Conformal and multi-band antennas for future mobile handsets and wireless sensors

Kimmo Rasilainen
Conformal and multi-band antennas for future mobile handsets and wireless sensors

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A doctoral dissertation completed for the degree of Doctor of Science (Technology) to be defended, with the permission of the Aalto University School of Electrical Engineering, at a public examination held at the lecture hall AS1 of the school on 29 September 2017 at 12.
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Abstract

Desired performance of handsets and wireless sensors used in communications systems of current and future generations places significant challenges also on designing the antennas used in these devices. The antennas must adhere to requirements of existing frequency allocations, and in the case of handsets, must be compatible with multi-band and multi-antenna implementations for enhanced performance. This thesis studies the operation of electrically small antennas for particular implementations and radio systems, and gives ideas for a conceptually new class of devices.

The first part of the thesis investigates the design and performance of harmonic transponders, wireless sensors that utilise two different, harmonically separated frequencies for communication. Compared to systems with the same transmission and reception frequencies, the harmonically separated case is beneficial for detection in clutter-rich environments. Two different techniques for implementing the required harmonic matching are studied and verified experimentally. General design rules and figures of merit are presented for the suitability of particular antennas and diodes for transponder applications.

Antennas for conformal sensing and handset applications are considered in the second part of the thesis. Efficient antenna performance requires taking into account the intended usage case and environmental effects already during the design process. Bending the handset is modelled using equivalent circuit models for better physical understanding, and the results show that the most significant effects take place at frequencies below 1 GHz. Case studies are also made for a conformal, wrist-worn handset application, in which the effect of the user is also considered.

The final part of the thesis deals with the design and implementation of well-performing multi-element (MIMO) antennas. The proposed antenna structures utilise combinations of fed, passive, and parasitic elements to obtain good impedance matching and efficiency across the LTE low band (698-960 MHz). Performed computational and experimental studies show that this can be achieved using antennas with large volume but moderate surface area and fully passive matching circuits. The last point is beneficial for Carrier Aggregation implementations. The concept of physical antenna diversity can help to improve the multi-antenna performance, especially at the low band.

Keywords Harmonic transponders, mobile antennas, MIMO systems, equivalent circuit

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<tr>
<td>Väitöskirjan nimi</td>
<td>Pinna mukaiset ja monikaistaiset antennti tulevaisuuden matkaviestimien ja langattomiin antureihin</td>
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<td>Kieli</td>
<td>Englanti</td>
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**Tiivistelmä**

Nykypäivityy ja tulevien sukupolvien langattomilta viestintäjärjestelmiltä haluttava suorituskyky asettaa vaatimuksia antennisuunnittelulle sekä matkaviestimien että langattomien antureiden osalta. Antennien tulee täyttää asetetut taajuuskaistajaottelut, ja erityisesti matkaviestimien tapauksessa niiden tulee olla yhteen sopivia monibaauju- ja moniantenniratkaisujen kanssa suorituskyvyn parantamiseksi. Tämä vääköskirja käsittää sähköisiä pienten antennien toimintaa ja mallinnusta eräiden sovellusten ja radiojärjestelmien osalta, sekä esittelee mahdollisia ratkaisuja aivan uudentyypisten pääetälaitteiden toteuttamiseen.


**Avainsanat**  Harmoniset transponderit, mobiiliantennit, moniantennijärjestelmät, vastinpiiri

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Författare
Kimmo Rasilainen

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Sammandrag
Prestationsförmågan som krävs av nuvarande och framtida trådlösa kommunikationssystem ställer krav på planeringen av antenner för både mobiler och trådlösa sensorer. Antennerna måste uppfylla fordringarna på frekvensallokering, och i synnerhet med mobilerna, skall de vara förenliga med flerbands- och flerantennssystem för att förbättra prestationssförmågan. Denna avhandling behandlar elektriskt små antenners funktion och modellering för somliga tillämpningar och radiosystem samt presenterar möjliga lösningar för att realisera alldeles nya typer av mobilapparater.


Antenner som används i mobiler och harmoniska transpondrar undersöks för deras brukbarhet i konforma tillämpningar i den andra delen av avhandlingen. Antennernas goda prestanda förutsätter att användningssituationen och -omgivningen beaktas i planeringen av antennerna. En tillämpning som granskas är vikbara mobilapparater; effekterna av denna vikning undersöks med elektriskt ekvivalenta kretsar. Resultaten visar att då man viker mobilapparaten, äger de mest betydande ändringarna rum vid frekvenser under 1 GHz. För mobilapparater som placeras runt handleden granskas prestationsförmågan tillsammans med effekten av användaren på antennens funktion.


Nyckelord
Harmoniska transpondrar, mobilantenner, flerantennssystem, ekvivalent krets


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Making science and research is in many aspects similar to doing police work. I base this observation on the words of the famous Finnish detective, Inspector Palmu, created by Mika Waltari. The leading principle of Palmu — collecting the facts, placing them in the right context, and making the necessary conclusions — to my mind captures the essential parts of scientific methodology as well.

The work resulting in this thesis has been carried out at the Department of Electronics and Nanoengineering (formerly the Department of Radio Science and Engineering) during March 2013 – April 2017. I would like to express my deepest gratitude to my supervisor and instructor, Prof. Ville Viikari, for allowing me to work with many interesting research topics over the years, and for numerous insightful discussions. Additionally, I would like to thank the late Prof. Pertti Vainikainen for hiring me to his research group as summer trainee back in 2009, and Prof. Antti Räisänen for supervising my Master’s Thesis.

I am grateful to my long-time colleagues and co-authors, Dr. Anu Lehtovuori, Dr. Janne Ilvonen, Dr. Jari Holopainen, Dr. Risto Valkonen, and Mr. Jari-Matti Hannula, for a highly fruitful and useful collaboration over the years. Through the work done together, I have learned many practical and theoretical things in radio engineering and circuit theory. Our stimulating conversations on all matters technical and non-technical remain a constant source of inspiration and encouragement in research and, more broadly, life in general. You have been great company on various business trips as well. Thanks also to Mr. Amine Boussada for successful collaboration in the summer of 2016. Furthermore, I want to thank Dr. Clemens Icheln for skilful project management over the years, as well as for organising the RAP discussions, a source for many a useful information.

I would like to thank the preliminary examiners of this thesis, Prof. Eva
Preface

Antonino Daviu and Prof. Fabien Ferrero, for a thorough and timely examination of the thesis.

I want to acknowledge the members of our research group as well as current and former colleagues of the department for creating a nice and friendly working environment. Help received from the service and technical personnel over the years is also highly appreciated.

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Finally, I would like to thank my parents for all the encouragement, help, and support I have received throughout my studies.

Espoo, August 30, 2017,

Kimmo Rasilainen
## Contents

1. Preface .............................. 1
2. Contents ................................ 3
3. List of Publications .................. 5
4. Author’s Contribution .................. 7
5. List of Abbreviations .................. 11
6. List of Symbols .......................... 13
7. 1. Introduction ............................ 17
   1.1 Background ............................ 17
   1.2 Objective of this work .................. 19
   1.3 Contents and organisation of the thesis .......................... 19
   1.4 Main scientific merits .................... 20
8. 2. Harmonic Radar and Transponders ....... 21
   2.1 General information on wireless sensors .............. 21
   2.2 Theory and applications of harmonic radar .............. 22
      2.2.1 Expression for the harmonic response .............. 25
      2.2.2 Figures of merit for the transponder antenna and diode 27
   2.3 Transponder implementation based on direct matching ...... 28
      2.3.1 Designing the transponder antenna .............. 29
      2.3.2 Characterisation of transponder performance ........ 33
      2.3.3 Transponder operation with different diodes ........ 35
   2.4 Transponder implemented using lumped-component matching circuits 39
      2.4.1 Diode selection .............. 39
      2.4.2 Antenna geometry .............. 40
List of Publications

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


Author’s Contribution


This work is the result of collaborative work, and it is based on an original idea by Prof. Ville Viikari, who also participated in deriving the theoretical basis in collaboration with the author and Dr. Anu Lehtovuori. Dr. Janne Ilvonen assisted in the antenna design and helped with the antenna measurements. Mr. Jari-Matti Hannula developed the measurement setup used and participated in the antenna measurements. The author had the main responsibility for preparing the manuscript.

Publication II: “Antenna Matching at Harmonic Frequencies to Complex Load Impedance”

This work extends the analysis of the antenna part of Publication I. The author had main responsibility for the contents of the paper. He designed, simulated and measured the antenna, as well as wrote the manuscript. Dr. Janne Ilvonen and Prof. Ville Viikari instructed and supervised the work.

Publication III: “Harmonic Transponders: Performance and Challenges”

This work extends the study performed in Publication I. The author had the main responsibility for the contents of the paper. Dr. Janne Ilvonen, Dr. Anu Lehtovuori, and Mr. Jari-Matti Hannula participated in analysing the results and commented on the manuscript. Prof. Ville Viikari participated in the theoretical basis development.
Author's Contribution

kari supervised the work.

Publication IV: “Designing Harmonic Transponders Using Lumped-Component Matching Circuits”

The author designed the matching-circuit based harmonic transponders with assistance from Dr. Janne Ilvonen. Mr. Jari-Matti Hannula participated in the measurements together with Dr. Janne Ilvonen. The author was responsible for writing the manuscript. Prof. Ville Viikari supervised the work.

Publication V: “Effect of Shape and Surroundings on Harmonic Transponder Performance”

The author had the main responsibility for the contents of the paper. He performed all simulations and calculations and wrote the manuscript. Prof. Ville Viikari supervised the work.

Publication VI: “Effect of Ground Plane Bending on Mobile Terminal Antenna Performance”

This work is based on the author’s Master’s thesis and on the methodology used in it. The author designed the antenna structures, performed all computer simulations, and also wrote the manuscript. Dr. Janne Ilvonen instructed the work. Dr. Jari Holopainen and Dr. Risto Valkonen assisted in implementing the equivalent circuit models. Prof. Antti Räisänen supervised the work.

Publication VII: “Investigation on Bendable Mobile Devices in the Presence of the User”

This work was inspired by the author's Master's thesis and continues on the topic of Publication VI. The author had the leading role in preparing the publication. He designed, simulated and analysed the antennas, as well as wrote the manuscript. Dr. Janne Ilvonen and Prof. Ville Viikari instructed and supervised the work.
Publication VIII: “LTE Handset Antenna with Closely-Located Radiators, Low-Band MIMO, and High Efficiency”

The author had the main responsibility in the work. He designed the proposed antennas and matching circuits, performed the simulations, and wrote the manuscript. Dr. Anu Lehtovuori instructed the work and Prof. Ville Viikari supervised the work.

Publication IX: “Carrier Aggregation Compatible MIMO Antenna for LTE Handset”

The author designed the antennas and matching circuits. He manufactured the prototypes, carried out the measurements, and wrote the manuscript. Dr. Anu Lehtovuori instructed the work. Mr. Amine Boussada assisted in the matching circuit design. Prof. Ville Viikari supervised the work.
<table>
<thead>
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<th>Full Form</th>
</tr>
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<tr>
<td>4G</td>
<td>Fourth Generation</td>
</tr>
<tr>
<td>5G</td>
<td>Fifth Generation</td>
</tr>
<tr>
<td>C</td>
<td>Capacitor</td>
</tr>
<tr>
<td>CA</td>
<td>Carrier Aggregation</td>
</tr>
<tr>
<td>CC</td>
<td>Component Carrier</td>
</tr>
<tr>
<td>CCE</td>
<td>Capacitive Coupling Element</td>
</tr>
<tr>
<td>CMOS</td>
<td>Complementary Metal Oxide Semiconductor</td>
</tr>
<tr>
<td>CPAM</td>
<td>Combined Parasitic-coupled Aperture Matched</td>
</tr>
<tr>
<td>CTIA</td>
<td>Wireless Association</td>
</tr>
<tr>
<td>(former Cellular Telecommunications Industries Association)</td>
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<tr>
<td>ECC</td>
<td>Envelope Correlation Coefficient</td>
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<tr>
<td>EM</td>
<td>Electromagnetic</td>
</tr>
<tr>
<td>FDTD</td>
<td>Finite Difference Time Domain</td>
</tr>
<tr>
<td>FDx</td>
<td>Full Duplex</td>
</tr>
<tr>
<td>FIT</td>
<td>Finite Integration Technique</td>
</tr>
<tr>
<td>FOM</td>
<td>Figure of Merit</td>
</tr>
<tr>
<td>GSM</td>
<td>Global System for Mobile Communications</td>
</tr>
<tr>
<td>ILA</td>
<td>Inverted-L Antenna</td>
</tr>
<tr>
<td>IoT</td>
<td>Internet of Things</td>
</tr>
<tr>
<td>ISM</td>
<td>Industrial, Scientific, and Medical</td>
</tr>
<tr>
<td>L</td>
<td>Inductor</td>
</tr>
<tr>
<td>LTE-A</td>
<td>Long Term Evolution Advanced</td>
</tr>
<tr>
<td>MEMS</td>
<td>MicroElectroMechanical Systems</td>
</tr>
<tr>
<td>MIMO</td>
<td>Multiple-Input Multiple-Output</td>
</tr>
<tr>
<td>PEC</td>
<td>Perfect Electric Conductor</td>
</tr>
<tr>
<td>PIFA</td>
<td>Planar Inverted-F Antenna</td>
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<tr>
<td>R</td>
<td>Resistor</td>
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<td>RF</td>
<td>Radio Frequency</td>
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List of Abbreviations

RFID  Radio Frequency Identification
RLC  Resistor, Inductor, and Capacitor
Rx  Receiver
SAR  Specific Absorption Rate
SISO  Single-Input Single-Output
SNR  Signal-to-Noise Ratio
Tx  Transmitter
UHF  Ultra High Frequency
WLAN  Wireless Local Area Network
List of Symbols

- $2f_0$: second harmonic frequency
- $C_{c1}, C_{c2}$: capacitance of the first and second order chassis wavemodes in the planar case
- $C_{c1,b}, C_{c2,b}$: capacitance of the first and second order chassis wavemodes in the bent case
- $C_{j0}$: small-signal junction capacitance
- $C_{mm}$: capacitance of the monopole mode
- $C_p$: parasitic capacitance
- $FOM_a$: figure of merit (antenna)
- $FOM_d$: figure of merit (diode)
- $F_1, F_2, F_3, F_4$: abbreviation terms in the expression for harmonic response
- $f$: frequency
- $f_r$: resonance frequency
- $f_{rc}$: diode threshold frequency
- $f_{rc,det}$: detector diode threshold frequency
- $f_{rc,var}$: varactor diode threshold frequency
- $f_0$: fundamental (first harmonic) frequency
- $I_j$: junction current
- $I_s$: saturation current
- $j$: imaginary unit
- $L_{c1}, L_{c2}$: inductance of the first and second order chassis wavemodes in the planar case
- $L_{c1,b}, L_{c2,b}$: inductance of the first and second order chassis wavemodes in the bent case
- $L_{\text{comp},f_0}, L_{\text{comp},2f_0}$: component losses at $f_0$ and $2f_0$
- $L_{\text{conv}}$: conversion loss
- $L_{\text{mm}}$: inductance of the monopole mode
<table>
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<th>Description</th>
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<td>$L_{\text{path, } f_0}$, $L_{\text{path, } 2f_0}$</td>
<td>path loss at $f_0$ and $2f_0$</td>
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<tr>
<td>$L_s$</td>
<td>series inductance</td>
</tr>
<tr>
<td>$n$</td>
<td>harmonic index</td>
</tr>
<tr>
<td>$n_1$, $n_2$</td>
<td>transformer coupling coefficient to the first and second order chassis wavemodes in the planar case</td>
</tr>
<tr>
<td>$n_{1,b}$, $n_{2,b}$</td>
<td>transformer coupling coefficient to the first and second order chassis wavemodes in the bent case</td>
</tr>
<tr>
<td>$P_{\text{in}}$</td>
<td>input power</td>
</tr>
<tr>
<td>$P_{\text{loss}}$</td>
<td>power dissipated in the antenna</td>
</tr>
<tr>
<td>$P_{r, 2f_0}$</td>
<td>calculated power received by the reader at $2f_0$</td>
</tr>
<tr>
<td>$P_{r, 2\omega_0}$</td>
<td>power received by the reader at $2\omega_0$</td>
</tr>
<tr>
<td>$P_{t, f_0}$</td>
<td>transmit power at $f_0$</td>
</tr>
<tr>
<td>$Q$</td>
<td>quality factor</td>
</tr>
<tr>
<td>$Q_a$</td>
<td>antenna quality factor</td>
</tr>
<tr>
<td>$Q_{a, \omega_0}$, $Q_{a, 2\omega_0}$</td>
<td>antenna quality factor at angular frequencies $\omega_0$ and $2\omega_0$</td>
</tr>
<tr>
<td>$Q_d$</td>
<td>diode quality factor</td>
</tr>
<tr>
<td>$Q_{d, \omega_0}$, $Q_{d, 2\omega_0}$</td>
<td>diode quality factor at angular frequencies $\omega_0$ and $2\omega_0$</td>
</tr>
<tr>
<td>$Q_j(V_j)$</td>
<td>charge stored in the diode junction</td>
</tr>
<tr>
<td>$Q_a$</td>
<td>quality factor calculated from input impedance</td>
</tr>
<tr>
<td>$R_a$</td>
<td>antenna resistance</td>
</tr>
<tr>
<td>$R_{a, \omega_0}$, $R_{a, 2\omega_0}$</td>
<td>antenna resistance at angular frequencies $\omega_0$ and $2\omega_0$</td>
</tr>
<tr>
<td>$R_d$</td>
<td>diode resistance</td>
</tr>
<tr>
<td>$R_j$</td>
<td>junction resistance</td>
</tr>
<tr>
<td>$R_{\text{loss}}$</td>
<td>loss resistance</td>
</tr>
<tr>
<td>$R_{\text{mm}}$</td>
<td>resistance of the monopole mode</td>
</tr>
<tr>
<td>$R_{\text{rad}}$</td>
<td>radiation resistance</td>
</tr>
<tr>
<td>$R_s$</td>
<td>series resistance</td>
</tr>
<tr>
<td>$R_{Rx}$</td>
<td>receiver distance</td>
</tr>
<tr>
<td>$R_{Tx}$</td>
<td>transmitter distance</td>
</tr>
<tr>
<td>$r_b$</td>
<td>bending radius</td>
</tr>
<tr>
<td>$V_j$</td>
<td>junction voltage</td>
</tr>
<tr>
<td>$W$</td>
<td>stored energy</td>
</tr>
<tr>
<td>$W_e$</td>
<td>stored electric energy</td>
</tr>
<tr>
<td>$W_m$</td>
<td>stored magnetic energy</td>
</tr>
<tr>
<td>$X_a$</td>
<td>antenna reactance</td>
</tr>
<tr>
<td>$X_d$</td>
<td>diode reactance</td>
</tr>
<tr>
<td>$Z_a$</td>
<td>antenna impedance</td>
</tr>
<tr>
<td>$Z'_a$</td>
<td>antenna impedance with series inductance</td>
</tr>
<tr>
<td>Symbol</td>
<td>Description</td>
</tr>
<tr>
<td>--------</td>
<td>-------------</td>
</tr>
<tr>
<td>$Z_D$</td>
<td>diode impedance</td>
</tr>
<tr>
<td>$Z_D(f)$</td>
<td>complex, frequency-dependent diode impedance</td>
</tr>
<tr>
<td>$Z_d'$</td>
<td>diode impedance with packaging</td>
</tr>
<tr>
<td>$Z_{in}$</td>
<td>input impedance</td>
</tr>
<tr>
<td>$Z_j$</td>
<td>junction impedance</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>inverse of diode thermal voltage</td>
</tr>
<tr>
<td>$\gamma$</td>
<td>profile parameter</td>
</tr>
<tr>
<td>$\varepsilon_r$</td>
<td>relative permittivity</td>
</tr>
<tr>
<td>$\tilde{\eta}_{\text{mux}}$</td>
<td>multiplexing efficiency</td>
</tr>
<tr>
<td>$\eta_{\text{rad}}$</td>
<td>radiation efficiency</td>
</tr>
<tr>
<td>$\eta_{\text{tot}}$</td>
<td>total efficiency</td>
</tr>
<tr>
<td>$\eta_{\omega_0}, \eta_{2\omega_0}$</td>
<td>antenna efficiency at angular frequencies $\omega_0$ and $2\omega_0$</td>
</tr>
<tr>
<td>$\lambda$</td>
<td>wavelength</td>
</tr>
<tr>
<td>$\lambda_0$</td>
<td>wavelength in free space</td>
</tr>
<tr>
<td>$\rho_e$</td>
<td>envelope correlation coefficient</td>
</tr>
<tr>
<td>$\sigma$</td>
<td>electric conductivity</td>
</tr>
<tr>
<td>$\Phi$</td>
<td>junction potential</td>
</tr>
<tr>
<td>$\omega$</td>
<td>angular frequency</td>
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1. Introduction

1.1 Background

Nowadays, the world around us is increasingly global and networked. Major enablers of this development are efficient and well-performing communications networks, and the amount of wireless communications is constantly increasing. Examples of this include the use of handsets and wireless local area network (WLAN) services, and also that of various wireless sensing systems.

A major change or development that has taken place, and will also occur even more prominently in the future, is a significant increase in both the number of networked devices with wireless connectivity and in the amount of data that is transferred wirelessly. Mobile devices of previous generations used voice and text as the main means of communications, and consequently, the resulting overall data transfer was at a moderate, even low, level. In comparison, the smartphone and tablet devices of today are used in increasing amounts for various data-hungry applications, such as web browsing and live video streaming. This development and the resulting communications requirements place constraints and challenges for both network and antenna designers.

One topic that has become something of a buzzword in the wireless community is the coming fifth-generation (5G) communications systems, which are expected to be launched around 2020 [1–3]. Sometimes, these systems are also referred to as Internet of Things (IoT), or even “Internet of Everything”. Some of the new features in these networks include the use of higher data rates and operating frequencies, and also a dramatic increase in the number of connected devices. As an example, devices such as future domestic appliances (refrigerators, washing machines, or TV sets)
are envisioned to feature wireless connectivity. In term of volumes, this might translate to several billions of new ‘users’ crowding the networks.

Despite an ever-increasing desire for more data traffic occurring simultaneously, faster, and for a larger amount of users, wireless communications is restricted by the availability of a fundamental ‘natural resource’: radio-frequency spectrum. All radio systems that are used must coexist with each other in certain frequency ranges or channels, exact frequencies of which depend in part on the system, and in part on frequency allocations in different parts of the world. A somewhat analogous situation to dealing with an increasingly crowded spectrum is coping with traffic volumes on the motorways. In order to accommodate more vehicles on the road, new lanes and other means of controlling the traffic flow are needed.

In the wireless world, utilising the electromagnetic equivalent of lanes — new frequencies and frequency bands — can only go so far. Limitations of spectrum availability have forced designers and engineers to develop more efficient techniques for using the spectrum, such as full-duplex (FDx) technologies that theoretically allow doubling the data rate by utilising the same frequency in transmission and reception (see, e.g., [4]). Other possibilities include the use of Multiple-Input Multiple-Output (MIMO) technology, in which several antennas are used to transmit and/or receive data at the same frequency band or different bands, and Carrier Aggregation (CA) technology utilising several narrow frequency channels across designated bands for enhanced communications.

Wireless technologies are not only used for communications. Another important area, whose global significance is also constantly increasing, is the use of wireless sensing. Examples of areas in which this technology is utilised include tracking of assets, objects and livestock, various medical applications, and obtaining information on different quantities or parameters of objects to which sensors are connected. Depending on the particular implementation, various technologies can be used. For the general public, one of the best-known technologies is radio-frequency identification (RFID) tags. By placing these tags on an object, information from the tag can be read wirelessly across a distance without physical contact to the target.

In all of the previously discussed cases, and in any wireless system, a device of fundamental importance for achieving efficient performance is an antenna [5–7]. The dimensions of devices used in typical applications, such as handsets, are such that the antennas used are physically rather
small. For the operation of these antennas at the desired frequencies, a rather unfortunate consequence is that the electrical size of the antenna also becomes small. An electrically small antenna has physical dimensions that are small compared to the wavelength: an antenna is defined to be electrically small if it can be enclosed inside a sphere with radius smaller than $\lambda_0/2\pi$, in which $\lambda_0$ is the wavelength in free space [8]. Subsequently in this thesis, the term ‘small antenna’ refers to an antenna that is electrically small. A well-known “rule of thumb” regarding small antennas is that they can have good simultaneous performance with respect to two of the following quantities: size, bandwidth, and efficiency. Designers of small antennas have to properly take this issue into account.

Even though the previous might seem to warn against the use of electrically small antennas, they are still among the first antennas ever used. Indeed, during the days of Marconi and other pioneers of radio engineering, the antennas that were used were physically large (due to low frequencies), but electrically — compared to the wavelength — they were still small. In some situations, the concept of infinite ground plane considerably simplifies calculations. In Marconi’s time, the ground plane was, for all practical purposes, infinite, as this part was played by planet Earth.

1.2 Objective of this work

The general framework of this thesis studies wireless connectivity and communications from the point of view of antenna design. Two main application areas for the antennas are considered, namely wireless sensors and handsets. The first approach considers the analysis and modelling of the properties and performance of a particular type of wireless sensor, the so-called harmonic transponder. The second part looks at antennas for handset applications, including bendable mobile devices, and at developing well-performing multi-antenna (MIMO) solutions using antennas with closely-located radiating elements.

1.3 Contents and organisation of the thesis

This thesis consists of an overview and nine publications [I]–[IX], and it is organised as follows. Chapter 2 presents an overview of wireless sensors, and provides a detailed analysis on the design and modelling of harmonic
transponders [I]–[IV]. In Chapter 3, the suitability of conformal antennas for wireless sensing and handset applications is investigated [V]–[VII]. Chapter 4 discusses the challenges of current and future handset antenna implementations, and presents novel design solutions to overcome some of these challenges [VIII]–[IX].

1.4 Main scientific merits

The main scientific merits of the thesis are as follows:

1. Analysis, design, and modelling of harmonic transponders, wireless sensors operating at two harmonically spaced frequencies, are performed theoretically, computationally, and experimentally. Two different ways of matching the transponder antennas to typical, complex load impedances at harmonically-spaced frequencies are presented [I]–[V].

2. Investigation on the suitability of harmonic transponders to conformal applications. Effects of the intended usage environment such as material parameters have to be taken into account in the transponder design to maintain good performance at harmonically-spaced frequencies [V].

3. Analysis, design, and modelling of handset antenna structures for bendable applications. Theoretical understanding of changes in the operation of a Capacitive Coupling Element (CCE) handset antenna, and a case study of utilising such antennas in wrist-worn applications are presented [VI]–[VII].

4. Design and implementation of a novel, efficient LTE handset antenna utilising a combination of active and parasitic antenna elements. By placing the radiators around or above the edges of the device, very good performance is achieved also at the low band with a fully passive implementation [VIII].

5. Design and experimental verification of a two-element LTE MIMO handset antenna with a novel Combined Parasitic-coupled Aperture-Matched (CPAM) main antenna and conventional CCE diversity antenna. The design has low- and high-band MIMO with good performance, and the design is CA compatible due to a fully passive matching implementation to cover both the low and high bands [IX].
2. Harmonic Radar and Transponders

The use of wireless communications for sensing and detecting different objects is gaining more and more attention, even though some of the basic building-block technologies have been used for a long time. One early implementation is the use of radio detection and ranging (radar) for tracking objects such as aeroplanes, and these systems date back to radar implementations such as the British Chain Home radar system during World War II. In most of the conventional radar applications, the transmitter transmits a signal, and the direction and time of arrival of the reflected or return signal provide an estimate of the direction and distance of the object being tracked.

This chapter presents general theory and basic principles of wireless sensors and harmonic radar in Sections 2.1 and 2.2, and explains the theoretical and experimental work performed on harmonic transponders in Sections 2.3 and 2.4. This type of sensor is sometimes also called harmonic RFID (e.g., in [9]). The performance and ease of implementation is considered for transponders based on different techniques to achieve the desired operating frequencies, and comparisons are made in Section 2.5.

2.1 General information on wireless sensors

Depending on the intended application and communication principle, different types of wireless sensors can be utilised. One well-known commercial example is the use of RFID tags, which are typically inexpensive, produced in large quantities, and allow positioning or obtaining other information of the object onto which the tag is placed [10–13].

In addition to the technique or method used for communication, wireless sensors can be categorised based on their power supply. Three main categories can be identified: active, semi-passive, and passive sensors. Each of
these sensor types has its own benefits and drawbacks. Generally, using a sensor with an internal battery tends to increase the size, complexity and cost of the sensor. Additionally, active implementations are limited by battery lifetime, and the presence of the battery may also prevent using the sensor in certain applications. Compared to active sensors that use their own battery to run an active radio transmitter, semi-passive sensors communicate back to the reader without external power. Even though these sensors also have a battery, it is used to operate, e.g., the memory of the chip, not for communications.

The current work considers wireless sensors that are fully passive. This means that they gather all of the energy required for their operation from the ambient electromagnetic fields, namely from the interrogation signal used to detect the sensor. Thus, fully passive sensors do not include any battery or other external power supply. Compared to active sensors, passive ones are simpler and more reliable, but the power levels required for obtaining a response from the sensor can limit the achievable read-out range [10]. In some cases, simplifying the sensor design may also limit the level of specificity with which sensing information can be obtained [14].

A common feature shared by all forms of wireless sensing is that the sensor, tag, or transponder is interrogated by a reader device (transmit or Tx signal), and in the case of successful interrogation, the reader receives a signal (receive or Rx signal) from the object under study. With different systems and communications schemes, the exact frequencies used vary, but typically both the Tx and Rx signals are located within a certain, continuous frequency band. For instance, the operating band specified for RFID systems is 865–870 MHz and 902–920 MHz in Europe and in the U.S., respectively [15, 16].

2.2 Theory and applications of harmonic radar

For conventional wireless sensors, utilising the same frequency band for transmission and reception can cause challenges for reliable detection in certain types of environments. For instance, if the sensor or radar is used in a location having strong environmental clutter, it may be complicated to distinguish between signals arriving from the actual target and those simply reflecting from the environment.

One solution to this issue is to use transmit and receive signals at different frequencies in such a way that these frequencies have a particular
Harmonic radar is based on this approach, and it has Tx and Rx signals that are harmonic, integer multiples of each other. Figure 2.1 illustrates this concept on a general level, and also provides a comparison to regular, linear radar. Due to the harmonically-spaced operating frequencies, a signal picked up at the expected harmonic multiple of the original interrogation signal is more probably caused by the object rather than by environmental reflections. At typical power levels used in wireless sensing, most natural and man-made objects can be considered linear, and they therefore do not convert the fundamental-frequency signal to higher-order harmonics.

The concept of harmonic radar was first introduced in [18], and many applications use a fundamental frequency \( f_0 \) and the second harmonic frequency \( 2f_0 \) [19]. Some works, such as [20, 21], consider the use of the third harmonic frequency \( 3f_0 \) in UHF RFID applications. The predominant use of the second harmonic is mainly due to three facts: 1) existing frequency regulations, 2) at smaller harmonics, the transponder response is less power-dependent (conversion loss tends to grow exponentially as the harmonic index \( n \) increases), and 3) path loss increases with frequency (consequently, the read-out distance may decrease).

Harmonic radar is a special case of the more general nonlinear radar, in which two closely-located frequencies \( f_1 \) and \( f_2 \) are transmitted, and nonlinearities in the object or transponder create a response signal at an

![Figure 2.1. Simplified, schematic representation of wireless sensing based on linear and nonlinear communications. Modified from [17].](image-url)
intermodulation frequency (e.g., $2f_1 - f_2$) [22, 23]. In some literature, this kind of radar is also called harmonic. Due to the closeness of $f_1$ and $f_2$, the response signal is also in their vicinity. This makes complying with existing frequency allocations easier than in the case of purely harmonic radar. The same principle of generating intermodulation signals is utilised in intermodulation-based wireless sensors, as explained in, e.g., [24].

Generation of the harmonic frequencies in the transponder relates to the more general phenomenon of nonlinearity. If a system or device is fully linear, its response is additive and homogeneous. This means that the output of a linear system is the sum of the input signal(s), possibly scaled by a constant (amplification/attenuation). In terms of frequencies, the output signal of a linear system contains exactly the same frequency components as the original input signal. At sufficiently high power levels, or with devices whose operation is based on nonlinearity (such as mixers), the nonlinearities create new frequencies to the output signal that did not exist in the input signal.

A general overview of harmonic transponders and their implementations can be found in, e.g., [17]. Examples of applications in which harmonic transponders have been utilised include tracking of flying and walking insects [25, 26] and detection of avalanche victims [27]. Harmonic and nonlinear radar has also been applied for studying the corrosion of steel reinforcements in concrete structures [28], as these are also capable of producing a nonlinear response to the incident signal. In [29], the implementation of a gas sensor using carbon nanotubes and harmonic radar has been investigated.

Operating frequencies in harmonic radar applications vary, and typically, three major frequency ranges are used. The first one operates at a fundamental frequency of 0.917 GHz, and it is based on the RECCO avalanche detector [27]. Another, maritime radar-based technology uses fundamental frequencies around 9.4 GHz (e.g., [25, 26, 30]). The unlicensed Industrial, Scientific, and Medical (ISM) band is used in the third approach, and this type of transponder has fundamental frequencies at 2.4, 5.8, or 5.9 GHz (e.g. [31,32]). Some studies, such as [33] have also considered harmonic transponders for millimetre-wave frequencies, which are generally not used for this purpose. One main reason is higher path loss, which will reduce the read-out range of high-frequency transponders.
2.2.1 Expression for the harmonic response

To describe the performance and response obtainable from a harmonic transponder, let us first consider the equivalent circuit model illustrated in Figure 2.2, which shows an antenna connected to a Schottky diode. A small-signal representation is used for the diode, and the antenna is represented with a Thévenin equivalent circuit. The antenna impedance is expressed as $Z_a = R_a + jX_a$, where $R_a = R_{\text{rad}} + R_{\text{loss}}$ consists of radiation and loss resistance, respectively, and $X_a$ represents the antenna reactance. This formulation does not make any assumptions on the way in which matching is implemented, and for the case of having an external matching circuit, its impedance can be included in $Z_a$.

For the Schottky diode, the circuit model includes a series inductance $L_s$, a parasitic capacitance $C_p$, a series resistance $R_s$, a small-signal junction capacitance $C_{j0}$, and a current source. The diode packaging and its effects are modelled with the terms $L_s$ and $C_p$ [34].

A very important property of the harmonic transponder related to its practical implementation and utilisation is the amount of backscattered power the transponder is able to generate. Operating at two distinct frequencies, the transponder scatters back towards the reader device a signal that has been converted in the transponder from $f_0$ to $2f_0$. In this work, the backscattered signal (power) is referred to as harmonic response, and it represents the signal received by the reader device. This quantity takes into account both the converted power scattered by the transponder and additional path loss due to the read-out distance.

The circuit model in Figure 2.2 is utilised in [I] and [III] to determine the harmonic response generated by a given combination of antenna and diode. These works derive and analyse an expression for the harmonic response.
Harmonic Radar and Transponders

response, which is given as

\[ P_{r,2\omega_0} = |F_1 F_2 F_3 F_4|, \]  \hfill (2.1)

where

\[ F_1 = \left( \frac{Z_{j,\omega_0}}{R_{s,\omega_0} + Z_{j,\omega_0}} \right)^4 \left( \frac{Z'_{d,\omega_0}}{Z'_{a,\omega_0} + Z'_{d,\omega_0}} \right)^4 \]

\[ F_2 = \left( \frac{Z_{a,2\omega_0}}{R_{s,2\omega_0} + Z_{a,2\omega_0}} \right)^2 \left( \frac{Z'_{d,2\omega_0}}{Z'_{a,2\omega_0} + Z'_{d,2\omega_0}} \right)^2 \]

\[ F_3 = \eta_0^4 R_{a,\omega_0}^2 \eta_{2,\omega_0}^2 R_{a,2\omega_0}^2 \]

\[ F_4 = \left( \frac{\omega_0 C_{j0} \gamma}{2\Phi} - \frac{j\alpha}{4R_j} \right)^2. \]

Here, the junction impedance is expressed as \( Z_j = \left( j\omega C_j + 1/R_j \right)^{-1} \), and the following expressions are used to simplify the above notations: \( Z'_a = Z_a + j\omega L_a \) and \( (Z'_d)^{-1} = j\omega C_p + (R_s + Z_j)^{-1} \). These follow the notation used in Figure 2.2 as well as in [I] and [III].

The above equation has been derived under small-signal conditions, and the calculation has been done in three parts. First, we determine the signal going from antenna to diode at \( f_0 \). Then, the diode provides frequency conversion from \( f_0 \) to \( 2f_0 \), and we calculate the (frequency-)modulated currents resulting from this conversion. In the third and final stage, the signal provided to the antenna by an equivalent current source at \( 2f_0 \) is calculated to get the actual harmonic signal transferred from the diode back to the antenna. Furthermore, it is assumed that the junction current and capacitance of the diode are

\[ I_j(V_j) = I_s \left( e^{\alpha V_j} - 1 \right) + \frac{d}{dt} Q_j(V_j) \]  \hfill (2.2)

and

\[ C_j = \frac{C_{j0}}{\left( 1 - \frac{V_j}{\Phi} \right)^\gamma}. \]  \hfill (2.3)

respectively. Here, \( Q_j(V_j) \) is the charge stored in the diode junction and \( I_s \) is the saturation current.

The terms \( F_1 - F_4 \) in Equation (2.1) can be categorised based on what kind of contribution on the overall response they describe. From them, a number of important observations can be made. Terms \( F_1 - F_3 \) include all the antenna-related properties of the response, including impedance matching and radiation efficiency, and the \( F_4 \) term includes the contribution of the diode mixing properties.
As can be seen in expressions $F_1$–$F_3$, the terms related to $f_0$ have a greater impact on the overall harmonic response than those related to $2f_0$, as the $f_0$ and $2f_0$ terms are proportional to the fourth and second power, respectively. This is true both matching and efficiency-wise, implying that for the overall transponder performance, it is beneficial to aim for as good a matching (and efficiency) performance at $f_0$ as possible.

The diode mixing properties term $F_4$ has two components: a frequency-dependent, capacitive nonlinearity (term $\omega_0 C_j \gamma / 2 \Phi$) and a frequency-independent, resistive nonlinearity (term $-j \alpha / 4 R_j$). Different diodes, irrespective of whether they are of varactor or detector type, have both types of nonlinearity. Depending on the frequency, one of them can have a considerably larger value, meaning that at that frequency, either the capacitive or resistive nonlinearity is mainly responsible for the frequency conversion. When equating the nonlinearity terms, it is possible to determine a threshold frequency ($f_{rc}$), above and below which it is beneficial to use a diode with good capacitive or resistive mixing properties, respectively. This calculation has been performed in [I] assuming parameter values typical for varactor and detector diodes. Based on the calculations, the threshold frequency for varactors is $f_{rc,\text{var}} = 1$ Hz and $f_{rc,\text{det}} = 848$ MHz for detector diodes. This means that at frequencies typically utilised for wireless sensing, such as at the allocated RFID bands and above, suitable diodes should be chosen based on their capacitive mixing properties.

### 2.2.2 Figures of merit for the transponder antenna and diode

As is presented in [I], the previously described transponder equivalent-circuit model can also be utilised to formulate certain figures of merit (FOM). These describe the suitability of particular antennas ($\text{FOM}_a$) and diodes ($\text{FOM}_d$) for a given transponder implementation.

Regarding efficient transponder performance, it is important to be able to accept as much of the incident power at $f_0$ as possible, to efficiently convert this power to $2f_0$, and to scatter back as efficiently as possible towards the reader device at $2f_0$. One key parameter that significantly affects the overall operation of the transponder is impedance matching at both frequencies.

For the antenna, a figure of merit is obtained by combining all antenna-
Harmonic Radar and Transponders

related terms of \(F_1\)–\(F_4\) in Equation (2.1) into the following expression

\[
FOM_a = \left| \frac{Z'_{d,\omega_0}}{Z'_{d,\omega_0} + Z'_{a,\omega_0}} \right|^4 \eta_{\omega_0}^4 R_{a,\omega_0}^2 \left( \frac{Z'_{d,2\omega_0}}{Z'_{a,2\omega_0} + Z'_{d,2\omega_0}} \right)^2 \eta_{2\omega_0}^2 R_{a,2\omega_0}^2. \tag{2.4}
\]

In this expression, \(Z'_d\) represents all diode-related quantities, and all other terms describe the properties of the antenna. With the help of the \(FOM_a\), it is possible to investigate and compare the suitability of different antenna structures when using a certain diode. This figure of merit also allows to see how, e.g., compromises between antenna matching and efficiency affect the overall transponder performance.

Compared to the \(FOM_a\) parameter, in which all diode-related properties are reduced to a single term, the opposite is done in the following expression for the optimal power received by the reader at \(2f_0\)

\[
P_{r,\text{opt},2\omega_0} \approx 16 P_{\text{in}}^2 \cdot \frac{Z_{j,\omega_0}}{R_{a,\omega_0} + Z_{j,\omega_0}} \left( \frac{Q_{d,\omega_0}}{1 + Q_{d,\omega_0}/Q_{a,\omega_0}} \right)^4 \Re\{Z'_{d,\omega_0}\}^2 \times \left( \frac{Z_{j,2\omega_0}}{R_{a,2\omega_0} + Z_{j,2\omega_0}} \right)^2 \left( \frac{Q_{d,2\omega_0}}{1 + Q_{d,2\omega_0}/Q_{a,2\omega_0}} \right)^2 \Re\{Z'_{d,2\omega_0}\}^2 \times \frac{\omega_0 C_{j0} \gamma}{2 \Phi} - \frac{j\alpha}{4 R_{j}} \right|^2. \tag{2.5}
\]

Assuming that the quality factor of the antenna is limited to \(Q_a\), and that the impedance level of the antenna can be chosen at will, the above expression depends entirely on the diode properties (except for \(Q_a\) and the input power \(P_{\text{in}}\)). The figure of merit for the diode becomes \(FOM_d = P_{r,\text{opt},2\omega_0}/16 P_{\text{in}}^2\). In this case, critical antenna matching is assumed, and that the diode quality factor \(Q_d = |\Im\{Z'_d\}|/|\Re\{Z'_d\}| \gg 1\). Under the given assumptions and conditions, the \(FOM_d\) term gives information about the suitability of different diodes for a given antenna design. From a particular selection of diodes, the one resulting in the largest \(FOM_d\) value should be chosen.

\subsection*{2.3 Transponder implementation based on direct matching}

Transponders can be designed for various applications, frequencies and purposes. Depending on the particular implementation, details such as
shape and materials may vary, but two basic building blocks need to be found in all transponders:

1. One or more antennas to take care of the communication between the transponder and transmitter/receiver.

2. A diode or other nonlinear element to provide the necessary frequency multiplication from the fundamental frequency to the required harmonic.

There are different techniques that can be used to match the antennas at the desired harmonic frequencies. The first alternative considered here is based on so-called direct matching. In this approach, the geometrical details of the antenna are modified in such a way that proper (ideally, conjugate) matching is obtained at the targeted frequencies without using dedicated matching circuits or components.

Direct matching is commonly used to implement conventional RFID tag antennas, both for single- and dual-band applications. Examples of this can be found in, e.g., [35,36]. As a matching problem, the impedance levels of regular RFID chips are somewhat comparable to those of typical diodes used in harmonic transponders. With dual-band RFID tags, the operating frequencies are not harmonically dependent, which simplifies the design. Studies carried out within the framework of this thesis investigate both varactor and detector diodes from Skyworks [37] and Avago (currently part of Broadcom Ltd) [38].

### 2.3.1 Designing the transponder antenna

In most applications, an antenna and its impedance bandwidth and matching level are characterised with respect to the common 50-Ω normalisation impedance. There are, however, some important applications where the antenna should work well in designs without a 50-Ω impedance level, and where the desired load impedance is complex and frequency-dependent. In these cases, taking into account the proper impedance characteristics at different frequencies is crucial for the operation of the system.

One example of such a system is an RFID tag operating at one or more frequency bands. In the tag, the antenna is connected to an RFID chip, whose impedance usually has a fairly small resistance and a large, capacitive reactance. Here, smallness of the resistance should be considered to be proportional to the reactance, as some chips can also have large (absolute) resistance values [39]. In harmonic transponders, diodes are
used instead of RFID chips to generate the required second harmonic frequency. Generally speaking, the diode and chip can be considered to have somewhat similar impedance characteristics, but the need to have harmonically spaced operating frequencies makes implementing the required matching in the transponder design more complicated than in the case of regular RFID tags.

The harmonic transponder developed in [I] and further analysed in [II]–[III] is illustrated in Figure 2.3. The starting point for this design was obtained by modifying an existing 915-MHz RFID antenna of [35] to better suit the targeted operation at a fundamental frequency of $f_0 = 1$ GHz. Consequently, the second harmonic frequency is $2f_0 = 2$ GHz. This frequency pair is chosen mainly for demonstration purposes and not to make the design highly compliant with particular existing frequency regulations. The 1 and 2-GHz frequencies have been applied by other research groups as well (see, e.g., [40]). The transponder antenna is designed and simulated using the FDTD-based EM simulator SEMCAD-X [41] and the FDTD/FIT-based CST Microwave Studio software [42].

In addition to the operating frequencies, also the corresponding diode impedance $Z_D = R_d + jX_d$ should be specified as the design goal to, in the ideal case, conjugately match the antenna ($R_a = R_d$ and $X_a = -X_d$).
The antenna design process contains five main steps or components, the effects of which on the antenna matching are depicted in Figure 2.4. These design steps are:

1. Implementation of the coupler/radiator at \( f_0 \) (Step 1).
2. Implementation of the coupler/radiator at \( 2f_0 \) (Step 2).
3. Adding a parasitic coupler (Step 3).
4. Cutting slots to the antenna (Step 4).
5. Slight lengthening of the meandered structures (Step 5).

When designing a direct-matched transponder, the largest antenna dimension is determined by the desired fundamental frequency. In the design of [I]–[II], the largest component is the meandered-dipole type structure (Step 1). Within it is situated the radiator responsible for the operation at \( 2f_0 \). This is a C-shaped strip (Step 2) that couples inductively to the meandered dipole. The antenna is fed from the middle of the \( f_0 \) radiator. In [35], the antenna feed is located in a structure somewhat similar to the C-shaped radiator, and the meandered section is left untouched.

After the first two steps, the matching levels at \( f_0 \) and \( 2f_0 \) are not very good, as seen in Figure 2.4, and further modifications are therefore performed. In Step 3, an additional, parasitic coupler is placed next to the meandered dipole. The Smith chart plot of Step 3 in Figure 2.4 shows that the impedances at the desired frequencies are quite well spaced, and that they need to be ‘rotated’ to the inductive side of the Smith chart. This is done by cutting slots in the different radiators (Step 4). The effect of the slots can be modelled as a series inductance [43]. Further fine-tuning to the two resonance frequencies is done in Step 5 by slightly extending the open ends of the meandered structure.

Figure 2.5 illustrates the details of the final transponder antenna described by Steps 1–5 and presented in [II]. It should be noted that even though some geometric details of the transponder of Figure 2.5 are different from those of Figure 2.3, the fundamental operating principles nevertheless remain the same. The initial transponder design of [I] operates slightly above 1 and 2 GHz, and the modifications described above help to nudge the fundamental and second harmonic frequencies more precisely to the desired ones.

Even though the designed transponder antenna is fully self-resonant, an additional inductor is placed next to the diode to properly zero-bias
Figure 2.4. An illustration of the different, principal design steps of the direct-matched transponder of [I]–[II]. Taken from [II] (© 2015 IEEE).

Figure 2.5. An illustration of the harmonic transponder following the design process of Figure 2.4. All dimensions are in millimetres. Taken from [II] (© 2015 IEEE).
it. In principle, the required inductance could also be implemented as a distributed component, much in the same way as the various capacitive and inductive modifications carried out during the antenna design. In practice, though, using a lumped component is a more straightforward approach, and it is also used later in Section 2.4 when implementing a transponder with an external matching circuit.

### 2.3.2 Characterisation of transponder performance

In the works of [I]–[III], the performance of the proposed transponder is investigated experimentally. This is done both for impedance matching and harmonic response. The actual manufactured transponder has the diode and inductor connected to the antenna. However, having these components present in the transponder is challenging from the point of view of measuring the actual antenna impedance. This is mainly related to the fact that the proposed transponder antenna does not have space for mounting the measurement cable (especially in the presence of the circuit components), and also because the cable itself used in the measurement easily becomes part of the radiating structure. Ways in which the cable affects the measurements as well as ways of reducing its effects are discussed in, e.g., [44, 45]. One typical approach is to place a suitable balun or cap structure on the outer surface of the cable to reduce the surface currents flowing on it, and thereby also the overall effect of the cable.

One further issue regarding wired measurements is the impedance level of typical measurement cables and equipment. They usually have a 50-Ω impedance, which is significantly different from that of the direct-matched transponder antenna. One possibility, which is utilised in [II] is to measure the input reflection coefficient of the transponder antenna in the 50-Ω environment, and to normalise the antenna matching to this impedance also in simulations. In this measurement, the diode and inductor were not connected. The antenna impedance could also be determined from measured backscatter data with the help of three known loads: open circuit, short circuit, and a known resistance [46].

As can be seen in Figure 2.6, the simulated and measured 50-Ω curves generally show quite similar characteristics. From this, it can be concluded that the manufactured transponder operates at the desired frequencies also with the actual, complex-impedance load. However, there are also discrepancies, e.g., in the location of the 50-Ω resonance at the second harmonic frequency. Potential reasons for this may be the sensi-
Figure 2.6. Transponder matching with different normalisation impedances. Taken from [II] (© 2015 IEEE).

...tivity of the antenna to the orientation of the measurement cable, or the effect of the connector. These effects were not modelled in the simulations, and the actual measurement was performed without a broadband balun to mitigate the cable effects. The curves of Figure 2.6 also show that normalising the reflection coefficient to the correct diode impedance only at $f_0$ or $2f_0$ mainly provides information of the matching around that frequency, and there remains some uncertainty as to the exact position of the other resonance.

To overcome some of the previous challenges related to characterisation of wireless sensors and other small antennas, an alternative is to use contactless measurements. In this approach, the desired antenna properties are investigated in similar conditions in which the sensors would actually be used or with the actual load connected to the antenna. Examples of using this type of technique include studying the radiation patterns and quality of matching of, e.g., RFID tags and harmonic transponders [47–50].

Figure 2.7 illustrates the harmonic response achieved with the transponder of Figure 2.3. The response is calculated based on the theoretical model given in Equation (2.1), and it is also measured in an anechoic chamber with different propagation distances. In the measurements, the Tx and Rx are positioned in the same direction, and their distance to the transponder under study is the same ($r_{Tx} = r_{Rx}$). Both calculations and measurements performed across different distances assume a transmit power level of $+16$ dBm. The calculated and measured curves are shown with respect to the fundamental frequency, and the corresponding second harmonic is seen by simple multiplication. Strongest transponder
response is achieved at the frequencies 1.01/2.02 GHz, which agrees with the simulated matching results achieved in [I] and illustrated later in Figure 2.8. Table 2.1 shows the calculated path and conversion losses across the different measurement distances, as well as the power actually picked up at the receiver.

When taking into account additional losses (measurement cables, filters, diode and inductor), the theoretical and measured responses of Figure 2.7 agree well, but especially at frequencies above 1.01 GHz, the measured power levels exceed the theoretical predictions. One of the reasons for the difference can be power received from directions other than line of sight. The theoretical model used in the calculations assumes a 'two-ray' propagation model, with one incident and one reflected ray, respectively. In practice, the harmonic signal can also experience multiple reflections, which can increase the angular range from which power is received. Even though the measurements were performed in an anechoic chamber, the power levels involved are relatively low, meaning that small environmental reflections or power leakage between Tx and Rx can appear in the measured response. One factor that causes additional uncertainty in the measurements done in the anechoic chamber is that the 1-GHz fundamental frequency falls a bit below the specifications of the absorber material used to line the chamber. The effect of this is most likely small, as the general agreement between theory and measurements is very good — as indicated by the curves of Figure 2.7.

An additional metric that is of general interest is the maximum read-out range of the transponder. The performed experiments did not allow measuring this, as distances above 5 m become challenging in the available anechoic chamber. With the currently used set-up, power levels and so on, one can estimate the read-out range to be around 6–7 m in the ideal conditions used in the current study.

### 2.3.3 Transponder operation with different diodes

Above, the harmonic transponder performance was investigated using a diode to which the antenna impedance is matched. A logical next step is to consider how this design is suited for or generalises to a case when the diode impedance changes, i.e., when switching to different diodes. Table 2.2 illustrates the impedance calculated for the transponder antenna of Figure 2.3, and for altogether six different varactor and detector diodes suitable for operation at this frequency range.
Figure 2.7. Measured and calculated harmonic response of the direct-matched transponder. Taken from [I] (© 2015 IEEE).

Table 2.1. Link budget for the theoretically calculated received power of the direct-matched transponder at different read-out distances. Taken from [I] (© 2015 IEEE).

<table>
<thead>
<tr>
<th>$r$ (m)</th>
<th>$P_{r, f_0}$ (dBm)</th>
<th>$L_{\text{path}, f_0}$ (dB)</th>
<th>$L_{\text{conv}}$ (dB)</th>
<th>$L_{\text{path}, 2f_0}$ (dB)</th>
<th>$P_{r, 2f_0}$ (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>12.4</td>
<td>29.4</td>
<td>12.0</td>
<td>32.7</td>
<td>-61.7</td>
</tr>
<tr>
<td>3</td>
<td>12.4</td>
<td>32.9</td>
<td>15.5</td>
<td>36.2</td>
<td>-72.2</td>
</tr>
<tr>
<td>4</td>
<td>12.4</td>
<td>35.4</td>
<td>18.0</td>
<td>38.7</td>
<td>-79.7</td>
</tr>
<tr>
<td>5</td>
<td>12.4</td>
<td>37.3</td>
<td>19.9</td>
<td>40.7</td>
<td>-85.6</td>
</tr>
</tbody>
</table>

The values of Table 2.2 show that between different diodes, the impedance levels may vary considerably. Of particular importance is to properly take into account the diode reactance in the matching. It is generally larger (in absolute values) than the resistance, and is furthermore frequency-dependent. With sufficient power levels, the nonlinearities in the diode make its resistance and reactance power-dependent. However, the calculations and modelling done in [I] and [III] utilise a small-signal approximation in which the impedance is only frequency-dependent. In light of the numbers in Table 2.1, this is a sound concept. Consequently, also the impedances of Table 2.2 assume no dependency on input power.

Figure 2.8 gives the impedance matching obtained with different diodes and the antenna of Figure 2.3. When looking at the matching curves, it
Table 2.2. Antenna and diode impedances at the targeted \( f_0 \) and \( 2f_0 \) frequencies. The impedances are calculated under small-signal conditions. Taken from [III] (reproduced courtesy of The Electromagnetics Academy).

<table>
<thead>
<tr>
<th>Case</th>
<th>( f_0 = 1 \text{ GHz} )</th>
<th>( 2f_0 = 2 \text{ GHz} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Antenna</td>
<td>( 8 + j54 )</td>
<td>( 10 + j16 )</td>
</tr>
<tr>
<td>SMV2019</td>
<td>( 4.5 - j69 )</td>
<td>( 4.5 - j34 )</td>
</tr>
<tr>
<td>SMV1430</td>
<td>( 2.5 - j128 )</td>
<td>( 2.5 - j64 )</td>
</tr>
<tr>
<td>SMV1231</td>
<td>( 1.6 - j69 )</td>
<td>( 1.6 - j34 )</td>
</tr>
<tr>
<td>SMV1405</td>
<td>( 0.64 - j60 )</td>
<td>( 0.64 - j30 )</td>
</tr>
<tr>
<td>HMPS-2820</td>
<td>( 7 - j206 )</td>
<td>( 7 - j94 )</td>
</tr>
<tr>
<td>HSMS-2860</td>
<td>( 2.4 - j492 )</td>
<td>( 2.06 - j238 )</td>
</tr>
</tbody>
</table>

is obvious that without modifications to the original antenna structure, simply switching to a different diode will be problematic for the transponder performance. Either the achieved matching levels can be weak and/or detuned, or the frequencies with good matching are not harmonically separated. As an example, good matching levels are obtained with the HMPS-2860 detector diode at 1.2 and 2.2 GHz, but this non-harmonic frequency pair does not perform well in terms of frequency conversion and harmonic response.

Calculations on the harmonic response achieved with the different diodes have been carried out, and the results are given in Figure 2.9. In these calculations, the distance between the transponder and Tx/Rx is assumed to be 2 m. Studies performed in [III] investigate the effects of using different diodes only theoretically and using simulations. The curves of Figures 2.8–2.9 show that both the matching and resulting harmonic response are less affected in the case of diodes whose impedance characteristics resemble those of the ‘reference’ SMV2019 case. In cases with detuned matching frequencies, the strongest response occurs near the best-matched frequency at \( f_0 \). This agrees with the more significant effect of the \( f_0 \) properties described in Section 2.2.1.

As the above results show, the antenna of Figure 2.3 should be modified to better match it to the impedance levels of diodes other than SMV2019 to improve the performance at the desired frequencies. In principle, this may appear to be a straightforward impedance matching problem, but in
practice, altering the antenna design can affect its $Q$, meaning that the FOM$_d$ ranking of individual diodes can change. As a consequence, some other diode than the one considered for the modified antenna could be better in terms of the FOM$_d$. Thus, achieving the most use of the FOM$_d$ quantity might require an iterative approach. Here, the diodes are initially ranked based on an assumption of $Q_a$, and the antenna design is implemented for this diode. Then, based on the actual $Q_a$, an updated ranking can be obtained. In the end, this may lead to a rather cumbersome design process, and a more practical way may be to just design the best possible antenna for the diode that stands out in the initial ranking.
2.4 Transponder implemented using lumped-component matching circuits

Comparisons made with different diodes and a direct-matched antenna showed that switching to another diode is problematic, especially if the transponder response is wanted at a certain level or at particular frequencies. In light of the strongly self-resonant-type matching of the previously discussed antenna, this might seem an obvious result. An alternative to direct matching is to use an antenna whose performance is as little dependent on geometric details as possible. This possibility is the subject of [IV]. With this kind of antenna, a separate matching circuit is used to create the necessary resonance frequencies.

Considering practical wireless sensing applications, such as RFID tags in mass production, utilising lumped-component based matching circuits is typically not a preferred approach. The main limiting factors relate to cost and challenges in fabrication [51], and each lumped component placed in the sensor also creates additional losses compared to a similar direct-matched implementation. However, in spite of the above issues, implementing the required harmonic frequencies with an external matching circuit instead of using direct matching can be beneficial in terms of overall design complexity, especially on the antenna side. For this reason, the suitability and versatility of transponders based on matching circuits compared to the previously presented direct-matched transponders is investigated in the following.

2.4.1 Diode selection

On a conceptual level, having an inherent mismatch between the antenna and diode impedances appears beneficial for utilising external matching circuits. This is due to the fact that the same antenna geometry may be suitable for use with several different diodes, provided that differences between their impedances are such that the diodes can be matched to the antenna using a circuit of sufficiently moderate complexity.

In the work of [IV], three different diodes are chosen for the study. These are the SMV2019 and SMV1231 varactor diodes and the HMPS-2820 detector diode previously used in [I] and [III]. Like the previous transponder, also the lumped-component design is intended to operate at 1 and 2 GHz, and the corresponding diode impedances are shown in Table 2.2.

Of the three diodes, the varactors have similar reactances but different
resistance, and the detector diode impedance differs in both quantities. Furthermore, the transponder designed in [IV] using the SMV2019 diode can be compared to the direct-matched design implemented for the same diode in [I]–[III]. This comparison can be made in a more or less straightforward way, even though antenna dimensions of the two transponders are not exactly the same.

2.4.2 Antenna geometry

Figure 2.10(a) gives an illustration of the general antenna geometry considered for use in the matching-circuit based transponder, along with a depiction of the possible matching component locations. The proposed antenna geometry can be described as solid metal sheet with a slot cut in the middle. Therefore, the antenna represents something of a thick dipole, whose inherent resonances \( (X = 0) \) are determined by the overall antenna dimensions and the position of the slot.

During the design process, different antenna dimensions were parametrised to determine a suitable size of the antenna. Even though the purpose here is not to use the antenna geometry for creating the impedance matching, having such dimensions that do not result in too extreme antenna impedances compared to those of the various diodes is beneficial. Having extremely different impedance levels would cause challenges to the matching circuit design, e.g., in terms of required circuit complexity.

The length and width of the antenna, as well as the length of the feeding gap were varied to see their effect on the antenna impedance. In all cases, the width of the feeding gap is equal to that of the antenna. For the study, the length of the antenna is varied from 20 to 80 mm with 10-mm steps, and the width is varied from 20 to 40 mm with 5-mm steps. Values for the feeding gap were considered from 0.25 to 1.5 mm with 0.25-mm steps.

Based on the studies, an \( 80 \times 40 \text{ mm}^2 \) antenna (length \( \times \) width) with a 0.5-mm long feed gap appeared to provide a suitable compromise in terms of impedance with respect to the three diodes studied. The overall area of the transponder is roughly comparable to the direct-matched case (for that transponder, the substrate area is \( 80 \times 35 \text{ mm}^2 \)), but here the actual antenna area is larger. The transponder of Figure 2.10(a) is implemented on a 0.79-mm thick FR-4 substrate \( (\varepsilon_r = 4.4) \).

For the antenna of Figure 2.10(a), the impedance at 1 and 2 GHz is \( Z_{a,f_0} = 4.5 - j8 \Omega \) and \( Z_{a,2f_0} = 21.3 + j89 \Omega \), respectively. When comparing these values to the diode impedances of Table 2.2, it is clear that the
impedance mismatch between the antenna and diode is so significant that either one or both of the $f_0$ and $2f_0$ frequencies will be detuned from the intended value. As was observed previously, especially mismatch at the fundamental frequency is detrimental for the operation of the transponder.

### 2.4.3 Matching circuit design

To create the necessary operating frequencies for the transponder, a matching circuit is connected between the diode and the antenna. The initial design consideration is to use matching circuits with as low number of components as possible with the different diodes. This is done to reduce the overall complexity of the design, and also to achieve lower losses. Of course, the exact amount of losses depends on the initial impedance mismatch between the antenna and diode, and also on the values of individual components and their position within the circuit.

For designing the matching circuits, Optenni Lab [52] and AWR [53] are used to determine suitable matching circuit topologies and component values. In the initial stage, the circuits are implemented with ideal induc-
Figure 2.11. Simulated impedance matching of the matching-circuit based transponder with the different diodes. Taken from [IV] (c⃝ 2017 IEEE).

stances and capacitances. Final implementations are based on realistic inductor and capacitor models available from Murata [54], and 0603-sized components from the LQW18 and GQM18 series are used. The design also takes into account the effect of mounting the components on the antenna, and the simulation model also features a pad network of realistic dimensions, to which the various components are placed.

The final matching circuit implementations for the transponders based on different diodes are illustrated in Figure 2.10(b)-(d). For the designs implemented using the varactor diodes (SMV2019 and SMV1231), the matching circuits have two and three circuit components, respectively. Different numbers of components were also considered, but the proposed designs provide a sufficient compromise in terms of achieved matching level and design complexity. For the HMPS-2820 detector diode, the contrast between antenna and diode impedances is greater than with the varactors, meaning that a larger number of components — in this case, four — is needed to achieve a proper matching performance.

2.4.4 Performance of the transponders

Figure 2.11 presents the impedance matching obtained in simulations with the different matching-circuit based transponder implementations. In all three cases, quite good matching levels can be achieved, and especially at the intended $f_0$ frequency, the resonance remains quite consistent with the different diodes. At the second harmonic frequency, around 2 GHz, the different cases have more significant variance in both matching levels and bandwidths.

All three proposed transponders were manufactured, and their opera-
Figure 2.12. Calculated and measured harmonic response of the matching-circuit based transponders with the different diodes. Taken from [IV] (© 2017 IEEE).

Table 2.3. Simulated diode and matching circuit losses of the matching-circuit based transponders at $f_0 = 1$ and $2f_0 = 2$ GHz. Taken from [IV] (© 2017 IEEE).

<table>
<thead>
<tr>
<th>Case</th>
<th>$L_{\text{comp}, f_0}$ (dB)</th>
<th>$L_{\text{comp}, 2f_0}$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Reference</td>
<td>0.53</td>
<td>0.57</td>
</tr>
<tr>
<td>SMV2019</td>
<td>4.5</td>
<td>0.45</td>
</tr>
<tr>
<td>SMV1231</td>
<td>3.2</td>
<td>0.11</td>
</tr>
<tr>
<td>HMPS-2820</td>
<td>7.1</td>
<td>0.85</td>
</tr>
</tbody>
</table>

tion was investigated using the same measurement set-up as in [I]. The resulting harmonic response results, both simulated and measured ones, are depicted in Figure 2.12. For comparison, the performance of the reference direct-matched transponder of [I] is also given. In this case, the distance from the transponder to the Tx and Rx was 2 m, both in calculations and in the measurements. The actually measured transponder responses in Figure 2.12 show that each of the three transponders is able to generate a clearly distinguishable response, but the results vary in terms of operating frequency and response level from the calculated ones.

Table 2.3 gives the calculated losses in the matching circuit components at $f_0 = 1$ GHz and $2f_0 = 2$ GHz for the different cases. When comparing the mutual order of the harmonic response curves at $f_0$, the results show that with higher losses at this frequency, the achievable response
decreases. It should be emphasised that the losses are calculated for a specific frequency pair, which does not necessarily correlate with, e.g., the frequencies giving the largest response for a certain case. A general observation on the results of Table 2.3 and Figure 2.12 is that the more complex the needed matching circuit is, the higher typically are the losses. In this case, though, the circuit topology and original contrast in antenna and diode impedances contribute to the overall losses, which helps to explain why, e.g., the two-component matching circuit of the SMV2019 diode has more losses at $f_0$ than the three-component one of the SMV1231 case.

One major source for the observed differences in the performance is the quality of the achieved matching, and also the exact frequencies at which the implemented matching circuits actually provide the resonances. The curves of Figure 2.12 show that the harmonic response of the designs with varactors are tuned upwards from the predicted values, and the detector-based one is tuned downwards. This indicates that the values of some of the components used in the matching circuits should be changed.

For this purpose, it would be interesting to study in greater detail the actual matching characteristics of the matching-circuit based transponders in a similar way as was done in [55] for the direct-matched design of [I]. For that antenna, the agreement between simulated and measured matching curves is good. This is somewhat expected, as the operating frequencies are largely determined by the antenna, whose performance matches that predicted by the simulations, as is evident from studies made on a similar antenna in [II]. The operation of the transponders of [IV] significantly depends on the matching circuits. This means that to improve or tune the matching characteristics, the values of particular circuit components need to be altered (in practice, iterated), and the effect of these changes on the transponder response need to be measured.

### 2.5 Comparison of different transponder implementations

The previous sections have presented and investigated two different techniques of implementing a harmonic transponder operating at desired $f_0$ and $2f_0$ frequencies. When comparing the direct-matched designs of Section 2.3 and the transponders implemented with external matching circuits in Section 2.4, a number of observations can be made.

Utilising an external matching circuit for creating the harmonic operating frequencies makes designing the antenna easier, as obtaining proper
Harmonic Radar and Transponders

matching to a complex, frequency-dependent load impedance does not have to be implemented through modifying the antenna geometry. In this case, the same antenna design can be applied to different diodes, as presented in [IV]. The chosen dimensions are a compromise between the properties of the three diodes used in the study, and are not really ideal for any individual diode. In practice, one might consider choosing a set of dimensions feasible for whatever diode is selected, in order to simplify both the matching circuit and the overall complexity of the design.

The main drawback of the transponders based on matching circuits is the challenge of verifying the quality of the matching of the assembled design. By measuring the transponder response using the intermodulation technique presented in [55], the reflection coefficient can be obtained, but this result does not directly reveal, which of the components should be changed and to what direction if the response is not what is expected. One solution could be to arrange a place to connect a measurement cable for a wired measurement.

Compared to this, the transponders based on direct matching are more challenging to implement due to geometric intricacies, and are not that well suited for switching to different diodes. On the other hand, they are closer in design to typical RFID tags that often just feature an antenna connected to an RFID chip. The requirements of a particular application, e.g., constraints on transponder size, shape and weight, largely determine which of the two matching approaches suits better to a specific design.

Considering the overall design, both of the transponders have used a single antenna — which in the direct-matched case is composed of several physically-separated, electrically-coupled parts. This approach is the most commonly used one in harmonic radar applications. In some works, such as [33, 56], separate patch antennas are used for the fundamental and harmonic frequencies, although both are connected and matched to the same diode. Compared to a single-antenna case, this approach increases the overall size of the design, even though it may be beneficial for matching the antennas.

One of the main benefits of using harmonic instead of linear radar is increased tolerance against environmental clutter. To improve the accuracy of the transponders, the use of two orthogonally-polarised antennas to transmit the signal back at $2f_0$ has been suggested in [57]. In this case, sensor information can be encoded in the amplitude and/or phase difference of the two branches.
This chapter investigates the design and performance of antennas that are intended to take a particular, conformal shape. For this kind of application, it is important to consider the intended shape of the antenna, both in cases with fixed, conformal surfaces, and in cases where the shape of the design can change in the intended usage case.

Section 3.1 describes general aspects related to conformal antennas, including example applications and possible requirements on the antennas set by the conformity. The suitability of the direct-matched harmonic transponder of the previous chapter for conformal sensing applications is investigated in Section 3.2. Analysis and modelling of antennas for bendable handsets is performed in Section 3.3, which describes an equivalent circuit model used to characterise the effects of bending. The operation of such handsets is considered in both varying and fixed bending cases. Section 3.4 looks at practical issues in implementing bendable handsets.

### 3.1 Preliminaries on conformal antennas

In many applications, a conformal antenna shape may be desired or even required. Examples of fields benefiting from the use of antennas that conform to their surroundings are aerospace and space applications, and cases where the antennas are mounted on non-planar surfaces of various vehicles [58–61]. For the framework of the upcoming 5G communications and its significantly increasing number of connected devices and base stations, conformal antenna solutions have been suggested for use in various wearable, mobile, and home devices [62], and also for integrating antennas or antenna arrays in the built environment [63]. Depending on the type and implementations, antennas may in these cases occupy a variety of different shapes. The shape can be, e.g., paraboloidal (such as in the
Conformal Transponder and Handset Antennas

reflector antennas in radio telescopes), conical (as in antennas mounted on the nose cone of an aeroplane), or the elliptical or (hemi)spherical geometries used in many lens antennas.

Another set of particular shapes or forms arises if the antenna has to follow the contours of an object, onto which it is to be positioned. In many cases, these would be cylindrical or spherical shapes and surfaces. As an antenna design problem, such ‘exotic’ shapes might not be a significant issue, as long as the features can be taken into account during the design process. On the other hand, situations in which the antenna should adapt to dynamic changes in shape during its use are significantly more involved, and they are also the main focus of the following sections.

Making the shape of an antenna conformal introduces changes in both its operation and various parameters, such as input matching, and consequently also the impedance bandwidth [64, 65]. The directive properties of the antenna can also change, and in the case of, e.g., conformal antenna arrays, the three-dimensional, conformal shape may complicate predicting the gain of such designs [58]. On a general level, the exact nature of the effects is affected by a number of different parameters, such as:

1. Operating frequency.

2. Amount of conformity (how much the shape changes with respect to the “natural” antenna shape).

3. Distribution of conformity along the shape of the antenna.

In certain cases, also other differentiating factors may apply. The first point applies to, e.g., handset antennas, with which at frequencies below 1 GHz the majority of the radiation is caused by the chassis rather than the antenna elements [66]. At higher frequencies, the relative contribution of the elements increases, and effects of changing the shape of the device can vary at different frequencies (see Section 3.3). On the other hand, the way in which a conformal shape is applied with respect to the device structure also has an impact on the effects, depending on whether the structure is bent evenly and symmetrically, or whether some parts are subject to more bending than others.
3.2 Conformal antennas for wireless sensing applications

In the previous chapter, an extensive amount of work was performed on characterisation of harmonic transponders. All those cases did not, however, accurately take into account the effect of the surroundings. Experimental verification of the proposed transponders took place in an anechoic chamber, with the transponders mounted on a Styrofoam platform. More realistic operating conditions feature harmonic transponders and other wireless sensors that are situated in the vicinity of an object or surface, which is the typical usage case of conventional RFID tags as well. Properties of the object as well as its material parameters affect the performance of the sensor, usually by detuning the antenna response to lower frequencies. For this reason, the effect of different materials or environments on sensor performance should be considered during the implementation to obtain a well-performing and sufficiently robust design.

On the other hand, if the performance of the transponder is strongly affected by environmental effects, it could be possible to use the transponder to sense some external parameter or quantity. This aspect is investigated in [V]. Possibilities of using RFID systems for sensing purposes have been studied in, e.g., [67, 68] in which impedance variations caused by bending the transducer phase modulates the backscattered signal, thereby enabling a strain sensor based on RFID technology. The work of [56] considers applying a harmonic RFID (transponder) for humidity sensing.

Two different environmental effects or modifications to the transponder are considered in [V]. Firstly, changes in the impedance matching and harmonic response are observed when the transponder is placed close to an object with varying dielectric constant. The dimensions of the object are $80 \times 160 \times 10 \text{ mm}^3$ (width $\times$ length $\times$ thickness), which are 'large' compared to the transponder size. Secondly, different degrees of bending are applied to the transponder both in free space and in the presence of a cylindrical object to study the sensitivity of the transponder response in the conformal case.

Both for the current conformal transponder case and for the handset antennas of Section 3.3, the parameter bending radius ($r_b$) acts as the measure of the amount of bending applied. It represents the radius of a cylindrical surface, about which the transponder or handset can be bent in a given state. At lower degrees of bending, the cylindrical shape approaches a planar one, and in the planar case, $r_b \to \infty$. Current studies
Conformal Transponder and Handset Antennas

Figure 3.1. Illustration of (a) the direct-matched transponder placed on a cylindrical surface (PVC cylinder), (b) the tightest bending applied ($r_b = 30 \text{ mm}$), and (c) the loosest bending applied ($r_b = 150 \text{ mm}$). Modified from [V]; © EurAAP; used with permission.

and analysis are only made with respect to dielectric materials, and transponder performance in terms of other environmental factors such as the proximity of metal bodies is not taken into account.

In the first case, integer values for dielectric constant from 1 to 10 are considered. Results show that with the current direct-matched transponder, obtaining good impedance matching at harmonically spaced frequencies becomes challenging when the dielectric constant $\varepsilon_r \geq 2$, especially if good matching at the fundamental frequency is desired. This performance is partially caused by the fact that the operation of the transponder antenna in the presence of particular dielectric materials was not considered during the design process. The effects observed in connection with the transponder placed on the dielectric block generally resemble those of traditional dielectric loading, a well-known concept in antenna engineering.

The second case considers the effects of bending the transponder about a cylindrical surface and in the presence of a PVC pipe ($\varepsilon_r = 3.19, \tan \delta = 0.0096$). Figure 3.1 illustrates the placement of the transponder on the cylindrical surface, as well as the highest and lowest amount of bending applied ($r_b = 30 \text{ mm}$ and $r_b = 150 \text{ mm}$). Additionally, an intermediate 90-mm bending radius is used.

Figure 3.2 presents the calculated harmonic response of the transponder in the different bending cases both in free space and with the PVC pipe. Calculations are done for the case of a 2-m distance between the transponder and Tx/Rx. For comparison, also the planar reference case from [I] is given. Compared to the planar case, the curves show that when the transponder is bent in free space, the frequency of strongest response de-
Figure 3.2. Achieved harmonic response with the direct-matched transponder in different bent cases with dielectric cylinder (dashed lines) and without dielectric cylinder (solid lines). Taken from [V]; © EurAAP; used with permission.

tunes upwards, and when the PVC pipe is added, the response detunes downwards. In the bent case (either in free space or with the object), the variance between the response levels obtained with different $r_b$ values is quite small. The main reason is that the matching levels and resonances at $f_0$ and $2f_0$ vary less than in the presence of the dielectric block discussed previously.

Based on the results presented above for the direct-matched transponder, the exact suitability of harmonic radar and transponders for the considered environmental effects remains somewhat unknown. In principle, the current design is quite suitable for conformal operations, apart from the used substrate material, whose thickness does not allow bending the prototype design used in [I] and simulated in [V].

In practical implementations, environmental effects and transponder performance in the intended application should be considered in a similar way as with RFID tags [69–71]. Two main usage cases can be considered for the transponder in the conformal case: either to utilise it in strain-sensor type applications where the response depends on the shape of the transponder, or to place the transponder on a particular curved surface and making the antenna sensitive to some external property or quantity (see, e.g., [72]).

For designing more realistic, conformal harmonic transponders, it is necessary to consider the choice of substrate material in greater detail. One option could be to use a thin, flexible FR-4 substrate, which might provide a truly bendable design, possibly also with robustness to sustain the repeated bending cycles and changes in shape during the operation of
the sensor. Considering the use of truly bendable or conformal transponders for harmonic radar applications, the intended platform on which the transponder is to be placed as well as the range of $r_1$ values across which the device would need to be bent should be taken into account during the design. This applies also to the type of sensing desired from the transponder, and to the changes required in the response of the transponder to distinguish between different sensor or bending states.

### 3.3 Antennas for bendable mobile handsets

The following section considers antennas for future-generation bendable handsets. A typically used antenna type in current handsets is the Capacitive Coupling Element (CCE) [66, 73], in which an electrically small metal body ("coupler") is used to couple to the resonant wavemodes of the chassis of the device. In traditional, resonant antenna types used in handsets, such as inverted-L antennas (ILA) or planar inverted-F antennas (PIFA), the desired operating frequencies are implemented by modifying the antenna geometry to obtain the wanted resonances. Therefore, the antenna size is comparable to the wavelength at the targeted frequency (typically being of size $\lambda/2$ or $\lambda/4$).

In comparison to resonant antennas, the CCE-based antenna designs are not inherently resonant at the target frequencies. Instead, an external matching circuit is used to cover the desired frequency bands. The matching-circuit based approach helps to simplify the design process by transforming an antenna design problem largely to a circuit design problem. Instead of having to implement all different frequencies by fine-tuning the antenna geometry, the primary design challenge becomes finding a well-performing matching circuit that meets the requirements set for the design.

Even though plenty of research has investigated various conformal antennas, such as wearable antennas (e.g., [74–86]) for different radio systems as well as the effects of conformality [87–91], there are not that many works that consider handsets featuring some flexibility in their shape. One example of this kind of research is found in [92], where the effects of both bending and twisting the handset are investigated using simulations.

In the present work, publications [VI]–[VII] investigate how bending the handset affects the antenna performance. These studies are made us-
ing electromagnetic and circuit simulations, as well as using equivalent circuit models to describe the operation of a CCE-based handset antenna system. Two main approaches or implementations are considered. In the first case, a handset with fixed dimensions is subject to different bending states (Section 3.3.2), and in the second case, handset dimensions are varied in a fixed bending state (Section 3.3.3). In the first approach, only the free-space case is analysed, but the second case also considers the effect of the user.

### 3.3.1 Equivalent circuit model for coupling-element based handset antenna

When considering practical and commercial handset antenna designs, the large amount of antennas, other electronics, and required frequency bands make the design problem so complex that simulations become an invaluable aid. Together with measurements, they provide valuable and concrete information on what happens in a particular scenario, but they do not always give a definitive answer to the question why.

One approach to answering both of the above questions is to characterise the operation of the antenna (or indeed that of any device or system) using a model that both takes into account the essential properties of the system it is describing, and is still sufficiently simple to use. The equivalent circuits used in the following are an example of such physical models, i.e., models that consider the underlying physical properties. Different kinds of behavioural models can also quantitatively explain the operation of a system, but these might not have any particular physical background.

Previous works, such as [93, 94], have shown that the performance of a CCE-based handset antenna can be quite accurately and simply characterised with an equivalent circuit model consisting of coupled series and parallel lumped resistor, inductor, capacitor (RLC) resonators. Figure 3.3 illustrates such a model, and this particular model has been used in [VI]. The required complexity of the circuit model depends in part on what is being modelled, and how large a frequency range is to be covered.

Based on the theory of characteristic modes [95, 96], it is known that metal bodies can support certain inherent resonant wavemodes (so-called characteristic modes) that can be excited above particular frequencies. This principle is also applied in CCE-based handset antennas, in which the (possibly small) capacitive coupler is used to couple to the chassis wavemodes. This is true especially at low frequencies, below 1 GHz, where
most of the handset radiation originates from the chassis rather than the antenna elements [66]. The way in which the characteristic wavemodes are distributed, e.g., the location of the strongest fields or currents, can be used to find locations for the antennas and their excitations [97–99].

Each of the wavemodes in the circuit model is represented by individual resonators, which are connected to each other using ideal transformers. The series RLC resonator describes the so-called monopole mode of the coupling element itself, and the parallel RLC resonators model the chassis wavemodes. Here, the analysis focuses on frequencies below 3 GHz, so that two of the lowest-order chassis wavemodes are enough to describe the operation of the antenna.

3.3.2 Handsets with fixed dimensions and varying degrees of bending

The first case to consider is to study the effects of bending on a handset whose dimensions (antenna element and chassis size) are fixed, and the structure is modified by changing the amount of bending applied to the device. This has been done in [VI], which is based on antenna designs and research methodology introduced in [100]. All analysis and modelling in this section are done for the free-space case only, and the effect of the user is not considered.

Two different device sizes are considered in the study, 105×55 mm$^2$ and 175×55 mm$^2$ (length×width). Figure 3.4 shows the different antenna
dimensions and locations considered (Cases 1–3), and both on- and off-ground antennas are investigated. In addition to studying various impedance and $S$-parameter based results with simulations, the work characterises the performance and the effects of bending using equivalent circuit modelling. The main focus is on the smaller-sized device, for which the circuit models have been calculated.

Figure 3.5 gives the quality factor calculated from the input impedance ($Q_z$) for different antenna and bending cases for the (a) 105-mm long chassis and (b) 175-mm long chassis. Taken from [VI]; © EurAAP; used with permission.
in $Q_z$ are most prominent at smaller $r_b$ values, i.e., when the chassis ends are brought closer to each other. At the two studied frequencies, the $Q_z$ increases and decreases at 920 and 1920 MHz, respectively, with reducing $r_b$ values.

In simplified terms, the quality factor is the measure of the capability of a resonator to store energy. $Q$ is proportional to the ratio of stored energy ($W$) and power dissipated ($P_{\text{loss}}$) in the antenna. This would indicate that as the chassis is subject to tighter bending, the amount of electric and/or magnetic energy ($W_e$ or $W_m$) stored in the antenna near fields increases compared to more planar cases. Due to the larger radiation contribution of the ground plane at lower frequencies [66, 102], the more significant effect at 920 MHz is somewhat expected. At 1920 MHz, most of the radiation originates from the antenna element, the relative position of which with respect to the chassis does not dramatically change due to bending. Consequently, the corresponding effects are also smaller than at lower frequencies.

Table 3.1 gives the values for the circuit elements in the monopole and ground plane wavemode resonators that have been calculated for the planar, unbent handset. The resulting antenna impedance (resistance and reactance) as well as the input matching with respect to 50 Ω are illustrated in Figures 3.6 and 3.7, respectively. For comparison, also the corresponding results given by EM simulations are given.

In addition to the planar case, Table 3.1 also gives the element values used in two different bending cases for the 105-mm chassis. These are chosen to be the most extreme degree of bending, $r_b = 20$ mm, and an in-between point at $r_b = 50$ mm, still representing a fair amount of bending. In the bent cases, the circuit element values have been optimised numerically to match the simulations. Figure 3.7 gives the corresponding impedance matching curves. Here, it should be emphasised that none of these results feature a matching network, and all of the effects seen in the curves are caused by the properties of the antenna alone.

What kind of changes or effects can take place when the handset chassis is bent? Resistive, capacitive, and inductive changes may generally occur, and also the coupling to individual chassis modes can be affected. In the following, a physical interpretation of the observed effects is provided. The modified equivalent circuit features additional R, L, and C elements parallel to the original resonators. This kind of approach is more insightful than just modifying the original element values, as the effect of
Table 3.1. Component values in the planar and bent cases for the equivalent circuit shown in Figure 3.3. Modified from [VI]; © EurAAP; used with permission.

<table>
<thead>
<tr>
<th>Wavemode</th>
<th>$R$ (Ω)</th>
<th>$L$ (nH)</th>
<th>$C$ (pF)</th>
<th>$f_r$ (GHz)</th>
<th>$Q$</th>
<th>Other components</th>
</tr>
</thead>
<tbody>
<tr>
<td>Monopole</td>
<td>0.6</td>
<td>4.5</td>
<td>2.38</td>
<td>1.54</td>
<td>72</td>
<td>–</td>
</tr>
<tr>
<td>Ground plane 1$^{st}$ order</td>
<td>8.05</td>
<td>0.5</td>
<td>42.7</td>
<td>1.09</td>
<td>2.3</td>
<td>$n_1 = 1$</td>
</tr>
<tr>
<td>Ground plane 2$^{nd}$ order</td>
<td>3.8</td>
<td>0.18</td>
<td>23.8</td>
<td>2.47</td>
<td>3.0</td>
<td>$n_2 = 1$</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$r_b = 50$ mm</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Wavemode</td>
<td>$R$ (Ω)</td>
<td>$L$ (nH)</td>
<td>$C$ (pF)</td>
<td>Other components</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Monopole</td>
<td>–</td>
<td>–</td>
<td>–</td>
<td>–</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Ground plane 1$^{st}$ order</td>
<td>–</td>
<td>2.85</td>
<td>7.17</td>
<td>$n_{1,b} = 1.29$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Ground plane 2$^{nd}$ order</td>
<td>–</td>
<td>1.79</td>
<td>1.64</td>
<td>$n_{2,b} = 0.89$</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$r_b = 20$ mm</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Conformal Transponder and Handset Antennas

Reactive changes caused by bending affect the capacitances and inductances of the circuit model. In the study of [VI], the position of the antenna element with respect to the chassis is assumed to remain more or less the same, and the bending effects are modelled for the chassis wavemodes only. As the bending radius decreases, the ends of the chassis approach each other, which introduces additional coupling between them. In the model, this coupling is described with a capacitance ($C_{c1,b}$ and $C_{c2,b}$ in Figure 3.3). Bending the chassis can also affect the surface currents and their distribution on the chassis, and these changes in the current paths are modelled using the $L_{c1,b}$ and $L_{c2,b}$ inductances of Figure 3.3. The antenna element and chassis are assumed to be Perfect Electric Conductors (PEC) in the simulations, meaning that they have no resistive losses. Therefore, any changes in antenna resistance due to bending relates to
As can be seen from the curves of Figure 3.6(b) and 3.7, bending the handset has the most pronounced effect in the vicinity of the lowest-order chassis wavemode \( f_r = 1.09 \text{ GHz} \), whereas the changes at the second-order wavemode are essentially negligible. This behaviour is in agreement with the previously mentioned contribution of the chassis at lower frequencies. Apart from the values of the additional components placed in the resonators, Table 3.1 also shows the changes in the coupling coef-
Table 3.2. Effective component values calculated from Table 3.1, which can be compared with the planar case values. Taken from [VI]; © EurAAP; used with permission.

<table>
<thead>
<tr>
<th>Wavemode</th>
<th>$r_b = 50\text{ mm}$, effective values</th>
<th>$r_b = 20\text{ mm}$, effective values</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$R (\Omega)$</td>
<td>$L (\text{nH})$</td>
</tr>
<tr>
<td>Monopole</td>
<td>0.6</td>
<td>4.5</td>
</tr>
<tr>
<td>Ground plane 1st order</td>
<td>13.4</td>
<td>0.71</td>
</tr>
<tr>
<td>Ground plane 2nd order</td>
<td>5.0</td>
<td>0.21</td>
</tr>
</tbody>
</table>

Efficient of the ideal transformers describing the coupling to the individual wavemodes.

In order to facilitate the comparison of various bending-related effects, Table 3.2 displays the 'effective' component values for the different resonators when the coupling coefficients are scaled to 1. The effective values also show that in both bending cases, the first-order chassis wavemode has the strongest contribution, and the additional changes in coupling and current affect the chassis wavemode capacitances and inductances compared to the planar case as can be expected from the underlying physics.

By further placing additional elements into the circuit model, it could be possible to model also the effect of the user. However, this analysis and modelling falls beyond the scope of this thesis, but some quantitative predictions can still be made. Due to the nature of human tissue at microwave frequencies (lossy dielectric material with low permeability), adding the user into the model would mainly constitute additional losses (changes in resistance), as well as changes in the resonant frequencies of the chassis wavemodes (capacitive changes). By using additional capacitances in the model, as done in [VI], the capacitive effect of the user has been modelled in the case of a regular handset in [94]. That work did not consider the resistive effects, as modelling them especially across wide frequency ranges requires different, more complex models.
In principle, similar kind of modelling could be done for resonant antennas such as ILAs or PIFAs as well. For these cases, interpreting the results may be more complicated, especially distinguishing between changes due to the ground plane shape and those due to changes in the antenna matching itself. In CCEs, an additional matching circuit is used to achieve desired operating frequencies, and most of the effects seen in the model and therefore in the antenna performance are caused by chassis effects due to the lack of a matching circuit. By including also a matching circuit, the effects of bending on the quality of the matching could be observed also with CCE antennas.

### 3.3.3 Handsets with varying dimensions in a fixed bending case

After investigating the bending of a handset with fixed dimensions and varying $r_b$ values, the next step is to study an opposite scenario. This aspect is considered in [VII]. Here, the bending radius is fixed to a value that is feasible for placing the device around a model representing the human wrist. By varying the dimensions (width and length) of the handset, the effects of device size on various quantities is analysed at different frequencies both in free space and with the user.

To study the proposed wrist-worn application, i.e., a case where the device is wrapped or rolled around the wrist of the user, a dielectric cylinder is used to model the human wrist. The radius of the cylinder is 28 mm, and based on anthropomorphic measurements, the resulting circumference is comparable to that of an average male wrist [103]. To properly take into account the properties of user tissue, the wrist phantom is modelled with frequency-dependent material parameters (relative permittivity $\varepsilon_r$ and electric conductivity $\sigma$ [S/m]) according to CTIA specifications [104].

Figure 3.8 depicts the different handset widths and lengths investigated, as well as the position of the handset with respect to the wrist phantom. The length of the handset varies from 30 to 170 mm with 20-mm increments, and the width ranges from 20 to 50 mm with 10-mm steps. The same coupling element antenna is used in all cases, and there is a constant 2-mm air gap between the wrist and the chassis. Antenna dimensions are determined in such a way that the 20-MHz instantaneous bandwidth requirement of LTE-A systems is met in free space, and the air gap takes into account the fact that a real device would have a casing or enclosure between the wrist and the antenna structure.
As the size of the device varies, discussion can be raised on two distinct features in this kind of application: combined effect of shape and dielectric tissue loading (large devices) and the effect of dielectric tissue loading alone (small devices). This relates to the fact that the smaller the design gets (approaching, e.g., a wrist watch), the amount of bending gets rather negligible, and the problem reduces to “well-known” aspects of the effect of the user, such as reduced RF power and efficiency, antenna detuning, and changes in antenna impedance and radiation pattern (e.g., [105–109]).

An important aspect to consider, in addition to the effect of the user on the antenna, is the effect of the antenna on the user. When human tissue is in the vicinity of the antenna, part of the RF power is absorbed into the tissue. This absorption is characterised by Specific Absorption Rate (SAR, [W/kg]). For handsets to be allowed on the market, they must fulfil the SAR limit, which is 1.6 W/kg normalised across a 1-g cubical mass of tissue used in the USA [110], or 2 W/kg normalised across a 10-g cubical mass of tissue applied in Europe [111]. Of these, the U.S. limit is tighter, and the SAR results of [VII] are calculated using the 1.6-W/kg limit.

Antenna performance is studied at four frequencies: 722, 920, 2045, and 2595 MHz, which correspond to calculated centre frequencies of particular LTE channels. In these cases, the bent handsets of different sizes are critically matched at the centre frequency, and the symmetric –6-dB bandwidth achieved with a two-element L-section matching circuit is calculated in free space. This bandwidth is compared with that obtained with the user, in which case also the relative resonance frequency shift is observed. In the presence of the wrist, the radiation efficiency ($\eta_{rad}$) and the SAR value are investigated as well.

Compared to the free space case, the resonance frequency mainly tunes downwards in the considered bending state when the wrist of the user...
Conformal Transponder and Handset Antennas

Figure 3.9. Simulated radiation efficiency ($\eta_{\text{rad}}$) as a function of chassis width and length in the presence of the wrist model at (a) 722 MHz, (b) 920 MHz, (c) 2045 MHz, and (d) 2595 MHz. Taken from [VII]; © EurAAP; used with permission.

is introduced. When going from planar to bent case in free space, the frequency tends to tune upwards. The observed downwards-tuning in the resonance frequency in the presence of the user agrees with the perturbation theory, which can be used to show the effect of dielectric and magnetic materials on the resonance frequency of wavemodes (such as those of a handset chassis) [112]. According to this theory, the presence of lossy dielectrics such as user tissue can only decrease the resonance frequency [94, 112].

In addition to frequency detuning, an important quantity to characterise the handset performance in the proposed bending case is the achievable bandwidth. Here, the bandwidth is defined with respect to the –6-dB impedance matching level. Generally, having a good bandwidth in the present usage case is not a problem at the highest frequencies considered (2045 and 2595 MHz) with the different chassis sizes. On the other hand, at the frequencies below 1 GHz, bandwidth starts to become an issue. Some chassis sizes do not even provide a –6-dB matching level. At the lowest frequencies, chassis lengths in the range of 130 mm are suitable in terms of the bandwidth.

As an example of the results, Figure 3.9 shows the obtained $\eta_{\text{rad}}$ performance with different handset sizes. At lower frequencies, where the contribution of the chassis on the overall handset radiation is more sig-
nificant, there is greater variety in $\eta_{\text{rad}}$ compared to higher frequencies. With handsets whose length is 90 to 110 mm and width 40 mm or less, the radiation efficiency is $-6 \text{ dB}$ or worse at 722 MHz. The 920-MHz frequency has two 'hot spots' around 90 and 150-170 mm chassis length that are problematic for the efficiency. At the two higher frequencies, the effect of handset size causes less variance on $\eta_{\text{rad}}$.

In terms of the SAR values, the graphs of [VII] show that the narrow chassis widths (20 and 30 mm) result in SAR up to $6 \text{ W/kg}$, which is clearly above the $1.6 \text{ W/kg}$ limit. This is mainly caused by field absorption occurring across a relatively small surface area on the wrist. To reduce the user exposure, the distance between antenna and wrist could be increased, or additional antenna shielding techniques might be applied.

When analysing the different quantities investigated in [VII], it is clear that with the chassis dimensions considered, no single combination of width and length is simultaneously optimal for all frequencies and quantities. Having a 120-mm long chassis is a reasonable compromise in terms of bandwidth at the lower frequencies (722 and 920 MHz), and when the width exceeds 30 mm, the frequency detuning remains moderate also with the wrist. From the point of view of SAR, a $60 \times 110 \text{ mm}^2$ chassis meets the specification at both low-band frequencies, making this size a fairly good compromise. The main drawback is efficiency, which is from $-3$ to $-6 \text{ dB}$. In this case, the high-band performance is sufficient, apart from the SAR values.

In light of current research, further studies are needed regarding the practical implementability of the concept presented in [VII]. One factor to consider is the quality of the matching as well as the exact location of the operating bands: the critical matching at calculated centre frequencies assumed in [VII] is not the best solution in terms of the multi-band operation required by current and future communications standards. Furthermore, the coexistence of the RF parts and other electronics should be studied, as both are required in an actual product. Whether it could be possible to truly implement the different usage scenarios considered in [VI]–[VII] with the same device remains something of an open question.

### 3.4 Realising bendable handsets in practice

Even though device manufacturers have presented — at least for promotional purposes — different concept designs promoting handsets with
varying degrees of flexibility in their shape, no truly bendable mobile devices have been available in the market. As was observed in the previous sections, realising this type of handsets in practice might not be dependent on the antenna characteristics. Rather, a more likely bottleneck relates to other main components or systems found in the device, such as implementing flexible electronics, battery technology, or probably more significantly, a flexible display module. Today's smartphones feature a large display, and producing this with a flexible form factor has been something of a challenge. Data storage in the flexible handset should also be taken into account, and flexible memory solutions have recently been reported [113].

Nevertheless, there is plenty of interest in implementing flexible or printable electronics [114], also for antennas and other communications applications (see, e.g., [115–119]). When considering the suitability of an overall design, taking into account the different parts and components of the device, also the robustness and reliability of the materials used in the device are key issues, e.g., in terms of the number of bending or folding cycles the structure should withstand over the lifespan of a typical handset [120, 121]. Development in nanotechnology-related fields can also provide important benefits for commercialising a completely new generation of handsets, and one promising solution seems to be the use of thin, flexible, and conductive films based on, e.g., carbon nanobud technology [122].
The antennas for bendable handsets considered in the previous chapter provide conceptual background for implementing a whole new generation of mobile devices. However, one important aspect that was not considered relates to achieving the desired operating frequencies across one or more frequency bands. More precisely, no particular attention was placed on the matching circuits, especially the component values, needed to match the antennas at the correct frequencies.

Current and upcoming fourth- and fifth-generation handsets feature many different radio systems, which all require antennas for their operation. Depending on the frequency and implementation, some systems can share the same antenna(s), thereby reducing the number of different antennas and occupied space [123]. Nevertheless, achieving good matching, efficiency, and isolation levels at the operating frequencies is of utmost importance. With an increasing number of antennas, their proper placement within the handset is important, in terms of overall operation and good performance with the user [124].

Today’s smartphones typically have a large touchscreen display, and compared to devices of previous generations, current devices have larger surface area. One might think that this makes antenna placement easier, but the devices have also become thinner. This, together with an increasing number of antennas means that the available volume for individual antennas may even have reduced. Therefore, it becomes increasingly inevitable to make compromises between the properties of different antennas. In the scope of the current work, implementing broadband and efficient antenna solutions for current and upcoming handset applications is studied in [VIII]–[IX].

This chapter is organised as follows: Section 4.1 explains some of the requirements and properties of multi-band and multi-antenna communi-
Sections 4.2–4.3 present two different MIMO handset antenna designs that provide good performance with fully passive implementations.

4.1 Properties of multi-band and multi-antenna communications

The number of different radio systems in the handsets of previous generations was low. For instance, second-generation devices could feature the low band (e.g., GSM900) and high band (GSM1800/1900/2100) and use single-antenna communications (SISO, Single-Input, Single-Output). The need to operate across a rather narrow range of frequencies meant that covering the required bands with a resonant antenna was relatively easy. Until the late 1990s, external antennas were used in GSM handsets, and they were helical, monopole-type, or a combination of the two [125].

In comparison, the constantly increasing number of frequency bands and radio systems also increases the number of antennas in the handset. These factors make implementing resonant antennas with good efficiency and bandwidth characteristics a challenging task. Therefore, an approach used in many of the current handsets is to utilise non-resonant CCE antennas that use external matching circuits to match the antenna at desired bands [66, 73]. The use of matching circuits means that antenna performance is less dependent on the electrical size of the coupler, but nevertheless, a larger antenna size is beneficial for the low-band operation.

When using external matching circuits together with non-resonant elements, finding a well-performing circuit topology becomes a very important issue. Apart from the different frequency bands that the antennas are expected to cover, also introducing additional antennas for MIMO communications causes challenges for co-designing the antennas and matching circuits. When the number of antenna elements in the handset increases because of MIMO, the inter-element distance decreases. This causes higher mutual coupling between different antenna elements, and additional attention must be paid to design the circuits in a way that the isolation levels remain sufficiently high.

With MIMO communications, the increasing number of antennas typically necessitates the use of smaller radiating elements. Making the physical size of the antenna smaller is detrimental for the performance of the antenna — especially at the low band (below 1 GHz), where the biggest
design challenges of handset antennas are typically observed due to the large wavelength compared to physical dimensions.

Decreasing physical and electrical size of the antenna also reduces the achievable instantaneous bandwidth. As an example, the bandwidth required to cover the entire LTE low band is 262 MHz (698–960 MHz). Covering this frequency range instantaneously becomes impossible if the antennas are made too small. In this case, a solution is to use tunable matching circuits, with which only a narrow part of the required band is covered at a time. By varying the state of the tunable component, operation across the desired frequency band can be obtained (see, e.g., [126–129]). Typically, capacitors are used as the tunable component, and different implementations based on, e.g., Complementary Metal Oxide Semiconductor (CMOS) or MicroElectroMechanical Systems (MEMS) technology can be applied.

Compared to fully passive designs with fixed matching circuits, introducing tunable components increases both the overall design complexity and losses in the matching circuit [130]. In this case, the efficiency and $Q$ of the tunable component become important parameters to consider in the design [131]. Additionally, the antenna impedance varies with different usage conditions (e.g., loading caused by the hand of the user), which can affect the power dissipated in the tuning circuit [132]. Antenna robustness against the effects of mutual coupling can increase when the antennas are made tunable and narrowband. This improves the overall isolation performance [124].

The LTE communication standard also includes the use of Carrier Aggregation (CA) techniques [133, 134]. Antennas supporting this approach communicate using multiple narrow (20–40 MHz) Component Carriers (CC) either within the same frequency band (intra-band CA) or over different frequency bands (inter-band CA). In this approach, having tunable matching networks can be challenging due to a large number of circuit states needed to cover all possible channel combinations. Additionally, the losses in the tuning networks are again an issue for the efficiency. Changing the state of a tunable component also involves some nonlinearities that can generate additional frequencies that may occur at undesired parts of the Tx or Rx band within the same or different transceiver.
4.2 Multi-element antenna with closely-located radiators

The following section describes the design of a two-element MIMO antenna that utilises multiple closely-spaced metal radiators or couplers [VIII]. By having the radiators close to each other physically and electrically, they inherently couple strongly to each other, and the antenna performance can be improved especially at the low band using suitable matching techniques.

4.2.1 Mutual coupling in handset antennas

Traditionally, when designing handset antennas, the designers want to have the level of inter-element mutual coupling sufficiently low to obtain well-performing designs in terms of, e.g., efficiency and correlation. If the different antennas have too strong coupling between them, suitable decoupling techniques need to be applied. Various ways to implement antenna decoupling have been reviewed in, e.g., [135]. Many publications have dealt with this issue, and possible decoupling techniques include:

1. Varying the spacing and orientation of the antennas.
4. Using defected or modified ground planes.
5. Applying neutralisation lines between the antennas.

Challenges related to these solutions are, e.g., that they can be fairly narrowband, and in some cases also bulky. The effect of the user should be taken into account not only when designing the handset antennas but also when implementing the decoupling structures. In some cases, making the decoupling structure tunable can be beneficial to improve the antenna performance with the user [136].

In the literature, there are also works that attempt to take advantage of mutual coupling. As an example, the placement of multiple feeding ports on a single radiating element has been investigated in [133, 137]. In this approach, having the feeds physically connected to a single metal piece results in a very high mutual coupling level, which, consequently, makes designing the matching circuits very difficult [138].
4.2.2 Utilisation of mutual coupling to improve antenna performance

A different approach utilising the mutual coupling between radiators that are located close to each other both physically and electrically is considered in [VIII]. In this case, the majority of the antenna elements are positioned around the edge of the substrate, as shown in Figure 4.1. For reasons related to both aesthetics and structural robustness, many current handsets have a metal rim around the edge of the device. Recently, one of the trends in antenna design has been to utilise this rim partially or in its entirety as an antenna (see, e.g., [139, 140]). One additional benefit of having antennas on the outer edge of the device is the possibility to make the antenna elements larger. This helps to improve the operation especially at the low band. With rim-positioned antennas, the antenna volume can be increased without making the occupied surface area unnecessarily large.

The antenna design of [VIII] is considered for a handset with overall dimensions of $160 \times 80 \times 5 \text{mm}^3$ (length $\times$ width $\times$ thickness), which represents a typical size for a current high-end smartphone. Initially, the fully-symmetric antenna element consisted only of a single metal piece above and around the substrate, with the possibility of connecting up to three ports to it. The applied design methodology starts from the low band. Similarly as in [133, 137], there is a clearance, i.e., metal-free area, between the antenna and chassis, but the previously-reported design only had two ports for the low- and high-band feeds. The idea of using three possible feed ports is to see whether it would be beneficial for the operation to use some of the ports simply to load the antenna with a lumped component (capacitor or inductor) instead of feeding all available ports.

4.2.3 Performance improvement through asymmetry and impedance matching

With the three-port single-element design, the mutual coupling levels between the different ports are very high, as could be expected, at around $-4 \text{ dB}$ across the entire LTE low band. A number of matching network alternatives were considered in Optenni Lab, but in all cases the same fundamental problem remained: getting the input matching and mutual coupling to a level providing sufficient bandwidth and good efficiency was practically impossible.
To get a better starting point for properly matching the antennas, the single-element design is modified in two principal ways: 1) making the open ends of the element asymmetric (different lengths) and 2) cutting a slot around the top part of the antenna element. Suitable antenna lengths and slot widths are determined using parametric studies in the EM simulator. These modifications, combined with separating the element around the substrate in two, result in two L-shaped and one C-shaped element of different dimensions. Having different-sized elements means that they have their fundamental resonances at different frequencies, which can help to implement well-performing matching circuits. Here, the term ‘fundamental resonance’ should be understood as the natural matching level of the unmatched antenna, which should not be confused with operation obtained together with actual matching circuits.

Figure 4.2 depicts the $S$-parameters (input matching and mutual coupling) of the modified three-element antenna and embedded radiation efficiency ($\epsilon_{\text{rad}}$) without any matching circuits. The $\epsilon_{\text{rad}}$ quantity is calculated from the $S$-parameters in the following way [7]:

$$\epsilon_{\text{rad},i} = 1 - |S_{ii}|^2 - \sum_{i \neq j} |S_{ij}|^2. \quad (4.1)$$

This expression considers both the power reflected at the input of the $i$:th antenna port as well as the power coupled between ports $i$ and $j$. 

Figure 4.1. Illustration of (a) the three-element antenna structure used as the main and diversity antenna and (b) the antennas in MIMO configuration. The pink prism depicts the metal block used to model the display and other electronics. Dimensions are in millimetres. Taken from [VIII]; © EurAAP; used with permission.
When looking at the $S$-parameter curves of Figure 4.2, one can see that at the low band, the three antenna elements have different matching and mutual coupling characteristics. Inherent matching levels are quite good (e.g., $S_{33} < -6$ dB across the LTE low band), but the high level of mutual coupling keeps the efficiency low (around 20–40%). To achieve good and efficient performance, matching circuits are designed using Optenni Lab.

The corner-fed, L-shaped antennas are considered for the low-band operation, and element 1 is chosen as the low-band antenna. The low band was implemented using the low-pass type circuit of Figure 4.3(a). The other corner-fed antenna, element 2, is chosen for high-band operation, and it is matched with the high-pass type network of Figure 4.3(b). The third, C-shaped element is not excited, and it is connected to the chassis through a grounding inductor, as shown in Figure 4.3(c). When determining suitable circuit topologies and component values for the matching circuits, both circuits were optimised together to achieve good matching at the desired passband and as perfect a mismatch as possible at the targeted stopband. Having one element matched and the other one mismatched at a given band reduces the mutual coupling between different
antennas and results in more power being radiated by the antenna, i.e., in higher efficiency.

All circuits used in the antenna of [VIII] are implemented with GQM18 and LQW18 series capacitors and inductors from Murata, based on \( S \)-parameter based component models available from the manufacturer. Although the antenna structure explained above has three, physically separate radiators, the analysis of its performance is more straightforward if this structure is considered as a single 'unit'. Subsequently, this unit is called the main antenna.

### 4.2.4 SISO and MIMO performance of the proposed antenna

The performance of the antenna concept proposed in [VIII] is investigated in three cases: 1) SISO case, 2) MIMO case, and 3) MIMO case in the presence of a large metal block. Figure 4.1 visualises these different cases. In the SISO case, only the main antenna is present in the structure (single-ended design), and the MIMO case is obtained by duplicating the main antenna to the opposite end of the handset (double-ended design).

In the second MIMO case, the suitability of the proposed antenna is considered from the point of view of including a simplified model of the display and other electronics found within the device. As all of the antennas are located around the edge of the device, they form an enclosure, within which all other components should in practice be placed. Figures 4.4 and 4.5 illustrate the matching and efficiency performance of the SISO and MIMO cases, respectively.

As the curves of Figure 4.4 show, the proposed main antenna can instantaneously cover the LTE low band with a matching level better than \(-6\) dB. In the high-band case, the matching is better than \(-4\) dB. This does not quite meet the traditional \(-6\)-dB condition, but the main focus in the
design was placed on the low band and on getting as good an overall efficiency across the low and high bands as possible. Generally, emphasising the efficiency rather than matching is done also in many commercial antenna designs [130]. At the low and high bands, the \( e_{\text{rad}} \) performance is better than 80% and 60%, respectively.

Figure 4.5 depicts the \( S \)-parameter and embedded radiation efficiency performance in the two-element MIMO case, when similar main and diversity antennas are used in opposite ends of the device. When going from single to multi-antenna case, the resulting main and diversity antennas are aligned symmetrically. This kind of antenna orientation has been found to be beneficial for the isolation [140]. Both antennas use similar matching circuit topologies as in the single-ended case, but some individual circuit element values are changed, as indicated by the values in Figure 4.3. The circuits are fine-tuned to get the MIMO antenna matching and mutual coupling performance closer to that of the SISO case.

Comparison of the results of Figures 4.4 and 4.5 reveals that the most significant change in performance occurs in the low-band efficiency. Its value drops from above 80% in the single-ended case to around 60% at
best in the MIMO case. The worst low-band mutual coupling is around –4 dB, which is problematic for practical MIMO implementations. These observed effects are chassis-related, as the main and diversity antennas share the same chassis.

In the second MIMO case, a large metal block is placed on the chassis to model the other electronics of the device. The block, shown in Figure 4.1(b), has dimensions that represent a worst-case scenario in terms of potential display size and overall device thickness. The metal block has the most detrimental effect on the antenna performance at the high band, where the minimum efficiency decreases from 60% to 40% when the block is introduced. This is mainly caused by additional coupling between the high-band antennas occurring across the block. Having additional metal near the antennas does not therefore destroy the overall feasibility of the antenna concept, and its effects can most likely be largely compensated by considering a sufficiently realistic handset model when designing both the antennas and the matching circuits.

Figure 4.5. Simulated S-parameters and embedded radiation efficiency ($\varepsilon_{\text{rad}}$) of the main and diversity antennas in the two-element MIMO case without the metal block. Taken from [VIII]; © EurAAP; used with permission.
4.3 Physical antenna diversity in MIMO handsets

The MIMO antenna proposed in [VIII] utilises perhaps the simplest possible approach to implement multi-antenna functionality: copying similar antennas to different parts of the handset. Even though this is an often-used approach, it has also some drawbacks. One main issue is that with similar antennas, also their radiation patterns resemble each other. This in part decreases the benefit of using antenna diversity techniques in MIMO communications. In the following, quantities and parameters used to characterise MIMO performance in the handset are described, and a multi-antenna implementation presented in [IX] utilising physically different antennas for enhanced performance is analysed.

4.3.1 Characterisation of multi-element handset antennas

The use of multi-element antennas in actual multi-antenna applications requires considering particular parameters or metrics, such as channel capacity, envelope correlation coefficient (ECC), diversity gain, and multiplexing efficiency ($\eta_{\text{mux}}$). Depending on the signal-to-noise ratio (SNR) of the communication environment, MIMO can be used either in diversity or in spatial multiplexing mode [141, 142]. In the diversity case, typically used in low-SNR environments, some of the fading effects can be remedied through the use of several antennas, and thus the quality of the communication link improves. Spatial multiplexing is the preferred technique in high-SNR environments, with the main purpose being the enhancement of the data rate.

In the scope of this work, including that of [IX], the main MIMO parameters considered are envelope correlation [7, 143] and multiplexing efficiency [144]. The ECC describes how similar (i.e., correlated) the radiation patterns of two antennas are, and its values range from zero (completely uncorrelated antennas) to one (completely correlated antennas). Typically, an ECC value of $\rho_e < 0.5$ is desired in practical implementations for efficient utilisation of diversity techniques.

The most general way of calculating the envelope correlation is to use antenna field patterns obtained either from simulations or measurements. In terms of computational workload, the fields-based approach is cumbersome especially across wide frequency ranges, but the results are valid for any antenna. An alternative technique is to calculate the ECC from $S$-parameters [145]. This method is fast and simple, but it is strictly valid
for lossless antennas. For antennas with high efficiency (and low losses), the $S$-parameter approach is a reasonable approximation [146, 147]. For antennas with higher losses, the $S$-parameter based results typically provide over-optimistic $\rho_e$ values especially at the low band.

Multiplexing efficiency characterises the degradation in SNR of antennas with particular efficiency, efficiency imbalance, and correlation in a given multi-antenna scenario compared to ideal MIMO antennas [144, 148]. Computationally, it is a rather simple metric that allows the designer to take into account the most significant properties when designing a well-performing MIMO system. The analysis of this work only investigates the free-space case, but more detailed scenarios should also consider the MIMO performance in the presence of the user [149].

### 4.3.2 Parasitic-coupled aperture-matched main antenna

The previously described three-element design for the main and diversity antennas provides good performance, but it also has some challenges related to its implementation. With separate low- and high-band feeds, eight matching components are needed per antenna, and the requirement of two feeds also complicates the design of the RF front-end needed to drive the antennas. Additionally, the identical main and diversity antennas of [VIII] result in very high ECC values at the low band, with $\rho_e$ up to 0.9 in the worst case.

For these reasons, the work of [IX] studies alternative, possibly more straightforward ways of implementing an antenna with comparable performance. Designing the antenna (subsequently called the main antenna) is done by co-optimising antenna geometry and the matching circuits in CST and Optenni Lab. As the design goal, an embedded radiation efficiency better than 80% was required at the low band (698–960 MHz) and better than 60% at the high band. The matching circuits were allowed to have at most four components per antenna element.

Figure 4.6 shows the dimensions and details of the optimised main antenna, and Figure 4.7 gives the matching circuits used. The best performance in terms of efficiency and impedance matching was achieved when the I-shaped metal strip located above the substrate is fully parasitic, and only one of the L-shaped corner elements is fed. The other L-shaped element is coupled to the chassis using an aperture-matching inductor. With this approach, the main antenna instantaneously covers the LTE low band with a matching level of $-4$ dB, and the total efficiency is better
Figure 4.6. Illustration and dimensions of (a) the Combined Parasitic-coupled Aperture-Matched (CPAM) main antenna and the CCE diversity antenna and (b) close-up of the CPAM antenna highlighting the different parts of the structure. All dimensions are in millimetres. Modified from [IX] (reproduced courtesy of The Electromagnetics Academy).

Figure 4.7. Matching circuits used in the main and diversity antennas both in simulations and measurements. Component values are those of actual Murata capacitors and inductors. Taken from [IX] (reproduced courtesy of The Electromagnetics Academy).

than –4 dB at both the low and high band (see Figures 4.8 and 4.9). Due to geometrical details, the main antenna of [IX] is referred to as a Combined Parasitic-coupled Aperture-Matched (CPAM) antenna.

4.3.3 Capacitive coupling element diversity antenna

To study the possible benefit of having different main and diversity antennas in the handset, the diversity antenna of [IX] is implemented using a more conventional approach. The symmetric nature of the ground clearances around the handset (see Figure 4.6) leaves a 80×7 mm$^2$ area in which to position the diversity antenna in the opposite end of the device.

Figure 4.6 illustrates the diversity antenna, for which a typical CCE design is used. It is well known that the performance of small handset
Figure 4.8. Impedance matching and mutual coupling of the main and diversity antennas in the two-element MIMO case. Port 1 = main antenna and Port 2 = diversity antenna. Taken from [IX] (reproduced courtesy of The Electromagnetics Academy).

antennas at the lowest operating frequencies benefits from increased antenna size. For this reason, the available volume is used efficiently, and the coupling element is made both as large and as simple as possible.

The design workflow of the diversity antenna follows a process presented in [150]. In this approach, the low-band operation is considered first, and here, the performance improves when the antenna volume is maximised. To achieve good operation at the high band, a high-pass type matching network is needed (shown in Figure 4.7). Four matching components provide good performance also at the low band. For enhanced high-band performance, the feeding pin of the diversity antenna is tapered into a triangular shape.

Compared to the CPAM main antenna, the coupling-element based diversity antenna has similar bandwidth and matching level at the low band, as shown in Figure 4.8. At the high band, the diversity antenna has better overall matching performance. In [IX], both the CPAM and CCE antenna are designed with the goal of emphasising efficiency rather than matching. From this point of view, the obtained designs are good and suitably simple.

4.3.4 Multi-antenna performance

A prototype of the two-element design of Figure 4.6 is manufactured to experimentally verify the operation. Figure 4.8 shows the simulated and measured $S$-parameters of the two-element antenna. The results show that especially at the low band, the simulated and measured input matching of the main and diversity antennas are in very good agreement. At
the high band, there are larger differences, which are mainly caused by the sensitivity of the main antenna to the position of the parasitic coupler. In the case of the diversity antenna, the challenges in connecting the antenna element to the chassis provide a source of inaccuracy and uncertainty compared to the simulations. The measured low-band mutual coupling is better than $-10$ dB, which is a sufficient level for practical MIMO implementations. In simulations, the mutual coupling of the antenna of [IX] is better than that of [VIII]. At the high band, antenna coupling is not an issue either in simulations or measurements.

Figure 4.9 presents the efficiency of the main and diversity antennas in the two-element MIMO case. Only simulated $\eta_{\text{rad}}$ values are given, but the $\eta_{\text{tot}}$ values are also measured. Including the second antenna to the handset reduces the low-band total efficiency of the diversity antenna approximately 1 dB, and the efficiency of the main antenna decreases more. The reduction in efficiency is partially caused by the decrease in $\eta_{\text{rad}}$ related to the mutual coupling. In the high-band case, the total efficiency is mainly affected by the matching, as the high-band radiation efficiency does not change considerably when adding the second antenna.

The efficiency achieved with the antenna of [IX] can be compared to other MIMO designs presented in recent literature, such as [126,140,151,152]. The work of [151] features a four-element MIMO implementation at the low band without tunable matching circuits, but there is no high-band MIMO. In [152], both low- and high-band MIMO are considered, but the low band starts from 750 MHz, whereas the design of [IX] covers
also the 700-MHz band. Low- and high-band MIMO are implemented with tunable matching circuits or switches in [126, 140], and even with tuning, the MIMO design of [140] does not operate at the most challenging LTE low-band frequencies. In light of this, the efficiency performance of [IX] is good, especially when considering that the matching levels are not the most optimal ones. Generally, the design of [IX] stands out from the literature by having both low- and high-band MIMO starting from 700 MHz with fully passive matching circuits. This makes it very suitable, e.g., for CA applications.

For the MIMO characteristics of the antenna in [IX], the envelope correlation and multiplexing efficiency are determined both from simulations and measurements. Here, ECC is calculated from the field patterns. Figure 4.10 presents the ECC and $\tilde{\eta}_{\text{mux}}$ results. The simulated ECC is quite high at the low band, and it falls below the typical requirement at around 850 MHz. At the high band, the correlation values meet the requirements. Using the field-based ECC calculations, the low-band performance represents typical values, as the main and diversity antennas have quite similar radiation patterns (even though they are physically different).

In order to make the design of [IX] more suitable for practical MIMO applications, the low-band ECC value should be reduced. This can be achieved using a number of different techniques, such as utilising antennas with orthogonal radiation patterns. Ways of achieving this include the use of electric and magnetic antennas, and suitable excitation of orthog-
onal characteristic modes of the chassis [153]. Other possible solutions include the use of various decoupling slot designs or common radiators between the antennas to control the isolation and ECC (see, e.g., [154, 155]). For the case of tunable antennas, a tunable coupling capacitor is suggested in [156] to achieve a similar kind of effect.

Of the previous approaches, the decoupling-based ones may be bulky and narrowband, which can limit their ability to provide proper ECC performance, e.g., across the entire LTE low band. On the other hand, solutions based on the inherent orthogonality would seem a proper approach. In this case, and indeed with all of the decoupling methods, the coexistence of the low and high-band antennas should be kept in mind. Attempting to aim for a very low LB ECC value can pose some challenges for the operation of the antennas meant for higher frequencies, forcing the designer to sacrifice the performance of the device somewhere.

The average $\tilde{\eta}_{\text{mux}}$ of the design in [IX] is from $-5$ to $-7$ dB in simulations and measurements at the low band, and at the high band, simulated and measured $\tilde{\eta}_{\text{mux}}$ values are from $-2$ to $-3$ dB and $-4$ dB, respectively. These values are comparable to, and in the low-band case, even better than those of the state-of-the-art antenna design in [126].

Based on the expression found in [144], the multiplexing efficiency performance essentially converges to the mean of the antenna efficiencies (in dB scale), and the antenna with lower efficiency has the most significant contribution to the overall $\tilde{\eta}_{\text{mux}}$. One way forward to improve the $\tilde{\eta}_{\text{mux}}$ performance of the current antenna would be to have better matching level. However, this can increase the coupling between the main and diversity antennas and also be detrimental for the envelope correlation.
5. Summary of Publications


Theoretical and experimental studies are performed on a harmonic transponder operating at frequencies 1 and 2 GHz. Using an equivalent circuit model, the performance of the transponder can be accurately predicted, and general figures of merit are derived to characterise the suitability of particular antennas and diodes for specific transponder implementations. The designed transponder shows a read-out range of at least 5 m in experiments.

Publication II: “Antenna Matching at Harmonic Frequencies to Complex Load Impedance”

This work describes the design of a direct-matched, self-resonant transponder antenna to a complex, frequency-dependent load impedance typical in harmonic radar applications. The design workflow and the contribution of different antenna parts are explained. In order to reliably characterise the two resonances at desired frequencies, the use of a proper normalisation impedance is crucial.

Publication III: “Harmonic Transponders: Performance and Challenges”

In this publication, the performance of a self-resonant, direct-matched transponder antenna is investigated theoretically using altogether six different varactor and detector diodes. The results of the work show that
for this type of antenna, having a significant impedance mismatch between the antenna and the diode especially at the fundamental frequency is detrimental for the transponder performance.

**Publication IV: “Designing Harmonic Transponders Using Lumped-Component Matching Circuits”**

This work presents harmonic transponder designs using lumped-component matching circuits instead of direct matching. The proposed approach considerably simplifies the antenna design process, and the same antenna geometry can be used with different diodes by varying the matching circuit. Theoretical and experimental studies are performed, and the agreement between the results is good.

**Publication V: “Effect of Shape and Surroundings on Harmonic Transponder Performance”**

The sensitivity of a direct-matched harmonic transponder to environmental effects is investigated in two cases: having a dielectric object in the vicinity of the transponder, and bending the transponder about a cylindrical surface with and without an object. The results show that the proposed transponder is not very well suited for operation with the dielectric object, but the transponder concept has potential to be used for strain or other conformal sensing applications.

**Publication VI: “Effect of Ground Plane Bending on Mobile Terminal Antenna Performance”**

Operation of a CCE handset antenna is studied using computer simulations and a resonator-based equivalent circuit model in a case when the chassis is subject to bending. The results demonstrate that compared to the planar case, resistive, capacitive, and inductive changes can take place, and that the most significant effects of bending occur at frequencies below 1 GHz, where the chassis has a large contribution to the overall handset radiation.
Publication VII: “Investigation on Bendable Mobile Devices in the Presence of the User”

This work investigates the performance of a CCE handset antenna in wrist-worn applications. Different handset dimensions (length and width) are considered at different frequencies and with respect to different quantities, including SAR. Based on the studies, none of the considered chassis sizes is optimal for all quantities and frequencies in this particular usage case, but a 60-mm wide and 110-mm long chassis is a reasonable compromise in most cases.

Publication VIII: “LTE Handset Antenna with Closely-Located Radiators, Low-Band MIMO, and High Efficiency”

In this publication, an LTE MIMO handset antenna is designed using antennas with separate metal radiators located in close proximity to each other. Matching circuits are used to achieve good matching and isolation performance both at the low and high bands. The antenna is studied using simulations. In the MIMO case, the proposed main and diversity antennas achieve up to 60% and better than 60% embedded radiation efficiency at low and high band, respectively.

Publication IX: “Carrier Aggregation Compatible MIMO Antenna for LTE Handset”

This work presents a novel main antenna design for LTE MIMO handset applications. The main antenna uses a combination of parasitically-coupled and aperture-matched elements to instantaneously cover the LTE low band with a fixed matching network. Together with a conventional CCE diversity antenna, the proposed design achieves good performance both in simulations and measurements. The fully passive implementation with both low- and high-band MIMO is compatible with the requirements of inter- and intra-band carrier aggregation.
6. Conclusions

This doctoral thesis has three main parts, and it studies the design and performance of harmonic transponders, conformal antennas for wireless sensors and bendable handsets, as well as LTE MIMO handset antennas. The different designs are investigated using theoretical calculations, computer simulations, and experimental verifications.

The wireless sensors that are investigated in the first part of the thesis are harmonic transponders that use harmonically separated operating frequencies at $f_0$ and $2f_0$ for communicating with the reader device. The use of different transmitting and receiving frequencies increases the robustness of the transponders to environmental clutter. Publications [I]–[IV] study two transponder designs implemented using different varactor and detector diodes. In order to match the transponder antenna to the diode at the required harmonic frequencies, both implementations based on direct matching and external, lumped-component matching circuits are considered. Both of the matching approaches have their benefits and drawbacks in terms of overall design complexity and suitability for operation with diodes having significantly varying impedance levels.

In the second part of the thesis, conformal antenna designs are analysed. Publications [V]–[VII] study antennas and their performance in future-generation, bendable handset applications, as well as the suitability of harmonic transponders for conformal applications. Performed investigations using computer simulations and equivalent circuit models provide both improved physical understanding of the effects of bending the handset chassis, as well as information on the performance and effect of the user in a particular, wrist-worn scenario. Due to the significant radiation contribution of the chassis at frequencies below 1 GHz, bending the handset has the most significant effect on antenna performance at these frequencies.
The final part of the thesis deals with multi-antenna (MIMO) designs for LTE handset applications. Publications [VIII]–[IX] utilise antennas with several closely-located and physically separated metal radiators that inherently have strong mutual coupling to achieve good low- and high-band performance with suitable matching circuits. Particularly the novel Combined Parasitic-coupled Aperture-Matched (CPAM) antenna used in [IX] provides good LTE performance with a simple and fully passive matching circuit. By placing most parts of the antenna elements around the corners of the handset, the antenna volume can be increased without occupying an excessively large surface area. The use of physically different main and diversity antennas (so-called physical antenna diversity) can improve the MIMO performance especially at the low band.

Even though the wireless sensor and handset antenna implementations featured in this thesis contribute to solutions and requirements of current generation communications schemes, upcoming standards and specifications for wireless communications will provide new aspects and challenges to take into account. The currently-developed and considered fifth-generation (5G) communications will include orders of magnitude more of basically everything: more devices and users, higher frequencies and data rates, broader bandwidths and so on, while still supporting the 'legacy' systems at frequencies below 6 GHz. Completely new concepts are needed for 5G communications, also on the antenna side, to meet the requirements. One suggested approach is to use beam-steerable antennas with narrow beams at millimetre-wave frequencies to provide the necessary performance [157]. Regardless of the way chosen in future communications systems, one thing remains certain: skilled antenna and RF engineers will be kept busy also in the decades to come.
References


References


References


Errata

Publication I

Equation (4) should have a minus sign in the denominator.

Publication V

In Table I, the directivity $D_{\text{max},f_0}$ should be replaced by the gain $G_{\text{max},f_0}$ also for the case of a PVC cylinder.

Publication VI

In Fig. 4, the first-order ground plane wavemode resonator should read $R_{c1,b}$ instead of $R_{c2,b}$ inside the dotted rectangle, and the second-order ground plane wavemode resonator should read $C_{c2,b}$ instead of $C_{c2}$ inside the dashed rectangle.
Designers of wireless communications systems face many challenges, including an increasing number of connected devices and users, a need for higher and higher data rates, new frequency ranges and radio systems. This thesis contributes to solving these issues by designing, implementing, and characterising new antenna solutions for wireless sensors and mobile handsets by using simulations, measurements and circuit models. Compared to their linear counterparts, the harmonic transponder sensors studied can be detected more reliably in clutter-rich environments, but their operating principle leaves some design intricacies to consider. On the handset side, antenna performance is considered both for future, bendable devices, and for current multi-antenna communications schemes. The research outcomes of the thesis provide valuable design rules and considerations for the developers of even better performing antennas of the future.