and location tracking. In the future generation wireless networks, location techniques need to be more accurate. For example, the Federal Communications
Commission defined in 2015 that the accuracy of network-based location has to be less than 50 m for 67% of indoor 911 calls [117]. For 5G networks, it is proposed that network-based localization in 3D space should have an uncertainty from ±10 m to less than ±1 m on 80% of the occasions and better than 1 m for indoor deployments [40]. This accuracy level cannot be reached with existing technologies. Thus, more accurate localization algorithms are needed. The significance of DoA estimation and directional communications will be further enhanced in future networks due to 5G UDNs, in which the probability of LoS conditions significantly increases.

There are lots of opportunities for high-accuracy location-based services and applications. For example, intelligent transportation systems [45], autonomous vehicles [118], and robotics [119] could all benefit from reliable, fast, and accurate localization. The use of automotive radars on vehicles will steadily increase in the future [120]. The most significant task of such radars is high-accuracy locating in all environmental conditions. This requirement is very difficult to achieve in small-sized automotive radars because the angular resolution directly depends on the size of the antenna [121]. However, high-resolution DoA estimation algorithms can overcome this size-related limitation. Furthermore, industrial and consumer systems in future Internet of Things (IoT) systems need to be able to track the direction and location of objects and persons with high accuracy. In summary, DoA estimation algorithms are effective solutions to enhance the performance of wireless communications, radar communications, and transmitter localization. In addition, a directional localization system would enable achieving the demands of future position-based services and applications and significantly enhance the radio resource management in the 5G radio network.
The primary goal of this thesis was to study DoA estimation algorithms for advanced directional antennas in future generation wireless networks. The contributions of this thesis consist of the development of the DoA estimation algorithms, as well as of the analytical, numerical, and experimental performance evaluation carried out in the thesis. The thesis particularly focused on developing DoA estimation algorithms for CRLH-LWAs and advanced antenna arrays. The main outcomes of this thesis are summarized below.

The thesis reviewed the relevant literature on advanced antennas in future generation small-cell networks. Directional antennas were presented and DoA estimation algorithms for antenna arrays and reconfigurable antennas were discussed. Novel DoA estimation algorithms for CRLH-LWAs and advanced antenna arrays were proposed and studied in this thesis.

The design of CRLH-LWAs differs from the design of conventional antenna arrays. Thus, conventional DoA estimation algorithms developed for antenna arrays cannot be directly applied to reconfigurable antennas. In this thesis, modifications were proposed to three DoA estimation algorithms to make them suitable for LWAs: modified MUSIC, LWA-based APPR, and LWA-based PPCC. The thesis showed that the spatial capabilities of traditional antenna arrays can be virtually formed with a one- or two-port antenna to receive $M$ impinging signals from $M$ different radiation patterns. The performances of these three algorithms were qualitatively compared with each other, as well as with a low-complexity PD. The DoA estimation capabilities of CRLH-LWAs were explored in additive white Gaussian noise channel measurements, experimental measurements carried out in an anechoic chamber at Drexel University, and in real-world indoor environment.

The thesis also presented improvements for the modified MUSIC and LWA-based APPR algorithms. The complexity of the algorithms was decreased by selecting a smaller number of received signals from different directions. The selection of a different radiation pattern configuration gives a different result, thus the selection of the configuration is very important.

In traditional phased array systems, all antenna elements of the array are used to adjust the direction of the beam and the shape of the radiation pattern. The algorithm
complexity significantly increases when using large antenna arrays. In this thesis, partially adaptive algorithm is studied to calculate beamforming weight coefficients only for a fraction of the array elements. The APPR DoA estimation algorithm was also applied to a smart antenna array with a reduced complexity in the calculation antenna beam steering. In addition, a smaller number of received signals from different directions was chosen in the same way as in the LWA case. By selecting only five out of 13 radiation patterns for the DoA estimation, the complexity of the APPR algorithm and storing time could be decreased. The results show that the antenna proposed for APPR DoA estimation provides the required DoA estimation accuracy. A Matlab 3D radiation pattern was modeled to demonstrate the changes of the radiation pattern in real-time during the DoA estimation measurement.

Based on the studies conducted for this thesis, it can be concluded that the proposed algorithms for DoA estimation, along with the planar and compact LWAs and advanced antenna arrays, are effective solutions for future generation wireless communications systems, where spectrum reuse, interference avoidance, and device localization play a key role. Furthermore, the contribution of the thesis constitutes a good starting point for future research on DoA estimation algorithms to enhance the performance of future generation wireless networks.
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DIRECTION OF ARRIVAL ESTIMATION ALGORITHMS FOR LEAKY-WAVE ANTENNAS AND ANTENNA ARRAYS