

Imperial College London
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Coherent ultra-wideband radar-on-chip for biomedical sensing and imaging

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I declare that this thesis has been composed solely by myself. Except where stated otherwise in the text, by footnote or reference, the work presented is entirely my own.

London, December 2019
Timo Lauteslager

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Abstract

Medical microwave imaging (MWI) has been investigated in the past two decades due to the attractive properties of microwave radiation: Electromagnetic waves at microwave frequencies can penetrate biological tissue and are non-ionising. Several MWI prototypes have been developed for breast cancer imaging and stroke detection, some of which are currently under clinical investigation. The recent innovation of coherent ultra-wideband (UWB) radar-on-chip (RoC) technology could enable MWI at a much lower cost and with portable or even wearable devices. Due to the ability to operate multiple RoC receivers simultaneously, multistatic radar data can be acquired at a much higher rate than when using conventional microwave techniques. This would allow for dynamic imaging of the cardiovascular system: An application which has been mostly out of reach for MWI systems.

This thesis describes the work undertaken to investigate the use of coherent UWB RoC technology for biomedical sensing and imaging, specifically of the cardiovascular system. Imaging hardware was developed by research partners University of Oslo and sensor company Novelda. A modular system of simultaneously operated RoC transceivers with body-coupled antennas was constructed. Original contributions were made by generating system requirements, characterising MWI hardware, developing signal processing and imaging algorithms, and experimentally validating the system in human participants. More specifically, microwave interactions with biological tissues along with the anatomy and physiology of the cardiovascular system were considered to generate system requirements. Developed hardware was tested and characterised experimentally and its feasibility for in-body sensing and imaging was assessed. Sensing experiments were performed with a single radar module, to demonstrate the ability of monitoring cardiovascular dynamics in the human body, and to explore the challenges associated with in-body radar signal processing. Finally, different antenna arrays with multiple synchronised radar modules were used to perform two-dimensional imaging: Dynamic imaging of both the femoral artery and the heart were demonstrated, as well as static imaging of a tissue-mimicking phantom. Simulated data were used to inform and optimise imaging algorithms.

To our knowledge, this work was the first to demonstrate imaging in the human body using UWB RoC technology, and the second work to demonstrate dynamic MWI. Although many challenges remain, this work signifies that RoC technology could potentially enable accessible

and low-cost devices for the assessment of cardiovascular function.

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Acronyms

1D one-dimensional.

2D two-dimensional.

3D three-dimensional.

BMI body mass index.

BP blood pressure.

CBV cerebral blood volume.

CMI confocal microwave imaging.

CMOS complementary metal-oxide semiconductor.

CW continuous-wave.

DAC digital-to-analogue converter.

DAS delay-and-sum.

DC direct current.

DMAS delay-multiply-and-sum.

DOA direction of arrival.

ECG electrocardiography.

EM electromagnetic.

FCC Federal Communications Commission.

FDTD finite-difference time-domain.

FFT fast Fourier transform.

fMRI functional magnetic resonance imaging.

GM grey matter.

HR heart rate.

ICL Imperial College London.

ICNIRP International Commission on Non-Ionizing Radiation Protection.

IQ in-phase and quadrature.

MDO module delay offset.

MIMO multiple-input and multiple-output.

MIST microwave imaging via space-time.

MWI microwave imaging.

P2P peak-to-peak.

PAT pulse arrival time.

PCA principal component analysis.

PSF point spread function.

PTT pulse transit time.

PWV pulse wave velocity.

RF radio frequency.

RMS root mean square.

RoC radar-on-chip.

Rx receiver.

SAR synthetic aperture radar.

SAR specific absorption rate.

SCR signal-to-clutter ratio.

SDO sample delay offset.

SIMO single-input and multiple-output.

SmCmR signal-maximum-to-clutter-maximum ratio.

SNR signal-to-noise ratio.

ToF time of flight.

Tx transmitter.

UiO University of Oslo.

ULA uniform linear array.

URA uniform rectangular array.

US ultrasound.

UWB ultra-wideband.

VNA vector network analyser.

Chapter 1

Introduction

1.1 Motivation

In developed markets such as the United Kingdom, the ageing population and changing care patterns have significantly driven up cost of healthcare, while funding has not kept up. Advances in medical technology may improve the quality of care, but do also prompt the demand for more and more costly diagnostic tests and interventions. Providing appropriate healthcare for our ageing population, despite scarcity of resources, is one of the challenges of our age. Innovation can provide relief, but should be directed towards areas that can improve healthcare outcomes considering limited resources: Preventive screening, personalised preventive care, promotion of health, remote patient monitoring, and an increased focus on primary care [1]. Technological advances in the fields of communication, consumer technology and data analytics allow for an infrastructure where ambulatory patient data are collected continuously for early diagnosis of disease or patient deterioration. An overhaul in society's views on privacy is required, as well as modernised regulations on patient data access and data integration. Simultaneously, technological advancements are required in the field of diagnostic sensing and imaging, to scale up data collection outside the hospital environment.

Since the accidental discovery of X-ray by Wilhelm Röntgen in 1895, medical imaging has revolutionised medical diagnostics. Nowadays, an MRI scanner can be used to image the human body with millimetre spatial resolution, whereas ultrasound (US) can be used in point-of-care imaging to generate real time videos of an unborn child. Despite these advances, there exists a constant need to improve existing technologies and investigate new methods of medical imaging. In light of the shifting focus of healthcare towards preventive diagnostics and treatment, new imaging modalities should be designed to be low cost, accessible, harmless, and portable. The Butterfly iQ (Guildford, United States) is a great example of this shift in medical technology:

A portable, hand-held US transducer that uses a mobile phone for processing and interfacing. Due to the use of complementary metal-oxide semiconductor (CMOS) technology [2], the device is pocket-sized and comes at the comparatively low cost of 2000 USD [3].

Microwave imaging (MWI) is a relatively new imaging modality, which also has the potential to fit these requirements. Imaging relies on the dielectric contrast that exists between tissues. Electromagnetic (EM) radiation at microwave frequencies penetrates biological tissue and is non-ionising. MWI has been researched for breast cancer imaging and more recently for stroke imaging. Several companies are in the process of commercialising MWI technology, and a number of devices are currently under clinical investigation. What most of the existing MWI systems lack, is the advantage of portability and low cost. The recent development of a CMOS implementation of an ultra-wideband (UWB) radar system has given rise to the question whether radar-on-chip (RoC) technology could be used for biomedical sensing and imaging. To the best of our knowledge, this research project is the very first effort to investigate the use of RoC technology for monitoring the dynamics of subcutaneous anatomical structures, and to develop an imaging system based on RoC. A system like this would have the advantage of very low cost and form factor: The radar chip itself has a cost in the order of 1 GBP, and is about the size of a human fingernail. Overall system cost and size will be application dependent, but considering advancements in microelectronics, cost and size could be orders of magnitude lower than imaging system based on conventional technology such as a vector network analyser (VNA). Due to the small size and low power requirements, a RoC MWI system would be portable, and even integration into wearable technology would be possible. Perhaps even more exciting, is the opportunity of dynamic imaging. Multiple RoCs can be run simultaneously to achieve high imaging speed without loss of sensitivity. Time-lapse imaging at a high rate would enable cardiovascular imaging or even functional neuroimaging: Applications which have been mostly out of reach for MWI systems based on conventional technology.

It is anticipated that the use of RoC will result in a unique set of challenges. Compared to conventional technology, RoC will have far less flexibility in transmitted signal properties, and it is yet to be seen whether RoC devices can achieve similar sensitivity as say a VNA. A portable or even wearable system would need to have some degree of flexibility in geometry, to accommodate for different body parts. This implies that the level of calibration which can be achieved with a fully static system will be out of reach. Conventional solutions for system calibration, signal leakage reduction and clutter rejection, might not be applicable for the RoC approach. Such challenges and others will be investigated as part of this research.

1.2 Objectives

This thesis describes the work undertaken in the past five years to investigate the use of coherent UWB RoC for biomedical sensing and imaging. The idea for the project was conceived through a collaboration with University of Oslo (UiO), and sensor company Novelda. A previous effort by Novelda had resulted in a RoC designed for subcutaneous heart inspection. The availability of this technology, combined with expertise at Imperial College London (ICL) in neural interfaces, led to the question: Can RoC be used for functional neuroimaging? Although functional neuroimaging would be extremely valuable from a clinical point of view, this goal was soon realised to be too ambitious, and many challenges would have to be overcome to achieve the required imaging sensitivity. The objectives set here, were thus constrained by the limitations of available technology. To demonstrate the use of UWB RoC for biomedical applications, cardiovascular imaging was identified as a feasible target. To achieve this, the following objectives were set:

- **Feasibility:** Consider the anatomy and physiology, as well as the physics of EM wave propagation through tissue to assess the feasibility of cardiovascular imaging, functional neuroimaging, and dielectric spectroscopy for blood glucose sensing.
- **System requirements:** Consider the design aspects of a MWI device, their interdependencies, and generate a first set of system requirements for a dynamic cardiovascular imaging system based on UWB RoC.
- **Hardware characterisation:** Characterise the available hardware as developed by collaborators UiO and Novelda AS. Assess their suitability for biomedical sensing and imaging.
- **In-body radar signal processing:** Consider the challenges of using conventional radar signal processing techniques for in-body sensing and develop biomedical radar signal processing recommendations.
- **UWB radar sensing of cardiovascular dynamics:** Investigate the challenges involved with RoC measurement of subcutaneous anatomical structures through 1D UWB radar sensing of cardiovascular dynamics.
- **Imaging algorithm design:** Consider anatomy and system specifications to inform imaging algorithm design and parameters.
- **UWB radar imaging of cardiovascular dynamics:** Provide the first demonstration of two-dimensional (2D) biomedical imaging using UWB RoC technology.

Project participants

This research project is a collaboration between different parties: Imperial College London, University of Oslo, and Novelda. A short introduction of the members of the project and their contributions follow below:

- **Timo Lauteslager**, Imperial College London, United Kingdom
Responsible for experimental characterisation of hardware, signal processing and image algorithm development, experimental validation of the developed system in human participants.
- **Mathias Tømmer**, University of Oslo, Norway
Responsible for antenna design, imaging device control system development, antenna array development.
- **Dr. Kristian Kjelgård**, University of Oslo, Norway
Responsible for radar hardware development, antenna design.
- **Novelda AS**, Kviteseid, Norway
Responsible for radar hardware development.
- **Prof. Tor ‘Bassen’ S. Lande**, University of Oslo, Norway
Supervision, conception and direction of work, expert knowledge on RoC.
- **Dr. Timothy G. Constandinou**, Imperial College London, United Kingdom
Supervision, conception and direction of work.

1.3 Thesis outline

- **Chapter 2 - Background:** Provides the required background information to understand the work described in this thesis and to place it in a frame of reference. It explains the fundamentals of MWI and EM wave interactions with biological tissues. The current state of the art of biomedical MWI is discussed, focusing specifically on hardware and system design. The opportunity of RoC for biomedical imaging is further elaborated.
- **Chapter 3 - System:** Discusses the anatomy of different imaging applications, assesses their feasibility for MWI and derives system requirements while considering MWI design aspects. Imaging hardware, as developed by collaborators at UiO and Novelda is characterised through experimentation. The System chapter can be considered a report on the current imaging system, which has evolved substantially over the past five years. Hardware was developed in iterations, with feedback provided through experimentation, and system requirements were refined through progressing insight into the feasibility of different applications.

- **Chapter 4 - Sensing:** Focuses on 1D sensing of the cardiovascular system using a single RoC device. Signal processing challenges associated with in-body radar sensing are discussed and experimental data from multiple subjects are presented. Work on 1D sensing was initiated before a multi-module imaging device was constructed, with the aim of exploring the challenges of in-body RoC measurements. The Sensing chapter is based on two different experiments on human participants, which were performed over the past three years.
- **Chapter 5 - Imaging:** Focuses on 2D imaging of the cardiovascular system using multiple synchronised RoC devices. Imaging algorithms and parameters are investigated through simulations and experimental data from a phantom and a single subject are presented. The Imaging chapter is based on simulation work which was performed in the past two years in preparation for experimental imaging datasets. The current imaging system was finalised at the start of 2019, and all imaging experiments were performed in a single data collection sprint. The fundamentals of beamforming, which are included at the start of Chapter 5, originate from a three-month internship at research collaborator Novelda, at the start of 2017.
- **Chapter 6 - Conclusions:** Conclusions and final thoughts on the use of UWB RoC for biomedical sensing and imaging are given, along with an overview of original contributions.

Parts of this research have been published in a journal and conference proceedings. An overview of publications is given in Appendix A.

Chapter 2

Background

2.1 Microwave imaging for biomedical applications

2.1.1 Principle of operation

MWI in the human body relies on the contrast in intrinsic impedance between biological tissues. As EM waves at microwave frequencies (300 MHz to 300 GHz, although various definitions exist) impinge on boundaries between tissues, scattering, reflection and transmission of EM energy will occur. The occurrence of scattering depends on the size and geometry of the anatomical structures relative to wavelength. The rate of reflection and transmission depends on the difference in intrinsic impedance, which is determined by the dielectric properties of the tissue. Biological tissues are dispersive: Their dielectric properties vary with frequency. Substantial attenuation of EM energy occurs when propagating through biological tissue, and signal attenuation increases with frequency. Unlike EM energy at optical frequencies, sufficient penetration depth can be achieved when using the lower end of the microwave spectrum. Importantly, EM radiation at microwave frequencies is non-ionising, and the only known adverse effect to health is heating of tissue at high power levels. These characteristics make microwave technology an interesting candidate for medical imaging applications. Image formation relies on radiation of microwaves into an anatomical structure using a transmitter (Tx) antenna, and capturing transmitted, reflected and scattered energy at various receiver (Rx) antenna locations surrounding the anatomy. Typically an array of antennas is used to acquire a spatially diverse dataset, but a single antenna may also be used for sequential scanning at multiple locations.

An example of an existing MWI system (in this case applied to breast cancer imaging) is shown in Fig. 2.1a. A circular array of antennas surrounds the breast tissue. A single Tx antenna transmits an EM waveform into the tissue, which is scattered by tumour tissue and subsequently

received by Rx antennas. Fig. 2.1b illustrates the type of image that can be obtained, in this case a cross-section of the breast with a contrast detected at the location of the tumour.

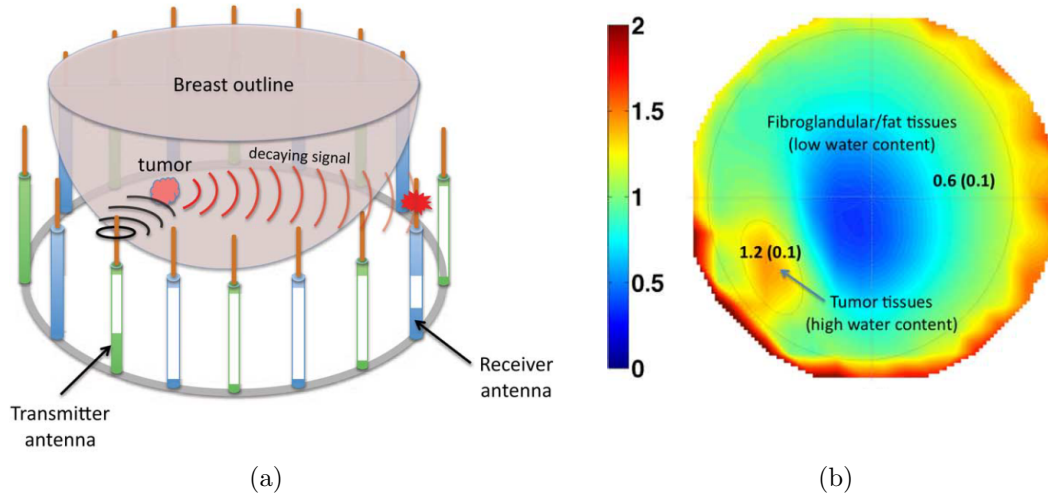


Figure 2.1: An example of a microwave imaging system, applied to breast cancer imaging (a). A circular array of antennas surrounds the breast tissue. A single Tx antenna transmits an EM waveform into the tissue, which is scattered by tumour tissue and subsequently received by Rx antennas. The obtained image shows a cross-section of the breast with a contrast detected at the location of the tumour (b). The depicted system uses a microwave tomography approach, but similar hardware could be used for UWB radar imaging. Figures from [4].

2.1.2 Microwave tomography and ultra-wideband radar imaging

Two approaches have dominated the field of MWI: Microwave tomography and UWB radar. Both approaches rely on image formation from transmitted and reflected EM energy at microwave frequencies, and the dielectric contrast between tissues. Microwave tomography attempts to reconstruct the dielectric profile of the anatomy of interest by solving the inverse scattering problem from measured scattering parameters. Often, a narrow frequency band or a set of discrete frequencies is used. UWB radar imaging on the other hand, utilises wideband radio frequency (RF) pulses to detect and determine the range of significant scatterers. By combining spatially diverse UWB radar data, the location of a scatterer can be determined in 2D or three-dimensional (3D) space. Although different definitions of UWB have been used, UWB radar often refers to short range, low power radar systems that operate in the band of 3.1 to 10.6 GHz, which was made available for unlicensed use by the Federal Communications Commission (FCC) in 2002. Microwave tomography and UWB radar have both successfully been applied in biomedical imaging applications. A disadvantage of microwave tomography is the computational complexity: To obtain the dielectric profile of the imaging object, an ill-posed inverse problem must be solved through an iterative approach. The complexity and high

computation time limit the use of high bandwidth. Radar imaging algorithms on the other hand are simple and robust, and the use of a wide frequency band provides better resistance against multipath interference.

2.1.3 The role of microwave imaging in medicine

MWI has been investigated intensively for biomedical applications, particularly in the past two decades. Although non-contact sensing of the human body using microwave technology has been described as early as in the 1970s, the first systematic description of the concept of probing the human body with microwave signals and imaging based on dielectric properties appeared in 1985 [5]. Technological advances in UWB radar in the late twentieth century caused an increase in interest of using UWB radar for biomedical sensing. Notably, a patent was awarded to Thomas McEwan in 1996 for the invention of the micropower impulse radar: A low power, simple and compact UWB radar, to be used for monitoring subcutaneous cardiac motion [6].

The principle of 1D sensing was quickly extended to 2D and 3D imaging, and although the first decade of the 21st century was marked mostly by simulation studies, the past decade has given rise to various MWI device prototypes. An overview of the state of the art in medical MWI is given in Section 2.3.

In comparison to existing medical imaging modalities, MWI possesses a unique set of advantages. Current techniques (such as X-ray CT, MRI, PET, SPECT), are accompanied by either high cost, poor temporal or spatial resolution, or require the injection of radioactive isotopes in the blood stream. MWI is affordable, completely non-invasive, and reasonable spatial resolution can be achieved. US shares a number of characteristics with UWB radar: The imaging principle is similar, and unlike for example MRI or CT, US is relatively inexpensive and portable. However, US requires an experienced operator and the cost is still prohibitive for widespread use in primary care. More importantly, the imaging contrast in US and MWI relies on different physical tissue properties, making the one modality potentially better suited for a specific application than the other. As an example, US does not penetrate well through bone tissue, whereas microwaves do. Clinical trials are currently ongoing for the use of MWI in stroke detection and monitoring [7]. The most commonly researched application of MWI is breast cancer imaging. When compared to X-ray based mammography, the increased level of patient comfort and absence of ionising radiation make MWI highly suitable, especially for preventive screening. Several clinical trials are currently under way [8]

2.2 Microwave imaging fundamentals

The following section covers some fundamentals of microwave imaging in the human body. Specifically EM wave propagation and interactions with biological tissue will be described, as they are essential to making correct assumptions in MWI and for informing system design, signal processing, and imaging algorithm parameters.

2.2.1 Electromagnetic wave propagation in biological tissue

EM wave propagation can be described using Maxwell's equations in lossless media such as air, but they are equally applicable in lossy media such as biological tissues. Wave propagation in a lossy dielectric is the most general form of wave propagation and a starting point before any simplifying assumptions are to be made. Section based on [9–12].

Considering a lossy dielectric medium that is charge free ($\rho_v = 0$), Maxwell's equations become:

$$\nabla \cdot \mathbf{E} = 0 \quad (2.1)$$

$$\nabla \cdot \mathbf{H} = 0 \quad (2.2)$$

$$\nabla \times \mathbf{E} = -j\omega\mu\mathbf{H} \quad (2.3)$$

$$\nabla \times \mathbf{H} = j\omega\varepsilon^*\mathbf{E} \quad (2.4)$$

where $\mathbf{E}$, $\mathbf{H}$ and ∇ are the electric field strength (V/m), magnetic field strength (A/m), and nabla operator, respectively, ω is the EM wave angular frequency (rad/s), μ is the permeability (H/m), and ε^* is the complex permittivity of the medium (F/m), further defined as:

$$\varepsilon^* = \varepsilon - j\frac{\sigma}{\omega} \quad (2.5)$$

where σ is the electrical conductivity of the medium (S/m). The real and imaginary parts of complex permittivity are often noted as $\varepsilon^* = \varepsilon' - j\varepsilon''$, in which case $\varepsilon' = \varepsilon$ and $\varepsilon'' = \sigma/\omega$. When discussing dielectric properties of biological tissues, often relative permittivity ε_r and relative permeability μ_r are used (both dimensionless). It should be noted that in all of the following equations ε and μ refer to absolute permittivity and permeability, which are obtained by multiplying relative values with the permittivity and permeability of vacuum: $\varepsilon = \varepsilon_r\varepsilon_0$ and $\mu = \mu_r\mu_0$, where $\varepsilon_0 \approx 8.854 \times 10^{-12} F/m$ and $\mu_0 \approx 4\pi \times 10^{-7} H/m$.

How lossy a medium is, and thus the amount of power losses, can now be described by the loss

tangent $\tan \theta$:

$$\tan \theta = \frac{\varepsilon''}{\varepsilon'} = \frac{\sigma}{\omega \varepsilon} \quad (2.6)$$

where θ is the loss angle. For materials with large $\tan \theta$ ($\sigma \gg \omega \varepsilon$), it can be said that the medium is a good conductor. If $\tan \theta$ is small ($\sigma \ll \omega \varepsilon$), one speaks of a good insulator, low-loss medium, or imperfect dielectric. The conductive properties of a medium are thus frequency dependent. Often a factor 100 is meant when stating ‘much greater/smaller than’, but usually biological tissues in the microwave frequency band are considered low-loss, even though $\tan \theta$ may be closer to 0.1 than 0.01.

The tissue dielectric properties allow us to describe how an EM wave behaves in any homogeneous medium. From the Maxwell’s equations above, an expression for the electric field intensity E as a function of propagation depth z (in m) can be derived:

$$E(z) = E_0 e^{-\gamma z} \quad (2.7)$$

where E_0 is the electric field intensity at $z = 0$, and γ is the propagation constant:

$$\gamma = \sqrt{j\omega\mu(\sigma + j\omega\varepsilon)} \quad (2.8)$$

Because γ is complex, we may write $\gamma = \alpha + j\beta$. The separate constants α and β describe the attenuation and phase propagation of the EM wave, respectively, and may now be obtained as:

$$\alpha = \omega \sqrt{\frac{\mu\varepsilon'}{2} \sqrt{1 + \left(\frac{\varepsilon''}{\varepsilon'}\right)^2} - 1} \quad (2.9)$$

$$\beta = \omega \sqrt{\frac{\mu\varepsilon'}{2} \sqrt{1 + \left(\frac{\varepsilon''}{\varepsilon'}\right)^2} + 1} \quad (2.10)$$

Attenuation constant α is expressed in Np/m , and phase constant β is expressed in rad/m . An attenuation of 1 Np corresponds to 8.69 dB .

The electric field strength for a uniform plane wave, propagating in direction z , through a homogeneous tissue with known dielectric parameters can now be determined using:

$$E(z) = E_0 e^{-\alpha z} e^{-j\beta z} \quad (2.11)$$

Phase velocity ν (in m/s) is given by the following equation:

$$\nu = \omega / \beta \quad (2.12)$$

and wavelength λ (in m) is given by:

$$\lambda = \frac{2\pi}{\beta} = \frac{\nu}{f} \quad (2.13)$$

where f denotes frequency (in Hz).

A final medium property of importance is a medium's intrinsic impedance η_c , which is a complex value with unit Ω :

$$\eta_c = \sqrt{\frac{\mu}{\epsilon'}} \left(1 - j \frac{\epsilon''}{\epsilon'} \right)^{-1/2} \quad (2.14)$$

From equations (2.5), (2.10) and (2.12) it follows that phase velocity ν is dependent on a material's conductivity σ and thus angular frequency ω . For many imaging applications this is inconvenient, hence it is often assumed that biological materials in the microwave frequency band are low-loss materials ($\epsilon''/\epsilon' \ll 1$). This offers the simplification that phase velocity can be assumed constant. For a low-loss medium, equations (2.9), (2.10) and (2.12) to (2.14) reduce to the following simplified form:

$$\alpha = \frac{\sigma}{2} \sqrt{\frac{\mu}{\epsilon}} \quad (2.15)$$

$$\beta = \omega \sqrt{\mu \epsilon} \quad (2.16)$$

$$\nu = \frac{1}{\sqrt{\mu \epsilon}} \quad (2.17)$$

$$\lambda = \frac{\nu}{f} \quad (2.18)$$

$$\eta = \sqrt{\frac{\mu}{\epsilon}} \quad (2.19)$$

As will be seen in section 3.5, the low-loss assumption results in a negligible error, particularly when compared to expected error due to uncertainty in exact tissue dielectric properties. For simplification purposes, biological tissues will therefore be considered as low-loss media throughout this thesis.

2.2.2 Electromagnetic wave interactions with biological tissue

Interactions with biological tissue include attenuation, which is described by equations (2.11) and (2.15). Other interactions are scattering, reflection and transmission, occurring at boundaries between tissues. Simple frequency-independent reflection of EM energy occurs in the optical region when wavelength λ is small relative to the object it impinges on. When considering a sphere (or a circular object in 2D such as an artery) with circumference roughly between 1 and 10 times wavelength λ , scattering in the Mie region occurs [13]. For this scenario, the radar cross section, which describes the object's ability to reflect EM waves in the direction of the receiver, depends on wavelength λ . Estimating exact radar cross section for anatomical objects such as an artery is challenging as it depends on many factors and exact dimensions are generally not known. The centre frequency of the RoC used in Chapter 4 is 3.8 GHz (details in Section 3.4.3). At that frequency, effective wavelength λ is approximately 1.1 cm, depending on the tissue's dielectric properties. Anatomical objects like arteries and bones will therefore fall within the Mie region: Reflected energy will be reduced and scattering in all directions is expected.

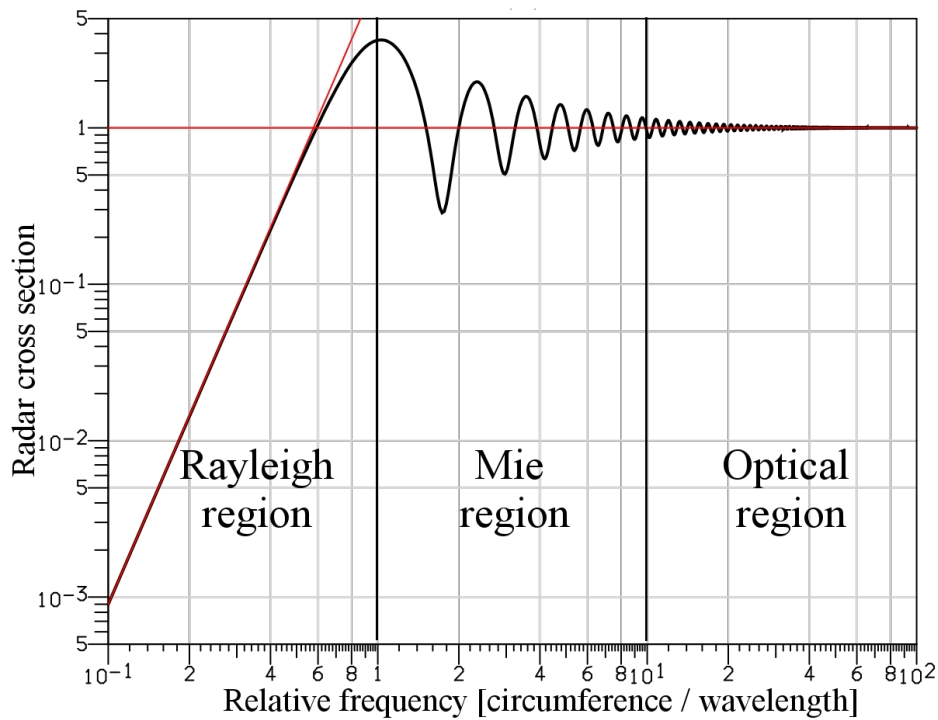


Figure 2.2: Radar cross section of a sphere, as a function of relative frequency (circumference / wavelength), illustrating the Mie scattering and optical region. At a centre frequency of 3.8 GHz, anatomical objects such as arteries and bones will typically fall in the Mie region. Figure adapted from [14].

Scattering effects may be ignored to calculate EM wave reflection and transmission. Although

not accurate in most biomedical scenarios, the reflection and transmission coefficient provide a clear indicator of contrast between tissues and may be used for simplified estimates of received signal strength. The rate of reflection depends on the dielectric contrast between two tissues, or the mismatch in impedance η (equation 2.19). When assuming normal incidence of an EM wave propagating through medium 1 onto the boundary with medium 2, the expressions for intensity of both the transmitted and reflected wave can be derived from the equations for transmission and reflection coefficient [15]. Transmitted intensity T is expressed as a ratio of transmitted and incident electric field strength (E_t and E_i), and reflected energy R as a ratio of reflected and incident electric field intensity (E_r and E_i):

$$T = \frac{E_t}{E_i} = \frac{2\eta_2}{\eta_2 + \eta_1} \quad (2.20)$$

$$R = \frac{E_r}{E_i} = \frac{\eta_2 - \eta_1}{\eta_2 + \eta_1} \quad (2.21)$$

where η is the intrinsic impedance of either medium 1 or 2 and $R + T$ equalling 1.

2.2.3 Dielectric properties of biological tissue

The complex permittivity, which is a measure for how a medium is affected by an EM field, is a function of frequency: materials cannot polarise instantaneously in response to an applied field. When polarisation time (and relaxation time after the EM field is removed) is in the order of magnitude of EM wave period, a material's polarisation lags behind the applied EM field, and it can intuitively be understood that a material's response to EM excitation is frequency dependent. This property is called dispersiveness. In general, permittivity decreases with increased frequencies, whereas conductivity increases. At least three major dispersion regions can be identified: alpha, beta and gamma dispersion. The main dispersion effect in biological tissues in the microwave range is gamma dispersion, which is due to polarisation of water molecules (dipoles) [11].

Dielectric properties for various biological tissues have been measured [16], and can be approximated across frequencies using a multi-pole Cole-Cole model [17]. Often when dealing with biological materials, relative values of permittivity are used. Complex relative permittivity ε_r^* is expressed as:

$$\varepsilon_r^*(\omega) = \varepsilon_\infty + \sum_{n=1}^N \frac{\Delta\varepsilon_n}{1 + (j\omega\tau_n)^{1-\alpha_n}} + \frac{\sigma_i}{j\omega\varepsilon_0}$$

Where N is the number of poles of the model. Model parameters ε_∞ (relative permittivity at high frequency limit), $\Delta\varepsilon = \varepsilon_s - \varepsilon_\infty$ (or, the magnitude of the dispersion, with ε_s the static

low frequency relative permittivity), τ (relaxation time constant; s), α (dispersion broadening parameter), σ_i (static ionic conductivity), and ε_0 (the permittivity of vacuum) for various tissues have first been published in [17], but parameters based on an extended dataset are available from [18].

Relative permeability μ_r of biological tissue is usually assumed to be 1, as will be assumed throughout this thesis.

2.3 Microwave imaging state of the art

Below follows an overview of the state of the art of MWI technology, organised by clinical application. Specifically, efforts by different research groups or companies that have developed a working MWI prototype are described. While focusing on systems based on UWB radar, a few notable microwave tomography systems have been included. Design aspects of a subset of prominent or interesting efforts are summarised in Table 2.4. Imaging algorithms as employed by different systems are not covered in this section: They will be discussed in Section 5.2.2.

2.3.1 Breast cancer imaging

Breast cancer detection using MWI technology has been researched intensively, as there exists a dielectric contrast between malignant tumour and healthy breast tissue, and the breast geometry lends itself well to microwave imaging. Typically, the device consists of a table on which the patient is asked to lie in prone position, with the breast inserted into an imaging cylinder. The cylinder either contains an antenna array or a moveable arm with a single antenna for mechanical scanning of the breast. Often, the cylinder is filled with a coupling liquid for efficient radiation into the tissue, and to absorb backscattered radiation. The majority of current prototypes use UWB radar technology. Scanning times of these systems are typically in the order of minutes.

Dartmouth College Researchers from Dartmouth college (USA) have developed a microwave tomography system for the detection of breast cancer [4]. The device consists of a circular array of 16 omni-directional monopole antennas in a coupling liquid. Antennas are adjustable in height, and a 3D image is created of the breast by scanning the array through a number of coronal slices. Data are acquired at multiple frequencies. The authors report 1.3 GHz as an optimal frequency, although this was found dependent on the patient's breast composition. Image acquisition takes 2 minutes and the image reconstruction time is under 20

minutes. Over 400 patients have been imaged using the system, and a study on chemotherapy monitoring shows changes in conductivity properties that correlate with overall treatment response for locally advanced breast cancer [19].

University of Bristol / Micrima Ltd. Based on initial research by the University of Bristol, the ‘MARIA’ (Multistatic Array processing for Radiowave Image Acquisition) breast cancer imaging device has been developed by Micrima Ltd. (Bristol, United Kingdom). Multistatic data collection is performed: Spatially diverse Tx and Rx antennas are used, typically to collect data from different combinations of Tx-Rx channels. Data collection is performed in the frequency domain through a VNA with switch array and a breast-shaped dome, populated with antennas. An earlier version of the device was tested in phantom studies and found to be able to detect tumours as small as 4mm diameter [20]. The first clinical trials proved to be unsuccessful [21]. The latest version of the device (MARIA 4, details in [22]) contains a higher number of antennas ($N = 60$), and clinical trials are currently ongoing. An interim presentation from 2016 mentions 287 patients enrolled to date with a 77 % sensitivity in breast cancer detection [23]. Fig. 2.3 illustrates the MARIA device.

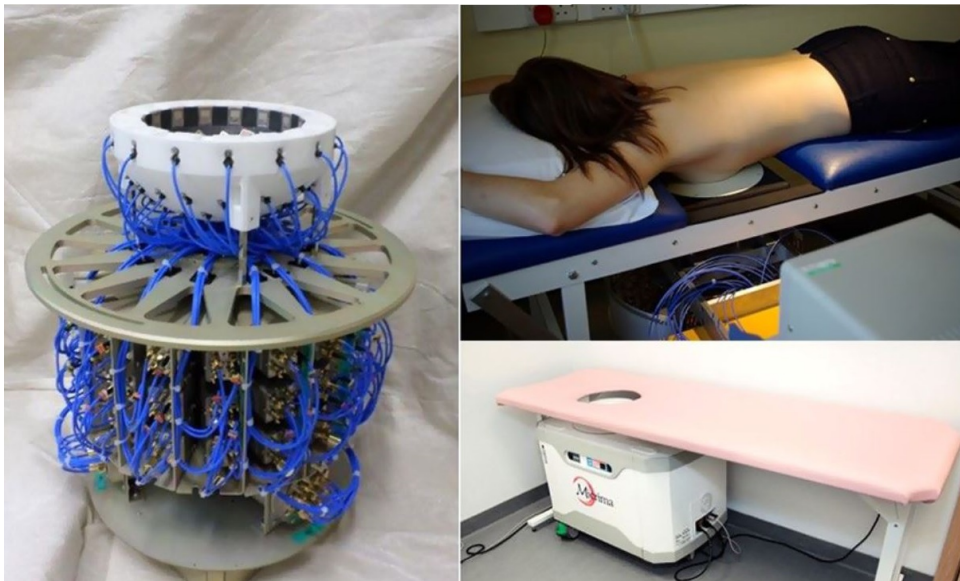


Figure 2.3: The ‘MARIA’ breast cancer imaging prototype by Micrima, based on UWB radar technology. The patient is asked to lie in prone position with the breast inserted in an imaging dome. The dome contains a coupling liquid and 60 antennas. A vector network analyser with switch array is used to collect multistatic radar data in the frequency domain. Figure from [22].

Universities of Calgary / Madison-Wisconsin A joint effort by the Universities of Calgary (Canada) and Madison-Wisconsin (USA) has resulted in the TSAR (Tissue Sensing Adaptive Radar) device. This breast cancer imaging device uses monostatic radar: The Tx and Rx

antennas are co-located, typically data are collected from a single Tx-Rx pair. A synthetic array is constructed by mechanically scanning a transceiver around the breast. Radar measurements are taken in the frequency domain using a VNA, and a single scan composed of 200 scanning locations is scanned in 30 minutes [24]. Eight patients have been scanned in a first patient trial, and although not all lesions could be detected, a number of tumours were correctly identified and localised [25]. A second device has been developed which uses transmissive measurements at a fixed path length with the sole purpose of assessing breast dielectric properties, to improve imaging assumptions and assist in accurate image formation. Clinical trials are currently ongoing [26].

McGill University A breast cancer imaging prototype using time domain pulsed UWB radar has been developed by a group at McGill University (Canada). The use of time domain radar as opposed to frequency domain carries the potential advantages of low cost equipment and shorter scanning times. A first clinical prototype was a table-based device and consisted of a hemispheric dome, housing 16 antennas. Multistatic data were acquired through a switch array. Tumours could correctly be identified in breast phantoms [27, 28]. Measurement repeatability between scans was tested on 13 healthy volunteers and was found to be high [29]. Challenges in accommodating to different breast sizes, in achieving optimal antenna orientation relative to skin surface, and high absorption due to coupling liquids, led to the development of a second generation device. A more portable and compact prototype was developed: A rigid array composed of 16 flexible antennas designed for direct skin contact were built into a bra [30]. The system was tested on 13 healthy volunteers for repeatability of measurements and comfort, with encouraging results. Predictions of breast permittivity were found to be comparable to predictions obtained using the table-based system [31].

Hiroshima University A handheld, portable breast cancer imaging system based on time domain radar has been developed by a group at Hiroshima University (Japan). Eight transmitting and eight receiving antennas are placed in a dome and multistatic data are acquired in the time domain through a switching matrix. A coupling shell is used to interface with the breast tissue, and a thin layer of coupling liquid between the coupling shell and the antenna dome allow for rotation of the array, thus synthetically increasing array density. Compactness was achieved by using a number of CMOS components: A mono-pulse generator and high speed switch system. Interleaved sampling is used to obtain the required sampling rate. Scanning time is still high at 14 minutes [32]. The prototype was tested on five patients. Obtained images were reported to be consistent with the clinical history of patients [33]. In addition, good results were acquired on imaging of excised breast tissue [34]. Fig. 2.4 illustrates the Hiroshima device and its intended use.

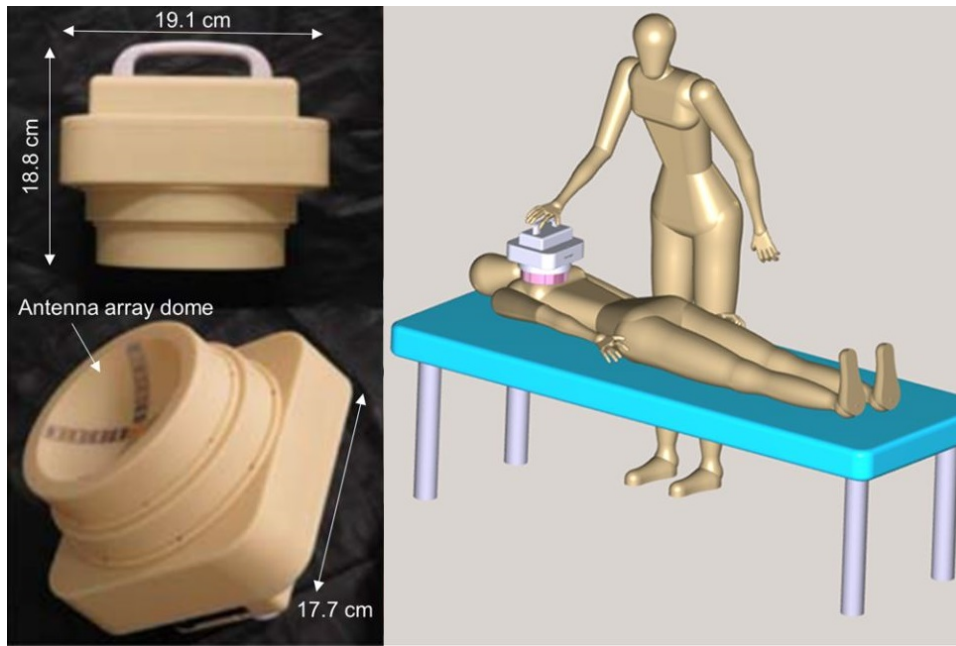


Figure 2.4: A handheld breast cancer microwave imaging device by Hiroshima University. By using a CMOS pulse generator and high speed switch array, multistatic radar data were collected in the time domain in a compact device. Figure from [33].

Shizuoka University A table-based breast cancer imaging device was developed, based on multistatic radar, by a group at Shizuoka University (Japan). The system was found to have insufficient sensitivity for differentiating between tumour tissue and other lesions [35]. In response to this finding, a combination of microwave tomography and radar imaging was implemented, in which the radar measurement provides a priori information to speed up the tomographic image formation [36]. Data are obtained in the frequency domain using a VNA, in combination with a real aperture and switch array. An aspiration system is used to suck the breast into a dome, thereby guaranteeing antenna-to-skin contact and avoiding the need to estimate breast shape. Domes with various sizes (and various numbers of antennas) were constructed to accommodate for different breast sizes.

Microwave Vision Group French company Microwave Vision Group (MVG; Paris, France) has developed a microwave breast cancer imaging device based on UWB radar, which has recently been deployed in Galway University Hospital for first in-patient testing. The device consists of a table with an imaging cylinder. The breast is submerged in a coupling liquid and scanned through a circular array of 18 antennas. Multistatic data collection is performed in the frequency domain, across the frequencies of 1 – 4 GHz. The circular array is scanned mechanically across the 3rd dimension. A secondary optical scanning system is used to estimate breast boundary and volume [37]. No clinical data have been published yet, but state-of-the-art signal processing techniques are being applied [38] which makes this an interesting effort.

Southern University of Science and Technology, Shenzhen A table-based radar imaging device for breast cancer has been developed by a research group in Shenzhen (China). The device collects monostatic data through a mechanically scanned antenna submerged in coupling liquid, with a VNA as data acquisition hardware. Little is published on this device, but authors mention it is to be tested on 100 patients in a first clinical trial [39].

2.3.2 Stroke and brain imaging

MWI of the brain is challenging due to low permittivity of the skull compared to surrounding tissues, which causes the skull to act as a dielectric shield. Unlike breast cancer imaging systems, brain imaging devices do not utilise coupling liquids as the head cannot simply be submerged. Efforts have concentrated on stroke detection. Being able to discriminate between haemorrhagic stroke (bleeding) and ischaemic stroke (blocked blood vessel) at an early stage (through pre-hospital assessment) is of high clinical importance.

Chalmers University of Technology A group at Chalmers University of Technology (Sweden) has developed a stroke detection helmet, using microwave technology. Transmission and reflection coefficients are measured through 12 patch antennas at a frequency band of 857 – 1493 MHz, which was found to have the highest transmission. Data are collected in the frequency domain using a VNA and switch matrix. Scanning time is not mentioned. Interestingly, no imaging is applied. Instead, a machine learning approach is taken to discriminate between the two types of stroke. Two small scale clinical studies have been carried out with the aim of discriminating between haemorrhagic stroke and ischaemic stroke. From 26 ischaemic stroke patients, 21 were successfully separated from the patient group. Initial data are interesting, but using machine learning at such an early stage may of course mask unknown factors. In addition, it will prove to be problematic while navigating the regulatory space as well as for obtaining clinical acceptance. A spin-off company named Medfield Diagnostics currently aims to commercialise the ‘Strokefinder’ product.

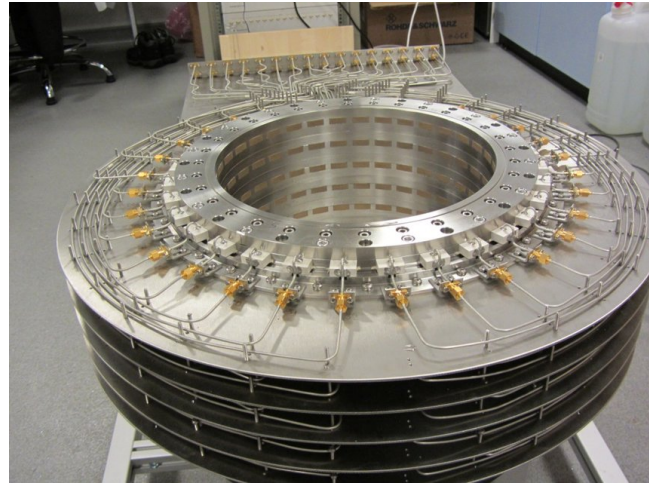
University of Queensland A stroke detection prototype has been developed by a research group at the University of Queensland (Australia). A monostatic UWB radar imaging system is operated in the frequency domain. The head is scanned at 100 positions using a single antenna, data are collected over a bandwidth of 1.1 – 3.2 GHz with a low-cost microwave transceiver. The device is designed for future use in an ambulance, thus a relatively light weight transceiver system was used. The system has not been tested on a clinical population, but using a 3D printed, realistic human head model, the group has demonstrated that a 1 cm³ blood clod can be detected [40]. Initial testing on two healthy human volunteers resulted in no

significant scatterers found apart from the skull, demonstrating the absence of false positives on haemorrhage detection [41].

EMTensor Vienna-based company EMTensor (Austria) has developed an EM tomography system for brain imaging, specifically stroke detection. The intended use is continuous monitoring of stroke and stroke imaging in an ambulance, as well as functional neuroimaging. The latest device iteration is the G4 (shown in Fig. 2.5), which is currently under clinical investigation. Details on a previous version, the G2, have been published [42]: 177 Antennas were mounted in a dome. A VNA and switch matrix were used to collect multistatic transmission coefficients. A single Tx-Rx channel was measured in 2 ms, which resulted in a scanning duration of 1.5 s for a subset of antennas lying on the same 2D plane. A frequency of 1 GHz is used. 2D experimental data collection of a human head phantom demonstrated that a large haemorrhagic stroke mimicking target could successfully be identified. However, in these phantom tests, the antenna dome was filled with a coupling liquid which was identical to the liquid inside the head phantom. Performance of the EMTensor device without coupling liquid in a realistic scenario is unknown.



(a)



(b)

Figure 2.5: Microwave tomography device for stroke imaging, developed by Austrian company EMTensor. A 3D render of the latest device iteration (G4) is shown here (a). The internals of one of the device earlier iterations, housing all antennas in a cylindrical structure, are shown in (b). Figures from [7] and [43].

2.3.3 Cardiovascular imaging

Radar technology has been applied to cardiovascular sensing (an overview of the state of the art is given in Section 4.2.1). Monitoring of cardiovascular health is of high clinical significance,

and affordable technologies to measure mechanical motion alongside electrocardiography (ECG) would be highly desirable. Research in this field has been mostly limited to 1D sensing: To perform imaging of heart motion in 2D or 3D a high frame rate is required, which is not easily obtained when using a VNA and a switch matrix to scan through different Tx-Rx channels.

University of Oslo / FFI University of Oslo (Norway) in collaboration with the Norwegian Defence Research Establishment (FFI) appears to have been the only group so far to report high speed dynamic MWI: 2D multistatic radar data are obtained at a frame rate high enough to track heart motion [44]. A linear array of eight Tx and eight Rx antennas is used in combination with a switch matrix. The antenna array was positioned on radar absorbent material, on top of a wooden table. Data were acquired in the frequency domain using a VNA. An impressive frame rate of 14.6 Hz was achieved, by scanning across a range of 0.75 – 2.27 GHz, using 21 frequency points. An even higher frame rate could be achieved when pruning redundant channels, but imaging performance dropped substantially. Results (shown in Fig. 2.6) illustrate that a moving scatterer was detected at a location which corresponds to that of the heart. Time series data at the hypothesised heart location indeed show periodic movement that corresponds to the heart cycle (verified using ECG). This study serves as a first proof of concept of dynamic UWB radar imaging. The use of a high cost VNA allowed for a high frame rate, but it is expected that sensitivity of the system dropped significantly as a result of the reduced scanning times. The low number (21) of frequency points directly affects the down range sampling in the time domain and will cause loss of imaging sensitivity and down range resolution.

2.4 Radar-on-chip technology for biomedical microwave imaging

A fully coherent UWB single-chip radar system has been developed by Novelda (Kviteseid, Norway): The XeThru X2 RoC. The X2 chip is mostly used in combination with aerial antennas for presence detection and for non-contact respiration monitoring. However, there has been an interest from Novelda to explore the opportunity of using the X2 for in-body applications, such as heart monitoring. A module has been developed using the X2, but with reduced centre frequency and added gain for improved tissue penetration. A collaboration between Novelda, UiO, and ICL was established to investigate in-body MWI using UWB RoC technology. A number of alternative companies produce UWB RoC transceiver systems. The majority of these are developed for the automotive industry, and use centre frequencies around 76 – 79 GHz (IMEC, Texas Instruments, NXP), which is too high for in-body applications due to excessive signal attenuation. A few RoCs are available that operate within the UWB emission mask as

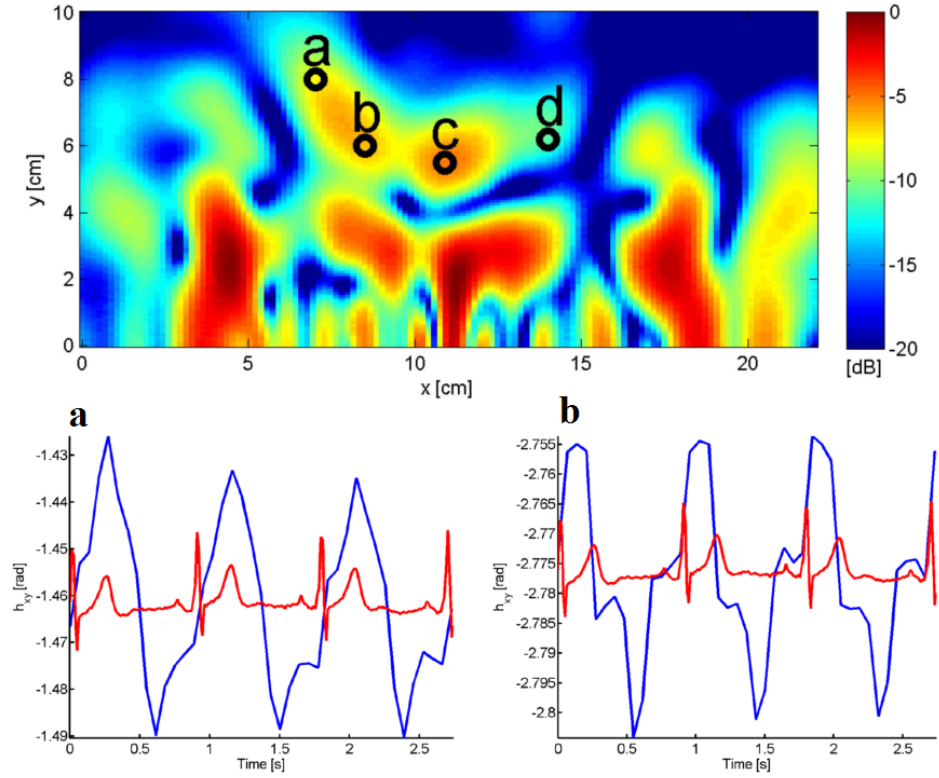


Figure 2.6: Results of dynamic radar imaging of the heart, performed using a VNA. A transverse cross section image of the chest is shown, with a moving scatterer at a location which corresponds to that of the heart. Time series data at the hypothesised heart location show periodic movement that corresponds to the heart cycle (verified using ECG). Figure from [44].

Table 2.1: System design aspects of a number of prominent or unique UWB radar imaging prototypes.

Group	Uni. Bristol/ Micrima	Uni. Calgary/ Wisconsin- Madison	Uni. Hiroshima	McGill Uni.	Uni. Queensland	Shizuoka Uni.	Uni. Oslo/ FFI Norway
Application	Breast cancer	Breast cancer	Breast cancer	Breast cancer	Stroke	Breast cancer	Cardiac imaging
Mono/ Multistatic	Multistatic	Monostatic	Multistatic	Multistatic	Monostatic	Multistatic	Multistatic
Aperture	Real (60 Ant)	Synthetic (200 Pos)	Real (8 Tx + 8 Rx) + synthetic (120 Rot)	Real (16 Ant)	Synthetic (100 Pos)	Real (6 – 40 Ant, breast size dependent)	Real (8 Ant)
Scanning	Switch array	Moving antenna	Switch array + Array rotation	Switch array	Moving antenna	Switch array	Switch array
Interface	Coupling shell + Coupling liquid	Coupling liquid	Coupling shell + Coupling liquid	Direct	Aerial	Direct	Direct
Time/Freq domain	Freq (VNA)	Freq (VNA)	Time (CMOS GMP)	Time (PG + Oscilloscope)	Freq (MW transceiver)	Freq (VNA)	Freq (VNA)
Freq range	3 – 8 GHz	50 MHz – 15 GHz	6 GHz BW at 6.7 GHz Fc	2 – 4 GHz	1.1 – 3.2 GHz	4 – 9 GHz	0.75 – 2.27 GHz
Scanning time	<5 min	30 min	14 min	6 min	1 min	5 – 200 s (breast size dependent)	68 ms
Feature	-	Optical system for breast contour	Handheld, portable	Wearable bra array	-	Breast fixed to dome by aspirator	Time-lapse imaging

Abbreviations: Ant: antenna, VNA: vector network analyser, Pos: position, Tx: transmitter, Rx: receiver, Rot: rotation angle, GMP: Gaussian monocycle pulse generator, BW: bandwidth, Fc: centre frequency, PG: pulse generator, MW: microwave.

Table 2.2: General properties of the envisioned dynamic microwave imaging system based on radar-on-chip technology.

Application	Dynamic imaging (cardiovascular, functional neuroimaging)
Mono/Multistatic	Multistatic
Aperture	Real
Scanning	Sequential Tx, Simultaneous Rx
Interface	Direct contact
Time/Freq domain	Time domain (radar-on-chip)
Freq range	3 – 4 GHz (-3 dB)
Scanning time	<0.1 s
Feature	Portable/wearable and modular

defined by the FCC (Vayyar, UMAIN), but come at a substantially higher price than the RoC of research collaborator Novelda. In addition, there is no indication that competing RoCs offer better sensitivity. For this project therefore, only Novelda RoC technology was considered.

RoC offers the advantage of extremely small form factor, low power, very low cost, and high imaging speed. These properties would be particularly interesting for developing a low-cost and portable (or even wearable) imaging modality, possibly for ambulatory diagnostics, preventive screening, or remote patient monitoring. When considering the design of most existing MWI devices, it is likely that a system based on portable RoC technology will not achieve the same level of signal leakage reduction, clutter rejection and calibration, as a fully static system embedded in an imaging table. However, the property of high imaging speed allows for imaging of time-varying signals while rejecting static clutter. Such a dynamic imaging system, provided sufficient sensitivity, could enable monitoring of cardiovascular dynamics (both the heart and arteries), functional neuroimaging, or even metabolic imaging. General properties of the envisioned RoC MWI system are given in Table 2.4.

Dynamic imaging using microwave technology has rarely been investigated. In all reported cases of static MWI, the imaging aperture was achieved by either mechanical scanning of a single antenna, or by employing a switch matrix to sequentially scan from different antenna pairs. Apart from one exception [44], scanning duration is high and systems are not suitable for imaging of dynamic anatomical structures, such as the cardiovascular system or functional neuroimaging. RoC technology has the advantage of simultaneous receiving of all radar systems, thus shorter scanning times are possible without loss of sensitivity.

Chapter 3

System design and analysis

3.1 Introduction

The recent innovation of UWB RoC technology provides the opportunity of developing a small, low power, and affordable MWI device. Although many system specific challenges exist, high imaging speed could be achieved with RoC technology. By imaging time-varying structures against a static clutter scene, additional processing gain is obtained through high-pass filtering. A first dynamic MWI prototype based on RoC technology was designed and developed to investigate the challenges and opportunities of RoC imaging.

The current chapter considers different potential target applications for dynamic MWI. By studying their anatomy and physiology, as well as microwave propagation through biological tissues, a first set of system requirements is generated. Feasibility of target applications is assessed considering the current state of technology, and different design options are considered. The hardware that was developed specifically for this research project (by research collaborators UiO and Novelda) was characterised and its suitability for the goal of sensing and imaging of the cardiovascular system was assessed. Finally, different models of EM wave propagation in biological tissue are compared to measured data, to test their suitability for incorporation into MWI algorithms.

3.2 Target applications and system requirements

The use of low-power, portable RoC technology provides unique opportunities but also poses specific challenges and limitations to the system imaging properties. A RoC medical imaging system will likely require application specific design choices. As a starting point for defining

a set of generic system requirements, two different potential biomedical imaging and sensing target applications will be considered: Cardiovascular sensing and imaging, and functional neuroimaging. These applications have been selected due to the dynamic nature of the signals of interest: The cardiovascular system will show periodic motion at the heart rate (HR) frequency (approximately 1 Hz), whereas the neurovascular signal will be of a slower nature, with changes occurring over the course of several seconds. While attempting to draft rough system requirements, the current state of technology and available hardware was considered for feasibility purposes. For completion, it must be mentioned that dielectric spectroscopy for continuous blood glucose measurement was considered as a potential target application. However, initial measurements on the permittivity of sugar-in-water mixtures, along with findings of substantial phase noise in the available RoC system, led to the decision not to pursue this application. An exploration into RoC for dielectric spectroscopy has been included in Appendix C.

3.2.1 Target applications and system requirements: Methods

Details on the geometry, anatomy and physiology of relevant body parts are obtained from various sources. Dielectric properties are assumed based on published data. A simple 1D linear EM propagation model will be used to provide rough estimates of dielectric loss and expected dielectric contrast between imaging target and surrounding tissues. More details on methods are given below.

Dielectric properties

All tissue dielectric properties used in this chapter are obtained from [18]. Throughout this section, a centre frequency of 3.8 GHz or 3.45 GHz will be considered (the centre frequencies of the RoC used in Chapter 4 and Chapter 5, respectively). For brain tissue, the average tissue properties of white and grey matter were used. Similarly, bone was represented as the average of cortical and cancellous bone. Average infiltrated fat (vascularised) was used to represent fat tissue. Inflated lung was used to represent lung tissue.

An overview of dielectric properties of relevant tissues at 3.45 GHz is given in Table 3.1. As illustrated in Fig. 3.1, dielectric properties of different tissues show similar dispersive behaviour and no significant changes in contrast occur across the frequency band of interest. This implies that conclusions drawn based on estimates at 3.45 GHz carry some validity for other frequencies in the band of interest.

Table 3.1: Overview of dielectric properties of relevant tissues at 3.45 GHz

Tissue	Relative permittivity	Conductivity [S/m]
Blood	56.6	3.52
Brain	41.2	2.20
Bone	14.2	0.89
Fat	10.5	0.41
Muscle	51.5	2.52
Lung	19.9	1.12
Skin	37.0	2.00
Blood vessel wall	41.3	2.15

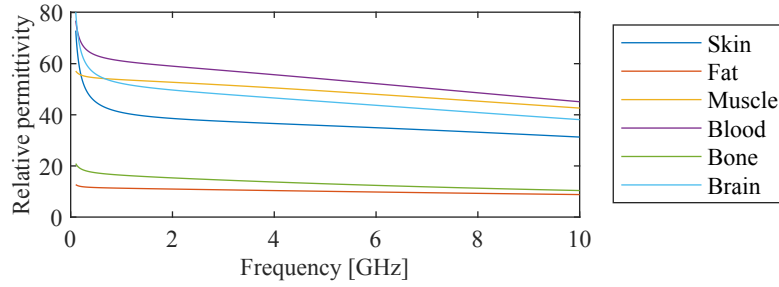


Figure 3.1: Relative permittivity for various tissues, across the frequency band of 0.1 to 10 GHz

1D Propagation model

A simple linear transmission model (proposed in [45], implementation described in [46]) can be used to illustrate EM wave propagation through layers of biological tissue. A 1D model is a coarse simplification and does not capture power density at target and receiver, the effective area of the reflector, antenna gain or scattering effects. However, it is a valuable tool for providing insight into EM wave propagation velocity and dielectric losses as a function of tissues, as well as the expected contrast between tissue boundaries. For anatomies of interest, a multi-layered tissue model is defined: Thickness and dielectric properties are determined for each layer based on published anatomical data (various sources) and dielectric property data (as described in previous section). For all tissues, low-loss media are assumed. The error resulting from this simplification is likely negligible compared to the error in dielectric property estimation as a result of inter-individual variation. Propagation velocity and attenuation are determined according to equations 2.11, 2.15 and 2.17. Reflection and transmission are determined according to equations 2.20 and 2.21.

3.2.2 Cardiovascular imaging

Both vascular sensing using radar, and imaging of the heart have been demonstrated before [44, 47, 48]. The current section explores a number of challenges which might be encountered

and attempts to generate rough requirements for cardiovascular sensing and imaging. The 1D propagation model is used as a starting point to discuss EM wave propagation. Here, the scalp is used as an example anatomy. In Fig. 3.2a, transmission of EM energy at 3.8 GHz is modelled

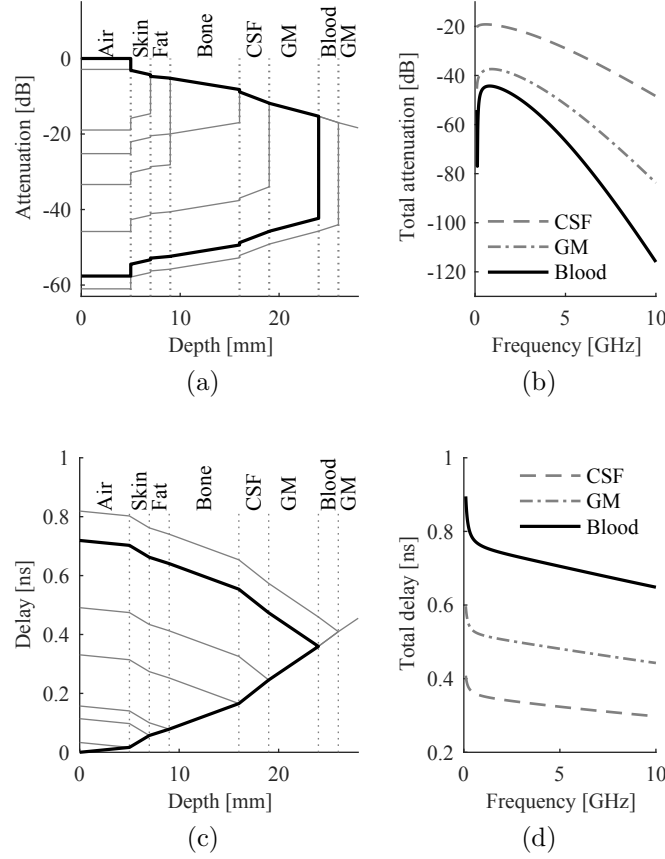


Figure 3.2: Results from a 1D signal propagation model, illustrating the interactions between electromagnetic (EM) radiation and biological tissues. EM energy propagating through the scalp and reflecting off a volume of blood in grey matter (GM) is modelled. Attenuation is plotted versus depth for all direct trajectories in (a). Total attenuation is plotted against frequency in (b) for three trajectories. Delay of reflected EM energy is plotted versus depth for all direct trajectories in (c). Total signal delay is plotted against frequency in (d). The trajectory of interest, caused by a layer of blood, is plotted in black. Figure has previously been published in [49].

through the scalp. The trajectory consists of the following media: Air (the matching medium), skin, fat, bone, cerebral spinal fluid, grey matter (GM) and blood. The head was modelled as multiple layers of planar tissue. The black line indicates the signal of interest: A reflection caused by a volume of blood in the brain, possibly an artery or a haemorrhagic stroke. In this simplified model, a total attenuation of 61 dB occurs for the blood volume reflection, while the air-skin boundary causes powerful reflections with a 2.9 dB attenuation. This illustrates the demand for high dynamic range of hardware and strong clutter rejection. Fig. 3.2b demonstrates the frequency dependency of total attenuation. It is known that higher frequency leads to higher

spatial resolution, but clearly there exists a trade-off between reduced attenuation (at low frequencies) and increased spatial resolution (high frequencies). To determine optimal centre frequency, anatomy geometry and depth must be considered as well as the sensor dynamic range and required spatial resolution. Fig. 3.2c displays the expected time delay of individual reflections: In order to resolve between targets in down-range, an extremely short pulse duration would be required. This is challenging to accomplish in hardware and the extremely wide frequency band would lead to strong pulse distortion due to dispersion. The figure implies that reflections of various tissue boundaries will most certainly overlap. Fig. 3.2d shows total time of flight (ToF) of received reflections as a function of frequency. Higher frequency radiation in the microwave band tends to propagate at higher velocity due to decreased permittivity. This signifies that waveform distortion may occur at increased imaging depths when using wideband signals, which should be considered for in-body signal processing. A final challenge which becomes apparent from Fig. 3.2 is the uncertainty of body composition. If coherent summation techniques are used for imaging (as in DAS beamforming), an assumption of anatomy and dielectric properties of different tissues is required to apply the correct signal delay. As EM propagation speed depends on tissue dielectrics, and both the anatomy and dielectric properties of tissue will vary strongly between individuals, exact coherent summation is impossible without a priori information on body composition.

Signal attenuation Signal attenuation depends on signal frequency, tissue composition and propagated path length, therefore the minimal system requirements will vary between different cardiovascular structures and between individuals. Dielectric losses can be estimated using the 1D model. For the thigh (containing a high fraction of muscle tissue), at 3.45 GHz, an average dielectric loss of 5.2 dB/cm is estimated, whereas for the chest (containing relatively more bone and fat, as well as lung tissue), a loss of 4.3 dB/cm was estimated. The 1D model does not account for path loss due to natural expansion of the wave front, as described by the Friis equation. In the far field, power varies inversely with the square of the propagation path length, but exact attenuation is difficult to estimate without a more detailed simulation of the geometries involved. For imaging applications with physically large antenna arrays, it must be considered that antenna pairs with high path length will receive significantly attenuated signals. For summation across channels in a DAS beamforming approach, path length dependent attenuation compensation should be considered.

Reflectors Strong reflections occur at large contrasts in permittivity and thus intrinsic impedance between tissue boundaries. In general, tissue water content is a good predictor for tissue permittivity: Tissues like blood, brain and muscle have high permittivity, whereas fat and bone have low permittivity. Superficial bone structures (such as the skull or the sternum bone) have been found to be beneficial for efficient radiation into tissue when using body-coupled antennas

(Section 3.3). However, strong reflections due to high dielectric contrast (skull to cerebrospinal fluid, sternum to muscle tissue) should be expected. Reflections and scattering at arteries may be attributed to the contrast between blood and blood vessel walls ($\Delta\epsilon \approx 15$), as well as the contrast between blood vessel wall and surrounding tissue (mostly muscle, $\Delta\epsilon \approx 10$). In the case of cardiac imaging, it is expected that both the movement of boundaries between cardiac muscle and surrounding tissue (lung, the ribcage), and the changes in compartmental blood volume will form significant signal contributions.

Scattering The 1D model does not account for the scattering effects which result from the exact target geometry. Due to small diameter relative to wavelength, and depending on orientation relative to antenna polarisation, arteries will reflect only a fraction of incoming EM energy back to a receiver. Due to Mie scattering, the radar cross section of an artery is frequency and diameter dependent and therefore challenging to predict without accurate a priori anatomical data. Particularly when scattering is expected (artery imaging), an antenna array surrounding the anatomy of interest should be considered.

Required dynamic range When setting dynamic range requirements, not only path attenuation and target reflection strength should be considered, but also the power of leakage signal (crosstalk and antenna backscatter) and strong static reflectors such as the skin. The leakage signal will not undergo heavy tissue attenuation and without appropriate shielding of antennas, received leakage signal will quickly distort or interfere with the signal of interest, and potentially saturate a receiver system. Adding gain on the transmit path can therefore only be done in combination with antenna shielding, and high dynamic range is required to differentiate between static clutter, leakage signals and attenuated signal of interest. Static clutter may be reduced by avoiding strong reflections at the skin interface through the use of a matching medium (more details in Section 3.3). When considering a centre frequency of 3.8 GHz, for structures at a depth of 3 to 4 cm, a reflected energy ratio of approximately -20 dB, and a coupling loss of around -10 dB, a dynamic range of at least 70 dB would be required.

Required spatial and temporal resolution Sensing or imaging of most larger arteries does not require high spatial resolution: When performing dynamic imaging it is assumed that static clutter is removed entirely, thus there will typically not be any co-located scatterers after high-pass filtering. Imaging of more complex arterial structures such as the basilar arteries would require near millimetre resolution. For cardiac imaging, spatial resolution in the order of centimetres would allow for differentiation between the left and right atria and ventricles, which would be a first step of clinical relevance. Sub-millimetre resolution as is achieved in US would allow for detailed imaging of the various structures in the heart anatomy. The temporal

resolution required depends entirely on the application of interest. US echocardiography can be performed at 50 frames per second, which is about double what the human eye can perceive. A frame rate of 20 – 50 fps would be desirable for assessing mechanical function. For accurate estimation of pulse arrival time (PAT) of the arterial pulse wave (for example for blood pressure (BP) estimation), sensing with higher temporal resolution would be required.

3.2.3 Functional neuroimaging

At an early stage of this research project, the question whether neural activity could be measured using MWI, arose. Neural tissue relies on a constant supply of oxygen and glucose by the cerebrovascular system. Because of this, there is a strict relationship between neural activity and local blood supply, called neurovascular coupling. Neural activation triggers vasodilation of local vessels, which leads to a regional increase in cerebral blood flow and volume. Additionally, an increase in local oxygenation is observed. In functional magnetic resonance imaging (fMRI), this rise in blood oxygenation is known as the BOLD (blood oxygenation level dependent) contrast. For functional MWI, different physiological mechanisms may be measured. Potential candidates are vasodilation, blood flow, blood volume and blood oxygenation. To come up with a prediction whether microwave functional neuroimaging would be in the realm of possibilities, the different neurovascular signals were quantified using available literature and their potential as a MWI target is discussed.

Vasodilation When considering a small volume of cortical GM, a number of blood vessels are involved in its blood supply: Arterioles, capillaries and venules. Arterioles are generally around 30 – 100 μm in diameter. Arterioles contract and dilate to regulate blood flow and maintain a pressure gradient. An increase in arteriole diameter of 20 % to a maximum of 50 % is observed upon neural activation [50–53]. Capillaries form a dense network of small vessels (8 μm in diameter) with a density of about 500 counts/ mm^2 [54,55]. Like arterioles, capillaries adjust diameter to regulate local blood flow [56,57]. Venules, which transport deoxygenated blood from capillaries to larger veins, have a diameter of 7 – 50 μm . Venules increase their diameter passively upon longer periods of neural activation [58].

Compared to the wavelength of radar signals in the microwave band, all vessels that could act as a marker for neural activity are extremely small. Applying the equations for Rayleigh cross section [59] demonstrate that the vessel density is by far not high enough to cause any significant scattering effect.

Blood flow As a result of local vasodilation, regional cerebral blood flow increases. According to estimates based on fMRI studies, blood flow locally increases 60 – 90 % [60–63]. In US

imaging, the flow of red blood cells in arterial blood causes a Doppler effect, which can be used for functional neuroimaging [64, 65]. Considering a propagation velocity of EM waves in tissue of around 5.5×10^7 m/s and a blood flow velocity of around 10 m/s, a Doppler frequency shift in the order of KHz would be observed. Frequency resolution of the used RF pulse will not be sufficient to detect a shift this small, hence changes in blood flow cannot be a suitable target for MWI.

Blood volume Cerebral blood volume (CBV) is known to change as a result of neural activation. A small volume of GM that contains a certain fraction of blood can be considered as a homogeneous medium. The fraction of blood will change depending on whether the volume is activated or at rest, thus change the average dielectric properties. CBV cannot be measured directly in vivo using fMRI, but has been estimated using various other techniques (PET, CT, MRI bolus tracking) in various species. Estimates vary between 4 and 10 % [63, 66–70]. A rough estimate of the volume based on the average diameter and total length of different vessels in GM results in a fraction of only 2 – 3 %. This does not take into account the complicated branching architecture and variability in vessel diameter, so it is likely an underestimation. A CBV fraction of 5 % will be assumed. Estimates for increase in CBV upon neural activation range from 10 – 30 % [70–77]. The majority of studies conclude that changes in CBV take place primarily in larger vessels, with an early increase in arterioles and a delayed increase in venules. A number of studies report the venous CBV to keep increasing upon longer stimulation (up to 30s), which could explain the large variance in CBV change estimates. A relative CBV increase of 20 % will be assumed.

The dielectric properties of GM at rest and while activated can be calculated as a weighted average of the GM and blood fraction. The relative permittivity for brain tissue at rest and while activated is estimated as 47.02 and 47.11, respectively. The contrast in intrinsic impedance would cause a reflection ratio of –66.4 dB. Considering scattering effects, path attenuation loss and coupling loss, and the fact that high scanning speed is also a requirement to capture this dynamic signal, this contrast is likely to be too small for existing hardware.

Blood oxygenation Blood oxygenation increases locally upon neural activation. Due to the fact that deoxygenated haemoglobin is paramagnetic and oxygenated haemoglobin is diamagnetic, a change in magnetic properties could be used as a biomarker for activity, as is done in fMRI. Magnetic susceptibility changes the intrinsic impedance of a material, so in theory a change in blood oxygenation could be measured using EM waves. Magnetic susceptibility of oxygenated and deoxygenated blood has been quantified [78, 79] and differences are of an extremely small scale. The resulting intrinsic impedance of oxygenated blood and deoxygenated blood is 50.4962Ω and 50.4963Ω , respectively. Considering that only the venous fraction of the

vascular compartment will be deoxygenated, and the total blood volume is again a small fraction of the cortical tissue, the contrast due to blood oxygenation can be considered negligible.

3.3 Design considerations

Now that some background on anatomical structures and dielectric properties has been provided for the main imaging targets of interest, general system design considerations will be discussed.

Radar type UWB radar offers the advantage of target ranging and time gating of unwanted noise sources for sensing applications. More importantly, it allows for imaging in 2 or 3 dimensional space when spatially diverse radar data are combined. Operating an UWB radar in the time domain offers less flexibility with regard to centre frequency and pulse bandwidth, but has the advantage of potentially lower scanning times [29] and thus increased spatial resolution for dynamic imaging. The current research focuses specifically on RoC technology. The small form factor and low cost of RoC devices make this an attractive option for applications where device mobility, availability, and ease of use is required. However, making a MWI device portable and even modular means there is significantly less control over the clutter environment, and more leakage (crosstalk, backscatter) is to be expected as compared to a fully static imaging system. When considering dynamic imaging, low scan times are required and RoC devices offer the possibility of simultaneously receiving from multiple (spatially diverse) devices. This means that the number of Rx channels can be increased (leading to increased signal quality) without the usual penalty in scanning time.

Spatial resolution UWB radar resolution using a single module is limited by the radar down-range resolution, determined as half the pulse length [80]. Effective pulse length in biological tissue is shorter than pulse length in air due to a slowing down of EM wave propagation in tissue by factor $\sqrt{\epsilon_r}$, where ϵ_r indicates the tissue relative permittivity. As previously shown in Section 2.2.1, propagation velocity in tissue is therefore given by $c/\sqrt{\epsilon_r}$, where c is the speed of light ($\approx 3 \times 10^8$ m/s). Down-range resolution ΔR (in m) is thus defined in tissue as:

$$\Delta R = \frac{c\tau}{2\sqrt{\epsilon_r}} \quad (3.1)$$

where τ is the pulse duration (in s). Spatial resolution can dramatically be improved by adding to the number of modules and taking an imaging approach. Whereas down-range resolution is limited by pulse length, cross-range resolution is limited by the array aperture and centre frequency of operation. From literature on focused synthetic aperture radar (SAR) with a

circular array arc (wideband, wide angle), it is known that cross-range resolution ΔR_c (in m) is determined by:

$$\Delta R_c = \frac{\lambda}{2\Omega} \quad (3.2)$$

where λ corresponds to the effective wavelength of the centre frequency in the propagation medium as defined in Section 2.2.1, and Ω is the angular range (or angular bandwidth) [81]. Note that in tissue, λ is a factor $\sqrt{\varepsilon_r}$ shorter than in air.

Transmitted waveform Due to the trade off between tissue penetration and spatial resolution, no clear answer exists as to what is the optimal centre frequency and bandwidth. High pulse bandwidth leads to higher down-range resolution, but provides challenges in hardware design and antenna matching. Wide bandwidth is easier to achieve at a higher centre frequency of operation. This also reduces the required antenna size and allows more antennas per unit area in a multistatic array, thus increasing the spatial sampling frequency, resolution, and array gain. However, a higher frequency of operation comes at the cost of significantly increased dielectric losses in tissues. Whether the transmitted waveform should be optimised for down-range or cross-range imaging, depends again on the imaging and array geometry. A circular array surrounding the anatomy of interest will rely mostly on cross-range resolution, whereas a rectangular array will require sufficient down-range resolution for target ranging. A final consideration for pulse bandwidth and centre frequency, is emission regulations. The FCC has authorized the unlicensed use of UWB emission between 3.1 and 10.6 GHz, with a power spectral density emission limit of -41.3 dBm/MHz . RoC systems have been developed to operate within this emission mask [82] and are particularly attractive due to their availability and low cost.

When operating within the FCC mask, the lower end of the spectrum is preferred due to the high signal attenuation at increased frequencies. The estimate of signal attenuation in tissue at 3.8 GHz (corresponding to the XeThru X2 chip) of approximately 5 dB/cm, would be as high as 16 dB/cm at a centre frequency of 7.29 GHz (corresponding to the XeThru X4 chip). Although there exists a trade-off with spatial resolution and array gain due to reduced antenna size, dynamic range of the X4 chip would likely not be sufficient for imaging at high depth.

Temporal resolution High frame rates can be achieved using RoC technology due to the use of multiple simultaneous receivers, but there exists a trade-off between temporal resolution and signal-to-noise ratio (SNR). During data collection of a single frame, the imaging scene must be assumed static. Frame rate must be sufficiently high to sample motion of anatomical structures, and is therefore application dependent. The Nyquist theorem can be used as a

guideline, and maximum signal integration within the available sampling duration should be applied.

Monostatic vs multistatic imaging A monostatic imaging system using a single, mechanically scanned transceiver offers the advantage of high array density and reduced size constraints on the antenna. However, the requirement of physical scanning of the antenna is not suitable for a portable and modular imaging system. The option of simultaneous receiving using multiple RoC devices make multistatic imaging an obvious choice, as adding to the number of receivers does not impose a penalty on scanning time. In addition, when considering small radar targets such as arteries, significant scattering will occur and monostatic radar using co-located Tx and Rx would capture only a fraction of scattered EM signal. To increase the robustness of the imaging system against scattering effects, spatially diverse receivers are preferred. Whether MIMO imaging using multiple transmitters is preferred or SIMO imaging using a single transmitter, is dependent on required temporal resolution and image quality. The virtual element density of a MIMO system is typically twice as high as the array density of a SIMO or monostatic system [83], which results in improved imaging quality through reduction of aliasing. In addition, a higher number of Tx-Rx pairs leads to improved SNR through coherent summation. However, unless orthogonal waveforms are used for simultaneous transmission (as in communication systems), a MIMO dataset is obtained by sequentially scanning through transmitters and is therefore time consuming.

Aperture A real (as opposed to synthetic) aperture is required for multistatic imaging and realised through a physical array of antennas. Increased aperture size tends to lead to increased spatial resolution, with the limitation of antenna directionality: Only the effective aperture is considered, which is limited to the set of antennas for which the target lies within the beamwidth [81]. For in-body imaging, not only antenna beamwidth is relevant to the effective aperture, but also signal attenuation due to dielectric losses must be taken into account. When signal attenuation exceeds the sensor dynamic range as a results of dielectric loss at high path length, a channel is no longer part of the effective aperture and does not contribute to increased spatial resolution.

Array density For any linear array imaging in the far field (where plane wave propagation is assumed), the inter-element distance d should not exceed $1/2\lambda$ in order to avoid grating lobes (spatial aliasing). In this, λ is the effective centre frequency wavelength in the propagation medium. In tissue, λ is a factor $\sqrt{\epsilon_r}$ shorter than in air. This would put considerable limitations on the antenna dimensions, unless when using matched antennas, in which case antenna dimensions scale accordingly. However, in MWI for biomedical applications, beamforming oc-

curs in the near field. Spatial aliasing in the far field is effectively direction of arrival (DOA) ambiguity. In the near field, spatial aliasing produces ambiguous target locations lying on an ellipse, which is unique for each Tx-Rx pair. This concept is illustrated in Fig. 3.3a (far field) and 3.3b (near field). In near field imaging with a sparse array, there is no coherent summation of aliasing as seen in the far field. Particularly in a circular array, subsets of antennas have diverse geometries relative to voxels, and coherent summation of aliasing is even less likely. A sparsely sampled array in the near field imaging will lead to increased clutter. No theoretical framework seems to exist to estimate the optimal array density for the imaging scenarios considered in this work. Simulations may be used to predict the point spread function (PSF) of an array design. Practically, in the field of biomedical MWI, the array density is limited by the physical dimensions of the antenna and by the need to avoid cross talk between antennas. An effective way to obtain a high density of virtual elements is to use MIMO radar [83].

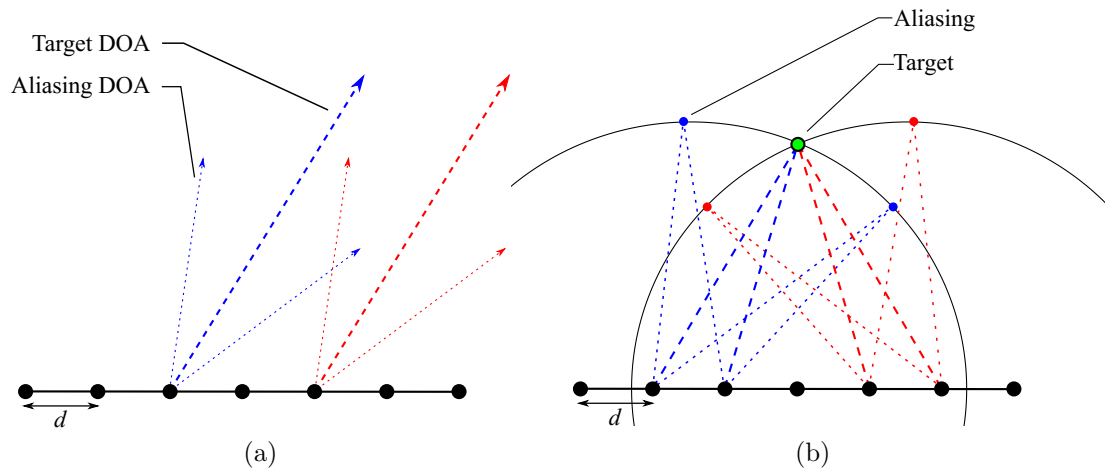


Figure 3.3: Illustration of the effects of an undersampled array in the far field (a) and near field (b). For the far field scenario, blue and red dashed lines indicate the target direction of arrival (DOA) and aliasing DOA for two different elements. As the DOAs are identical, coherent summation will occur. For the near field, aliasing targets lie on ellipses around element positions, and aliasing targets from different Tx-Rx pairs do not overlap.

Antenna As found in Section 3.2.2, in-body MWI is challenging due to high attenuation of EM energy. In addition, the air-to-skin interface causes strong reflections and backscattered signal may be modulated by any motion in the vicinity of the measurement setup, including chest motion caused by the heart beat. The signal of interest in many cases is dynamic, with known frequency content (in case of cardiovascular monitoring), but SNR is extremely low compared to static clutter. In the microwave breast cancer imaging community, this problem is often solved by using a liquid matching medium between antenna and skin [22]: The breast is suspended in a lossy fluid which reduces the impedance mismatch of EM radiation entering

the tissue, and absorbs backscattered signal. This tissue and antenna submersion however is often not practical. An alternative solution consists of the use of body-coupled antennas that are designed to operate while in direct contact with the skin. By matching the antenna to the intrinsic impedance of the tissue of interest, coupling efficiency of EM radiation into tissue is maximised [84, 85]. The air-to-skin interface reflections and backscattered radiation into air are reduced. An additional advantage of using matched antennas is that the reduced effective wavelength (scaled down by factor $\sqrt{\epsilon_r}$) also reduces the required antenna dimensions, allowing for a denser array. A challenge with operating antennas in direct contact with biological tissue, is the occurrence of high losses in the reactive near field. To avoid this, a low loss dielectric spacer may be used between antenna and tissue. Such insulating layers between antenna and lossy biological tissues have been found to greatly increase radiation efficiency in MWI [84, 85] as well as in the field of implanted antennas [86]. Simulations of body-coupled antennas as used in the current work have also shown increased performance when combined with dielectric spacers¹ [87].

Shielding As body-coupled antennas will be used in a modular and portable imaging setup, no coupling liquid will be present to aid in absorption of backscattered and crosstalk signal. This puts additional requirements on the level of shielding of antennas and RoC modules. Signal attenuation due to dielectric losses and low target contrast may be as high as 40 – 60 dB. Although static clutter (including leakage signals) may be reduced through high-pass filtering, artefact removal and spatial filtering, shielding should be a top priority in a RoC-based MWI device.

Power Radiated power for a biomedical imaging or sensing system is first and foremost limited by patient safety. EM radiation at microwave frequencies is non-ionising and the only known adverse effect is potential heating of tissue at high power levels. Radiation exposure is defined in levels of specific absorption rate (SAR), and guidelines such as those by the International Commission on Non-Ionizing Radiation Protection (ICNIRP) may be followed to set an exposure limit. A second consideration is radiation into air, which is limited by emission masks as provided by the FCC, or the European equivalent: The European Telecommunications Standards Institute. The RoC technology used in the current work is a commercially available device and complies with international standards on unlicensed UWB emission. With that, the radiation exposure for patients is far below ICNIRP guidelines (safety considerations provided in Appendix F). Transmitted power could further be increased as long as leakage signals are contained or attenuated.

¹Simulations performed by Mathias Tømmer (UiO.)

3.4 Hardware characterisation

The final system design as employed in the current research project was a result of a collaboration with UiO and Norwegian sensor company Novelda AS. In order to develop a working prototype and perform measurements on human participants within the time constraints of this project, a module based on existing commercially available coherent UWB single-chip radar was used. The current section aims to characterise used hardware and performance relevant to biomedical imaging and sensing applications.

3.4.1 System description

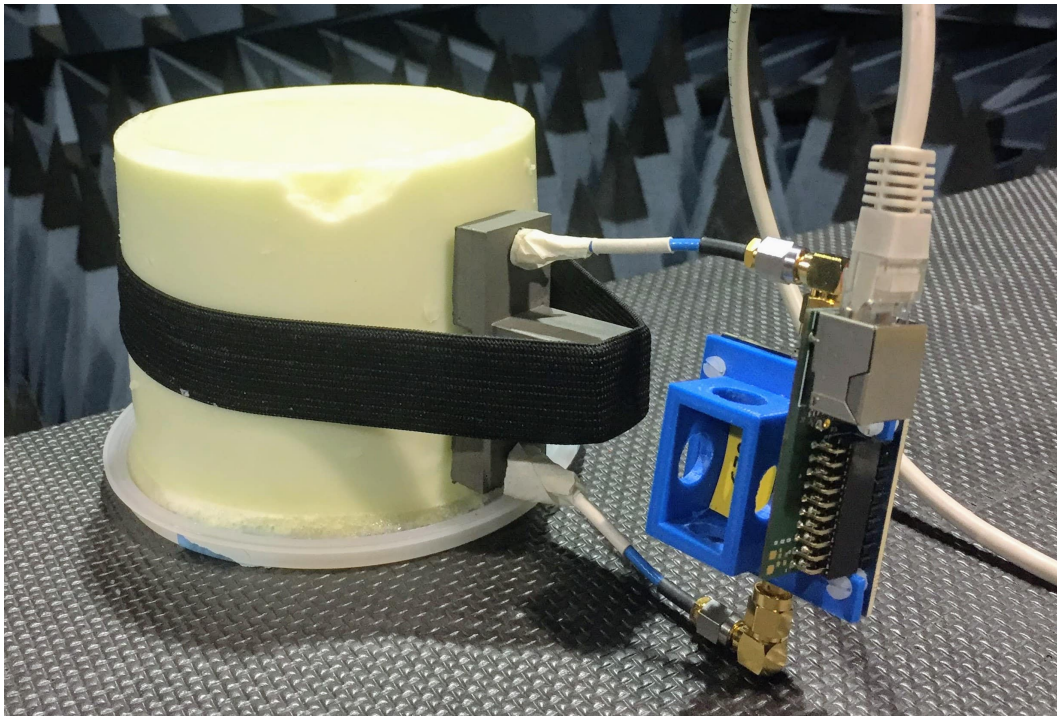
In this work, up to nine coherent UWB RoC devices were operated simultaneously. A BeagleBone system was used to control all radar modules. Data were transferred from BeagleBone to data acquisition computer over Ethernet cable. Each transceiver module was connected to a Tx and Rx antenna. Body-coupled wideband monopole antennas with low-loss dielectric spacers were designed to be in direct contact with the skin. Antennas and modules were held in place using 3D printed structures. Modules could be programmed to transmit sequentially, all modules could receive simultaneously. Earlier versions of the radar modules and BeagleBone system (as used in Chapter 4) are shown in Fig. 3.5. Array designs are described in detail in Chapter 5. Radar modules were developed by Dr. Kristian Kjelgård (UiO) and Novelda AS, Norway. The synchronised multi-module radar control system, antennas, module shielding, and arrays were developed by Mathias Tømmer (UiO).

3.4.2 Hardware characterisation: Methods

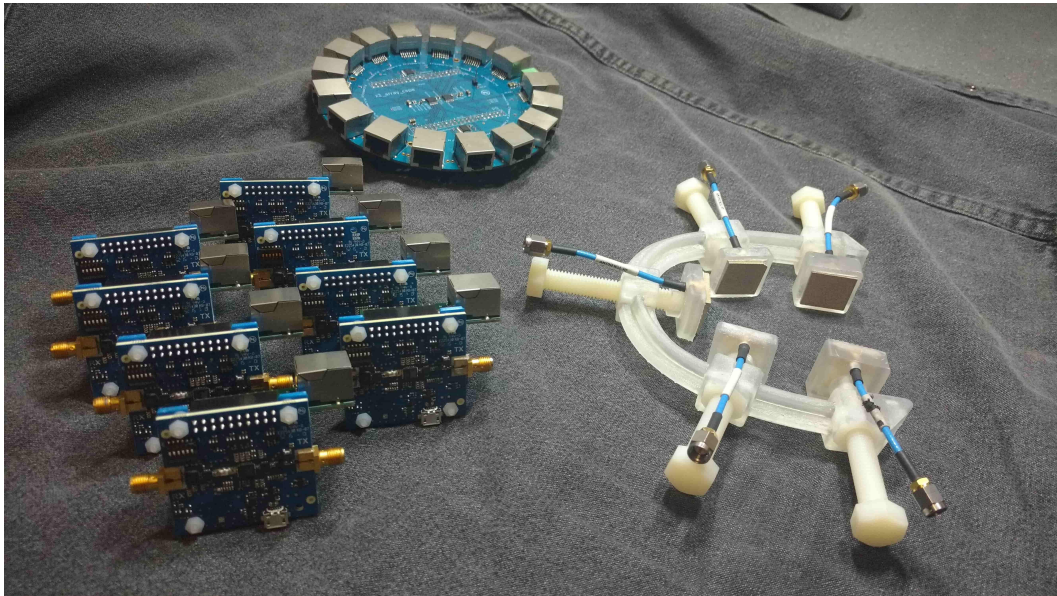
Phantoms mimicking the dielectric properties of various biological tissues were developed to test antenna coupling and radiation, signal transmission, and target detection in lossy media. A liquid phantom was developed so that antennas could be suspended for rotational and translational freedom, and a semi-solid phantom was created to obtain a more realistic model for antenna coupling. Phantoms were designed to mimic dielectric properties (prioritising permittivity over conductivity) in the frequency band of interest, including their dispersive nature.

Liquid phantom

A mixture of sugar in water was used for the liquid phantom, based on previously published recipes and dielectric measurement results [88,89]. Methods have been included in Appendix D.



(a)



(b)

Figure 3.5: Photographs of early versions of the imaging hardware: A single radar module with body coupled antennas held in a 3D printed structure, pressed against a semi-solid cylindrical phantom with dielectric properties similar to tissue (a). The radar module is controlled through the BeagleBone system. Eight modules and BeagleBone system are shown, along with a 3D printed bracket designed to position body-coupled antennas in an array around the arm (b).

Multiple samples of sugar-in-water phantom were created and their properties were assessed using a VNA (Agilent E8361A) and dielectric probe (Agilent 85070E). A liquid phantom was created that approaches the dielectric properties of GM tissue at a centre frequency of 3.8 GHz, although conductivity was found to be too high. Dielectric measurement results of the created sugar-in-water phantom, along with the dielectric properties of GM tissue are given in Fig. 3.4. Throughout the duration of this research project, permittivity of the phantom decreased due to evaporation, and occasionally water was added to increase permittivity to the desired level.

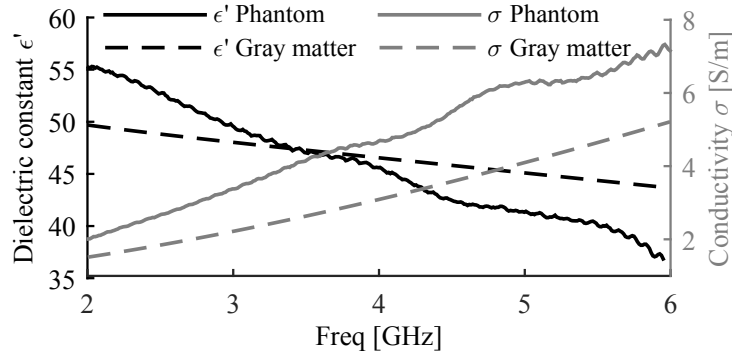


Figure 3.4: Dielectric measurement results of the sugar-in-water phantom, along with dielectric properties of grey matter tissue [18].

Semi-solid phantom

An oil-in-gelatin dispersion was created, based on previously published recipes and dielectric measurement results [90–93]. The exact ingredients and methods are included in Appendix E. Multiple samples of semi-solid phantom were created and their dielectrics were measured using a VNA and dielectric probe. In Fig. 3.6 measured dielectric properties of four different samples are given, along with the properties of GM and skin. The created samples mimic high water-content biological tissue (skin, GM) very well, and the exact properties can be fine-tuned by varying the oil content. Synthesis of a high oil-content (thus, low permittivity) samples has proven to be more challenging. Different recipes (such as ceramic-polymer composites [94]) ought to be explored for mimicking low water-content tissues (fat, bone).

3.4.3 Radar module

A custom radar module (dubbed the Ventricorder) was used, which was a result of a Novelda funded research effort on heart inspection. The module consists of the XeThru X2 single-chip radar (Novelda AS, Kviteseid, Norway) with reduced centre frequency and additional adjustable

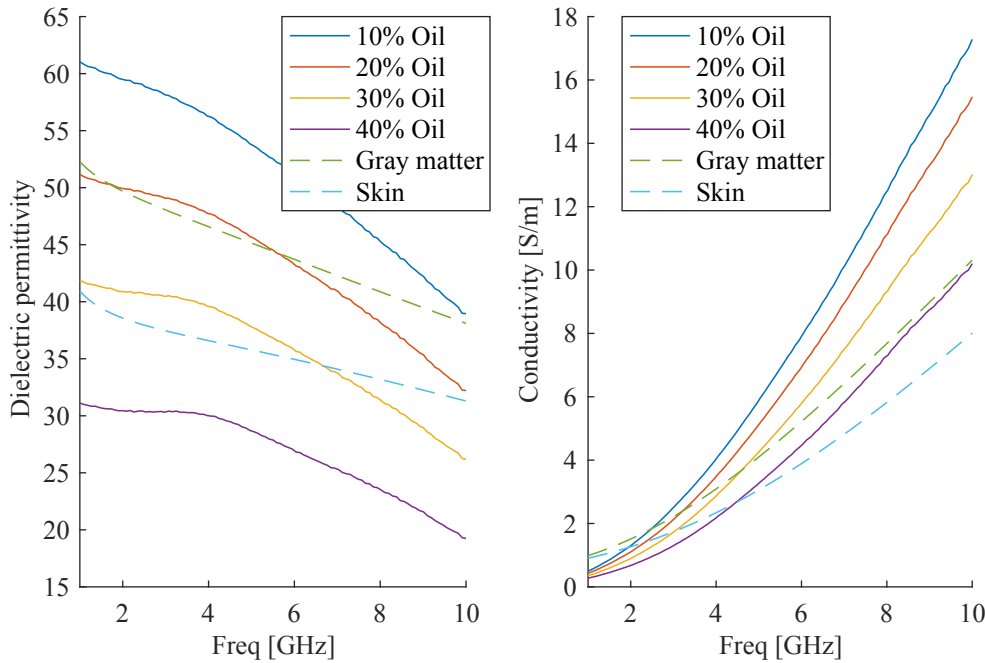


Figure 3.6: Dielectric measurement results of four semi-solid phantom samples, along with dielectric properties of grey matter and skin tissue [18]. Relative permittivity is shown left, conductivity is shown right.

amplification on both the transmit and receive paths. The architecture of the Ventricorder module is shown in Fig. 3.7. The X2 architecture overview is shown in Fig. 3.8. Use of the XeThru X4 chip was considered but deemed infeasible for measurements at high depth: A four-fold signal attenuation in tissue can be expected when using the X4 centre frequency of 7.29 GHz (Section 3.2.2).

Transmitter

Transmitted waveform The transmitted waveform is a Gaussian modulated sinusoid, with adjustable centre frequency and a -10 dB bandwidth of up to 2.5 GHz. Early versions of the radar module (as used in Chapter 4) had a lowest centre frequency of 3.8 GHz, whereas for later versions (as used in Chapter 5) centre frequency was lowered even more to 3.45 GHz. Transmitted pulse centre frequency is controlled through the module PG setting. Pulse amplitude spectra² for PG settings 0 to 9 are given in Fig. 3.9a. Pulse waveform at PG=0 (with centre frequency of 3.45 GHz) is shown in Fig. 3.9b. The X2 chip is designed to operate with radiation emission levels in compliance with FCC regulations (-41.3 dBm/MHz). To compensate for reduced supply voltage and achieve higher SNR, additional amplifiers were added on both the Tx and Rx paths (Fig. 3.7). The transmitted power level was increased by 8 dB. Pulses are

²Transmitted amplitude spectrum measurements performed by Mathias Tømmer (UiO).

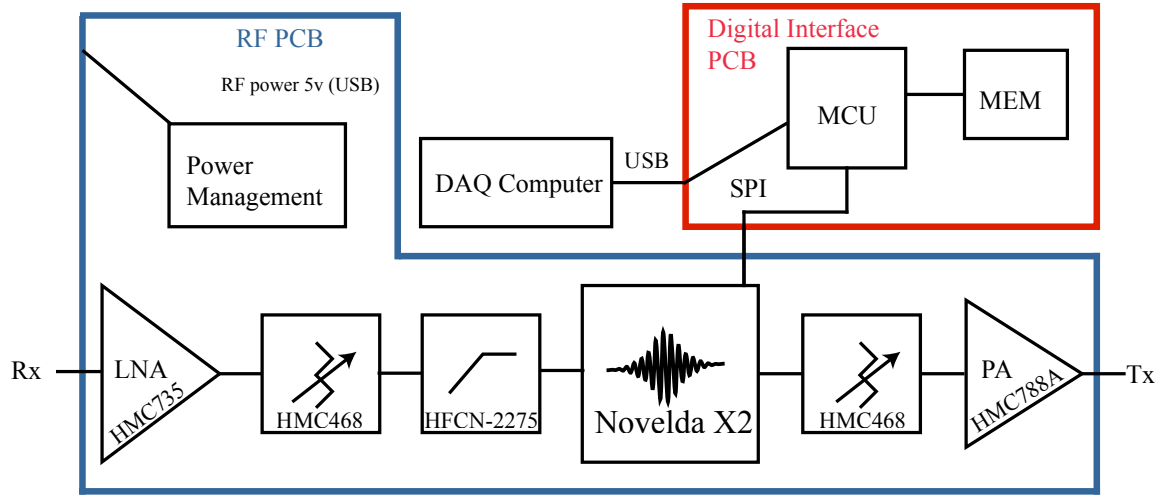


Figure 3.7: Overview of the Ventricorder radar module architecture. The radar module is shown in blue, the digital interface between radar module and computer is shown in red. Figure adjusted from [95].

transmitted at a frequency of 100 MHz, but repetition frequency can be lowered to increase the maximum unambiguous range. The -3 dB pulse width τ is measured at 1 ns for PG=0. In brain, muscle, and fat tissue respectively, an effective down-range resolution of 2.3, 2.1, and 6.6 cm can be expected at this frequency and pulse width. Considering that effective wavelength λ of a 3.8 GHz centre frequency in muscle tissue (at $\epsilon_r = 51$) is 1.1 cm, it is likely that sub-centimetre cross-range resolution can be achieved, depending on geometry of the body part and antenna array, as well as antenna beamwidth in tissue.

Receiver

Processing gain The X2 receiver consists of a delay line of 256 elements and a threshold which is swept from below to above the range of the input signal, while received signals repeatedly enter the receiver delay line. By integration over a high number of received pulses, a full signal sweep is obtained [82]. Various parameters control the threshold sweep. The digital-to-analogue converter (DAC) range of the swept threshold, number of pulses per sweep step, DAC step size, and number of iterations determine the overall processing gain. Better SNR can be obtained at the cost of reduced frame rate. The clockless design achieves an effective sampling rate of approximately 40 GHz. The radar outputs frames of 256 depth samples, corresponding to a frame length of approximately 6 ns. A sampling delay between pulse transmission and sampling onset can be set to control depth range. If longer frame length is required, multiple frames can be stitched together.

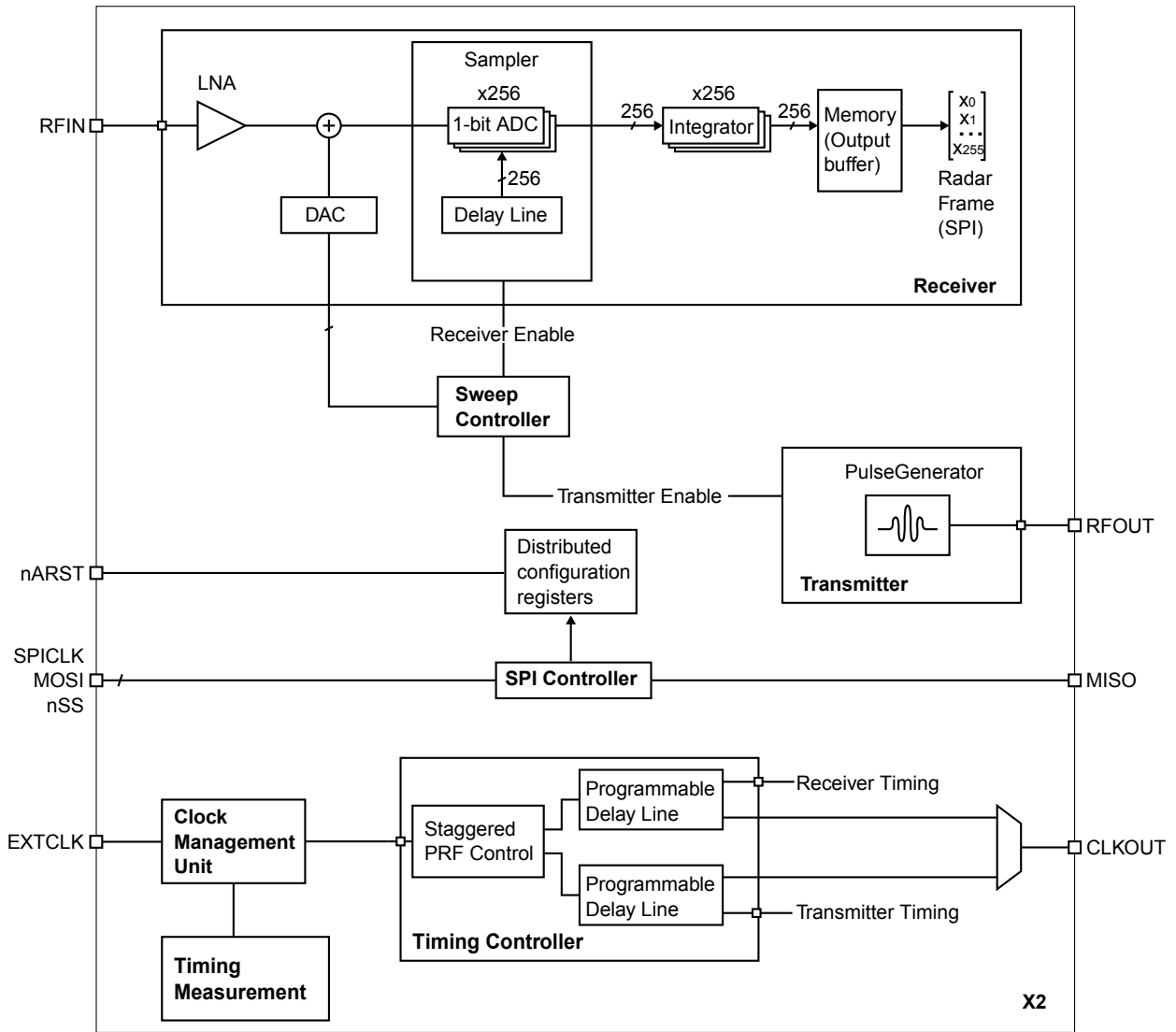


Figure 3.8: Overview of the XeThru X2 radar-on-chip architecture. Figure from X2 Datasheet, available at [96].

Dynamic range Dynamic range of the radar module was measured using a jumper cable between Tx and Rx paths, with additional attenuation to avoid signal truncation. For all measurements, a high level of on-chip processing gain was applied, resulting in a frame rate of 10 Hz. Dynamic range was measured within a single module and between two modules, to assess on-module leakage. Received data were high-pass filtered and the root mean square (RMS) was obtained. When using two modules (one for Tx, one for Rx), a noise floor was measured at -70 dB with 35 dB attenuation, resulting in a dynamic range of 105 dB. Different levels of attenuation showed that noise floor was static. When using a single module, the noisefloor was raised by up to 20 dB in the first 5 – 6 ns, due to on-module signal leakage. On-module attenuation did not affect leakage signal.

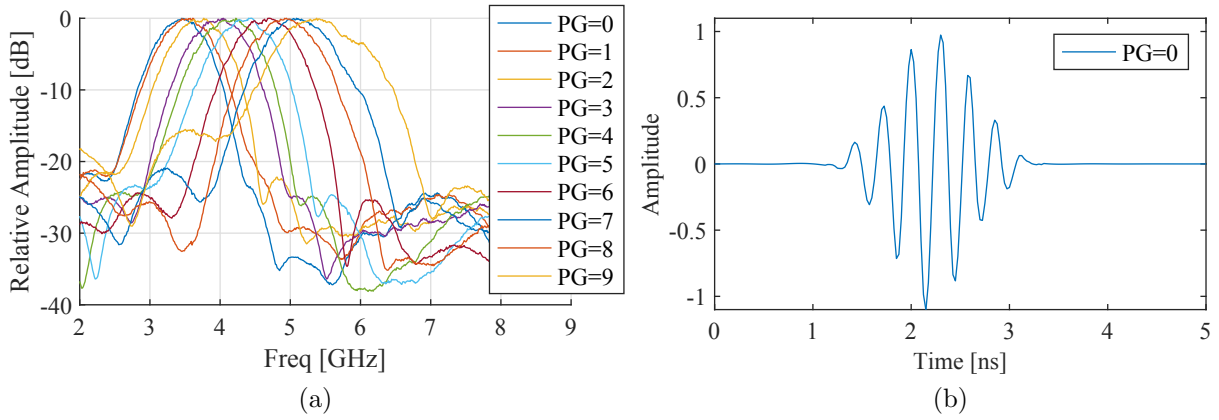


Figure 3.9: Measured amplitude spectra for different PG settings on radar module (a) and transmitted waveform at PG=0 (b).

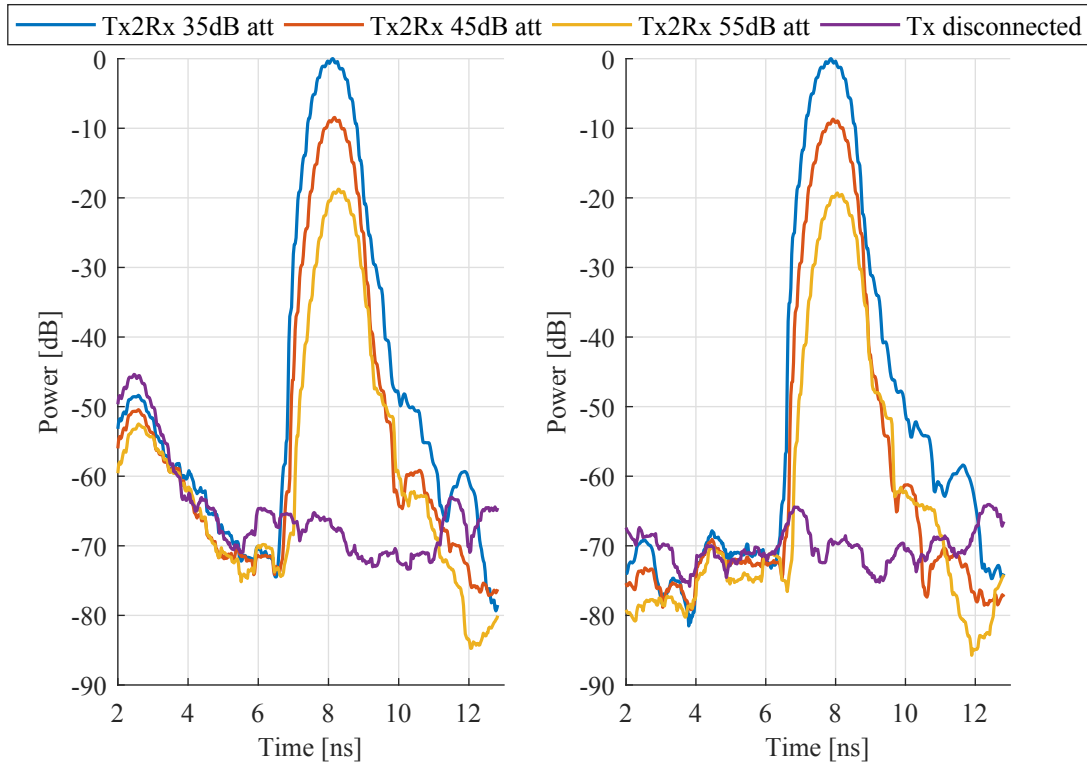


Figure 3.10: Dynamic range of the modified X2 radar module, measured by connecting Tx directly to Rx using a jumper cable and various levels of attenuation. Dynamic range was evaluated on a single module (a), and for one module acting as transmitter and a second module acting as receiver (b). For the single-module measurement, on-module leakage is clearly visible early in the frame, raising the noise floor by approximately 20 dB.

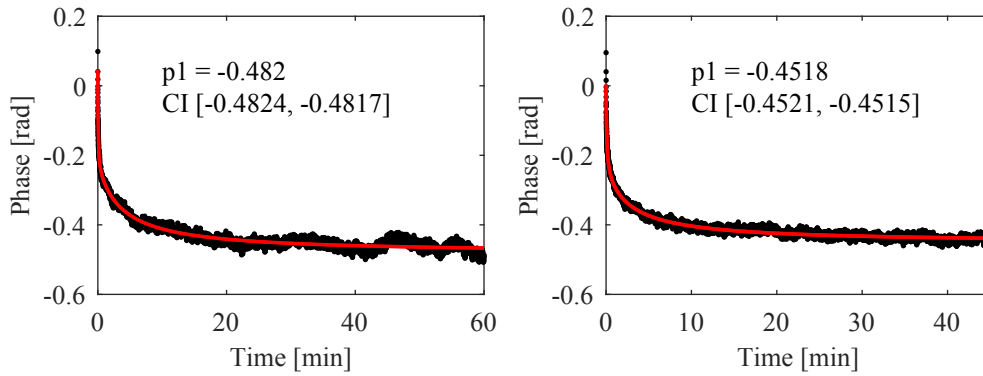


Figure 3.11: Measured phase drift of the X2 receiver, with fitted rational function, for 60 and 45 minute recording (a and b, respectively). Function parameter p_1 describes the model horizontal asymptote.

Phase drift Both the sampling delay lines and the delay elements that dictate sampling onset are sensitive to noise. Variations in supply voltage and temperature cause phase jitter on the receiver as well as variations in sampling rate. Particularly the heating of the radar chip during operation causes significant phase drift.

Receiver phase drift is tested by connecting Tx to Rx using a jumper cable, with additional attenuation to avoid signal truncation. The radar was operated for 45 and 60 minutes at low frame rate (8 and 10 fps respectively), with maximum signal integration for the chosen frame rates. Phase drift was measured over time using the following processing steps: Individual frames were detrended and zero-phase bandpass filtered (2.5 – 5.5 GHz). A baseband conversion with arctangent demodulation was applied. Signal peak was determined at maximum amplitude, phase was determined at signal peak. Obtained phase drift is modelled, for characterisation, comparison and compensation purposes. A rational function is chosen to capture the steep initial phase drift as well as the phase stabilisation which is expected at high run times. A nonlinear least squares approach is used to find model parameters. The horizontal asymptote is given by p_1 .

Phase drift and fitted model are given for two individual measurements in Fig. 3.11. Values for p_1 and 95% confidence interval (CI) are given in the figure. p_1 values differ slightly between measurements, making a generic phase drift compensation model problematic. In addition, for both measurements phase did not stabilise, even after 60 minutes of recording. Results imply that passive preheating of radar systems might not be sufficient.

Phase jitter High frequency phase noise is generally referred to as phase jitter. Its characterisation is of importance for identifying the phase noise floor at different frequencies. Phase jitter may or may not mask phase variation caused by the target of interest. To analyse phase jitter, modelled phase drift was subtracted from previously recorded Tx to Rx datasets. Results

for two individual measurements are given in Fig. 3.12 in blue. Obtained standard deviation is used as a descriptive metric, but can also be used as statistical noise floor for sensing applications where small phase variations are expected. Power spectral density (shown up to 2 Hz) is particularly relevant for sensing and imaging of dynamic structures: The HR is typically near 1 Hz, hence high jitter around that frequency would be problematic.

Within-frame phase jitter compensation Low frequency jitter components are substantial and will prove to be problematic for sensing applications where slowly varying signals are expected. The X2 receiver line consists of 256 samplers, corresponding to 6.4 ns. If phase noise is coherent across samplers, phase noise calibration could be performed within frames by taking the phase of a static signal as calibration signal. For that scenario, signal of interest and calibration signal must be separated in time, possibly using frame stitching.

To test this, a 60 minute recording was performed with Tx and Rx connected. Instead of using a jumper cable, two antennas were taped together, facing each other. The intentional poor contact and alignment between antennas caused reverberations in the received signal, ensuring that an oscillatory signal was present across the 6 ns receiver line of the X2. The phase signal was tracked for 11 probes spread across the sampling line, resulting in a 0.5 ns interval between probes. Compensation of phase at a certain probe was performed by subtraction of phase at a different probe. Results are shown in Fig. 3.12, with compensation using two adjacent probes (0.5 ns apart) in the left column and compensation using remote probes (4.0 ns) in the right column. Signals after compensation (orange) are compared to signals before compensation (blue). Great results are obtained when using within-frame, adjacent probe jitter compensation: An average gain of 20.27 dB/Hz. Performance drops when increasing the time delay between phase probes. A heatmap of jitter compensation between all 11 probes is shown in Fig. 3.13. It is apparent that phase noise compensation performance drops as the delay between compensation probe and test probe is increased.

Between-frame phase jitter compensation An alternative to within-frame jitter compensation is the use of a switch matrix which alternates between receiving the signal of interest and a static calibration signal. In this case, jitter compensation is performed between frames. The previously recorded dataset is used to test this hypothesis without the use of a switch matrix, by calibrating using a probe at identical sample delay but in the preceding frame. Results are displayed in Fig. 3.12 (left column). Successful suppression of low frequency jitter occurred, but high frequency jitter appeared independent between frames. An average gain of 14.18 dB/Hz was measured, indicating that between-frame calibration is still highly beneficial, particularly when the signal of interest is composed of low frequency components. No attempt was made to build a physical switch matrix. Instead, the (assumed static) on-module leakage,

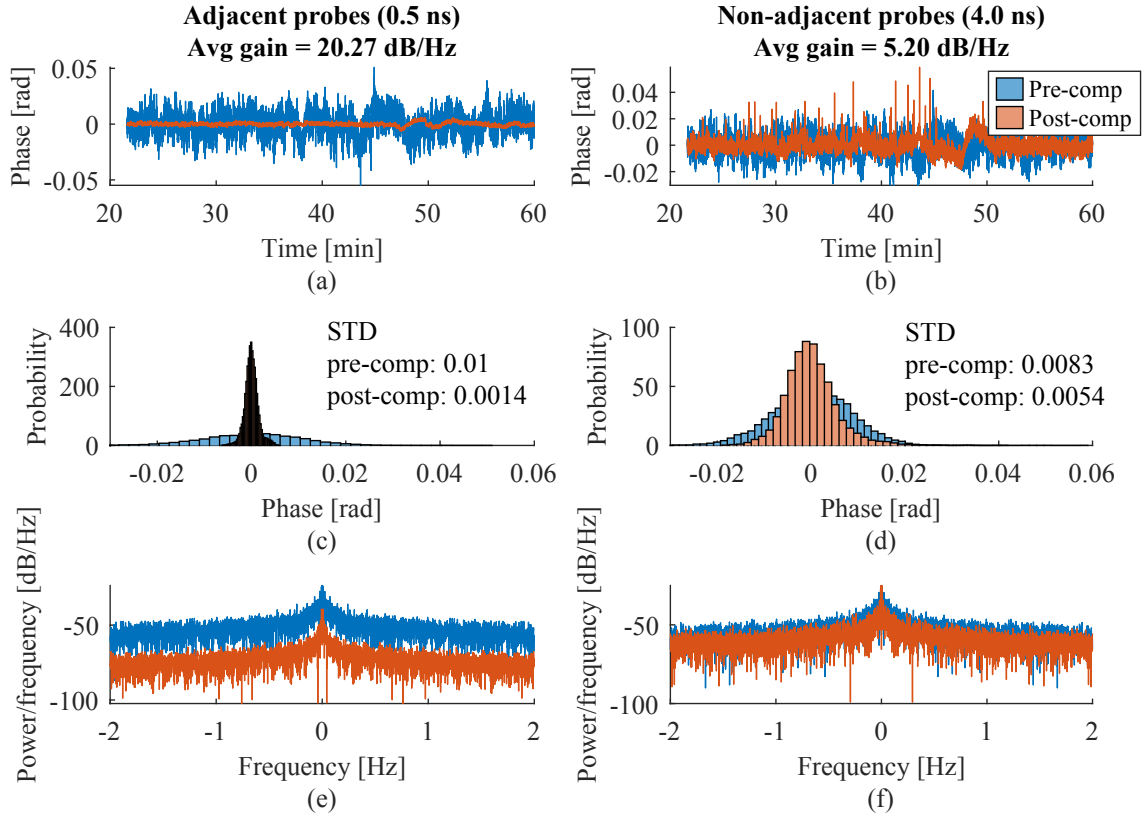


Figure 3.12: Within-frame phase jitter compensation performance for adjacent and remote phase probes (left and right column, respectively). Phase data before compensation (blue) are compared to phase data after compensation (orange). From top to bottom, graphs show phase time series, phase histogram along with standard deviation (STD), and power spectral density at frequencies up to 2 Hz.

which appears very early in the receiver line, was tried as a calibration signal. If successful, this would provide a simple jitter calibration solution requiring no hardware modifications. A recording was performed, in which the sample delay was alternated each 100 ms between one of two settings: 0 ns sample delay (capturing on-module signal leakage), and 7.6 ns (pulse received through jumper cable). The jumper cable pulse phase was calibrated on the leakage signal phase. Results are given in Fig. 3.12 (right column). No improvements in phase jitter were observed for the leakage signal compensation approach: Average gain was measured at -1.59 dB/Hz, which is a slight reduction in performance.

3.4.4 Antennas

Antennas were designed to comply with the following requirements: Efficient radiation and flat phase response across the bandwidth of interest, and small form factor to achieve sufficiently high array density. Challenges were the dimension being dictated by the lower cutoff of the

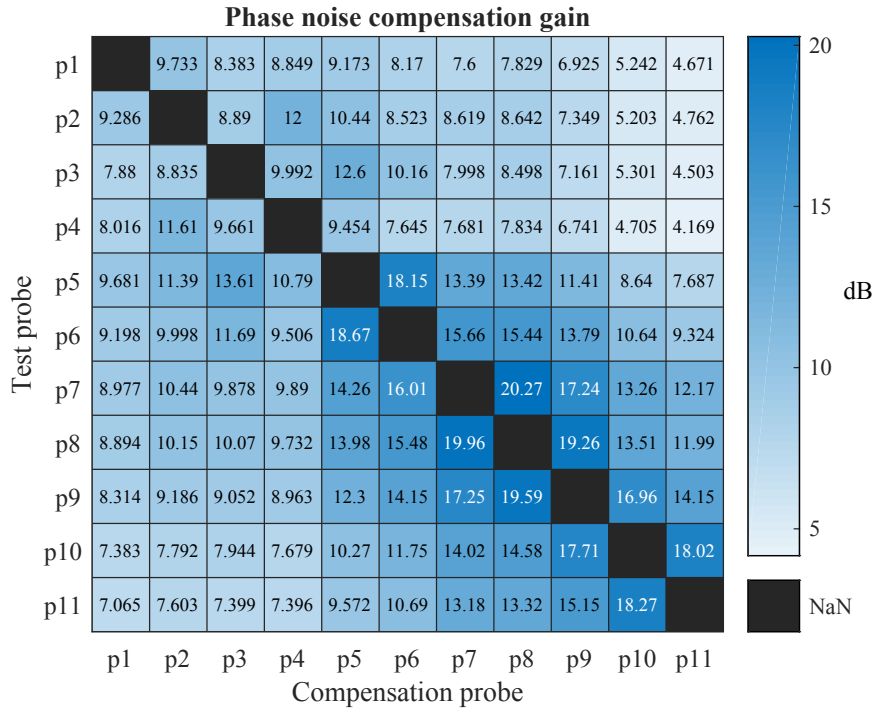


Figure 3.13: Average within-frame phase noise compensation gain (in dB/Hz), for combinations of all 12 phase probes.

bandwidth, and the large contrast at the skin-to-air interface reducing radiation efficiency and resulting in high backscattered signal. For the anatomies of interest and the design requirement of a portable system, no coupling liquid could be used to absorb backscattered signal. Instead, body-coupled antennas were developed to operate in direct contact with the skin. By operating in contact mode, antenna geometry could be scaled down by factor $\sqrt{\epsilon_r}$ due to reduced effective wavelength at high permittivity. An additional advantage is the lack of reflections at the skin interface at perfect antenna coupling. The final antenna design (pictured in Fig. 3.18a) was a wideband monopole antenna with 2 by 2 cm dimensions, designed for coupling to materials with a permittivity of $\epsilon_r = 30$. Further details on the antenna design and properties can be found in [87].

Dielectric spacer

Challenges associated with body-coupled antennas are high signal absorption experienced in the reactive near field due to lossy tissue, as well as altered antenna properties when in contact with different tissues. To counteract these effects, dielectric spacers were developed: Ceramic insulators with permittivity close to that of the skin at 3.8 GHz but low loss. Improvements in signal transmission as a result of the addition of dielectric spacers were tested on various early antenna designs, both through liquid phantom and through the forearm of a human participant.

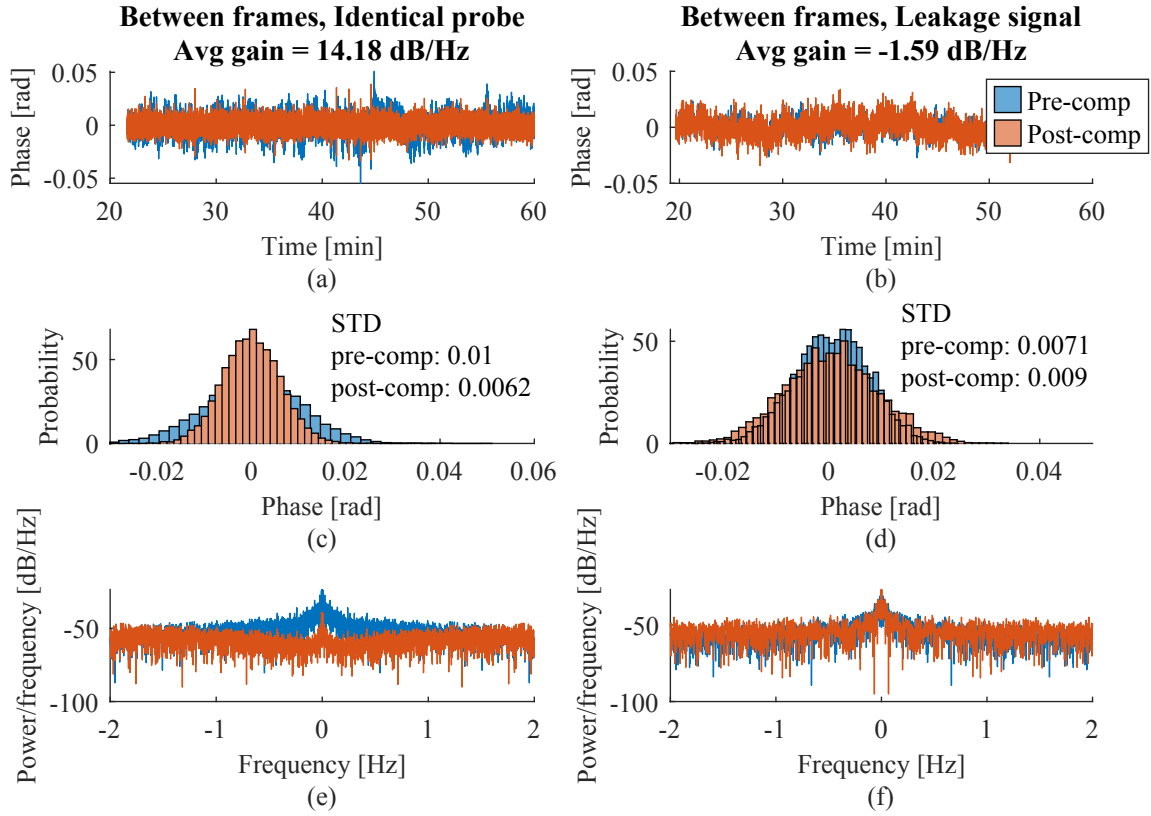


Figure 3.14: Between-frame phase jitter compensation performance for identical probes on subsequent frames, and for calibration using the on-module leakage signal (left and right column, respectively). Phase data before compensation (blue) are compared to phase data after compensation (orange).

Methods For the phantom measurements, a container was filled with 5 litre of liquid phantom. Two wideband elliptical patch antennas were wrapped in a thin, latex layer to provide a protective seal against the liquid phantom. A foam structure was used to maintain an air cavity behind the antenna, to mimic body surface contact. Using a VNA, it was found that the return loss (S_{11}) of the antenna with air cavity while completely submerged in the phantom was almost identical to the return loss while pressed against the phantom (antenna back in air). A Lego structure was used to position two antennas opposite to each other, while submerged in the phantom liquid. The Lego structure allowed for variable inter-antenna distance, with 8 mm increments. A VNA was used to measure scattering parameter S_{21} magnitude (considered the reciprocal of insertion loss) at different antenna distances. Fig. 3.15b shows a picture of the setup. A comparison was made between antennas with and without dielectric spacer (10 mm Eccostock HIK500F, $\epsilon_r = 30$). Spacer thickness was compensated for in inter-antenna distance.

For the forearm measurements, antennas were pressed directly onto the skin, at medial and lateral side of the arm. Antennas were held in place using a table vice. Measurements were performed at three positions, with varying path length. S_{11} magnitude was measured using a

VNA. Signals were inspected in the time domain and gating was applied to isolate the signal transmitted through tissue. Experimentation with absorbing foam was done to verify that no air paths were being included. Five recordings were taken for each of the three positions, with and without dielectric spacer.

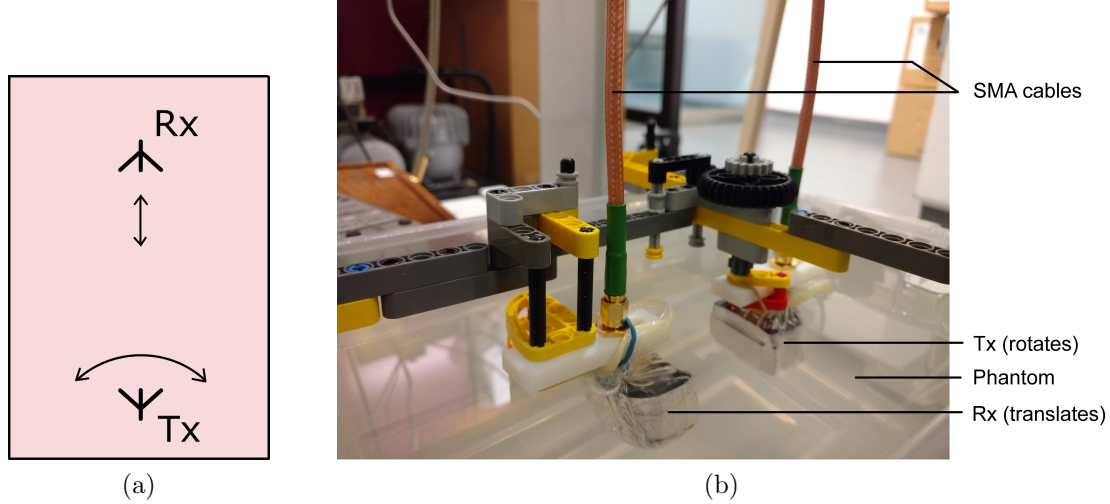


Figure 3.15: Experimental setup for measuring transmission through a liquid phantom. Antennas were held in a Lego structure, allowing for rotation of one antenna and translation (to vary antenna distance) of the other antenna (a). Picture of the setup shown in (b).

Results An example recording on the forearm at 44 mm antenna distance, comparing S21 magnitude across the frequency spectrum for the spacer versus no spacer scenario is shown in Fig. 3.16a. S21 Magnitude was determined for all recordings at 3.8 GHz and plotted in Fig. 3.16b. A linear model was fitted to each dataset to derive performance parameters. The data roughly showed linear attenuation, as expected. The y-intercept is the theoretical S21 magnitude at a range of 0 mm, which is an estimate of coupling loss. For both the forearm and phantom transmissive measurements, an improvement in coupling loss was found when adding a dielectric spacer to the body-coupled antennas, of 13 and 18 dB, respectively. Later iterations of the body-coupled antennas led to reduction in size and reduction in spacer thickness for increased array density and reduced weight. An overall transmissive signal improvement of 6 dB was found on forearm measurements [87].

Return Loss

The S11 magnitude parameter (reciprocal of return loss) of the final antenna design with dielectric spacer, while held in a 3D printed fixture as used in Chapter 4, was measured using a VNA. Antennas were placed in direct contact to various positions on the human body. Results

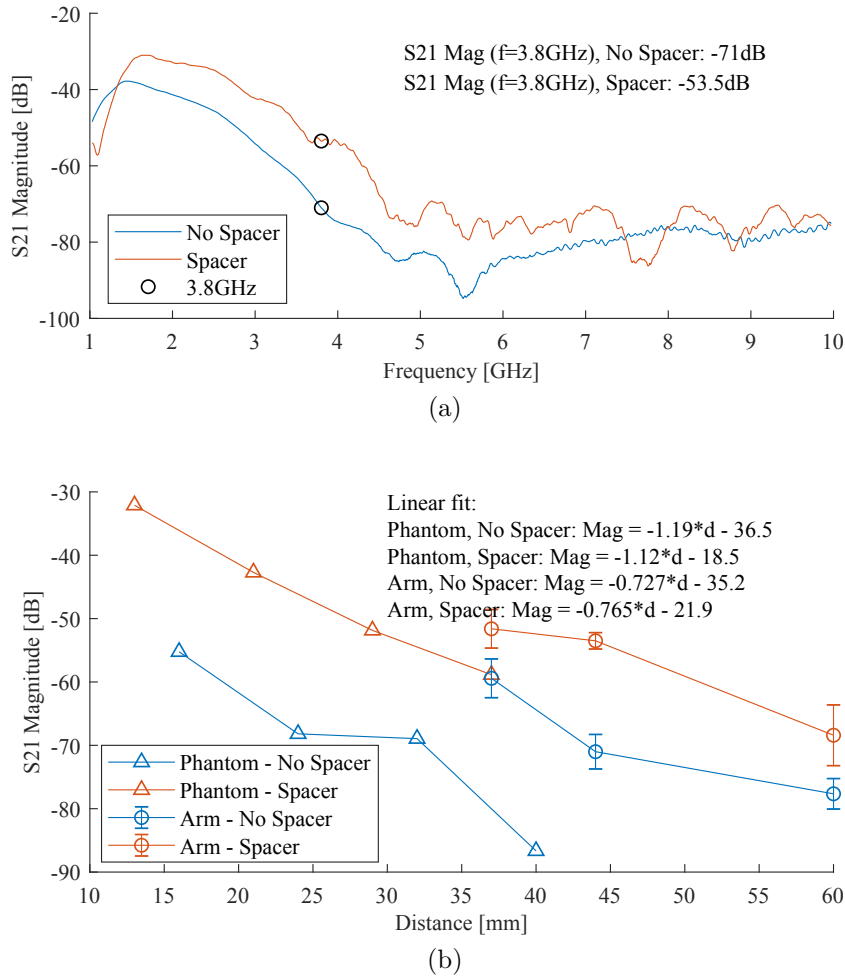


Figure 3.16: Results from transmissive measurements on liquid phantom and forearm, comparing body-coupled antennas with and without dielectric spacer. Example data from a forearm recording show reduction of insertion loss across the frequency band of interest when using a dielectric spacer (a). Data from all recordings are summarised at 3.8 GHz, and linear models are fitted to estimate coupling loss (b). A clear reduction in insertion loss is observed for the dielectric spacer scenarios.

are shown in Fig. 3.17 for a subset of four positions on the body. For all recordings, good return loss performance was found across the bandwidth of interest (roughly 2.5 – 5 GHz). It was found that positions with similar underlying tissues showed highly similar results. Best performance was observed on muscular positions (calf, bicep, forearm). Worse performance was found on bony positions (forehead, shin, sternum). Antennas were designed for a permittivity of $\epsilon_r = 30$, and the average tissue volume of skin, fat and muscle was likely to provide a better match than skin, fat and bone. An important realisation is that, although return loss is a valuable antenna performance metric, low values for S11 magnitude do not guarantee that delivered EM energy is in fact radiated into tissue.

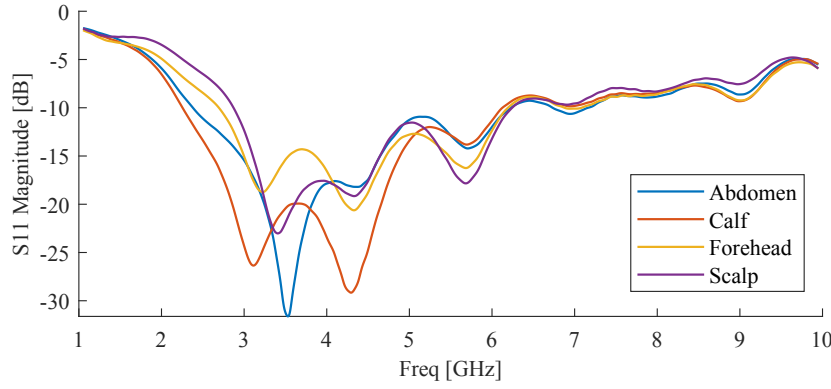


Figure 3.17: Measured S11 magnitude (reciprocal of return loss) of the final antenna design, while in direct contact with various positions on the human body.

Radiation pattern

Antenna radiation patterns are of relevance for imaging applications, where the target of interest may be at various DOAs relative to the antenna. The radiation pattern in liquid phantom is the best attainable approximation of radiation in biological tissue. For this purpose, two antennas (final design, with dielectric spacer), were positioned using a Lego structure with 3D printed antenna fixture components, and suspended in liquid phantom. Inter-antenna distance was held constant at 7 cm, which is roughly where the antenna far field (Fraunhofer) region is expected to begin. The Lego structure allowed for rotation of the Tx antenna, from -85° to 85° with 5° increments. The antenna fixture also allowed for a 90° rotation around a longitudinal z-axis, which facilitated rotation pattern measurements in the horizontal and vertical plane, as well as measurements of transmission when antennas were at a 90° offset, effectively measuring antenna polarisation. A small air cavity was maintained behind the back of the antenna to approximate on-body measurements. Conventional antenna theory dictates 10 wavelengths of air behind the antenna during characterisation. Unfortunately this is entirely impractical for a submerged antenna. Return loss measurements of antennas in contact with phantom with and without submersion showed little difference. A coordinate system relative to the antenna is illustrated in Fig. 3.18a. Definition of positive and negative rotation in the y-z plane relative to antenna is illustrated in Fig. 3.18b. Obtained radiation patterns in the H-plane (x-z plane) and E-plane (y-z plane), at five distinct frequencies, are given in Fig. 3.19 a and b, respectively. The radiation pattern in the H-plane was expected to be symmetrical. Radiation pattern measurements indicate imperfections in the (manual) manufacturing process. Transmission measured at a 90° rotational offset around the z-axis (polarisation misalignment) resulted in a reduction of signal transmission in excess of 35 dB.

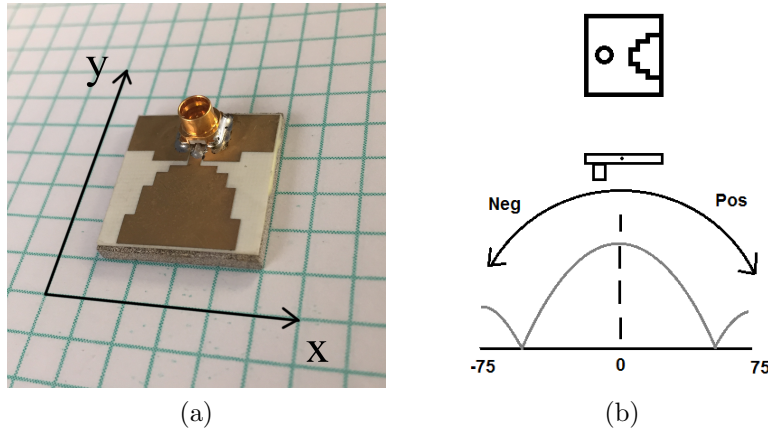


Figure 3.18: Coordinate system relative to antenna geometry for antenna radiation pattern measurements (a), and definition of positive and negative rotation in the y-z plane (b).

Coupling loss

Insertion loss is composed of signal attenuation as a result of dielectric losses, as well as losses occurring at the antenna-tissue interface. An antenna's return loss only accounts for signal energy that is not being delivered at the antenna, but does not measure the loss occurring due to radiation into air. An attempt to estimate coupling loss was made by measuring transmitted signal energy through liquid phantoms at different inter-antenna distances, and estimating the coupling loss component from the intercept of a fitted linear model. In addition, transmissive recordings were performed through media with well defined properties, so that signal attenuation due to dielectric losses could be compensated for. Various recordings were performed using the X2 RoC device. Peak-to-peak (P2P) amplitude was used as a measure of signal strength. In order to relate measured transmissive signal strength to transmitted signal strength, signal transmission of the X2 module was measured with a jumper cable

Various recordings were performed, with estimated coupling loss varying from 12 to 17 dB. A number of assumptions are being made to arrive at these estimates, and coupling loss will likely vary with exact medium dielectric properties. However a rough estimate is helpful for predicting signal amplitude in imaging applications.

3.4.5 Module synchronisation

For imaging applications with multiple simultaneously operated RoC modules, radar sampling synchronisation is paramount. By connecting Tx to Rx of various modules using jumper cables (with additional attenuation to avoid signal truncation), sampling synchronisation could be verified. Results for recordings within and between two different modules are shown in Fig. 3.20. The following must be concluded: 1) Sample delay offset (the time delay between

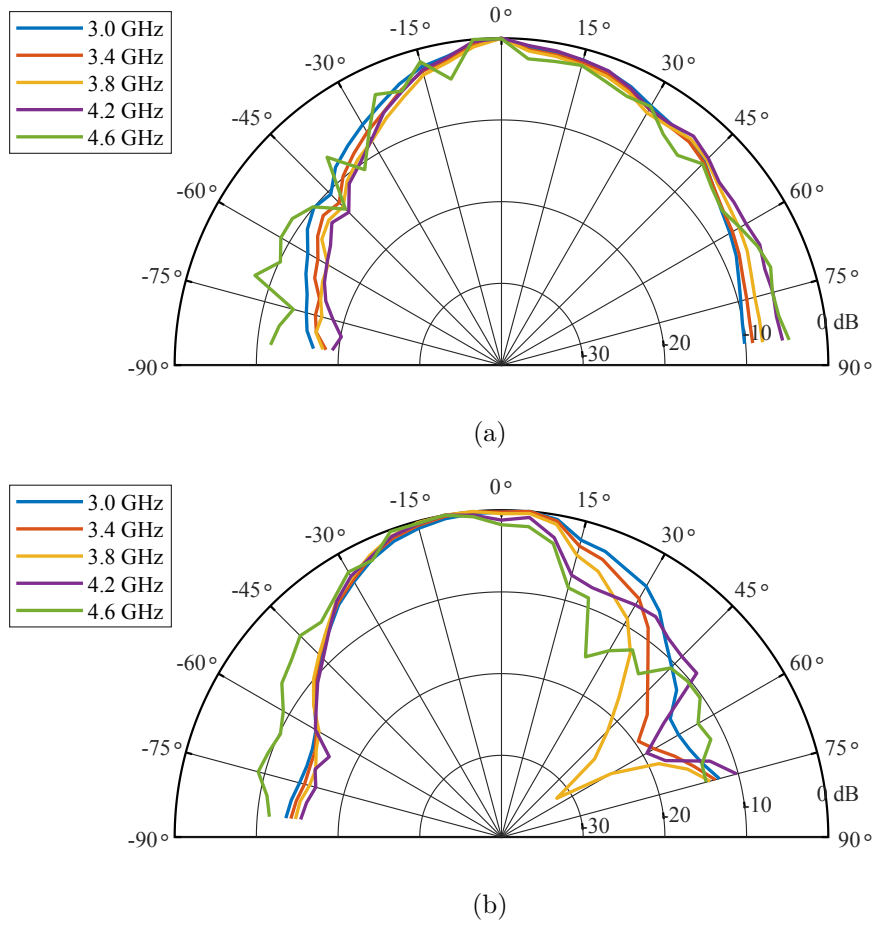


Figure 3.19: Measured radiation pattern in x-z plane (H-plane) (a) and y-z plane (E-plane) (b) for five frequencies in the band of interest.

signal transmission and sampling onset) differs between modules, and 2) module delay offset (the time delay between signal transmission on one module and sampling onset on another module) differs between modules. Both are problematic for imaging applications based on coherent signal summation, and require calibration. Module calibration is further described in Chapter 5.

3.5 Pulse propagation model

Correct estimation of EM signal propagation through biological tissues is a fundamental aspect of an imaging system based on coherent summation. The most common approach in the MWI community for compensating path dependent signal delay in DAS imaging, is the assumption of constant dielectric properties across the frequency band of interest, and pulse propagation velocity estimated by equation 2.17 (low-loss medium assumption). Phase velocity is assumed constant across frequencies and no pulse distortion occurs. If predicting (or compensating for) signal attenuation, a constant attenuation factor is applied. The microwave imaging via

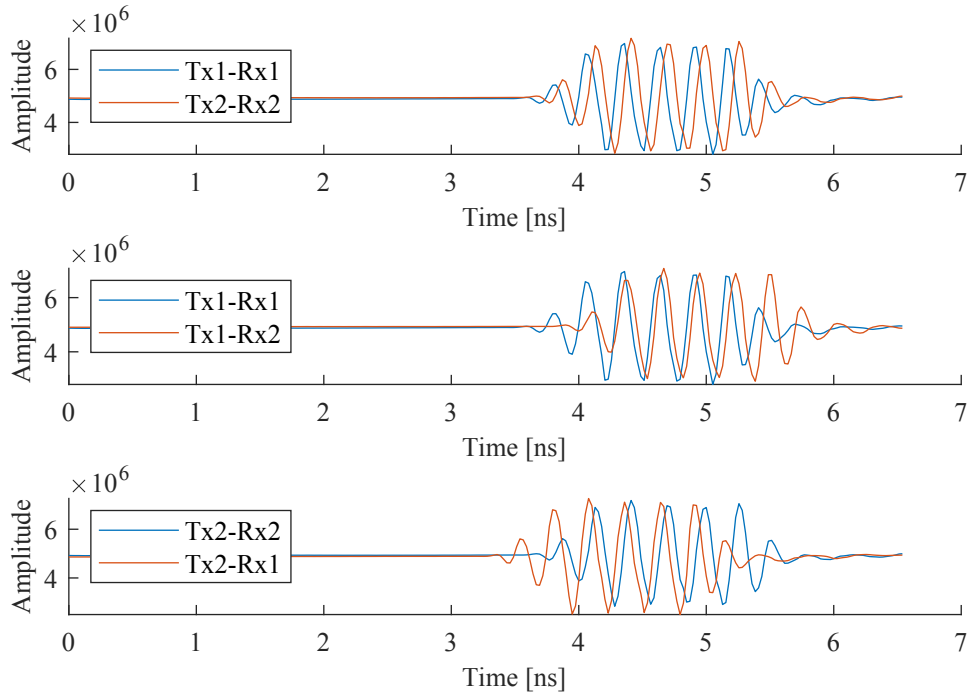


Figure 3.20: Experimentation results with module synchronisation. It is evident that both a variable within-module sample delay offset exists, as well as a variable between-modules delay offset. Both require calibration for imaging based on coherent summation.

space-time (MIST) beamforming algorithm, developed by [97], considers path length dependent dispersion and attenuation (lossy medium assumed) and uses finite impulse response filters to achieve sub-sample delay precision. However, permittivity and conductivity are assumed constant across the bandwidth of interest (media assumed non-dispersive) and evaluated at the spectral peak of the transmitted signal.

To test whether the assumption of low-loss media and constant dielectric properties are valid for the bandwidth and tissues of interest, the current section aims to compare empirical data to analytical models for RF pulse propagation. A second aim is to verify the correctness of finite-difference time-domain (FDTD) simulations, as these are used throughout this project to draw conclusions on signal processing and imaging algorithm parameters. The following exercise was performed:

- X2 signal transmission was measured through a dispersive liquid phantom, at various inter-antenna distances (ranging from 16 to 80 mm).
- Liquid phantom dielectric properties were measured using a VNA.
- The pulse measured at the smallest inter-antenna distance (16 mm) was considered the transmitted signal. Signal loss and distortion resulting from antenna coupling to medium were therefore accounted for in the transmission models.

- An FDTD simulation was run to simulate signal propagation. Phantom dielectric properties across the frequency band were modelled using a two-pole Debye model, with parameters obtained from fitting to measured data (details on the simulation methods in Appendix B).
- Analytical models for signal propagation were run to estimate signal delay and waveform at various path lengths.
- Measured pulse propagation, FDTD simulated pulse propagation and analytical model estimated pulse propagation were compared.

Two analytical models were tested:

- **Dispersive model:** The transmitted signal is decomposed into a set of discrete oscillatory components using the fast Fourier transform (FFT). A frequency-dependent phase shift and attenuation is applied in the frequency domain. Frequency-dependent behaviour was predicted from equations 2.9, 2.10 and 2.11 (lossy medium assumed). After phase shift and attenuation have been applied in the frequency domain, the signal is synthesised using the inverse FFT.
- **Constant model:** Constant group delay and linear phase response are assumed. A signal delay was accomplished by shifting the entire signal by n samples, with n determined by the path length and constant propagation velocity according to equation 2.17. Attenuation is determined according to equation 2.15. Constant permittivity and conductivity are assumed and evaluated at transmitted pulse spectrum peak amplitude.

Results of the measured and predicted signals are given in Fig. 3.21 at four different path lengths, both in the time domain (left column) and frequency domain (right column). Signal attenuation (measured as P2P amplitude) versus path length is given in Fig. 3.22. The FDTD simulation is the closest approximation of measured data, and pulse propagation is predicted almost perfectly as visible both in the time domain and frequency domain. An early high frequency component (visible at high path length) could not be predicted: This is likely a leakage signal occurring on the radar module or at the SMA connectors. The centre frequency down-shift which is observed in measured data, is perfectly captured by the dispersive analytical model. Some differences in signal attenuation occur at high path lengths, likely due to the (incorrect) assumption of unidirectional radiation of power in the analytical model. The constant model suffers from the expected inaccuracies: Attenuation at the higher end of the spectrum is reduced relative to measured data and pulse centre frequency remains constant. The constant model shows linear signal attenuation across depth, and amplitude differences start to occur at high path length. Despite these inaccuracies, it must be noted that the pulse distortion in measured and simulated data did not cause the pulse peak to shift relative to the

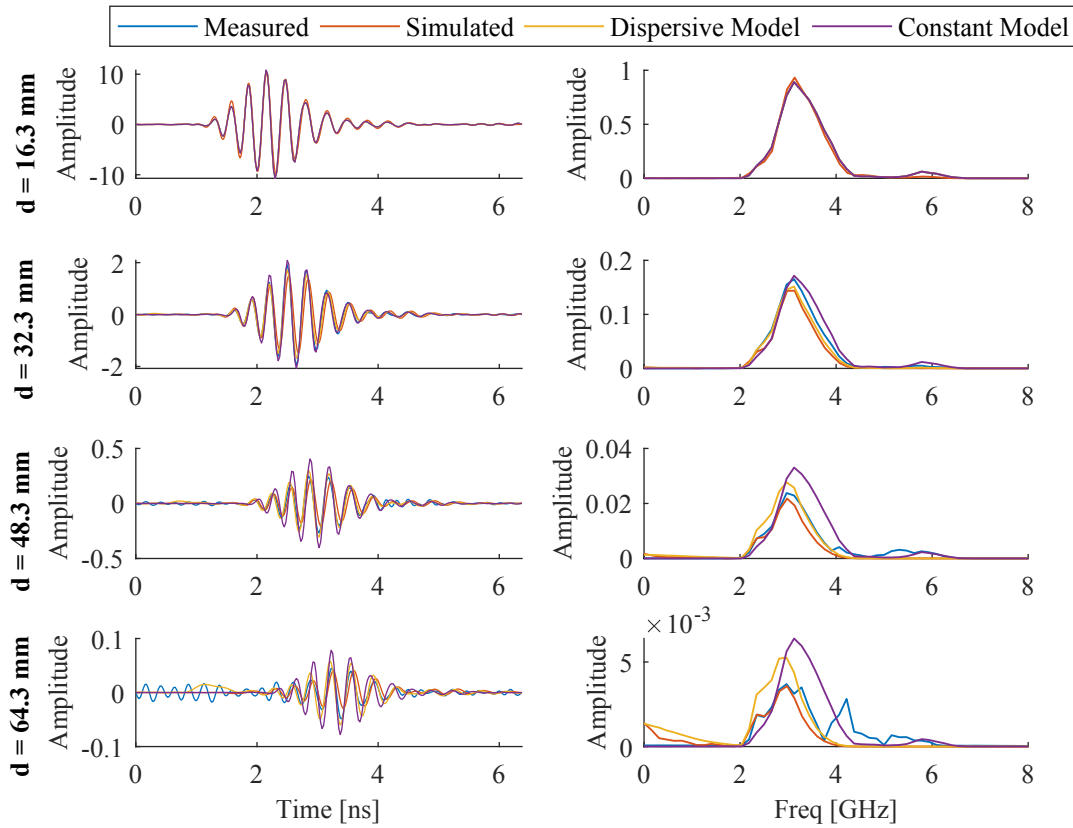


Figure 3.21: Measured pulse propagation through liquid phantom was compared to simulated pulse propagation and propagation as predicted by two analytical models. Results are given for four different path lengths, both in the time domain (left column) and in the frequency domain (right column).

constant model. This suggests that, up these path lengths, the constant model is sufficient for beamforming.

3.6 Conclusions

In the current chapter, anatomy and physiology for different imaging and sensing applications were considered. Various design parameters for a RoC based MWI system were discussed. Key hardware components for the developed system were characterised and finally, pulse propagation models were tested for their aptness in imaging based on DAS beamforming. The following conclusions can be drawn:

- System requirements will depend greatly on the imaging application. For sensing or imaging of cardiovascular structures at a pulse centre frequency of 3.8 GHz, it is likely that a dynamic range of at least 70 dB is required. Frame rates of at least 20 fps would be

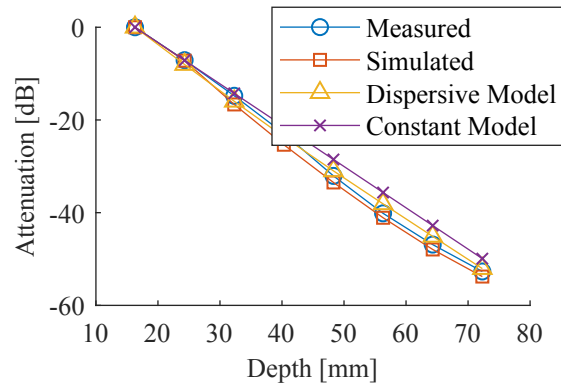


Figure 3.22: Empirically measured pulse attenuation as a result of propagation through liquid phantom was compared to pulse attenuation in simulated data and attenuation as predicted by two analytical models.

required, with desired frame rates of 50 fps. A spatial resolution in the order of centimetres would be a first acceptable step for cardiac imaging.

- Functional neuroimaging using radar is likely out of reach. Although theoretically possible, extremely high dynamic range would be required along with near perfect suppression of static clutter. Considering that there are constraints in scanning time due to the dynamic nature of the signal, it is estimated that the current state of technology (both RoC and VNA) is insufficient for functional neuroimaging.
- Hardware design considerations and the trade off associated with different system parameters are discussed. The use of simultaneously operated RoC technology offers a unique advantage of high multistatic frame rate without sacrificing sensitivity. It is clear however that there is no such thing as a free lunch, and the challenges of high signal leakage, pulse overlap, heavy dielectric losses, and limitations to array density will have to be addressed for successful imaging.
- The X2 RoC was found to have an excellent dynamic range of 105 dB and high potential frame rate. Phase drift was found to be substantial, and phase jitter is likely to cause complications when small phase variations are expected in the signal of interest.
- Phase jitter compensation could be performed, but for successful application slight hardware modifications would be required. High phase stability would likely be beneficial for dielectric spectroscopy applications.
- The developed antennas performed well in contact with the human body, and dielectric spacers were found to reduce insertion loss. Body-coupled antennas form a viable alternative to the commonly used antennas in coupling liquids.
- Sampling delay onset as well as module delay onset varies between modules. Calibration is required for imaging applications based on coherent summation.
- FDTD simulations were found to correspond to empirical data, which increased confidence of simulations as an appropriate tool for informing imaging algorithm parameters.

- Although a constant EM wave propagation model did not capture frequency dependent signal attenuation, pulse distortion effects due to dispersive propagation were found minimal at the considered path lengths and a constant propagation model is likely sufficient for DAS beamforming.

Chapter 4

In-body UWB radar sensing

4.1 Introduction

Since the FCC has authorised the unlicensed use of UWB emission between 3.1 and 10.6 GHz in 2002, radar has become a frequently researched sensing modality for non-contact health monitoring: Particularly cardiopulmonary monitoring [98–100], sleep monitoring [101], and fall detection for the elderly [102] have been investigated. Radar properties are of interest for sensing inside the body as well. Recent innovations in RoC technology [82] have made radar devices with a small form factor and low power consumption available at low cost. The current chapter investigates the use of RoC devices for biomedical sensing, specifically of the cardiovascular system. The focus is limited to in-body sensing using body-coupled antennas. The role of radar sensing within this research project was primarily to act as a precursor to RoC-based imaging, and to identify challenges and opportunities for a future imaging system. In this chapter, some of the fundamentals of radar sensing will be discussed, signal processing of in-body radar is being analysed through simulated and experimental data¹, and two experiments on in-body sensing of the cardiovascular system are being described: The first focusing on intracranial HR detection², the second on measurements of cardiovascular dynamics³.

4.2 Background

The current section explores some of the fundamentals of radar sensing in biological tissues. In addition, it aims to provide an overview of the state of the art of in-body radar sensing of the

¹Previously published [49]

²Previously published [95]

³Previously published [49, 103]

cardiovascular system.

4.2.1 Radar sensing fundamentals

Radar types

In general, radar systems are being used to detect either the absolute range or small variations in range of a target. A radar system works by transmitting EM energy and measuring the ToF of reflected signals. Reflections are caused when an EM wave impinges on a boundary between media with different dielectric properties.

CW Doppler radar Small (sub-wavelength) variations in ToF can be measured by emitting a single frequency, as done in continuous-wave (CW) Doppler radar. Variations in phase θ of the reflected signal indicate target motion, according to the following relationship:

$$\Delta\theta = \frac{4\pi\Delta x}{\lambda} \quad (4.1)$$

Where $\Delta\theta$ corresponds to the phase change (in *rad*), Δx is the reflector displacement (in *m*), and λ is the effective wavelength of the transmitted signal in the propagation medium (in *m*), as defined in Section 2.2.1. Due to the periodicity of the received signal, a CW Doppler radar is only suitable for tracking motion of up to $1/2\lambda$. This implies that a CW Doppler radar system cannot ascertain the absolute range of a reflector.

Pulsed radar A pulsed radar system emits short RF pulses. Boundaries between different media will cause signal reflections, and the ToF of reflections are used to determine absolute reflector range:

$$R = \frac{\nu \text{ToF}}{2} \quad (4.2)$$

Where R is the range of the reflector (in *m*), ν is the EM propagation velocity in the medium (in *m/s*) as defined in Section 2.2.1, and ToF is the time-of-flight from signal transmission to reception (in *s*). Accuracy in determining target range depends on the ability to accurately determine reflection arrival time.

Coherent UWB radar When the transmitted RF pulse consists of a brief oscillation with centre frequency f_c and known phase θ , one speaks of coherent UWB radar. A coherent UWB radar allows for radar pulse-Doppler processing: The range R of the target is determined from

ToF , and sub-wavelength motion Δx of the target can be tracked from $\Delta\theta$. The combination of CW Doppler sensitivity to motion, and the ranging abilities of pulsed UWB radar, makes coherent UWB radar a very attractive choice.

In-body radar sensing

The principle of in-body radar sensing is identical to radar sensing through air, with a number of additional challenges. The propagation velocity of EM waves is dependent on tissue dielectric properties: $\nu = c/\sqrt{\epsilon_r}$. The reduced propagation velocity leads to increased spatial resolution. As described in Sections 2.2.1 and 2.2.2, dielectric contrast between different tissue will cause reflections and scattering of EM pulses. As seen in Section 3.2.1, received reflections from different anatomical layers will overlap, and attenuation of the signal of interest is likely to be high. In the current work, the focus is mainly on measuring cardiovascular dynamics in the human body. This provides the advantage of dealing with a dynamic signal: The static clutter originating from non-moving structures can be removed by high-pass filtering. However, the unknown multipath environment in combination with high signal attenuation puts extra demands on the signal processing techniques applied. In addition, specific hardware is required to facilitate efficient radiation into tissue and to reduce backscattered leakage signals.

A schematic overview of a recordings setup for in-body radar sensing of the cardiovascular system is given in Fig. 4.1, illustrating a single radar module, body-coupled antennas, and the arterial pulse wave. In this case, radar is being used to detect the transient arterial dilation caused by the blood pressure wave propagating from the heart to extremities. Currently, no effort will be made to determine the depth of measured anatomical structures. The UWB property of the X2 RoC is therefore not fully utilised. However, this investigation into radar sensing serves as a precursor to radar imaging (Chapter 5), for which UWB radar is a requirement.

4.2.2 Radar sensing state of the art

Research on radar sensing inside the human body has focused primarily on the cardiovascular system. With cardiovascular disease being the number one cause of deaths globally, the search for early biomarkers of disease and affordable diagnostic tools is ongoing. Echocardiography, which uses US technology to detect motion, is frequently used to measure the mechanical functioning of the heart and arteries. However, US requires an experienced operator and equipment is relatively expensive. The ability of microwave frequencies to penetrate biological tissue allows for studying the mechanical functioning of the cardiovascular system using radar. Radar could form a relatively low-cost alternative to US, and would act as a complementary technique to bioelectric activity as measured through ECG. Thus far, radar sensing of the

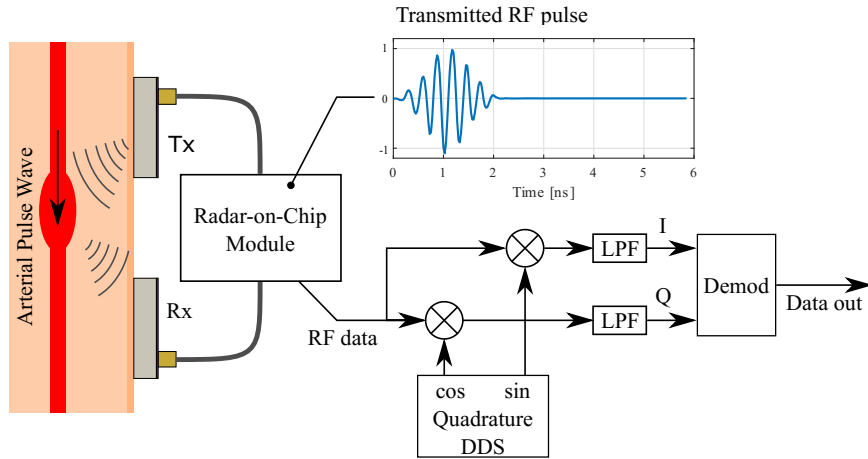


Figure 4.1: Schematic of the arterial pulse wave measurement setup, indicating the body-coupled antennas (indicated by Tx and Rx) and underlying tissue (including pulsating artery). RF data is digitised on chip, and digital down conversion (DDS) of the raw radar data is performed to obtain IQ baseband data, with subsequent lowpass filtering (LPF) and IQ demodulation. The transmitted waveform is displayed in blue. Figure from [49].

cardiovascular system has focused primarily on detection of PAT, with potential applications in arterial stiffness estimation and cuffless BP monitoring. The PAT is defined as the time at which the arterial pulse wave arrives at a specified location, following the heart contraction. The time of heart contraction is known through ECG measurement (R wave), and by combining ECG data with measurements of the arterial pulse wave, the pulse transit time (PTT) and pulse wave velocity (PWV) can be estimated. In particular PWV is of interest as a clinical marker for cardiovascular health and specifically blood pressure [104, 105].

CW Doppler radar is the most common type of radar investigated for cardiovascular measurements. High accuracy has been reported on PWV measurements using a portable UWB Doppler radar device that measured arterial pulsation from the foot and upper arm, for estimation of arterial stiffness [106]. CW Doppler radar has been proposed for blood pressure estimation from carotid PAT [48], and cardiac motion could successfully be monitored using a CW radar embedded in a patient table, for heart-cycle synchronised CT scanning [107]. Using a CW Doppler system in a hand-held device it was found that mechanical cardiac activity could be monitored from various locations near the heart [108].

Coherent UWB radar sensing

The aforementioned studies used radar sensing technology with no possibility of collecting spatial information. For future imaging applications and to take advantage of time-gating, range information is required. UWB radar, which does allow for ranging, is usually achieved through the use of bulky VNAs. A VNA was used in an attempt to estimate aortic diameter

variations, as a predictor for BP. It was concluded that measurement of PWV would be more suited for BP estimation than only aortic diameter variation [47].

Coherent UWB radar sensing using radar-on-chip technology

The recent innovation of a CMOS implementation of a coherent UWB radar [82], provides the opportunity of in-body radar imaging with extremely small and low-cost electronics. As opposed to CW Doppler radar, UWB radar allows for sensing at different depths in the body and thus for localization of scattering sources in space. This allows for time-gating of undesired scattering sources which may aid sensing from selective anatomical structures. Better spatial discrimination can be achieved through 2D or 3D imaging when multiple UWB radar sensors are combined. To date, coherent UWB RoC systems in combination with body-coupled antennas have been used to demonstrate HR detection through the back of a seat [109] and heart wall velocity sensing [110]. In this chapter, intracranial HR detection is described, as well as the measurement of arterial pulsation and cardiac dynamics at various locations in the body.

4.3 Signal processing of in-body radar data

When applying traditional radar signal processing techniques to in-body coherent UWB radar data, errors may occur due to a set of assumptions which are not valid for the in-body scenario. The current section briefly explains conventional radar signal processing techniques, and then proceeds to study the in-body case in more detail. Finally, a set of recommendations is given for processing in-body radar data⁴.

4.3.1 Conventional coherent UWB radar signal processing

Both in CW Doppler radar and in coherent UWB radar, the received RF pulse may be converted to complex in-phase and quadrature (IQ) baseband data by mixing with the carrier frequency at a 0° and 90° phase shift, and subsequently arctangent demodulating into amplitude and phase data. This approach is usually taken in contactless cardiopulmonary monitoring, as in [98]. When the received pulse amplitude is constant by approximation, for example in the case of a moving chest wall, phase data can be used to track sub-wavelength variations in ToF. Phase variation $\Delta\theta$ relates to target displacement Δx as described in equation 4.1.

⁴Current section was previously published [49]

With a moving target at distance d_0 , the outputs of the IQ baseband conversion become:

$$BB_I(t) = A \cos \left(\frac{4\pi d_0}{\lambda} + \frac{4\pi x(t)}{\lambda} \right) + DC_I \quad (4.3)$$

$$BB_Q(t) = A \sin \left(\frac{4\pi d_0}{\lambda} + \frac{4\pi x(t)}{\lambda} \right) + DC_Q \quad (4.4)$$

where BB_I and BB_Q are baseband I and Q data, A is the amplitude of received data, and $x(t)$ corresponds to the time varying scatter source location. For simplicity, residual phase noise is ignored. In addition, a digital down conversion is assumed, with no IQ imbalance or direct current (DC) offset which usually occur when performing down conversion in the analogue domain. DC_I and DC_Q here refer to the static DC component caused by overlapping received pulses from various stationary scatter sources. In a sensing scenario with multiple scatter sources, DC can be described as:

$$DC_I = \sum_{n=1}^N A_n \cos \left(\frac{4\pi d_n}{\lambda} \right) \quad (4.5)$$

$$DC_Q = \sum_{n=1}^N A_n \sin \left(\frac{4\pi d_n}{\lambda} \right) \quad (4.6)$$

where N is the number of static scatter sources, A_n is the amplitude of signal reflected from scatter source n , and d_n is the distance of scatter source n . In order to use phase data for monitoring of target motion, A must be constant by approximation, and the DC component must be compensated for. In the case of non-contact monitoring of the cardiopulmonary system, DC_I and DC_Q can be found and subtracted by fitting a circle or ellipse to received data in the IQ plane [100]. After DC compensation, received data of a moving target will describe an arc with radius A around the origin in the complex IQ plane. ToF variation is then derived from the phase angle θ as:

$$\theta = \tan^{-1} \left(\frac{BB_Q}{BB_I} \right) \quad (4.7)$$

To illustrate DC offset removal by ellipse fitting, respiration data has been recorded using a coherent UWB radar with 7.29 GHz centre frequency (XeThru X4, Novelda AS, Kviteseid Norway). A volunteer was seated in a chair, while radar measurements were being taken at a range of 1 m. RF data were converted to baseband IQ data, and the obtained phase signal was calculated using the arctangent method. Due to the presence of a DC component, arctangent demodulation did not result in a successfully extracted respiration signal (Fig. 4.2a and b). However, after ellipse fitting and transformation of the ellipsoid data to circular data centred around the origin, the arctangent assumption held true and a respiration waveform could be

obtained (Fig. 4.2c and d).

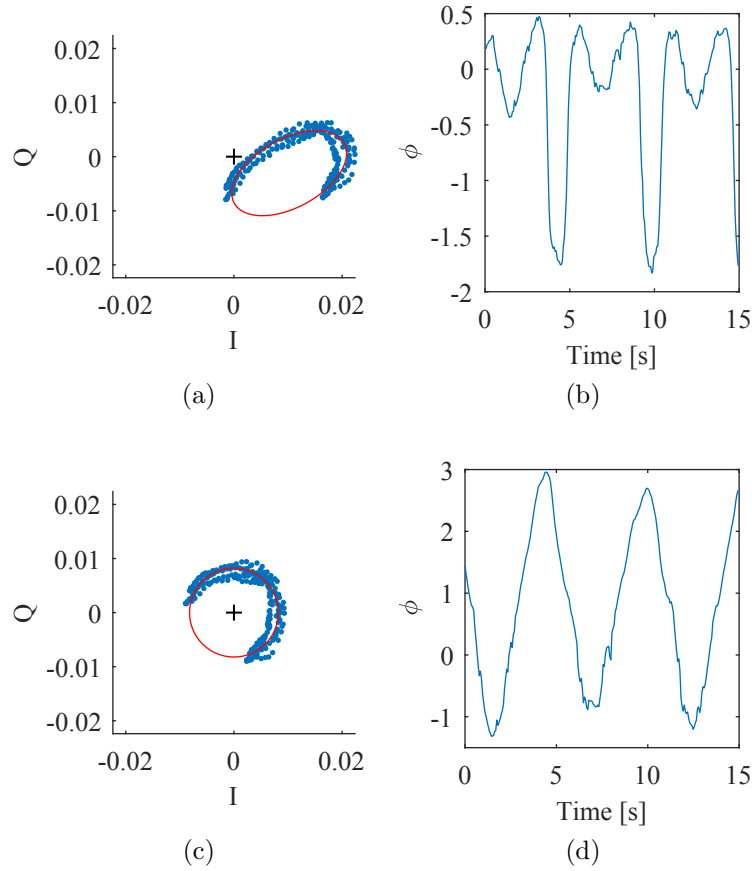


Figure 4.2: DC component removal of respiration data using conventional radar signal processing: IQ baseband data (a) and phase data (b) before ellipse fitting and DC removal, and IQ baseband data (c) and phase data (d) after DC removal.

4.3.2 Challenges specific to in-body radar signal processing

When performing in-body sensing of an artery however, the assumptions made for arctangent demodulation of IQ data are generally not valid. Arterial pulsation, effectively a transient displacement of tissue boundaries as well as a temporary increase in artery diameter, will interact with EM radiation at microwave frequencies in the following ways: The tissue boundary displacement will cause a slight decrease in ToF of received RF pulses, and the increased target diameter will cause a variation in received pulse amplitude.

Arctangent demodulation of IQ data When attempting arctangent demodulation on arterial pulse IQ data, a problem arises: Due to overlap and interference of strong static clutter signals DC_I and DC_Q (reflections from multiple tissue boundaries, as was found in Section 3.2.1) and the relatively weak dynamic signal of interest (51.37 dB dielectric loss at 3.8 GHz was

estimated for an intracranial volume of blood), a strong DC offset will be present, which must be compensated for when using arctangent demodulation. The circle or ellipse fitting technique (as in [100]) will fail for the following reasons: First, the artery diameter is small relative to carrier wavelength. The brachial diastolic artery diameter is around 4.2 mm [111], whereas the larger carotid and femoral arteries have diastolic diameters of 6.3 [112] and 9 mm [113], respectively. Considering a carrier wavelength of around 11 mm in tissue (muscle at 3.8 GHz), Mie scattering will occur and amplitude of received scattered data will vary with artery diameter. Thus, A in equations 4.3 and 4.4 is time varying: $A(t)$, which complicates circle fitting for DC offset removal. Secondly, the displacement of tissue due to artery expansion Δx is very small relative to wavelength λ . Radial expansion of the brachial, carotid and femoral artery have been measured at 3.5, 13, and 8% respectively [111–113]. Considering the carrier wavelength in tissue, the expected phase variations $\Delta\theta$ in ToF according to equation 4.1 are 0.07, 0.50 and 0.41 rad respectively. The arc length is therefore far too small for circle fitting and DC estimation. A final complication is that the increased absorption of higher frequencies leads to pulse distortion, specifically centre frequency down shifting. Reflected waveforms with a higher path length will have lower centre frequencies than signals reflected off superficial structures or crosstalk leakage. Baseband conversion applied to mixed frequency signals has shown incorrect arctangent demodulation in simulated data.

Simulated data To explore the above issues, the open-source Python-based FDTD simulation software gprMax [114] was used to simulate a simplified limb (composed of skin, fat, and muscle tissue) with a round object inside (simulating an artery filled with blood). The dielectric properties of the different tissues were modelled using a two-pole Debye model. Debye model parameters were obtained by fitting to known tissue dielectric properties (as described in Appendix B, with dielectric properties obtained from [18]). The artery diameter was step-wise increased from 0.4 mm to 10 mm, to study EM scattering across a potential range of artery diameters. A simulated transmitter was excited using a waveform identical to the one used in the here presented experimental work. Note that for the simulated scenario, DC data were simulated as an absence of the artery, and could entirely be removed. Simulated received RF data were converted to complex baseband data and shown in the IQ plane in Fig. 4.3a. In blue, red and yellow, the range of arterial expansion as a result of the heart beat is indicated for three arteries: Brachial, carotid and femoral, respectively. Neither of the three segments are long enough for circle fitting. The amplitude signal, visible as the radius in Fig. 4.3a or plotted versus artery diameter in Fig. 4.3b (top) after arctangent demodulation is not a constant, but shows a typical Mie scattering pattern. Considering the inability to correctly estimate DC offset in an experimental setting, arctangent demodulation of IQ data proves to be unsuitable.

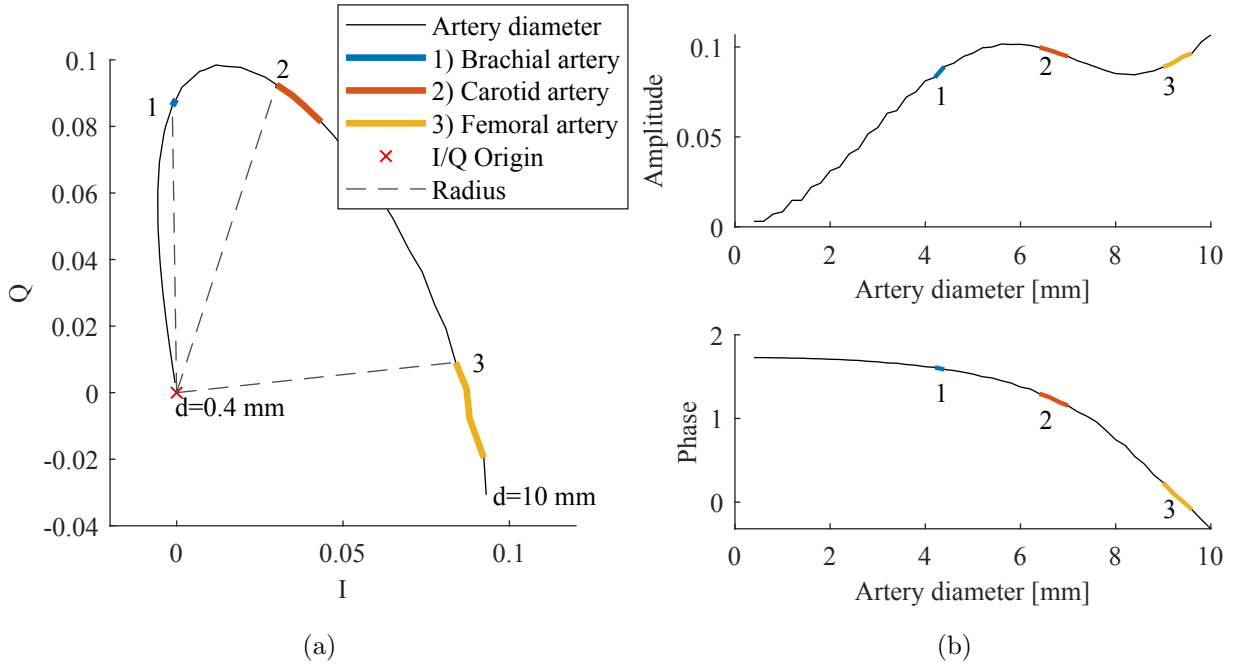


Figure 4.3: Simulated data illustrating the challenges of demodulating IQ data from in-body recordings. From simulating a theoretical range of artery diameter (0.4 – 10 mm), it becomes apparent that the arc length due to arterial pulse is too short for ellipse fitting (a). In addition, amplitude of reflected signals (obtained using arctangent demodulation) show a Mie scattering pattern, which further complicates ellipse fitting (b top). Phase data (b bottom) could be obtained because DC was known and subtracted. For real-world data, DC is unknown and arctangent demodulation might not be valid. The coloured segments in (a) and (b) indicate the diameter ranges due to pulse expansion for three arteries.

Alternative approaches DC cannot be removed prior to arctangent demodulation by high-pass filtering across slow time, as it removes factor A in equations 4.3 and 4.4 entirely, rendering correct arctangent demodulation impossible. Potential alternatives to baseband signal processing are tracking of amplitude at individual samples, or tracking of zero crossings. Both approaches have the disadvantage of not allowing for straightforward integration of coherent movement across the pulse length, and have been found to suffer strongly from unpredictable behaviour with interference from static signals with a different centre frequency. An alternative that has been applied in non-contact HR monitoring using radar, is linear demodulation of IQ data [99]. When the arc length is sufficiently short, obtained IQ data can be described by a linear function with little error. By approximation, variation along the linear function is representative of variation in pulse phase $\Delta\theta$.

4.3.3 In-body radar signal processing recommendations

When received pulse amplitude is not likely to be constant, and arc length is short due to small ΔToF relative to λ , linear demodulation of IQ data is recommended as a method to obtain a valid metric for Δx . Visualisation of high-pass filtered movement in the IQ plane is recommended to assess the suitability of linear fitting. Slow time high-pass filtering of RF data destructs the carrier waveform and leads to unpredictable down conversion, therefore filtering should be applied to I and Q data independently, prior to demodulation. Linear demodulation can be performed using linear regression, or using principal component analysis (PCA). If PCA is being used, the second principal component may be analysed in addition to the first component, as it likely reflects changes in pulse amplitude which may be of interest.

Fig. 4.4a shows experimental in-body arterial pulsation data, recorded from the femoral artery using a coherent UWB radar with 3.8 GHz centre frequency (modified XeThru X2, Novelda AS, Kviteseid Norway; experimental methods described in Section 4.5.2). It is evident that no circle can successfully be fitted to the small arc length, and variation in the data does not occur along the arc of a circle centred around the origin. In this measurement scenario, arctangent demodulation would be incorrect and the obtained phase data would account for only a fraction of the observed signal variation. In fact, due to the DC offset being unknown, using the origin of the IQ plane for arctangent demodulation would be entirely arbitrary. Instead, a linear function is fitted using PCA. The first principal component (PC1) captures most of the variation in the data and is thus a better measure for ToF variation in case of short arc length. Fig. 4.4b shows the obtained arterial pulsation waveform. A challenge that arises due to unknown DC component that cannot be overcome, is the polarity of the PC1 signal: The polarity of PCA components is arbitrary and it cannot be determined what the direction of $\Delta\theta$ is.

4.4 Intracranial heart rate detection

4.4.1 Introduction

With the ultimate goal of cardiovascular imaging using UWB RoC devices in mind, an initial verification was sought that arterial pulsation could be measured inside the body. Specifically, the vision at the time of the design of the experiment was brain imaging using UWB radar. For this reason, an attempt was made to detect intracranial HR. Demonstrating that HR could be measured from inside the head using RoC technology would serve as a first step in further imaging of cerebrovascular activity. Intracranial blood flow is commonly measured using transcranial Doppler US. Main targets are the cerebral basal arteries. However, there has been no previous report of detecting the HR intracranially using microwave techniques.

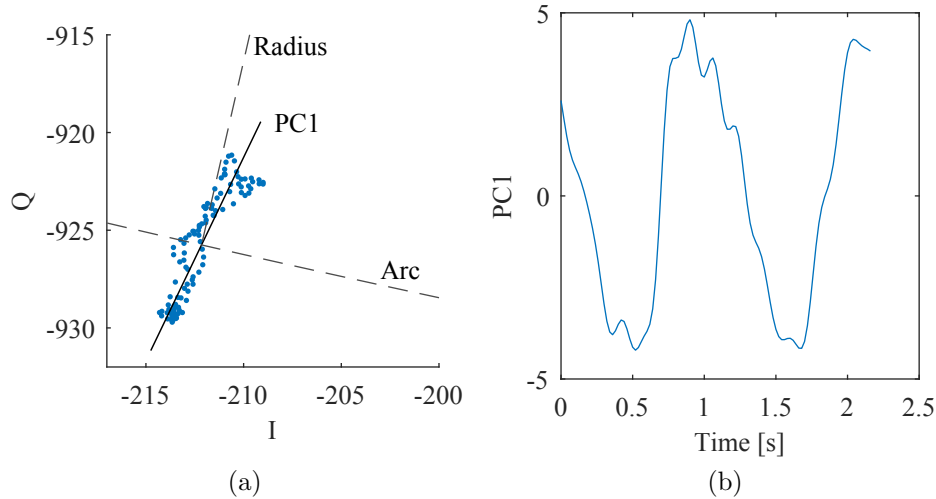


Figure 4.4: Experimental in-body data of femoral artery pulsation. Short arc length prevents DC offset removal, therefore arctangent demodulation assumptions are invalid (a). Linear demodulation results in the correct waveform (b).

The work described in the current section investigates whether intracranial HR can successfully be detected using an UWB single chip radar and body coupled antennas. Experiments were performed on a human participant. Transmissive radar signals were recorded from various antenna placements on the forehead and side of the head. ECG data were recorded simultaneously. The received radar data were analysed in the frequency domain and compared to ECG measurements. Simulations of the measurement scenario were used to predict the ToF of the signal of interest, to validate that the measured HR signal originated from within the skull⁵.

4.4.2 Methods

The experimental setup is illustrated in Fig. 4.5 with hardware consisting of a radar module⁶ (blue) and an ECG device (green). A digital interface between the radar module and data acquisition computer (red) controlling the radar module and two body-coupled wideband antennas⁷ located on the forehead.

Experimental setup

The Ventricorder radar module (described in detail in Section 3.4.3) was used for intracranial sensing. The module consisted of the XeThru X2 RoC (Novelda AS, Kviteseid, Norway),

⁵Current section was previously published [95]

⁶Radar module was developed by Dr. Kristian Kjelgård (UiO) and Novelda AS, Norway.

⁷Antennas were developed by Mathias Tømmer (UiO).

⁹Original figure was created by Mathias Tømmer (UiO).

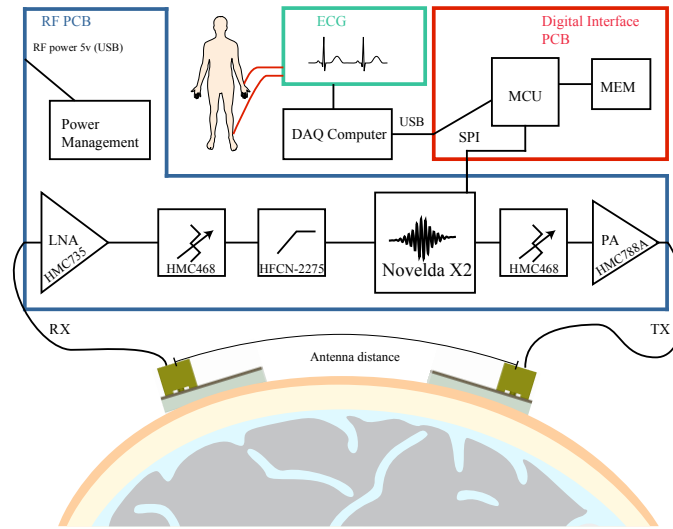


Figure 4.5: Block diagram of the experimental setup for intracranial heart rate detection⁹. Figure from [95].

with reduced supply voltage to achieve a 3.8 GHz centre frequency with 2.5 GHz -10 dB signal bandwidth. The X2 chip was designed to operate with radiation emission levels in compliance with FCC regulations (-41.3 dBm/MHz). To compensate for supply voltage reduction and achieve higher SNR, additional amplifiers were added on both the Tx and Rx paths (Fig. 4.5). The transmitted power level was increased by 8 dB.

The antennas used in the experiment are body coupled wideband monopole antennas with a -10 dB bandwidth of 2.5 – 7 GHz, with the antennas placed on the forehead. Using a low loss dielectric spacer with permittivity matching that of the skin ($\epsilon_r \approx 30$), increased body coupling and reduced performance sensitivity to underlying tissues were achieved. Details on the design, simulations and measurements of the antenna are given in [87]. More details and characterisation of radar modules and antennas are given in Section 3.4.

The ECG setup consisted of an Arduino with Olimex EKG shield and Olimex electrodes, controlled through Python. ECG data were recorded at 250 Hz, from the left to right wrist, with right leg drive.

Experimental protocol

Measurements were done on a single male participant, 26 years old, with a body mass index (BMI) of 20.5. Participant risk was carefully considered but none was perceived. General safety considerations of using the current module for in-body measurements are detailed in Appendix F. Even while using the most conservative assumptions, SAR safety limits as set forth by the ICNIRP would not be exceeded. Simulation studies using similar radar technologies have been published. In a study that considered an implanted transmitter (hence 100 % energy

absorption), it was found that 42.2 dB gain (relative to FCC mask) could be added to a highly similar pulsed UWB transmitter before SAR safety limits would be reached [115].

Recordings were performed from four different antenna positions on the forehead and side of the head. Using an elastic band the two antennas were pressed against the skin with the following inter-antenna distances: 60, 80, 100 and 200 mm. For the 200 mm scenario the antennas were positioned anterior to the ear and superior to the zygomatic arch (cheek bone), which is known as an acoustic window in transcranial Doppler US, where the skull thickness is sufficiently thin to image some of the basal cerebral arteries [116].

Data analysis

The X2 radar receiver consists of 256 consecutive samplers with an inter-sampler delay of 25 ps, resulting in a fast-time sampling frequency of 40 GHz. Pulses are transmitted at a pulse repetition frequency of 100 MHz and the receiver integrates each sampler to produce a slow-time frame rate at 80 fps. In fast-time the received signal is a superposition of pulses propagating through multiple paths between the Tx and Rx antennas, and due to the complexity of the cerebrovascular system no strong hypothesis on the effect of cardiac activity on the received radar signals in slow-time existed. Cardiac activity causing arterial dilation inside the brain could result in movement of cerebral tissue. Such a pulsating movement can be observed at the anterior fontanelle, a gap in an infants skull often called the soft spot. A movement of dielectric boundaries within the cranium would give rise to periodic changes in the received radar signal, both in terms of amplitude and phase. Another hypothesis would be that periodic changes in blood volume within the brain result in a periodic change in dielectric permittivity, causing a change in ToF (and therefore phase) of the transmitted signal.

The first step in analysing the radar data, is to determine the samplers of interest. That is, the samplers that correspond to a delay in fast-time, at which the signal of interest (path through the brain) can be expected. For this, the HR frequency was determined from ECG data. Next, the slow-time amplitude spectrum was calculated for each sampler, using the chirp Z-transform for increased frequency resolution. The amplitude at the HR frequency was obtained from the sampler's amplitude spectra to find at which samplers the HR was most prominent. HR amplitude was then plotted across fast-time. By visual inspection, the latest prominent peak was identified as the window containing the signal of interest. 40 Consecutive samplers within this window were chosen as samplers of interest.

Two approaches were then used to analyse the received radar data at the samplers of interest; average amplitude variations in slow-time and phase variations of the carrier frequency in fast-time. Amplitude and phase metrics as obtained from arctangent demodulation of frequency

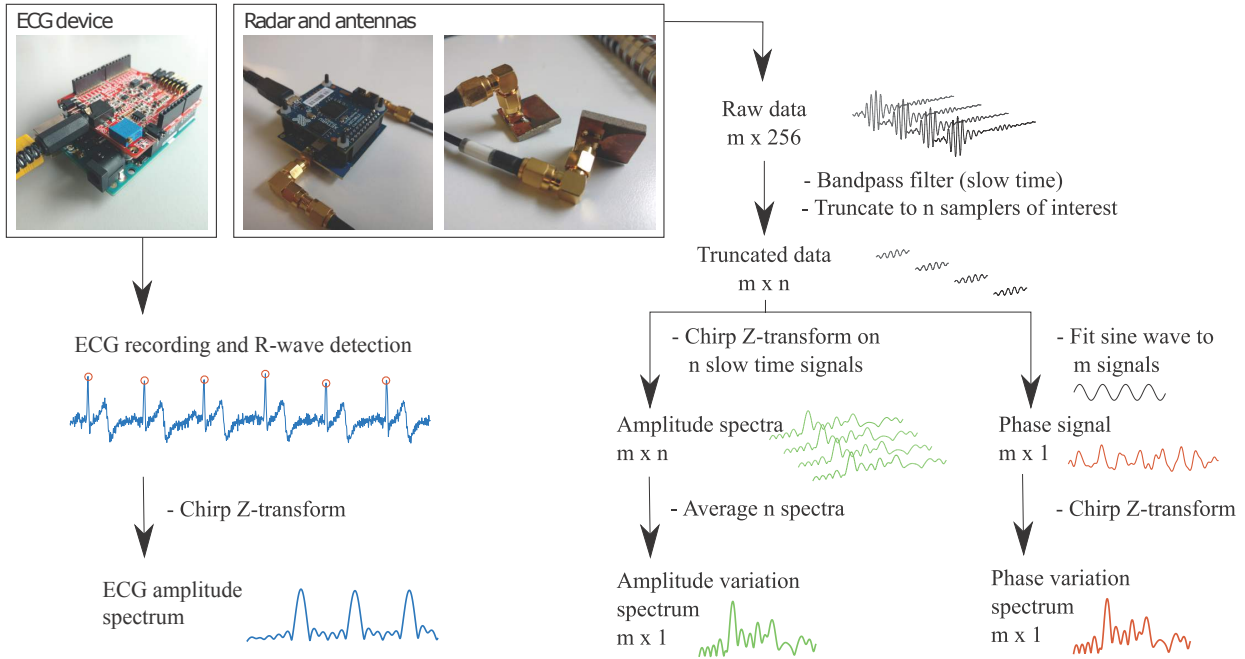


Figure 4.6: Signal processing pipeline: The amplitude spectra of the amplitude variation (green) and phase variation (red) spectra are determined from raw radar data, at a set of samplers of interest. An amplitude spectrum is obtained from ECG data (blue).

domain data suffered from the issues described in Section 4.3.2. Because no time-series data of the vascular signal were analysed, but only frequency analysis was performed, the error resulting from arctangent demodulation was deemed acceptable. A detailed view of the processing scheme is illustrated in Fig. 4.6. The slow-time data were bandpass filtered using a zero-phase shift filter with a pass band of 0.8 – 4 Hz. The bandpass filtering removes strong static signals and high frequency noise, leaving in place only the signal that could physiologically be explained by cardiac activity. The slow-time amplitude spectrum was calculated using the chirp Z-transform for each sampler of interest, followed by averaging over the resulting spectra, yielding an average amplitude spectrum. To study the phase variation a segment of the radar signal was considered as a carrier (single frequency), and its phase was determined as in CW Doppler radar. For each slow-time frame a window containing the samplers of interest was selected. Within this window, the relative phase of the carrier frequency was obtained by computing the arctangent of the inner product between the recorded data and a sine wave with a 3.8 GHz centre frequency and fixed phase. After unwrapping the resulting phase signal, the chirp Z-transform was used to obtain the amplitude spectrum of the phase variation of the data.

The amplitude spectrum of the ECG data was calculated, and compared to the amplitude spectra from the amplitude variation and the phase variation of radar data. Because the ECG signal was found to be very noisy, R wave detection was performed and the heartbeat signal

was used for frequency analysis instead of raw ECG data.

Time-domain simulations

To validate that ToF of HR modulated pulses corresponded to propagation through the head, and were not in fact a backscattered leakage signal, EM simulations were performed. A simulation of one of the measured scenarios (200 mm inter-antenna distance) was carried out using a simplified multilayered tissue model of the subject's forehead with dimensions based on the distance between temples and the arc length over the forehead¹⁰. The tissue model consisted of a layer of skin (4 mm), cortical and cancellous bone (1.45 and 3.45 mm), and a heterogeneous mixture of gray matter, white matter, cerebrospinal fluid and blood, where layer thickness data were gathered from multiple sources and the dielectric properties from [16]. The transient solver in Ansys HFSS was used and the Tx antenna was excited with the same pulse as the one transmitted by the radar module. To analyse contributions from different propagation paths (air, skin, skull, brain), half distance field monitors were placed in different layers of the geometry. The simulation geometry is illustrated in Fig. 4.7a.

4.4.3 Results

Fig. 4.7b shows the measured HR frequency amplitude as a function of time delay for the 200 mm inter-antenna distance, as well as results from the time-domain simulation of the 200 mm scenario, with the envelope of the signals added for clarity. The simulated received signal (Fig. 4.7b, middle) is a superposition of multiple paths that the radar pulse propagates through: The low velocity, high attenuation path directly through the brain, the high velocity path over air, and several other smaller contributions from the skin and skull layers of the model. The initial hypothesis was that the first signal component (1.5 – 3 ns) mainly constitutes the air wave, followed by signal contributions from paths through the tissue. To verify that the pulse arriving between 4 – 4.5 ns is the path going directly through the brain, field monitors were placed in the middle of the different layers (at half the total path length) of the simulation model. This way the multiple contributions seen at the receiver were isolated, making it possible to predict the delay of individual paths. The amplitude measured at the field monitors in the brain and in air are shown in Fig. 4.7b, bottom. Note that the range of the time axis is half that of the top and middle figure, as only half of the path length was considered. Comparing these simulation results to measurement results (Fig. 4.7b, top), it was found that all three graphs predict two distinct peaks. The measured HR amplitude shows a strong peak at 4 – 5 ns, which corresponds well to the simulated brain wave in the field monitors, but is slightly later than

¹⁰Simulations were performed by Mathias Tømmer (UiO).

the second major component in the received signal. The earlier peak in HR amplitude (2.5 – 3.5 ns) appears to be well explained by the air wave in simulation results.

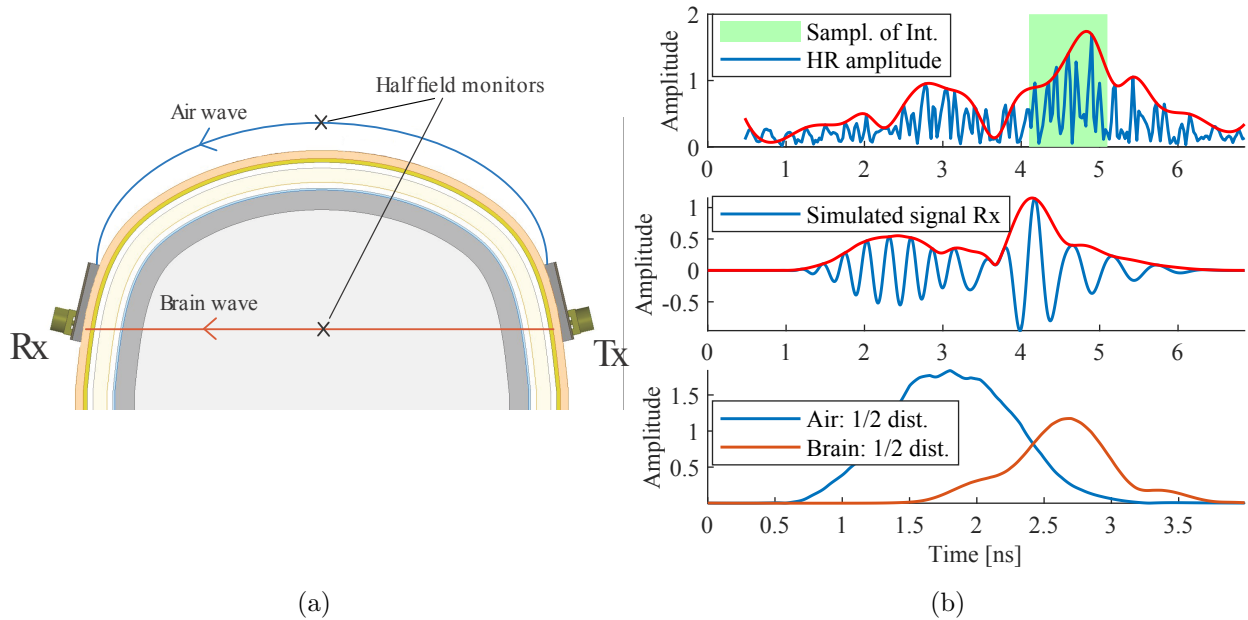


Figure 4.7: Simulation geometry indicating EM wave propagation through the brain (red) and through air (blue) and half field monitors (a). Experimental heart rate amplitude data (b, top) and simulated static signal amplitude data (b, middle) are plotted versus fast-time, with envelopes in red. Simulated signal amplitude envelopes at half distance field monitors are shown for the brain and air paths (b, bottom).

Repeating the measurement with the antennas separated with a distance of 60, 80, and 100 mm, the HR amplitude as a function of time delay was calculated to estimate the samplers of interest, which were required for further analysis. Results are shown in Fig. 4.8 with the samplers of interest marked in green (data at 80 mm inter-antenna distance not shown). Obtained amplitude spectra from radar amplitude variation and radar phase variation at samplers of interest are given for different antenna positions in Fig. 4.9, along with the measured ECG spectra. In general, there is excellent correspondence in dominant frequencies between radar and ECG data, with the phase variation spectrum at antenna distance of 60 mm being the only exception. Strong harmonics in the ECG spectra are present due to the use of a heart beat signal instead of raw ECG data.

4.4.4 Discussion

The results suggest that it is possible to measure intracranial HR using a UWB RoC in combination with body coupled antennas. With the antennas located on the temples, the received

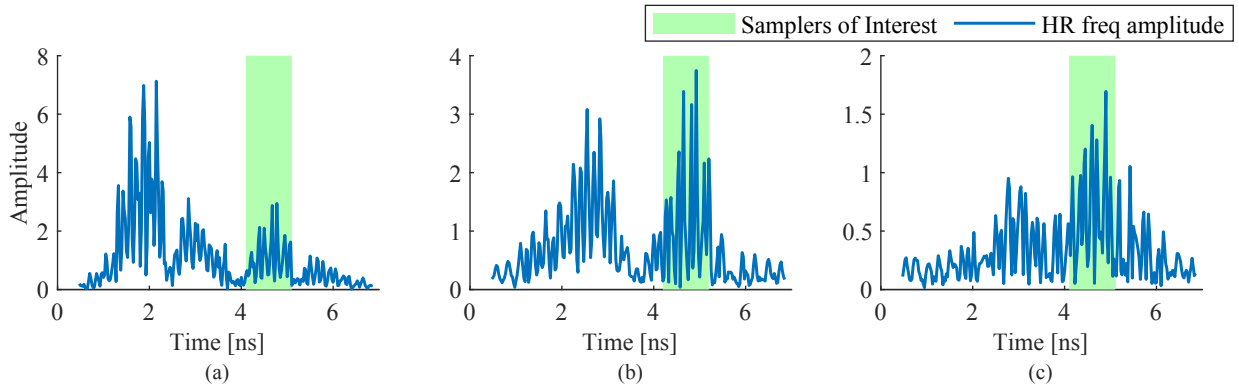


Figure 4.8: Amplitude of the heart rate frequency (~ 1.2 Hz) from radar data, for antenna distances of: (a) 60 mm; (b) 100 mm; and (c) 200 mm. The 40 samplers of interest are marked in green, and correspond to the latest component where the heart rate frequency is prominent.

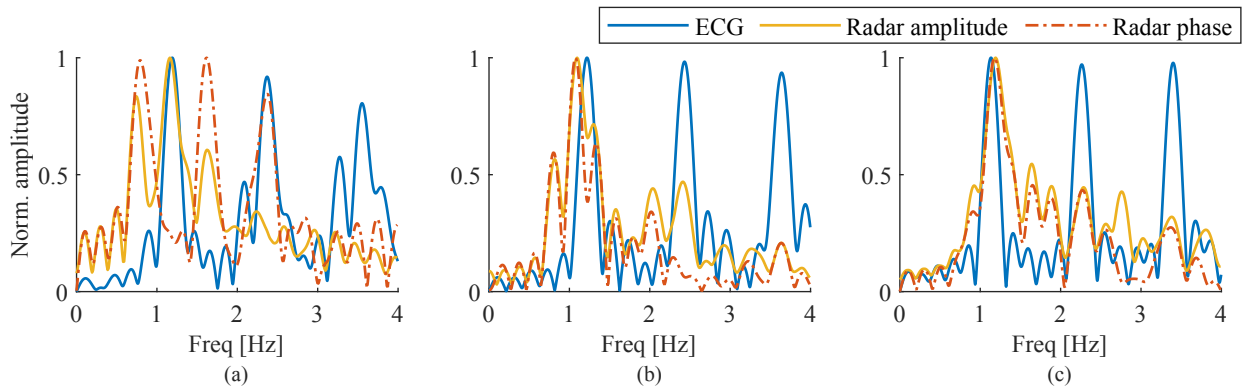


Figure 4.9: Normalised amplitude spectra of ECG, radar amplitude variation and radar phase variation for antenna distances of: (a) 60 mm; (b) 100 mm; and (c) 200 mm. With one exception, all data show a dominant peak at the heart rate frequency (~ 1.2 Hz), and excellent correspondence between the ECG and radar data. Strong harmonics in the ECG spectra are present due to the use of a heart beat signal instead of raw ECG data.

signal passing through the brain gets both amplitude and phase modulated by cardiac activity in slow-time. Both effects occurred at HR frequency, as validated by ECG. In the current experiment, two different analyses were applied on the raw radar data, based on different hypotheses of how cardiac activity influences the radar signal. Although both methods were shown to be effective, Section 4.3.2 has pointed out the assumptions taken for arctangent demodulation are not valid. The derived amplitude and phase signals may show temporal variation at the HR frequency, but the method is not necessarily robust (depending on arbitrary DC offset position) and the time-series are unlikely to be representative of arterial dilation.

The simulated data indicate that the strongest peak in Fig. 4.7b likely originates from transmission through the brain. The earlier observed peaks can most likely be attributed to the path through air, which has a higher propagation velocity. As can be observed in Fig. 4.8, changing the inter-antenna distance increases the path length through air, thus the time delay of the first

component. The hypothesised brain signal seems to occur at a consistent delay, which implies that in case of the 60, 80 and 100 mm antenna distance, reflected energy is being measured.

A question that remains unanswered is how the air wave gets modulated at the HR frequency. One possible explanation is that parts of the air wave creep into the skin. The skin contains arteries, which could cause both skin movement and a change in blood fraction, changing the dielectric permittivity of the skin. Through interference of the path through air and paths propagating partially through tissue, the air wave could be affected by cardiac activity.

A remaining concern is that leakage signals in the recording system are higher than anticipated. Backscattered signal leakage was not quantified in Section 3.4, but signal modulation by nearby motion has been observed in experiments using body-coupled antennas on tissue and phantom materials. The current experiment was not performed in an anechoic chamber for practical reasons. After conclusion of the experiments, it was observed that motion in the environment of the radar could be picked up in the received signal. This modulation of leakage signal is concerning, because the chest is known to show a slight (sub-mm) displacement as a result of the heartbeat. It was concluded that further experimentation was required to provide conclusive evidence that intracranial cerebrovascular activity was being measured.

4.5 Sensing cardiovascular dynamics

4.5.1 Introduction

In Section 4.4 it was demonstrated that the HR signal recorded from the scalp, likely originated from modulation of transmitted radar signals by pulsation of the cerebral vasculature. Simulated data and ToF analysis support this hypothesis. To provide conclusive evidence of the source of recorded HR signals, data must be studied in the time domain by measuring arterial PAT.

The work described in the current section aims to investigate the use of coherent UWB RoC technology for monitoring the cardiovascular system in the human body. The arterial pulse wave is recorded at various locations in the body, on four participants. Antennas with improved shielding have been used, and improved signal processing methods were applied to extract a robust and valid metric for arterial pulse. By assessing the arterial pulse dynamics, the source of UWB radar measured HR signals is being verified. In addition, propagation of the blood pulse through arteries is demonstrated, as well as precise PAT measurements. Finally, it is shown that ventricular volume and heart motion can be sensed using radar¹¹.

¹¹Current section was previously published [49, 103]

Experiments focus on 1D sensing of the cardiovascular system using a single radar module. Research that contributes to the development of low-cost and portable diagnostic tools for assessing cardiovascular function needs little justification. However, this work primarily serves as a preparation for 2D and 3D imaging using an array of UWB RoC devices.

4.5.2 Methods

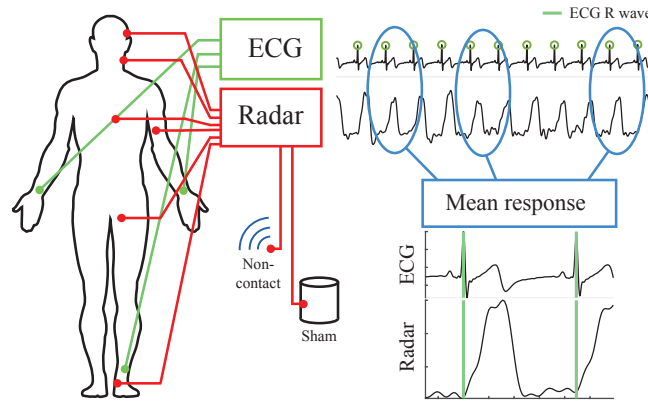


Figure 4.10: Schematic of experimental setup illustrating the radar and ECG measurement positions. Radar measurements included the neck, head, upper arm, lower back, chest, thigh and foot, as well as two off-body control measurements. Recorded radar data were synchronised to ECG data and multiple heart cycles were averaged to study the temporal characteristics of the arterial pulse wave and cardiac motion.

In-body recordings were performed using a coherent UWB RoC module^{12,13} and body coupled antennas¹⁴, on four participants. Recordings were made at various locations on the body and synchronised to ECG data. Radar data were analysed in the time-domain to validate that measured HR modulation originated from a pulsating subcutaneous structure (in most cases: an artery), and was not an artefact caused by leakage signal modulated by chest displacement. Whereas the chest displacement signal is expected to have its waveform onset coincide with ventricular contraction (QRS complex), the arterial signal is expected to be delayed in time by: 1) the cardiac pre-ejection period, and 2) the PTT through blood vessels. Due to different artery lengths, a difference in PAT is expected for the various measurement locations. A diverse set of locations was selected after initial experimentation. Unsurprisingly, large and superficial arteries proved to be robust measurement targets. Locations were chosen on the left side of the body for consistency with typical BP measurements. Radar signals were collected from:

- Carotid artery (left side of neck)

¹²Radar module was developed by Dr. Kristian Kjelgård (UiO) and Novelda AS, Norway.

¹³Radar control system was developed by Mathias Tømmer (UiO).

¹⁴Antennas were developed by Mathias Tømmer (UiO).

- Brachial artery (medial left upper arm)
- Femoral artery (medial left thigh)
- Posterior tibial artery (medial left foot)
- Intracranial arteries (left temple)
- Lower back (medial, directly superior to iliac crest)
- Heart (sternum)

To further validate the source of HR modulation, two off-body recordings were performed:

- Non-contact measurement of the chest (50 cm range)
- Sham recording on a phantom object with dielectrics resembling mean characteristics of the human body

As the RF propagation time through air is negligible (order of ns) compared to the blood pressure wave propagation through arteries, the non-contact measurement is expected to show a clear timing difference with local contact measurements. The sham recording is designed to test whether HR modulation by chest displacement is present in leakage signals, as the antenna coupling with the phantom closely resembles antenna coupling with real tissue.

Experimental setup and protocol

The experiment was performed on a convenience sample of healthy volunteers with no self-reported cardiovascular disease. Four subjects, of which two females, participated in the experiment. Mean age ($\pm$ SD) was 25.0 ($\pm$ 2.2), BMI ($\pm$ SD) was 20.7 ($\pm$ 1.6). The sham recording and non-contact recording were included to validate the source of HR modulation, and were thus only performed on a single subject. The recording from the lower back was added as a proof of concept and performed on two subjects. Approval for the study was obtained from the Imperial College London institutional ethics committee (letter of approval attached in Appendix G).

The experimental setup is illustrated in Fig. 4.10, showing the different on-body radar measurement locations, two off-body control measurements and ECG. The radar measurements were carried out using a custom radar module, which is a result of a Novelda funded research effort on heart inspection. The module consists of the XeThru X2 single-chip radar (Novelda AS, Kviteseid, Norway) lowered centre frequency and added amplification in both transmit and receive paths for increased sensitivity. The transmitted waveform (shown in Fig. 4.1) is a Gaussian modulated sinusoid, with 3.8 GHz centre frequency and a -10 dB bandwidth of 2.5 GHz. Antennas designed specifically for in-body sensing were used in the experiment. Details on the

design, simulations and measurements of the antenna are given in [87]. More details and characterisation of radar modules and antennas are given in Section 3.4. Antennas were positioned adjacent to one another, with 2 cm inter-antenna distance. Placement of the antennas was directly on the skin superficial to where the artery of interest was expected to be (illustrated in Fig. 4.1). During the experiment the antennas were held in place using a 3D printed structure and elastic band (photograph of radar module and antennas in fixture pressed to a semi-solid phantom in Fig. 3.5a). The antenna structure was spray painted with conductive paint to reduce signal leakage. A layer of radiation absorbent materials covering the back of the antennas was used to reduce leakage signals further. The non-contact control measurement used conventional antennas. ECG data were recorded from the left and right wrist, with the left leg acting as ground. ECG data were sampled at a rate of 250 Hz, using an Arduino and Olimex EKG shield. The validation phantom was a semi-solid phantom with dielectric properties similar to average tissue in a human limb. Phantom synthesis is described in E.

All measurements were performed in an RF anechoic chamber to reduce possible multi-path propagation of leakage signal. Participants were comfortably seated, and given time to relax before each recording. Participants were asked to minimise movement during the recordings. For each measurement, two 60 s recordings were made: One in which participants were asked to hold their breath for as long as they could up to 60 s, and one in which subjects were asked to breathe freely. For all further analysis except for the carotid artery measurement, the breathing datasets were used: The chest motion caused by respiration proved not to be a problem. In the carotid artery recordings, respiration signals were strongly present. Separating respiration from HR signals is challenging due to interfering harmonics. Because this is beyond the scope of this work, non-breathing recordings were used for the carotid artery measurement.

Data analysis

A schematic of the recording setup, indicating the body-coupled antennas and underlying tissue, as well as digital down conversion performed on the raw radar data, is given in Fig. 4.1. Subsequent signal processing steps are illustrated through Fig. 4.11.

Individual datasets consisted of 60 s of radar data collected at 64 fps. Each frame consisted of 256 range bins, sampled at approximately 40 GHz. The fast-time sampling frequency corresponds to a inter-bin range of 7.5 mm in air, or ~ 1 mm in tissue. The frame rate was chosen to be high enough to capture the time-domain characteristics of the arterial pulse waveform. Raw radar frames (Fig. 4.11a) were first converted to complex baseband data, filtered with a matched filter, and down-sampled by a factor 4. Obtained complex IQ baseband data were detrended and zero-phase shift bandpass filtered: Radar frames contain both static and dynamic signals which overlap in fast-time. As the arterial pulsation signal is purely dynamic and

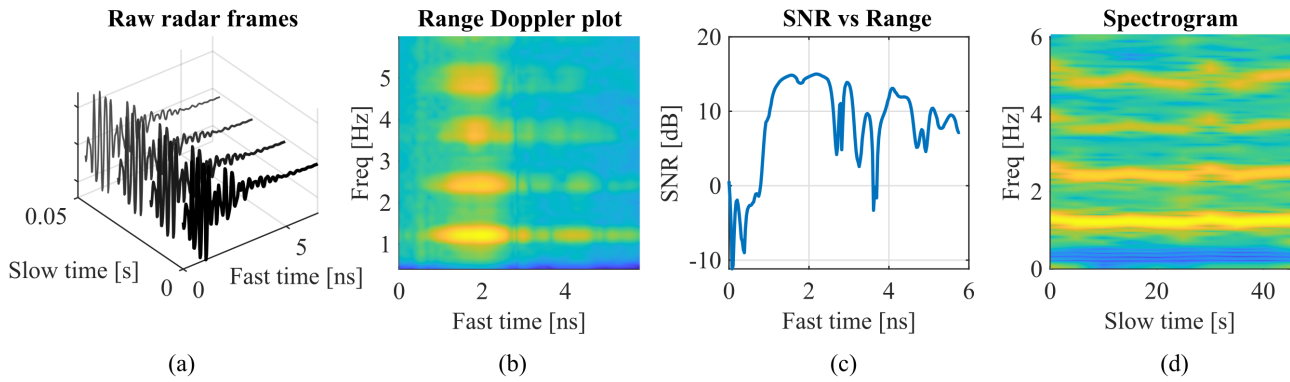


Figure 4.11: Illustration of raw radar data analysis steps: Raw radar frames (a), Range Doppler plot showing spectral content at different range bins (b), SNR of HR signal across range bins (c) and spectrogram at peak SNR (d).

occupies a limited bandwidth, data were bandpass filtered in slow-time to extract the signal of interest and to remove static clutter as well as the respiration signal. For this, an infinite impulse response filter with passband of 0.6 – 8 Hz was used. Next, linear demodulation was applied to filtered IQ baseband data for each of the range bins, to obtain a valid measure for ToF. PCA was used to find a function along which the data showed the highest statistical variance in the IQ plane. The first principal component (PC1) was used for further analysis, the second principal component was discarded. A disadvantage of the PCA approach is the ambiguity of polarity of the obtained signal. To ensure that polarity was matched across range bins, polarity was detected and signals were inverted if necessary.

Spectral analysis of radar data was performed to verify that HR measured over time corresponded to ECG data. Amplitude spectra of detrended and windowed PC1 data was obtained for all range bins in sliding 6 s windows (in which HR may be assumed stationary) using a chirp Z-transform. Frequency spectra of these 6 s windows of data were averaged to obtain the mean range-Doppler plot (Fig. 4.11b). The sampler containing the most prominent HR signal was found by estimating SNR for each sampler as the ratio between the power at the fundamental HR frequency and its harmonics, to non-harmonic spectral content (Fig. 4.11c). The sampler with peak SNR was used to produce a spectrogram of PC1 radar data (Fig. 4.11d) which was used for direct comparison of radar to ECG data.

Subsequent analysis was aimed at studying the time-domain characteristics of the measured cardiovascular signals. First, an accurate timing reference of the heart cycle was required. For this, the R wave of the QRS complex was chosen as fiducial ECG marker, for its robust detectability. ECG data were zero-phase shift filtered using a low-pass infinite impulse response filter. R wave detection was achieved by finding the local maximum in ECG data directly preceding a peak in its first approximate derivative. PC1 data at peak SNR was used for all radar signal processing. To improve SNR in the radar time-series, data from both radar

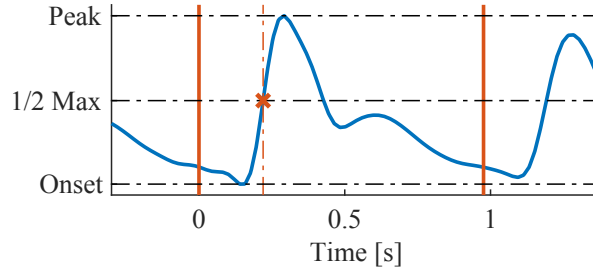


Figure 4.12: Pulse arrival time (PAT) was defined as the occurrence of half the maximum of pulse height. Half Max PAT is clearly identifiable, even in absence of a sharp pulse onset or pulse maximum.

and ECG were aligned using the fiducial ECG markers and radar data over multiple heart cycles were averaged. Once an averaged radar response was obtained for each of the measured locations, pulse onset and pulse peak were determined from zero-crossings in the derivative signal. Both pulse onset and pulse maximum may be unreliable metrics of PAT when no sharp onset or peak is observed. As illustrated in Fig. 4.12, the timing of half the pulse maximum was used as a temporal marker of PAT.

For PTT analysis between multiple measurement sites (for example: carotid to femoral), only heart cycles with similar duration were used. As recordings were performed from only one position at a time, individual heartbeats were matched between recording sites based on similar R-R intervals: Individual heartbeats were grouped based on R-R interval using a 16 ms bin width, and the bin with the highest number (generally 10–15) of heart cycles was selected for processing. Heart beats with the selected R-R interval were averaged and a valid comparison in PAT could be made between recording sites.

4.5.3 Results

The heart beat was successfully measured using radar, at all on-body positions across four participants, with the exception of the upper arm measurement on participant 3. However, in four recordings (indicated in Table 4.1), the ECG equipment failed. Although HR is clearly visible through spectral analysis, no time domain analysis can be performed on these recordings due to the absence of a time reference.

Average SNR values of obtained HR signals across participants ranged from -3.2 dB (tibial artery), to 3.2 dB (carotid artery). Comparing SNR values obtained using linear demodulation to SNR values from arctangent demodulation (amplitude and phase), shows that linear demodulation is more robust across measurements. Median SNR and median absolute deviation ($\pm$ MAD) of SNR across all recordings were: 0.25 dB (± 4.89) for linear demodulation, -2.15 dB (± 5.07) for amplitude arctangent demodulation, and -3.75 dB (± 3.74) for phase arctangent

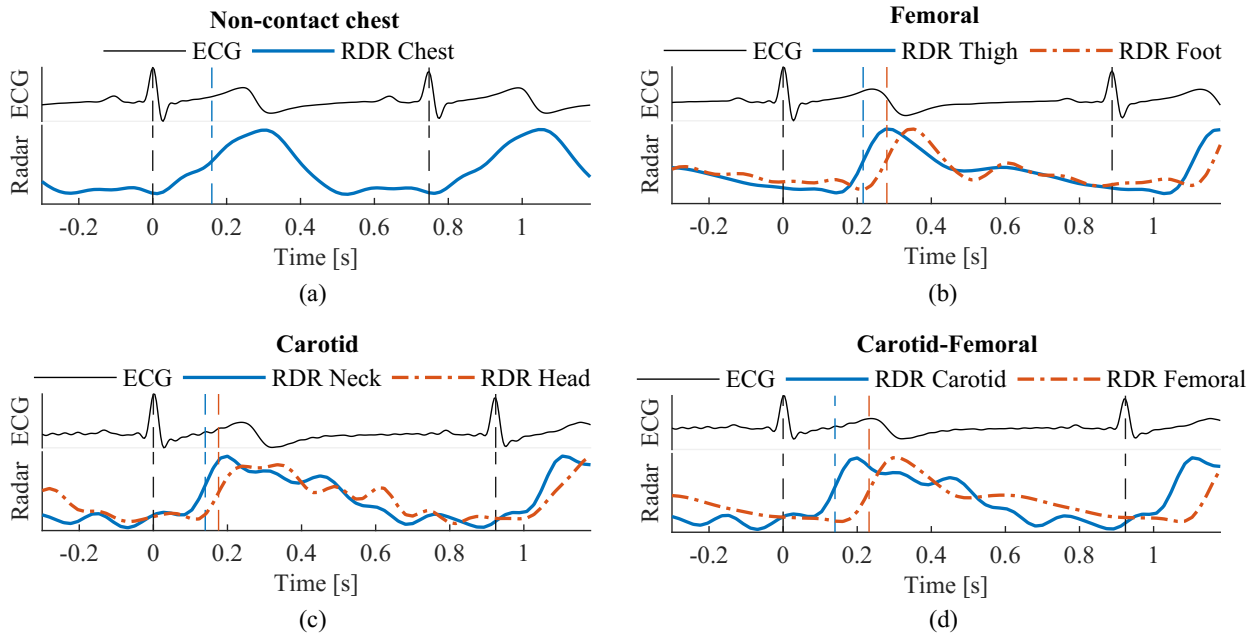


Figure 4.13: Comparison of pulse arrival time (PAT) between the non-contact chest measurement (a) and in-body measurements (b – d). Pulse wave velocity is demonstrated through the femoral artery (b), through the carotid artery (c), and through the aorta (d; as measured from carotid to femoral artery). The blue traces represents the more proximal measurement point, orange the more distal measurement point. Half Max PAT is indicated by coloured dashed lines, ECG R wave is indicated by a black dashed line.

demodulation. A paired-samples sign test (data distribution was neither symmetric nor normal) with Bonferroni correction for multiple comparisons resulted in $p < 0.05$ and $p < 0.001$ when comparing linear demodulation to amplitude, and linear demodulation to phase, respectively.

PAT measured at different sites validate that the source of the HR signal, measured using body-coupled antennas and pulsed UWB RoC, are indeed pulsating subcutaneous structures. Data from participant 1 (Fig. 4.13) illustrate that a clear timing difference is apparent between in-body measurements and the contactless measurement: The onset of chest displacement in Fig. 4.13a coincides with the ECG R wave (10ms delay) and reaches a peak at the offset of the T wave. The rise period covers the entire systole of the heart. The arterial pulse onsets shown in Fig. 4.13b–d however, are delayed due to a limited blood pressure pulse propagation velocity through arteries (delay of 73 – 212 ms, depending on location) and pulses show much shorter rise times. Even when searching at different range bins in the received radar signal, it has not been possible to detect a waveform with similar time domain characteristics as the non-contact chest wall signal. The sham recording (not shown in any of the figures) did not contain a detectable HR signal. This indicates that no HR modulation of leakage signal, was caused by either chest displacement or any other superficially moving structure.

PWV was demonstrated in the femoral artery (Fig. 4.13b), by a clearly distinguishable delay

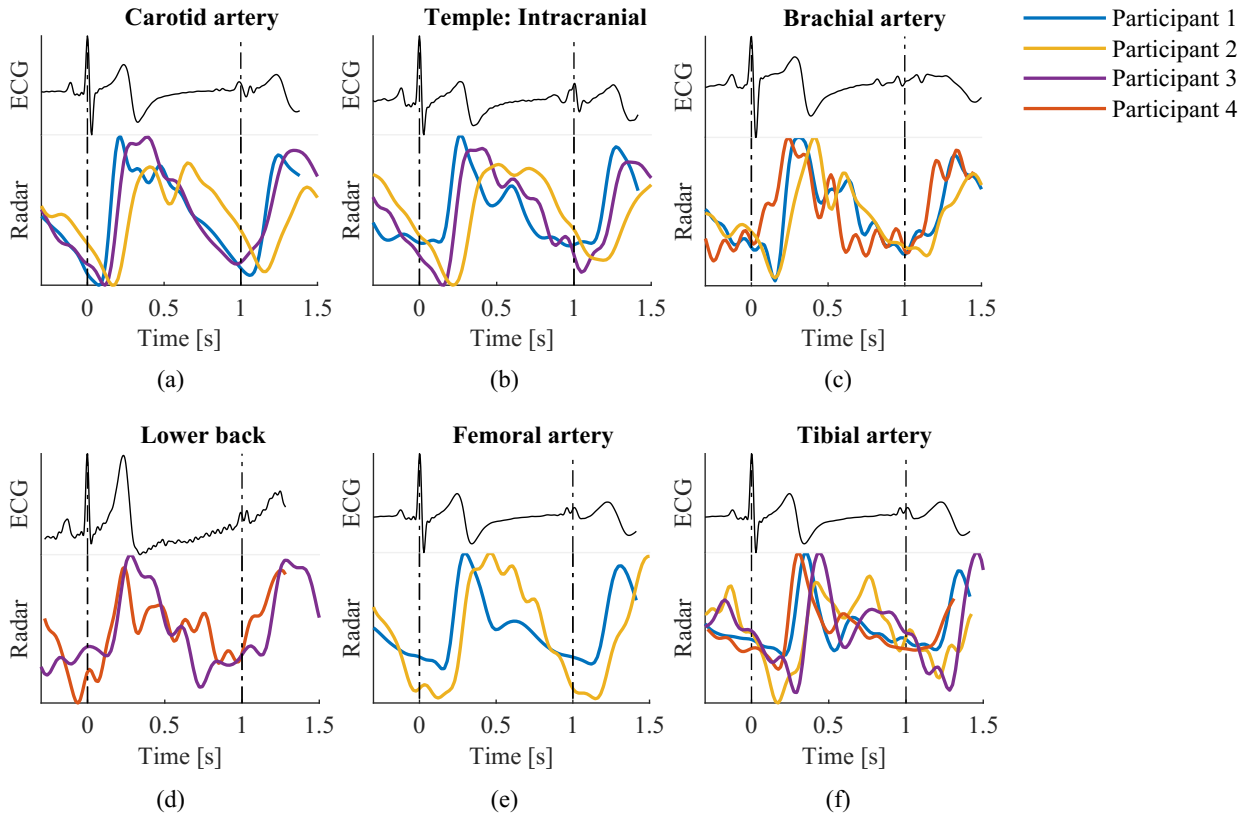


Figure 4.14: Arterial pulse wave at different measurement locations in response to the heart cycle, for all four participants: Neck (a), temple of the head (b), upper arm (c), lower back (d), thigh (e) and foot (f). All heart cycle durations were normalised to allow for waveform comparison. Radar time series are colour-coded for the different participants, ECG data are shown in black. Black dashed lines indicate the ECG R wave.

in PAT between the thigh and foot measurements. By comparing the Half Max PAT, a difference of 64 ms was measured. This translates to a PWV in the femoral artery of 12.5 m/s (artery length roughly estimated at 80 cm). A delay in PAT between the head and neck was observed (Fig. 4.13c), but PWV is harder to define due to the unknown exact path length of the intracranial blood pressure wave. Finally, PWV velocity through the aorta, as measured between the carotid and femoral arteries, is shown in Fig. 4.13d and measured at 92 ms. The radar-measured arterial pulse wave in response to the heartbeat, across measurement locations for all participants, is shown in Fig. 4.14. For illustration purposes and to allow for waveform comparison, heart cycle duration was normalised between participants. Fig. 4.14 is therefore not suitable for a direct comparison of PAT between participants. Instead, it demonstrates that arterial pulse wave measurements were robust and repeatable. Arterial pulse waveforms were found to vary between participants which is probably due to the limited control over exact measurement location. Half Max PAT is given in Table 4.1 for a number of recording sites, along with each participant's height and average HR. The chest and lower back measurements are not included because PAT of these sites are of less interest. As expected, PAT increases with increased distance of measurement location to the heart. PAT at identical locations varies

Table 4.1: Pulse arrival time of individual measurements

		Participant			
		1	2	3	4
Half Max PAT [ms]	Height [cm]	177	168	164	177
	HR [bpm]	62.5	70.8	76.2	59.8
	Mean ($\pm$SD)	(± 3.2)	(± 6.9)	(± 3.9)	(± 4.2)
	Neck	0.148	0.204	0.152	*
	Intracranial	0.200	0.256	0.176	*
	Upper arm	0.192	0.240	*	0.152
	Thigh	0.220	0.248	**	*
	Foot	0.284	0.332	0.300	0.256

Measurement locations: (1) Neck: Carotid artery; (2) Intracranial: Temple; (3) Upper arm: Brachial artery; (4) Thigh: Femoral artery; (5) Foot: Tibial artery. *Half Max PAT* denotes pulse arrival time measured at half of pulse maximum. *Data excluded due to ECG device failure. **No HR signal measured in radar data.

strongly between participants. This can partly be accounted for by differences in height and differences in arterial stiffness. In addition, differences in blood pressure may have contributed to differences in PWV between participants. Blood pressure directly affects PWV, and it is not unlikely that blood pressure during the recording deviated from standard values: No particular blood pressure assessment protocol was followed, and participants might have had elevated stress levels. In the current study, no blood pressure measurements were performed.

Results for the in-body recording at the sternum, measured directly above the heart, are shown in Fig. 4.15, for participant 1. Radar data and ECG data are shown along with a Wiggers diagram, which shows the theoretical relationship between the heart's electrical activity (ECG) and blood volume in the ventricle of the heart. Interestingly, but not surprisingly, the radar data closely resemble the ventricular volume in the Wiggers diagram, showing start and end of ventricular systole, as well as the atrial systole that occurs during the ECG P wave.

4.5.4 Discussion

Experimental work in the current section demonstrates that cardiovascular dynamics can be measured using coherent UWB RoC systems in combination with body coupled antennas.

A major challenge in in-body sensing and imaging, with low-power microwave equipment operating at a wide frequency band, are leakage signals. Due to the high signal attenuation in tissue, leakage signals are likely to be much stronger than the signal of interest, and may be modulated by chest displacement due to HR and respiration. By employing the temporal characteristics of measured waveforms, leakage signal as a cause for measured HR signal could be ruled out. Blood pressure PTT through arteries is in the order of tenths of seconds, whereas signal delay

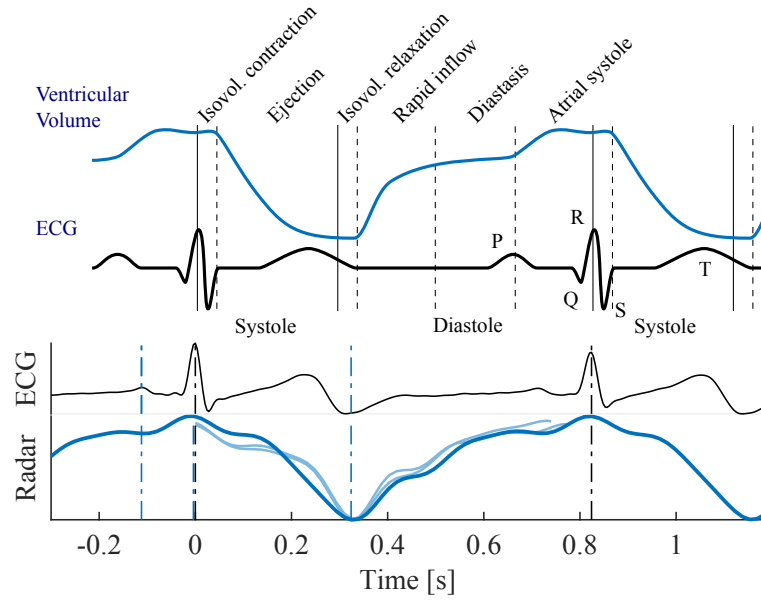


Figure 4.15: Wiggers diagram (adjusted from [117]) illustrating the relationship between the heart's electrical activity (ECG) and ventricular volume (top), and in-body radar recording from the heart, measured at the sternum of participant 1 (bottom). Local and global minima and maxima in radar data are indicated using blue dashed lines. Radar data closely resembles ventricular volume.

due to propagation through air is in the order of ns. A comparison of PAT between non-contact chest measurements and arterial pulsation therefore validates the source of obtained signals. As a final assurance, an attempt was made to measure HR modulation in leakage signal from a phantom object with dielectric properties matching those of average tissue in the limb. No HR modulation was found, even though proximity of the sensor to the chest was identical as in the upper arm recording. The more powerful respiration signal could occasionally be detected in contact recordings, particularly in the recordings from the neck. Future work will seek to further shield antennas from leakage signals. In addition, effective signal processing strategies are required to distinguish the HR signal from respiration harmonics. In non-contact UWB RoC monitoring of HR and HR variability, the combination of singular value decomposition and wavelet filtering has proven to be effective for this purpose [118].

Through ECG-aligned averaging of multiple heart cycles of radar data, the arterial response to the heartbeat could be studied even in low SNR scenarios. The arterial pulse waveform was demonstrated for arteries in the neck, head, upper arm, lower back, thigh and foot. For recordings from the temple and lower back, it is unknown which artery caused the signal modulation, as no attempt at imaging was made. For the other measurements however, recordings were taken at sites superficial to major arteries. Depth of the arteries was in the order of cm, but due to the limited down range resolution of the used system, no attempt was made at exact depth estimation.

Because of the use of linear demodulation, a valid measure for sub-mm ToF variation was obtained, despite the inability to perform DC offset removal. The significance of this is that the obtained arterial pulse waveform is not distorted by incorrect arctangent assumptions. Arterial pulse waveform analysis is a field of interest to the medical community and can be used to predict aortic pressure from waveform features [119]. Obtained waveforms varied strongly between participants, but in general did display fast systolic upstroke, dicrotic notch and diastolic runoff, as expected. Variation between participants could be caused by motion artefacts and sub-optimal sensor positioning, as well as signal distortion due to averaging of heart cycles with unequal length. Comparison against US or tonometry recordings would be of interest.

Half Max PAT was found to be the most robust marker for a comparison of PAT between recording sites, as it was not affected by the smoothness of pulse onset or pulse peak. From PAT measurements, PTT and PWV could be estimated for a number of arteries. For brevity, this was demonstrated only for a single participant. As an extra precaution against signal blurring, and to eliminate the effect of HR on the arterial response, only radar data from heart cycles with similar length were included in averaging. Measured PTT and estimated PWV lie within physiological range. Artery length estimation would be required for accurate PWV estimation, and US Doppler recordings would have to be performed to validate PWV accuracy, which is beyond the scope of this paper.

From the heart recordings at the sternum, it was found that a robust signal could be measured that closely resembles theoretical ventricular volume, as well as cardiac signals measured using CW Doppler radar [108]. The signal maximum occurs slightly before R wave, on the onset of ventricular systole. The signal minimum aligns with the end of T wave (end of systole). A local minimum was found to coincide with P wave, and second local minimum (although not always present) is hypothesised to indicate the start of ventricular ejection. These results indicate that UWB radar could potentially serve as a powerful diagnostic tool: Determination of ventricular volume, as well as derived metrics such as ventricular ejection fraction, are essential for cardiac function analysis. Because UWB radar measures mechanical motion as opposed to electrical activity, radar could act as a complimentary modality to ECG in a clinical setting. Future work using an array of radars and an imaging approach will explore this in more depth.

In Section 4.3.2 it was argued that linear demodulation is more suitable than alternative methods for tracking the arterial pulse wave, considering the size of the arteries targeted, the arterial expansion factor, and the effective wavelength of the system's centre frequency in tissue. A statistical analysis of SNR values obtained using different demodulation approaches confirmed the theory. A future challenge is how linear demodulation techniques can effectively be combined with DAS beamforming methods. ToF variation data obtained from linear demodulation measured at multiple sensors does not necessarily lie along the same principal component, so coherent summation must be performed either on raw RF data, or after a correction step in

which ToF variation data between sensors get aligned and inverted if necessary.

The current work focuses on 1D in-body sensing using UWB RoC. Although alternative modalities are available for cardiovascular sensing (US, tonometry, photoplethysmography), radar has a number of potential advantages: Hardware is affordable and comes at a small form factor. Whereas optical methods sense blood volume in superficial tissues, radar has a much higher penetration depth. This allows for better tissue penetration and thus sensing of deeper lying arteries. As opposed to ultrasound, transmission through bone is possible, enabling sensing and imaging inside the skull. Finally, radar is comfortable to use without the need of applying pressure (required for applanation tonometry) or a coupling gel (required for ultrasound).

As pointed out in Section 4.2.2, cardiovascular sensing has previously been performed using CW Doppler radar. However, the use of coherent UWB radar is particularly of interest due to the option of time gating, and the imaging capabilities when extending to a multiple-module rig with an array of antennas. Dynamic MWI using a system of synchronised RoC modules and arrays of body-coupled antennas will be investigated in Chapter 5.

4.6 Conclusions

Based on the here presented analysis and experimental work on UWB radar sensing in the human body, the following conclusions can be drawn:

- When choosing radar signal processing techniques for in-body sensing of cardiovascular activity, the size of the artery, the arterial expansion factor, and the effective wavelength in tissue should be considered.
- Linear demodulation provides a more robust and valid alternative than arctangent demodulation when the target structure and motion are small relative to effective wavelength.
- Using transmissive measurements, cardiac activity could be measured from the temples of the head with UWB RoC and body-coupled antennas. ToF analysis of simulated data suggests that the source of HR modulation is intracranial.
- Observations of leakage signal modulation by motion in the vicinity of the test setup emphasises the importance of further shielding and the challenge of sensing or imaging when the environment is not controlled for.
- The arterial pulse wave can be measured at various sites on the body, using coherent UWB RoC system. Measurements of PAT from pulse half maximum was found robust and could be used to estimate arterial PWV.
- From the pulse wave dynamics, the subcutaneous source of the measured signals was verified.

- ECG aligned averaging of multiple heart cycles with similar duration proved to be a valuable tool in increasing SNR of pulse dynamics time series.
- UWB radar measurements from the heart closely resemble theoretical ventricular volume.
- Coherent UWB RoC technology could potentially be used as a low-cost diagnostic tool in assessing cardiovascular function, complementary to existing technologies such as ECG and US.
- The current investigation into in-body sensing serves as an early exploration of the challenges and opportunities of in-body measurements using UWB RoC technology, and therefore provides the fundamentals necessary to develop RoC imaging systems.

Chapter 5

UWB radar-on-chip for biomedical imaging

5.1 Introduction

In the past two decades, biomedical MWI has been investigated, mostly for the applications of breast cancer imaging and stroke imaging. Existing research has predominantly used bulky and high cost radar hardware, and only a single research effort has focused on dynamic imaging. The use of RoC technology for MWI offers the unique opportunity of developing a low-cost, small form factor, portable, and high speed imaging system. In the current work, dynamic and static MWI in the human body is performed using a system of synchronised RoC devices. Arrays of body-coupled UWB antennas are used for multistatic radar data collection. A MIMO dataset is formed by sequentially scanning through transmitters with all radar modules receiving simultaneously. Simulated data are used to test various imaging algorithm parameters. Both differential imaging and the Woody average for skin artefact removal are compared. The DAS and DMAS beamforming techniques are used, with an exploration into compensation of pulse distortion and attenuation due to propagation through dielectric media. Experimental imaging data are obtained from a tissue-mimicking phantom, from the human heart, and from the femoral artery in the human thigh.

5.2 Background

The current section explores the fundamentals of beamforming for biomedical imaging applications. In addition, it aims to provide an overview of the state of the art in radar-based MWI, specifically focusing on data processing and imaging algorithms.

5.2.1 Radar imaging fundamentals

Beamforming notation

Radar imaging, either in 2D or 3D, relies on coherent summation of transmitted and received waveforms, in order to achieve constructive interference for signals arriving from the direction of interest. Spatial filtering, or beamforming, is achieved by applying appropriate delays to data received across the different sensors in a real or synthetic array. Some theory is considered here (based on [120, 121]) to aid in understanding of how different parameters of the imaging system contribute to image quality.

Consider a uniform linear array (ULA) with M elements, and element spacing d (in m). When the ULA is being excited by a complex sinusoid signal $s(t)$ with angular frequency ω (in rad/s):

$$s(t) = e^{j\omega t} \quad (5.1)$$

The received signal $x(t)$ at sensor m will be:

$$x_m(t) = s(t)e^{-j(m-1)\xi} \quad (5.2)$$

With $\xi = \frac{\omega d}{c} \sin \theta$, c the speed of light, and θ the angle of the scattering source relative to broadside. The obtained dataset from the ULA at sample n then becomes:

$$\mathbf{x}[n] = [1, e^{-j\xi}, \dots, e^{-j(M-1)\xi}]^T s[n] \quad (5.3)$$

To achieve coherent summation for data in matrix $\mathbf{x}(\mathbf{n})$, in direction θ_0 , a steering vector is used to compensate for time delays at each of the M sensors:

$$\mathbf{w} = [\alpha_0, \alpha_1 e^{-j\zeta}, \dots, \alpha_{M-1} e^{-j(M-1)\zeta}]^T \quad (5.4)$$

With $\zeta = \frac{\omega d}{c} \sin \theta_0$, and α_m the amplitude weight for sensor m . The output of the beamformer is now given by:

$$y[n] = \mathbf{w}^H \mathbf{x}[n] \quad (5.5)$$

The maximum response of this beamformer, with steering vector $\mathbf{w}$ is expected in direction θ_0 . Sensor amplitude weights α_m can be used to control sidelobe levels and beamwidth. Sensor weights may be fixed (conventional beamforming) or may be data-driven (adaptive beamforming).

Beampattern

A beamformer can be characterised through its beampattern: the response for a point target across a range (-90 to 90 degrees from broadside) of DOA. The beampattern is a product of the array factor and element factor. The element factor is the antennas radiation pattern and will for now be assumed isotropic. Array factor and beampattern can thus be used interchangeably. The beampattern P for a certain beamformer $\mathbf{w}$ can be obtained by plotting the magnitude of y as a function of signal arrival angle θ . P can be found by:

$$P(\xi) = \sum_{m=0}^{M-1} \mathbf{w}^H e^{-jm\xi} \quad (5.6)$$

With $\xi = \frac{\omega d}{c} \sin \theta$. The output of the above equation is a function of ξ , but may be converted so that it becomes a function of DOA θ . Interestingly, the above equation is of the same form as the discrete Fourier transform, which allows for a number of insightful analogies:

- The FFT can be used in beamforming, to transform a spatially sampled signal (array input) into the spatial frequency domain (output as a function of DOA θ).
- The Nyquist theorem applies to spatially sampled signals: A minimum of 2 samples are required per frequency cycle, to avoid aliasing. The highest spatial frequencies occur at $\theta = -\pi$ and $\theta = \pi$. At high DOA, projection of the waveform onto the array has the highest wavenumber k , relative to the sampler spacing.
- From the previous statement, we can derive that the minimum sampler spacing to avoid spatial aliasing (grating lobes) needs to be $\lambda/2$.
- Analogues to time domain Fourier theory, the total aperture size ($d * M$) determines the angular resolution of DOA.
- Just as windowing is performed in time domain Fourier analysis, sensor weights α_m can be used to reduce sidelobes in the spatial frequency domain.
- The beampattern magnitudes are determined for a uniformly sampled range of spatial frequencies ξ , ranging from $-\pi$ to π . However, due to the conversion to the physical angle of arrival (θ), resulting magnitude values on the θ axis are no longer uniformly sampled. This causes beam broadening when the scan angle off broadside increases.

Scanning architecture and virtual array density

From the above section it becomes apparent that a larger aperture generally contributes to increased spatial resolution, and a higher array density reduces aliasing. A large aperture can be achieved either by constructing a real array of antennas or by mechanically scanning an antenna through a range of positions, thereby constructing a synthetic array as is done in SAR

imaging. When a single radar system is used with co-located transmitting and receiving elements, monostatic data are obtained, consisting only of reflective measurements. An advantage of using a real array, is the access to multistatic data: scattered EM waves are collected from multiple, spatially diverse, receiving antennas. When a single transmitter is used in combination with multiple receiving elements, one speaks of SIMO radar. A full MIMO dataset can be obtained synthetically from multiple SIMO datasets, when sequentially scanning through different transmitters. Real MIMO is achieved when multiple transmitters and multiple receivers operate simultaneously, thus generating a large dataset in a short duration. In this case, orthogonal waveforms must be used for simultaneous transmission.

When considering steering vector $\mathbf{w}$ in equation 5.6, only the path from source to receiving element is considered. For a MIMO dataset, the number of unique channels (Tx-Rx combinations) is M^2 , as opposed to M for a SIMO or monostatic dataset. The real array sensor density does therefore no longer determine the system beampattern, but instead the virtual elements must be considered to assess the spatial sampling properties of an array [83]. The virtual element position $\mathbf{pos}_{\mathbf{ve}} = (x, y, z)$ for channel m, n depends on the physical positions of transmitting and receiving elements $\mathbf{pos}_{\mathbf{tx}}(m)$ and $\mathbf{pos}_{\mathbf{rx}}(n)$:

$$\mathbf{pos}_{\mathbf{ve}}(m, n) = \frac{\mathbf{pos}_{\mathbf{tx}}(m) + \mathbf{pos}_{\mathbf{rx}}(n)}{2} \quad (5.7)$$

Where $m = 1..M$ and $n = 1..N$, representing the indices of real transmitting and receiving elements, respectively. When a ULA is used to collect a MIMO dataset, the density of virtual elements is therefore twice as high as the density of physical sensors in the SIMO case. This is a highly desirable property, as beamforming of a spatially undersampled dataset will result in aliasing.

Beamforming for biomedical imaging

Beamforming techniques originate from military and communication applications, in which the scatterer or signal source is usually located in the far field. For imaging in the far field, plane wave propagation is assumed, and a set of delays can be applied without consideration of the target range. For biomedical imaging applications, the plane wave assumption is generally not valid and the beamformer delays are dependent on both DOA and target range. This requires an imaging system to focus on individual voxels in the target geometry. The term confocal microwave imaging (CMI) is generally used to describe beamforming in the near field for biomedical applications. The simplest form of a DAS beamformer can be written as [122]:

$$y(\mathbf{r}) = \left[\sum_{m=1}^M s_m(\tau_m(\mathbf{r})) \right]^2 \quad (5.8)$$

Where M now represents the total number of Tx-Rx channels, and τ_m is the time delay corresponding to the path length from transmitting element $\mathbf{pos}_{\text{tx}}$ to focal point $\mathbf{r} = (x, y, z)$ and back to the receiving element $\mathbf{pos}_{\text{rx}}$, for channel m . Instead of summing s_m only at a sample corresponding to delay τ_m , a window around this sample may be chosen. Increasing the window length results in signal smoothing, which may add to imaging robustness. A second difference between traditional radar imaging applications and medical radar imaging is the propagation medium. EM wave propagation velocity is tissue dependent, EM wave absorption is high, and biological tissues are dispersive causing distortion of the transmitted waveform. Since CMI relies on coherent summation, correct estimation of signal delay $\tau_m(\mathbf{r})$ is paramount. Compensation of signal attenuation may be performed to avoid under-representation of channels with a high path length, and compensation of signal distortion may be required for correct UWB signal processing. To perform any of these operations, a correct model of the anatomy and dielectric properties of the imaging target are required.

5.2.2 Confocal microwave imaging state of the art

The current section provides an overview of the state of the art of CMI for biomedical applications. Unlike Section 2.3, the focus lies on algorithms used for image formation as opposed to imaging hardware, and only considers radar-based imaging. This section does not cover MWI based on tomography or holography.

Preprocessing

Typical preprocessing steps are passband filtering of received radar frames, jitter compensation, and reduction of clutter signals. Particularly the latter is challenging, as signal attenuation in biological tissue is high (~ 7 dB/cm at a centre frequency of 3.45 GHz) and clutter signals (skin boundary, crosstalk) may be many times higher in amplitude. Through hardware improvements, crosstalk may be suppressed. Some clinical applications allow for differential imaging, in which a recording without target may be subtracted from a recording with target, such as in [29]. If this is not an option, the skin artefact is often estimated by averaging a set of channels, under the assumption that the skin artefact is constant but the signal of interest varies across channels. The skin artefact estimate may then be subtracted to increase signal-to-clutter ratio (SCR). For breast imaging, neighbouring channels are often grouped for averaging [20, 123]. For head imaging, symmetrical channels may be chosen [124]. Better results may be obtained by estimating the skin artefact as a filtered combination of signals in other channels [97, 125]. The subtraction approach relies on coherence of the skin artefact signal, which is complicated by the irregularities in body shapes and hardware imperfections. An improved version of the

average subtraction approach is the Woody average: Signals are time aligned based on either correlation or least squares method before averaging, and a skin estimate with appropriate delay is subtracted [32, 41]. More recently, independent component analysis has been proposed as a method to estimate and remove clutter components [38, 126].

Assumed tissue properties

EM wave propagation velocity is tissue and frequency dependent. For correct use of equation 5.8, delay τ_m must be estimated with reasonable accuracy. Therefore, some model of the anatomy must exist. Dielectric properties are often taken from [16] or an expanded database of tissue properties [18]. An average geometry is usually estimated from various sources, after which the imaging volume (including the coupling liquid, if present) is considered as a homogeneous medium, represented by average dielectric properties. This method introduces a lot of error, but an individualised model of geometry and tissue properties is usually not feasible. For brain imaging, a more refined approach is required, as the head geometry is highly inhomogeneous (skin, fat and bone form a thin outer layer with significantly lower permittivity than brain tissue) and air is used as a coupling medium. A depth-dependend model for effective permittivity was used in [41]. A different approach is to attempt to estimate tissue properties through measurement, either by using the CMI system [29, 30], or by using a separate measurement device [26, 37]. An algorithmic approach has also been taken with success to optimise permittivity iteratively [8, 127].

Propagation model and attenuation compensation

Signal attenuation and propagation velocity in biological tissue is frequency dependent, causing significant waveform distortion when using UWB signals. When considering a certain focal point, signal amplitude from a channel with high path length will be more attenuated than signal from a channel with nearby antennas. To compensate for this inequality, attenuation compensation may be performed before channel summation takes place. Many CMI systems opt for not applying attenuation compensation [24, 29, 32, 41], as it does not necessarily improve SNR locally and at high path length (and particularly high frequencies) noise will be amplified to extreme levels. A data-driven gain compensation approach is employed in [44], which equalises contributions from all channels for individual focal points, without considering path length. Attenuation compensation based on path length is implemented in [20, 37]. The MIST beamforming algorithm uses a set of filters to compensate for both attenuation and signal distortion, but assumes constant dielectric properties across the frequency band [97].

Beamforming algorithm

The DAS beamforming algorithm is used most commonly for image formation. Signal delays are usually applied in the time domain by truncating or concatenating received radar frames. Fine, sub-sample delays may be applied through upsampling or filtering, but improvements to imaging performance were found to be minimal [97]. Several improvements and variations have been published on the DAS algorithm. Increased performance was obtained by applying heuristic weights, representing signal quality, to different channels for each focal point before summation [128]. The robust Capon beamforming algorithm is again a data-adaptive improvement on DAS beamforming, and has been applied for multistatic datasets in [129]. More variations on data-adaptive algorithms exist, but in essence they are quite similar. The DMAS algorithm, originally used in US imaging but now often applied for CMI, uses multiplication between channels to increase the number of signals before summation, thereby strongly reducing clutter [130]. The MIST beamforming algorithm is in essence DAS, but uses filters to compensate for signal attenuation and distortion as a result of propagation through tissue [97, 131]. In a direct comparison study DMAS was found to outperform both DAS, robust Capon beamforming and other data-adaptive algorithms [132]. More recently, the super-resolution beamforming technique time-reversal multiple signal classification was implemented, which was developed specifically to resolve targets much smaller than the system wavelength. Encouraging results were obtained on breast phantoms [38].

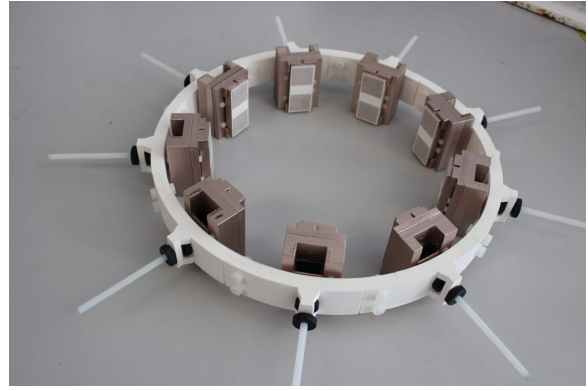
5.3 Methods

5.3.1 Overview

The developed system of synchronised RoC devices for biomedical MWI was tested on a human participant and on a tissue-mimicking phantom. Three different physical antenna arrays were constructed: 1) A circular array, 2) a dense semicircular array, and 3) a rectangular array. Details for each of the arrays are given in Fig 5.1. Imaging on the human participant was performed on the chest and on the thigh. Dynamic imaging targets were the heart and the femoral artery. The femur bone, as well as steel targets in the phantom acted as static imaging targets. The imaging scenarios were simulated in simplified form, to test various imaging algorithm parameters and to predict imaging system performance.

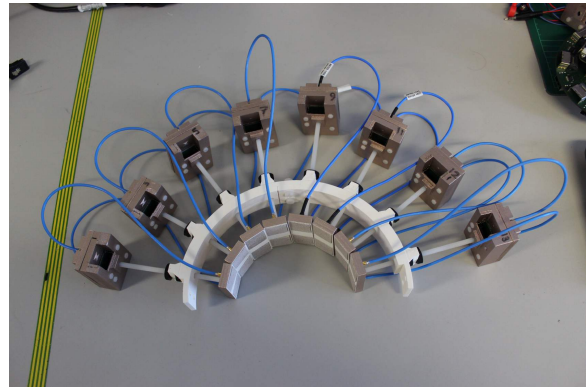
Circular array

- 9 Modules
- 9 Tx and 9 Rx body-coupled antennas
- 360° Coverage
- Diameter range: 11.0 – 16.0 cm
- Minimum inter-antenna distance: 3.8 cm
- Shielded modules with integrated antennas



Dense semicircular array

- 8 Modules
- 8 Tx and 8 Rx body-coupled antennas
- 144° Coverage
- Diameter: 11.0 cm
- Modules separate from antennas, SMP cable connectors
- Minimum inter-antenna distance: 2.0 cm



Rectangular array

- 8 Modules (2 x 4 modules)
- 8 Tx and 8 Rx body-coupled antennas
- Array dimensions: 8.7 x 12.0 cm
- Inter-module distance: 4.0 cm and 5.7 cm
- Shielded modules with integrated antennas

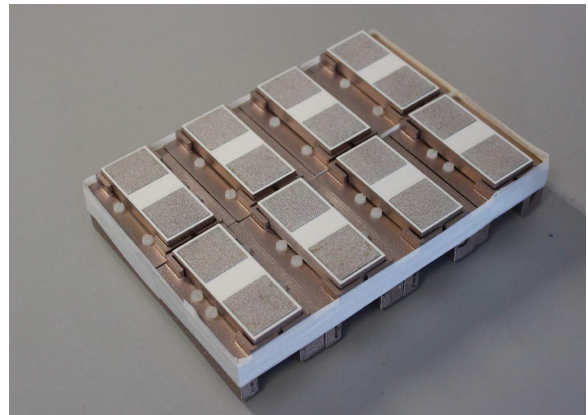


Figure 5.1: Overview of the three different antenna arrays

5.3.2 Assumed medium properties

In the current work, imaging objects (phantom, human thigh and chest) will be considered homogeneous. The geometry of the phantom is well defined, and dielectric properties have been measured across the frequency band of interest. For the in-human testing, estimates have been made of anatomy, and dielectric properties have been taken from [18], assuming a centre

frequency of 3.45 GHz. More details are given for each imaging object below:

Phantom properties Dielectric properties of the semi-solid phantom were measured using a VNA (Agilent E8361A) and dielectric probe (Agilent 85070E). It was expected that the volatile components of the phantom close to the surface would evaporate, thereby changing the phantom properties. After imaging measurements were finalised, the phantom was cut in half and its dielectric properties were measured from the surface of the cross section. A two-pole Debye model was fitted to measured data, which allowed for incorporation of the phantom dispersive properties in numerical simulations. Measured and fitted data are shown in Fig. 5.2.

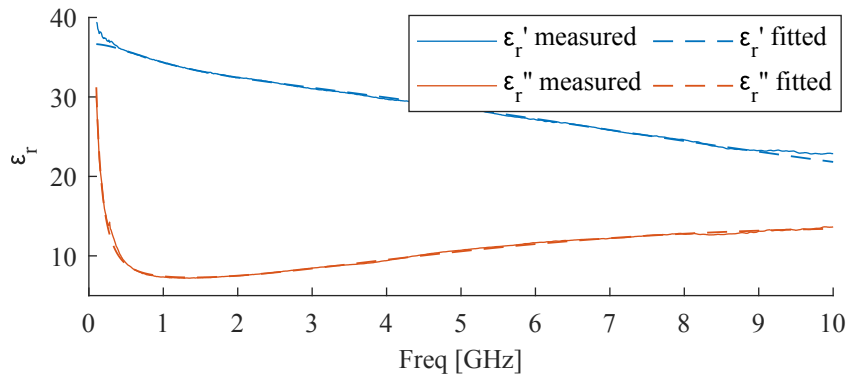


Figure 5.2: Measured dielectric properties of a semi-solid phantom (solid line), along with fitted two-pole Debye model (dashed line). The real part and the imaginary part of relative permittivity are denoted as ϵ_r' and ϵ_r'' , respectively.

Chest properties Fig. 5.3 illustrates what a cross section of the human torso may look like. The heart lies directly against the ribcage and is laterally surrounded by lung tissue. A strong dielectric contrast is present at the heart wall boundary. Peripheral channels will image through lung tissue, whereas central channels will contain more bone tissue. For cardiac imaging, a 1-dimensional model of the path up until the heart wall boundary will be considered. As a starting point, values for layer thickness from [133] were considered, which were based on the dissection of a pork rib. The addition of lung tissue (to account for peripheral channels) and reduction of the fat layer resulted in the values presented in Table 5.1. Taking dielectric properties for the relevant tissues from [18] resulted in effective relative permittivity and conductivity of 25.1 and 1.33 S/m, respectively.

Thigh properties Imaging was performed on two locations on the thigh: The ‘Upper thigh’, with a measured circumference of 48 cm, and the ‘Lower thigh’, with circumference of 40 cm. Thigh composition was estimated from published data. Cross-sectional area measurements of male thighs [135] were used as a starting point. A 2 mm skin layer was added (final estimates in

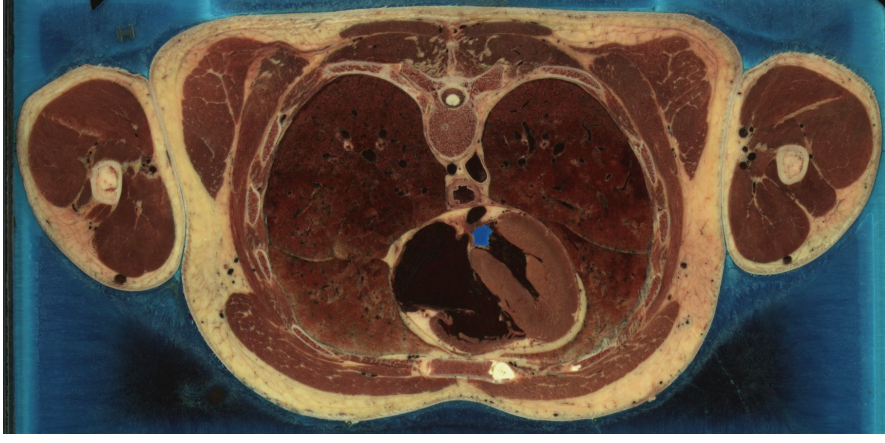


Figure 5.3: Transverse cross section of human torso, illustrating the size and location of the heart [134].

Table 5.1: Estimated tissue fractions for the chest and thigh.

Tissue	Chest		Thigh
	Thickness [mm]	Fraction [%]	Cross-sectional area fraction [%]
Skin	2	5	5
Fat	5	12	15
Muscle	10	24	75
Bone	15	36	5
Lung	10	24	-

Table 5.1), and a weighted average was used to estimate the relative permittivity and conductive as 42.8 and 2.1 S/m, respectively.

5.3.3 Beampattern characterization

The beampattern of a ULA or uniform rectangular array (URA) is well defined in the far field and can be characterised by an analytical model. Although plane wave propagation cannot be assumed in the case of CMI, beampattern analysis will aid in the understanding of characteristics of the URA used in the chest measurements. In order to obtain the beampattern of a 2-dimensional URA in both x and y direction, an adjusted version of equation 5.6 is used, which makes the delay factor dependent on both the polar coordinates azimuth ϕ and elevation θ :

$$P_x(\theta, \phi) = \sum_{m=0}^{M-1} \mathbf{w}^H e^{-jm \frac{\omega d}{c} \sin \theta \cos \phi} \quad (5.9)$$

$$P_y(\theta, \phi) = \sum_{n=0}^{N-1} \mathbf{w}^H e^{-jn \frac{\omega d}{c} \sin \theta \sin \phi} \quad (5.10)$$

With M and N the number of elements in x and y direction, respectively. The total array beampattern is then defined as [136]:

$$P(\theta, \phi) = P_x(\theta, \phi)P_y(\theta, \phi) \quad (5.11)$$

The polar coordinate system (θ and ϕ) with its relation to the Cartesian system (x and y) is illustrated in Fig. 5.4.

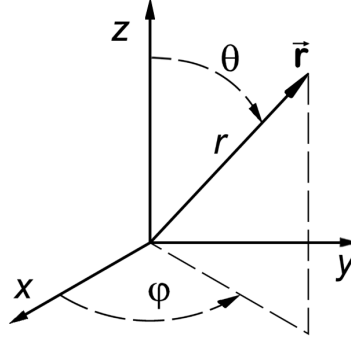


Figure 5.4: The polar coordinate system with its relation to the Cartesian system, for a uniform rectangular array in the xy plane. Elevation θ and azimuth ϕ are illustrated. Direction z indicates broadside: The direction normal to the array (figure from [137]).

Instead of probing the beamformer with a narrowband complex sinusoid (equation 5.1), the waveform as transmitted by the X2 was measured (at PG0) and used as an input to beamformer $\mathbf{w}$. To ensure fine (sub-sample) steering, signal delays were applied in the frequency domain. Because the output of a beamformer probed with the X2 pulse has a certain duration, beamformer response P was defined as the RMS of the resulting signal. Positions of Tx and Rx were defined by the physical array geometry as used in the chest imaging experiments. A MIMO mode of operation was considered, therefore the virtual element positions were computed (according to equation 5.7) and used for beampattern analysis. An average chest permittivity of $\epsilon = 36$ was assumed. EM propagation velocity was set to $c/\sqrt{36}$, and held constant across frequency (dispersive properties ignored). Different values for permittivity were tested for their effect on the resulting beampattern. The 2-dimensional beampattern is visualised in the UV coordinate system: A projection of a semi-spherical space onto a 2-dimensional plane, with the following relation to elevation and azimuth:

$$U = \sin \theta \cos \phi \quad (5.12)$$

$$V = \sin \theta \sin \phi \quad (5.13)$$

Because the final imaging application will likely only utilise a 2-dimensional imaging slice (considering depth), the 1-dimensional beampattern is computed across elevation angles, at fixed azimuth. Beamformer performance is assessed by visual inspection of the generated beampat-

tern. In particular, the amplitude of sidelobes will be considered along with the elevation angle at which grating lobes occur.

The obtained beampattern will allow for further analysis of the following three parameters:

- **Element weights:** A common method of suppressing sidelobes in the beampattern involves multiplying individual elements (virtual or real) with a weight (α_m in equation 5.4), analogues to the windowing operation which typically precedes the Fourier transform. A 2-dimensional window grid is defined and values for α are determined for each virtual channel location using interpolation. Chebyshev windowing (for sidelobe reduction at the cost of beam broadening) will be compared to a rectangular window.
- **Virtual element pruning:** The URA has a high number (44%) of redundant virtual channels (virtual channels at identical or near identical positions). An experimentation with channel pruning is performed.
- **Imaging plane rotation:** Whereas a typical 2-dimensional imaging plane might be chosen along the x or y direction of the array, the plane may be rotated to any azimuth. The effect of azimuth angle on the resulting beampattern will be investigated.

5.3.4 Numerical simulation

To predict imaging system performance and test the effect of various imaging algorithm parameters, FDTD simulations are performed in gprMax: An open source, python-based EM simulation package [114]. The phantom scenario (circular array) is being simulated in 2D, and the chest imaging scenario (rectangular array) is being simulated both in 2D and 3D. Antenna positions were defined in accordance with the array geometries used for real measurements. For 2D simulations, the real-world 3D geometry was projected onto a 2D plane, resulting in co-location of Tx and Rx elements in case of the circular geometry. For all simulations, a homogeneous medium was created with dielectric properties resembling real world measurements. Dimensions of the phantom simulation corresponded to dimensions of the real-world semi-solid phantom (diameter of 11.3 cm). For the chest geometry, a plane of 12 x 24 cm and a space of 12 x 16 x 10 cm were created for the 2D and 3D simulation, respectively. Point reflectors were placed in the simulated geometry so that the system's PSF could be obtained. For the phantom scenario, up to three point targets were positioned in accordance with the target position in the real world measurement (specified in Fig. 5.6). Simulations were also performed without a target present, to serve as an 'empty room' recording for differential clutter removal. Antennas were simulated as a point source or receiver: Modelling the antenna behaviour was beyond the scope of this simulation and crosstalk of simulated perfect geometries can easily be removed if desired. Transmitters were excited using a waveform identical to the signal being transmitted

by the X2 radar. Simulations were run both with media having constant dielectric properties, and with media having dispersive dielectric properties. Dispersive properties were modelled using a two-pole Debye model with parameters obtained by particle swarm optimisation fitting to either measured dielectric properties (phantom scenario) or properties obtained from published data (chest scenario) [18]. Details on assumed medium properties are given in Section 5.3.2. Details on used simulation methods and dispersive model fitting are given in Appendix B.

Obtained simulated datasets were processed in the same manner as data obtained in imaging experiments. Details on data preprocessing, beamforming, assessment of performance and result visualisation are given in Section 5.3.7.

Simulated datasets were used to investigate the following questions:

- **Window length:** Which window length is optimal?
- **Calibration and permittivity error:** What is the effect of module calibration error and permittivity estimation error on the PSF?
- **Clutter rejection:** Can the skin artefact and cross talk be successfully eliminated by the grouped averaging approach?
- **Propagation model:** Which signal propagation model is most suitable
- **Attenuation compensation:** Does signal attenuation need to be compensated for?
- **Beamforming algorithm:** Does the DMAS beamforming algorithm indeed result in better performance than DAS?

And for the rectangular array specifically:

- **Array density:** What is the effect of array density and does the rectangular array have a suitable element density?
- **Imaging plane rotation:** Can system performance be improved by rotating the imaging plane (adjusted azimuth)?
- **Array tapering and pruning:** Can system performance be improved by applying element weights or by pruning redundant virtual channels?

5.3.5 Experimental data collection

Imaging was performed using a system of synchronised coherent UWB RoC modules¹ and body-coupled antennas in three different arrays², both on the human body and on a phantom³. More

¹Radar modules were developed by Dr. Kristian Kjelgård (UiO) and Novelda AS, Norway.

²Radar control system, body-coupled antennas, module shielding and antenna arrays were developed by Mathias Tømmer (UiO).

³Recordings were performed in collaboration with Mathias Tømmer (UiO).

details and characterisation of radar modules and antennas are given in Section 3.4. Recordings were made at a number of locations on the body, and ECG data were collected simultaneously. The purpose of the phantom recording was to test the system's static imaging ability in a controlled environment. Chest and thigh recordings were performed to test the dynamic imaging ability, with the aim of imaging the participant's heart and femoral artery, respectively. The three different array configurations (circular, dense semicircular, and rectangular), as well as their properties are detailed in Fig. 5.1. The three different measurement scenarios (phantom, thigh, and chest), as well as the array placements relative to the imaging object and the different recording conditions are detailed in Figs. 5.6, 5.7 and 5.8.

Experimental setup

In this work, up to nine radar modules were operated simultaneously. Radar modules consisted of the XeThru X2 single-chip radar (Novelda AS, Kviteseid, Norway), with lowered centre frequency and added gain. The transmitted waveform is a Gaussian modulated sinusoid, with adjustable centre frequency and a typical -10 dB bandwidth of 2.5 GHz. Adjustable amplification of up to 8 dB was added both on the transmit and receive paths for increased sensitivity. A BeagleBone system was used to control all radar modules, data was transferred from BeagleBone to data acquisition computer over Ethernet cable. The system was programmed and controlled through Python.

Each radar module connected to one Tx and one Rx antenna. Body-coupled wideband monopole antennas of 2 by 2 cm were used. Low-loss dielectric spacers of 2.5 mm thickness and permittivity matching the skin ($\epsilon_r \approx 30$) reduced losses in the near field and improved radiation into tissue [87].

Each radar module was enclosed in a 3D printed box, constructed from conductive graphene filament and spray-painted in conductive paint to improve shielding. The antennas were fixed in a separate 3D printed enclosure, also conductive spray-painted, and lined on the inside with thin microwave absorbing materials to suppress standing waves. The module box and antenna fixture are shown in Fig. 5.5. The distance between Tx and Rx antenna (centre to centre) was 3 cm. For the circular array and the rectangular array, the antenna fixture fitted into the module box, with antennas connecting to the radar module via SMP snap connectors. Coaxial cables were avoided because previous experimentation had shown that cables were prone to leakage, and cable movement caused significant recording artefacts. Because of the physical dimensions of the radar module box (4.0 by 5.7 cm), a dense array could not be created with the module box directly adjacent to the antenna fixture. For the dense semicircular array, antenna connectors were extended using SMP cables.

Radar modules boxes and antenna fixtures were fixed in 3D printed arrays. The circular array

consisted of a ring with nine inwardly protruding plastic bolts to which the module boxes were attached. The bolts allowed for an adjustable inner array diameter, ranging from 11.0 to 16.0 cm. The minimum inter-antenna distance was 3.8 cm. The dense semicircular array used a similar bolt positioning system, with only the antenna fixtures attached, so that an inter-antenna spacing of 2.0 cm could be achieved at an array diameter of 11.0 cm. Eight modules could be mounted into the rig, spanning an angular coverage of 144° . The rectangular array consisted of eight module boxes with antenna fixtures, attached in a four by two grid. Array density was limited by module box dimensions. Total array aperture (measured at centres of outermost antennas) was 8.7 by 12.0 cm.

ECG data were recorded at a sample rate of 250 Hz, using an Arduino and Olimex EKG shield. Electrodes were placed on the chest, with the left and right electrodes slightly inferior to the clavicles and medial to the shoulders, with the ground electrode placed on the left lower rib cage.

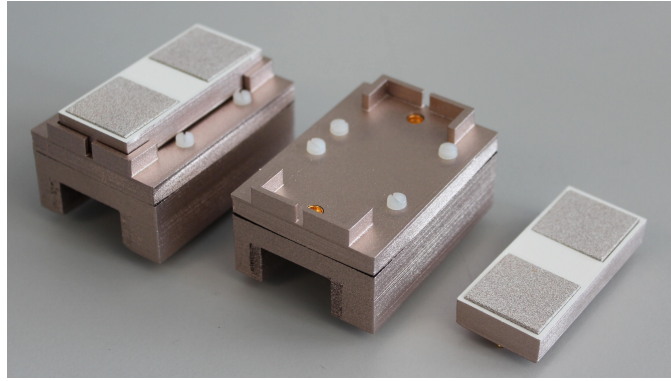


Figure 5.5: Radar module box and antennas (left), with separate module box (middle) and antennas in fixture (right).

Measurement protocol

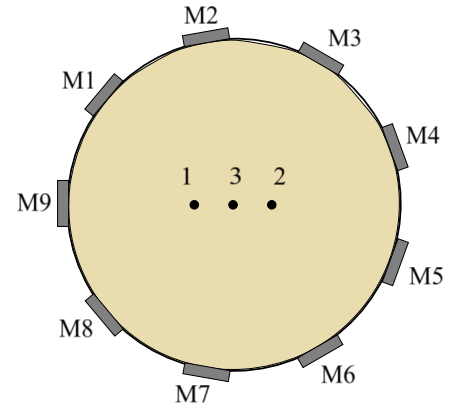
All in-human recordings were performed on a single volunteer. The male participant was 29 years of age, had a BMI of 20.5, and reported no cardiovascular disease. Approval for the study was obtained from the Imperial College London institutional ethics committee (letter of approval attached in Appendix G). For phantom imaging, a cylindrical semi-solid phantom with diameter of 11.3 cm was used. Phantom dielectric properties were measured, and are presented in Section 5.3.2.

Phantom imaging The circular rig array and dense semicircular arrays were used for phantom measurements. Antennas were firmly pressed against the phantom by tightening the module adjustment bolts. In the case of the dense semicircular array, an elastic band was used

Circular array: Phantom

Semi-solid cylindrical phantom with steel bolt targets (5 mm diameter, height as phantom). Inter-target distance was 13 mm. Module 9 was positioned along target axis. Full MIMO data were recorded across all PG settings. Recordings were performed under the following four conditions:

- Empty (no targets)
- 1 Target: 1
- 2 Targets: 1 & 2
- 3 Targets: 1 & 2 & 3



Circular array: Thigh

Transverse cross section of left thigh. Module 2 was positioned most anterior. Full MIMO data were recorded at the lowest PG setting, at an image frame rate of 13.0 fps. SIMO data were also recorded with only module 1, or only module 9 acting as a transmitter, at 142.9 fps. Recordings were performed at two different positions:

- Position 1: 22.5 cm superior to patella
- Position 2: 6.5 cm superior to patella

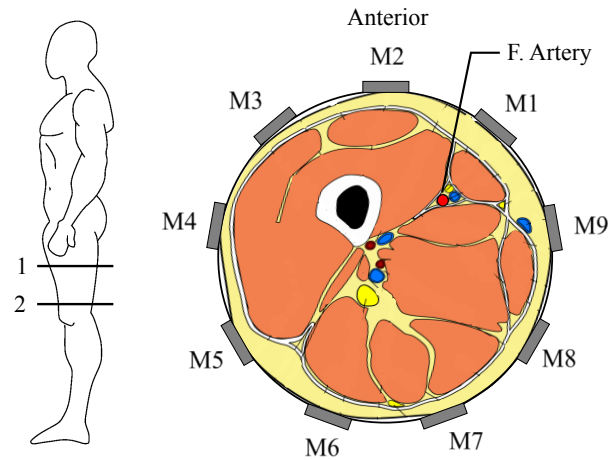


Figure 5.6: Overview of measurements performed using the circular array. Radar modules are indicated with an ‘M’. Thigh cross-section figure modified from [138]. Human outline figure modified from [139].

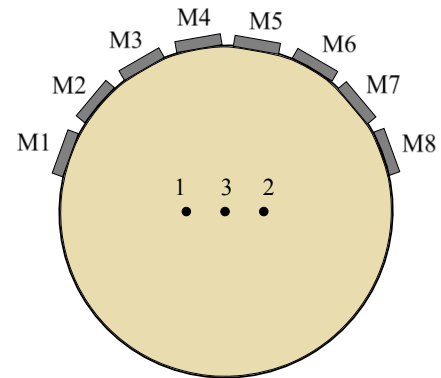
to ensure that antennas were pressed on the phantom surface. Metal bolts acted as targets and were carefully inserted in the phantom (circular array) or extracted from the phantom (dense semicircular array) between recordings, as illustrated in Figs. 5.6 and 5.7. Between recordings, it was ensured that nothing in a 1.5 m range of the system was altered or moved, to prevent changes in the clutter scene which could be picked up by backscattered signal.

Thigh imaging The circular rig array and dense semicircular arrays were used for thigh measurements, at two different heights. Antennas were firmly pressed against the skin by tightening the module adjustment bolts. In the case of the dense semicircular array, an elastic band was used to ensure that antennas were pressed on the skin. Array orientation relative to anatomy is illustrated in Figs. 5.6 and 5.7. For each condition, 60 s of data were recorded, both while the participant was breathing and while holding breath. During recordings, the

Dense semicircular array: Phantom

Semi-solid cylindrical phantom with steel bolt targets (5 mm diameter, height as phantom). Inter-target distance was 13 mm. Array centre was positioned on axis perpendicular to target axis. Full MIMO data were recorded across all PG settings. Recordings were performed under the following four conditions, but in reverse chronological order (targets were extracted):

- Empty (no targets)
- 1 Target: 2
- 2 Targets: 1 & 2
- 3 Targets: 1 & 2 & 3

**Dense semicircular array: Thigh**

Transverse cross section of left thigh. Module 6 was positioned most anterior. Full MIMO data were recorded at the lowest PG setting, at an image frame rate of 17.4 fps. SIMO data were also recorded with only module 4 acting as a transmitter, at 171.5 fps. Recordings were performed at a single position:

- Position: 6.5 cm superior to patella

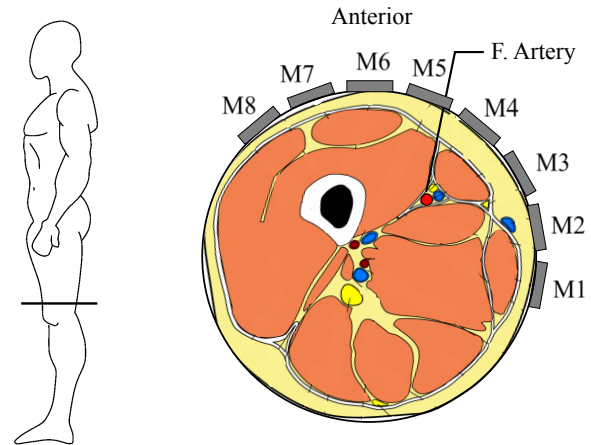


Figure 5.7: Overview of measurements performed using the dense semicircular array. Radar modules are indicated by an ‘M’. Thigh cross-section figure modified from [138]. Human outline figure modified from [139].

participant was comfortably seated and asked to minimise movement. Apart from MIMO data, SIMO data were also recorded with the module which was hypothesised to be nearest to the femoral artery acting as transmitter.

Heart imaging The rectangular array was used for chest measurements. Array positioning is specified in Fig. 5.8. The antenna array was placed on a flat surface, antennas facing up, with the participant lying on the antenna array in prone position. This ensured good antenna-skin contact, despite the irregular shape of the human chest. For each condition, 60 s of data were recorded both while the participant was breathing and while holding breath. During recordings, the participant was asked to minimise movement.

Rectangular array: Heart

Rectangular array was placed on the subject chest, allowing for cardiac imaging in the sagittal and transverse plane. The sagittal imaging plane is illustrated in green in the figure on the right. The array centre was placed lateral to the subject central axis, with transmitters of modules 5 – 8 positioned in the medial plane. The most superior modules (4 and 5) were positioned 11.5 cm inferior to most superior portion of the subject sternum (manubrium). Full MIMO data were recorded at the lowest PG setting, at an image frame rate of 17.3 fps. MIMO data were also recorded with only modules 1 and 5 acting as transmitters, at 69.4 fps.

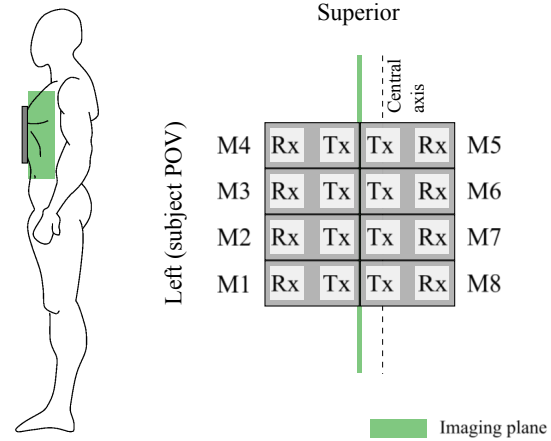


Figure 5.8: Overview of measurements performed using the rectangular array. Radar modules are indicated by an ‘M’. Human outline figure modified from [139].

5.3.6 System calibration procedure

It was found that delay offsets exist between radar modules (Section 3.4.5). Calibrating of all delay offsets in the system is paramount for coherent summation and thus beamforming. In the current set-up, the following delays were found to require calibration:

- **Sample delay offset (SDO):** The absolute delay within an individual module, from pulse transmission to onset of sampling. Denoted as τ_r in Fig. 5.9.
- **Module delay offset (MDO):** The absolute delay between the system (BeagleBone) master clock and individual modules, which dictates the timing of pulse transmission. Denoted as τ_1 and τ_2 in Fig. 5.10, for modules 1 and 2, respectively.

Delay offsets were obtained and calibrated for as specified below:

Sample delay offset calibration The sample delay offset (SDO) was analysed for each module individually by measuring the direct path from Tx to Rx with a jumper cable. Coarse SDO calibration was performed prior to data collection, to ensure that the full sample duration (6.4 ns) was utilised for signal collection. Fine SDO calibration was done offline: Peak detection of the transmitted pulse was performed, a matrix was constructed with SDO values for all modules across all PG settings, and compensation was done prior to beamforming.

Module delay offset calibration In a multistatic recording, an module delay offset (MDO) of the Tx module results in a delayed received signal, whereas an MDO of the Rx module

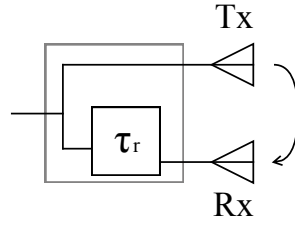


Figure 5.9: Sample delay offset: The absolute delay from pulse transmission to sampling onset.

results in an advanced received signal. In each Tx-Rx pair, MDO of both modules need to be compensated for. Measuring absolute MDO would be a challenge. Instead, relative MDO between modules was measured. To find relative module delays, an arbitrary module (module 1) is taken as reference, and outbound and return signals are collected between module 1 and each of the remaining 8 modules. Data are recorded using the circular rig in which antennas are fixed, therefore distance of outbound and return paths are equal. Received signals are compensated for SDO prior to MDO calibration.

Pulse arrival time for each signal was determined by finding an arbitrary fiducial point on the received signal: The first peak that exceeds a certain fraction of maximum signal amplitude. When considering the outbound (S_A) and return (S_B) signal between modules 1 and 2, the measured delay of the outbound signal τ_A and return signal τ_B are given by:

$$\tau_A(Tx = M1) = \tau_c + \tau_2 - \tau_1 \quad (5.14)$$

$$\tau_B(Tx = M2) = \tau_c + \tau_1 - \tau_2 \quad (5.15)$$

Where τ_c is some constant delay, τ_1 is the MDO of module 1, and τ_2 is the MDO of module 2. If we consider $\tau_\delta = \tau_A - \tau_B$, we can express τ_δ as a function of MDO:

$$\tau_\delta = 2(\tau_2 - \tau_1) \quad (5.16)$$

Because M_1 is assumed the reference module, we choose $\tau_1 = 0$, therefore:

$$\tau_2 = \tau_\delta / 2 \quad (5.17)$$

The module delay is obtained for each module and PG setting.

Using SDO τ_s and MDO τ_m , any recorded signal can be corrected. For signal S_{raw} recorded from the Tx-Rx combination $Tx = M1$, $Rx = M2$, the received signal is corrected as follows:

$$S_{corrected}(t) = S_{raw}(t - \tau_{m1} + \tau_{m2} + \tau_{s2}) \quad (5.18)$$

Outbound and return signals, before and after calibration, are shown in Fig. 5.11 for an arbitrary

Tx-Rx combination.

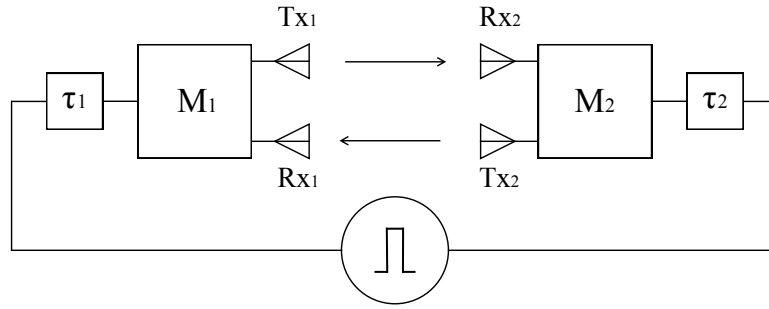


Figure 5.10: Module delay offset: The absolute delay between the system master clock and individual modules.

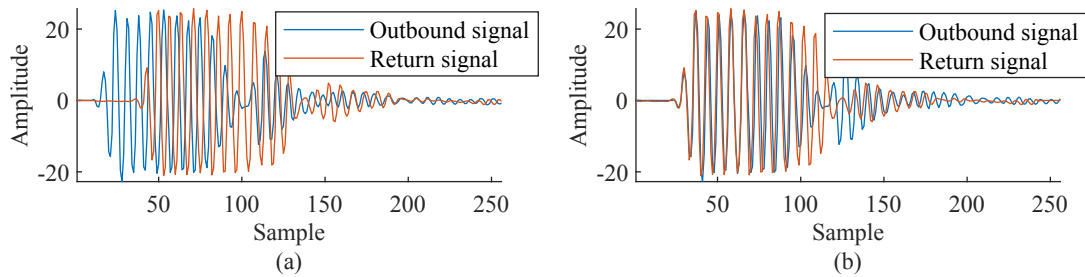


Figure 5.11: Signal delay calibration, verified on an arbitrary Tx-Rx pair. Before (a) and after (b) calibration.

5.3.7 Imaging Algorithm

The following section describes the algorithms that were applied to collected data (both simulated and experimental) for final image formation. A schematic overview of the signal processing pipeline is given in Fig. 5.12.

Preprocessing

Delay offset calibration Time calibration was performed to a precision of 1 sample (~ 26 ps), as described in Section 5.3.6. Single sample precision was found to be sufficient: Simulations were run in Matlab to assess the expected attenuation of coherently summed sinusoids, with $n=10$ and $n=100$ input signals. Input signals were given a random jitter ranging from -1 to 1 sample. Simulations were executed 100,000 times, and the median of the obtained probability density function was used to predict attenuation as a result of calibration imprecision. An attenuation of 0.13 dB was found, which was considered acceptable. By increasing the calibration precision, attenuation could be reduced but this was deemed unnecessary.

Filtering Obtained radar frames were zero-phase bandpass filtered (across fast time), to eliminate DC offset, signal trends, and out-of-band noise. The passband was determined by measuring the -30 dB bandwidth of transmitted signals across all PG settings. A challenge was encountered when applying attenuation compensation in the beamforming step: FFT filtering was found to smear out the carrier wave across the radar frame, which became noticeable at very path lengths (where high attenuation compensation was applied). Ultimately, wavelet filtering was found to be immune to this effect as the continuous wavelet transform does not assume signal stationarity. A wavelet filter bank was designed, using the analytical Morse wavelet.

Skin artefact removal Two approaches were taken to remove the skin artefact and crosstalk signals. When an ‘empty room’ dataset was recorded (in case of the phantom measurements), differential imaging was performed whereby the empty dataset could be subtracted from the dataset containing a target. When differential imaging was not possible, the skin artefact was estimated by averaging a group of channels with similar geometry. The Euclidean distance between Tx and Rx element was determined for all channels, and k-means clustering was applied to group channels according to their geometry. Skin artefact removal was performed on each channel by subtracting the group average. Because array imperfections could lead to jitter in the skin artefact or cross talk signal, the Woody average was applied: Cross-correlation between signals was optimised for, by iteratively changing a small delay applied to the input signals. The set of lags that resulted in the strongest cross-correlation was used to align and subtract the obtained skin artefact estimate.

Beamforming

Window length A window length of 5 samples was used, which corresponds to 130 ps, or approximately $1/2\lambda$. This window length was found to be a good trade-off between SNR and robustness to error in the dielectric estimate. Analysis of the effect of window length in Section 5.4.1.

Propagation model From Section 3.5 it was found that, at the bandwidth of interest and in the media considered in this study, the dispersiveness of the dielectric properties could be ignored. Pulse distortion is minimal and does not affect the peak amplitude of the transmitted waveform at the propagation path lengths considered. In addition, dispersive attenuation compensation poses significant challenges for high frequency noise. Only permittivity and conductivity at centre frequency were considered. As a result of this, steering delays could be applied in the time domain by truncation and concatenation of received radar frames. Compensation

for signal attenuation was found to improve imaging performance in case of the rectangular array, but not in case of the circular array. Attenuation compensation was therefore only applied to data obtained using the rectangular array.

Beamforming algorithm For this work, a simple and intuitive beamforming algorithm was chosen for ease of analysis and transparency. The DMAS beamforming algorithm provides superior clutter rejection when compared to DAS, at the cost of increased computational time. DAS was used for beamforming on dynamic data (heart and artery imaging in the chest and thigh measurements, respectively). For static imaging, consecutive frames could be averaged prior to beamforming, reducing the amount of data that needed processing. DMAS was therefore used for static imaging. All 2D images were computed with 2 by 2 mm voxel size.

Imaging plane The 2-dimensional rectangular array allowed for 3-dimensional imaging. To aid in visualisation and avoid unnecessary processing, only 2D slices were imaged. It was found that a rotated imaging plane (adjusted azimuth) resulted in more desirable beamformer characteristics (Section 5.4.2). It is also known that the human heart is slightly slanted: The apex (most inferior part) of the heart is pointed towards the left downside. Two imaging planes were therefore chosen to align with the heart's long and short axis, at 23 and 113° counter-clockwise from the sagittal plane, respectively. To aid in understanding, the concept of imaging plane rotation is illustrated in Fig. 5.20.

Element weights Array tapering by applying weights to individual channels was not found to improve imaging results. No element weights were used.

5.3.8 Data analysis and performance evaluation

Static imaging analysis The obtained images and PSFs were plotted as amplitude heatmaps on a logarithmic scale. In an analogy to US imaging, these 2D maps may be referred to as B-mode images. B-mode images were analysed by visual inspection. The -6 dB contour (denoting an amplitude reduction of 50 %) was used to aid in visual inspection. In addition, the following metrics were used to quantify performance:

- The SCR is defined as the logarithmic ratio of signal strength at a predefined voxel or set of voxels, to signal strength at all other voxels.
- The signal-maximum-to-clutter-maximum ratio (SmCmR) is the ratio of maximum signal strength at a predefined voxels, to maximum signal strength across all other voxels. The SmCmR metric is useful for finding the contrast between competing signal peaks.

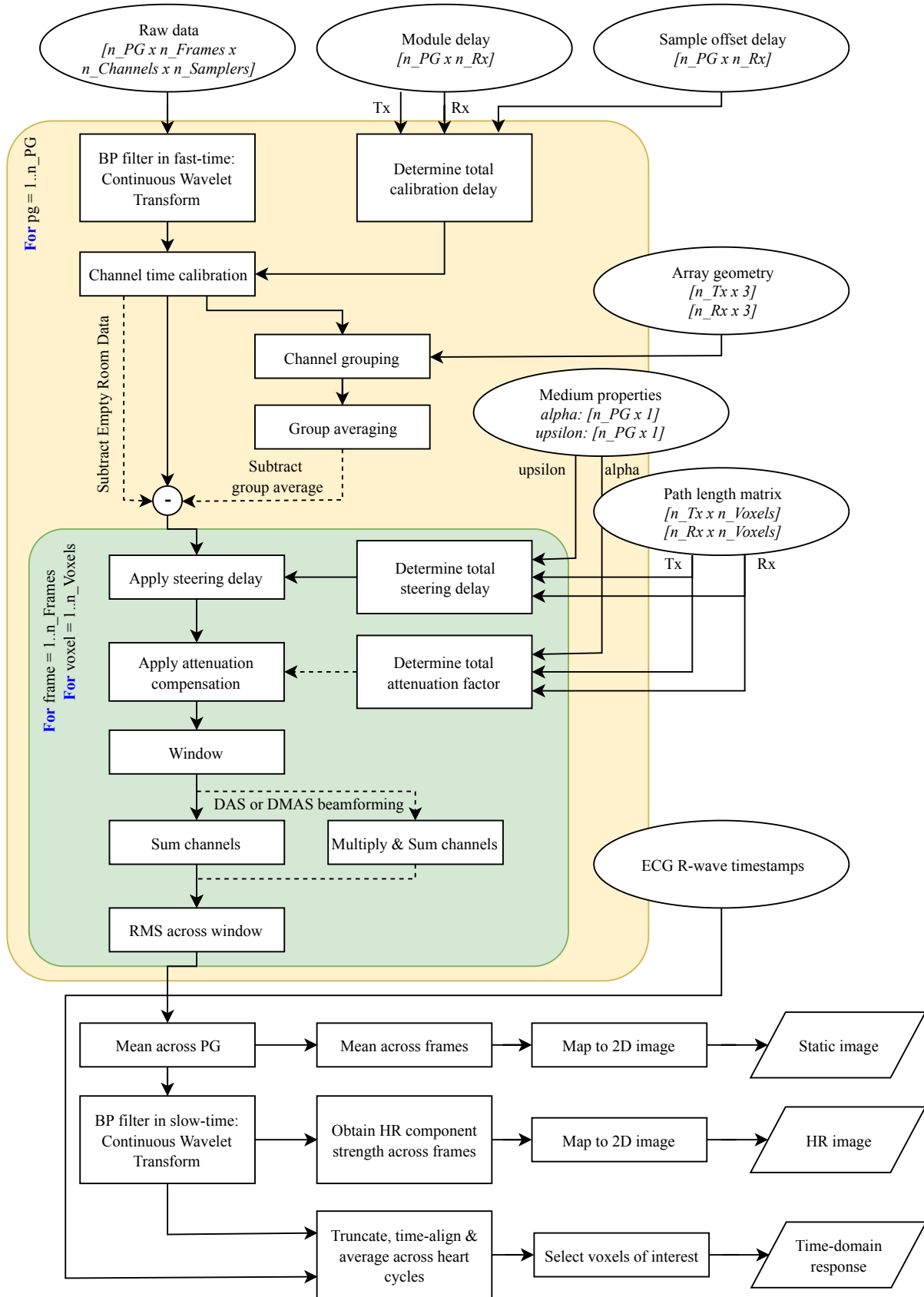


Figure 5.12: Schematic pipeline of imaging algorithm. Ellipses, rectangles and parallelograms denote data, processes and outcomes, respectively. Dashed lines denote an optional process.

- The location error is defined as the Euclidean distance between the heatmap maximum and a predefined position, typically the known target location.

Dynamic imaging analysis To increase SNR, radar imaging frames and ECG data were aligned using the fiducial ECG markers (R wave), and radar data over multiple heart cycles were averaged. A subset of heart cycles with similar inter-beat-intervals were chosen to avoid signal smearing. Visualisation of obtained video data was done by generating heatmaps of signal amplitude at the HR frequency, which was aimed to identify voxels with maximum cardiac activity. Besides the B-mode images, M-mode (nomenclature from US imaging) plots were used to visualise time series of data at specific voxels.

5.4 Results

5.4.1 Circular array simulation

The circular array and phantom were simulated in 2D, both with and without targets. As a result of the projection of a 3D geometry onto a 2D plane, the Tx and Rx elements were co-located. The simulation software could not deal with this, therefore the Tx=Rx channels were excluded from further beamforming.

Skin artefact removal Different methods of skin artefact removal and clutter rejection were compared. For all comparisons, the DAS algorithm was used with window length of 5 samples, a constant propagation model and no attenuation compensation. Without any form of skin artefact removal, the targets could not be identified (result not shown). When employing the ‘empty room’ recording for perfect clutter rejection (differential imaging), the target was identified and an SCR of 19 dB and SmCmR of 8 dB were obtained for the 1 Target condition (Fig. 5.13a). The location error for this scenario was 0. When using the grouped average subtraction approach, SCR and SmCmR dropped to 12.5 dB and 2.9 dB, still with correct target identification (Fig. 5.13b).

Window length The effect of window length on SCR was tested. A short wavelength was expected to have better spatial filtering (less smoothing) properties. On simulated data, with the dielectric properties of the simulation and beamforming being identical, a window length of 1 sample resulted in highest SCR. A longer window length was hypothesised to be more robust to error in steering delay estimation and potential hardware or calibration imperfections. SCR is shown as a function of window length in Fig. 5.14a, for the 1 Target condition. Apart from

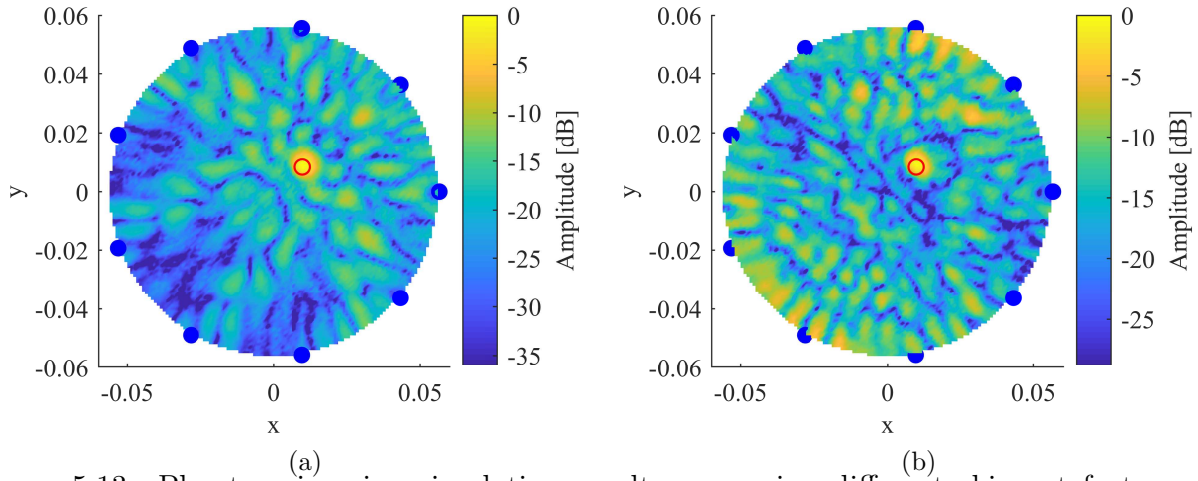


Figure 5.13: Phantom imaging simulation result, comparing different skin artefact removal methods: Differential perfect clutter rejection (a) and grouped average subtraction (b). Radar modules are indicated by blue dots, target location is denoted by a red circle.

the ideal (no error) scenario, SCR was tested with a module calibration error of 3 samples, as well as with overestimated (+10%), and underestimated (-10%) permittivity. A higher window length was found to be more robust against error in the beamforming assumptions. For the 3 Targets condition, an identical pattern was observed, with SCR levelling out at 11 dB (not shown). Examples of DAS beamformed data (1 Target condition) with a -10% permittivity error is shown for a window length of 1 (Fig. 5.14b) and a window length of 10 (Fig. 5.14c).

Propagation model Different models for EM wave propagation were tested in a DAS beamformer. By default, the phantom was simulated with dispersive dielectric properties. A second simulation was run using a phantom with constant dielectric properties. In the beamforming step, the constant velocity model assumed constant propagation velocity across the pulse bandwidth, whereas the dispersive model took the medium's dispersive permittivity into account to delay each frequency at the appropriate velocity (executed in the frequency domain). If constant attenuation compensation was applied, a path length dependent gain factor (constant across frequencies) was used, which was based on dielectric properties at the pulse centre frequency. For dispersive attenuation compensation, a frequency specific gain factor was determined and applied, based on the medium's dispersive dielectric properties. Performance metrics SCR, SmCmR and location error are given in Table 5.2 for the 1 Target condition, across various combinations of simulation, velocity model and attenuation compensation options. Any attenuation compensation resulted in incorrect target identification: Sensor noise at high path length would be amplified to disproportionate levels, as illustrated by Fig. 5.15a. Dispersive attenuation compensation was completely unsuccessful, as high frequency noise would be amplified to even more than in the constant assumption. Little difference existed in performance between the constant and dispersive velocity models, with the constant model performing slightly better

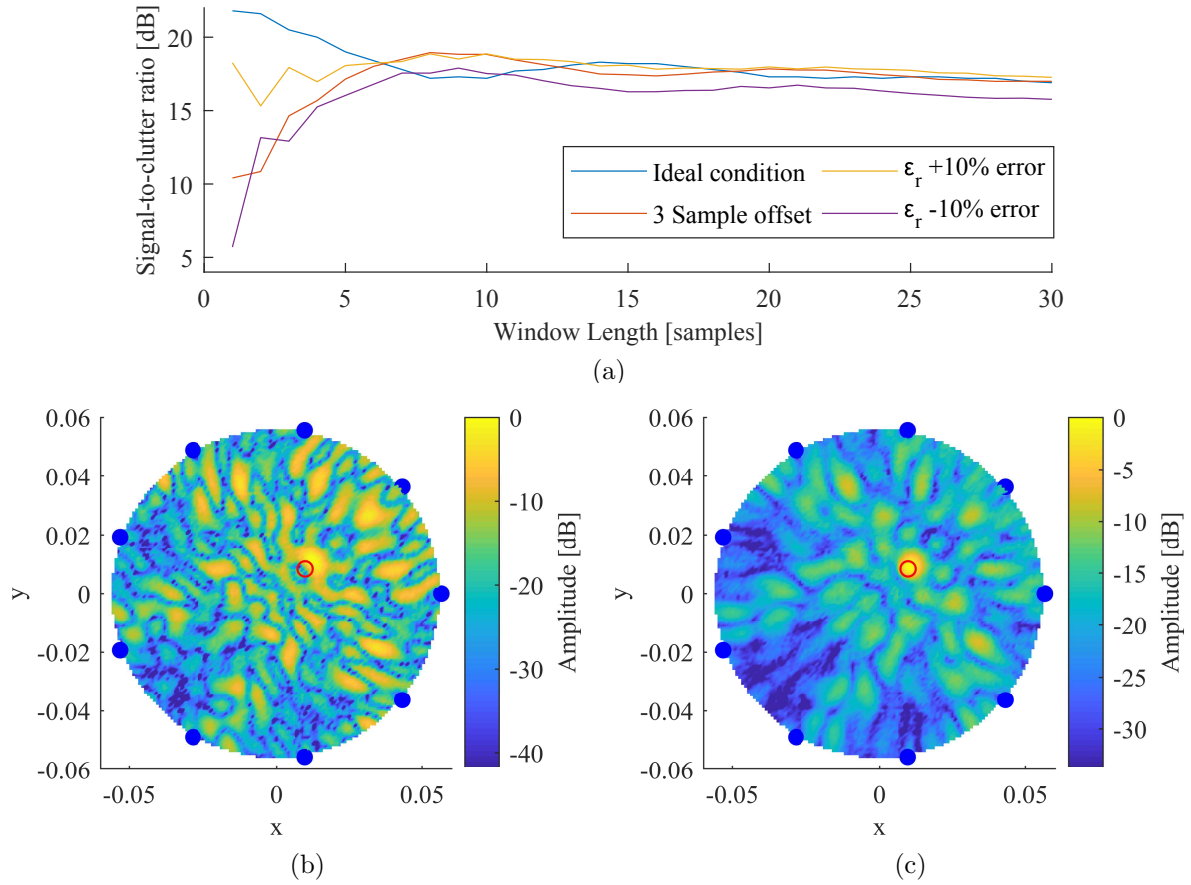


Figure 5.14: Signal-to-clutter ratio versus window length, for ideal beamforming and beamforming with errors on the delay calibration or permittivity assumption (a). Examples of beamforming results using a 1 sample window (b) and using a 10 sample window (c). The target position is denoted by a red circle.

(result shown in Fig. 5.15b).

Beamforming algorithm The DAS beamforming algorithm was compared to the DMAS algorithm. In addition, the effect of window length on SCR and SmCmR was analysed for the 1 Target and 2 Targets conditions. Figs. 5.16a and 5.16b show that the DMAS algorithm is superior in terms of clutter rejection, for all conditions. The obtained relationship between imaging performance and window length is similar to the one from Section 5.4.1, and shows that a shorter window length leads to increased imaging performance. Examples of imaging results of the 2 Targets condition, using DAS and DMAS, are shown in Figs. 5.16b and 5.16d, respectively.

Table 5.2: Imaging performance across various EM wave propagation model options.

Simulated medium	Velocity	Attenuation	SCR [dB]	SmCmR [dB]	Location error [mm]
Dispersive	Constant	None	17.3	7.2	0.65
Dispersive	Dispersive	None	16.7	7.0	0.65
Dispersive	Constant	Constant	11.1	-2.0	55.6
Dispersive	Dispersive	Constant	9.6	-3.2	54.9
Dispersive	Dispersive	Dispersive	-55.0	-72.0	66.6
Constant	Constant	None	12.5	2.9	0.65
Constant	Dispersive	None	11.7	2.2	1.64

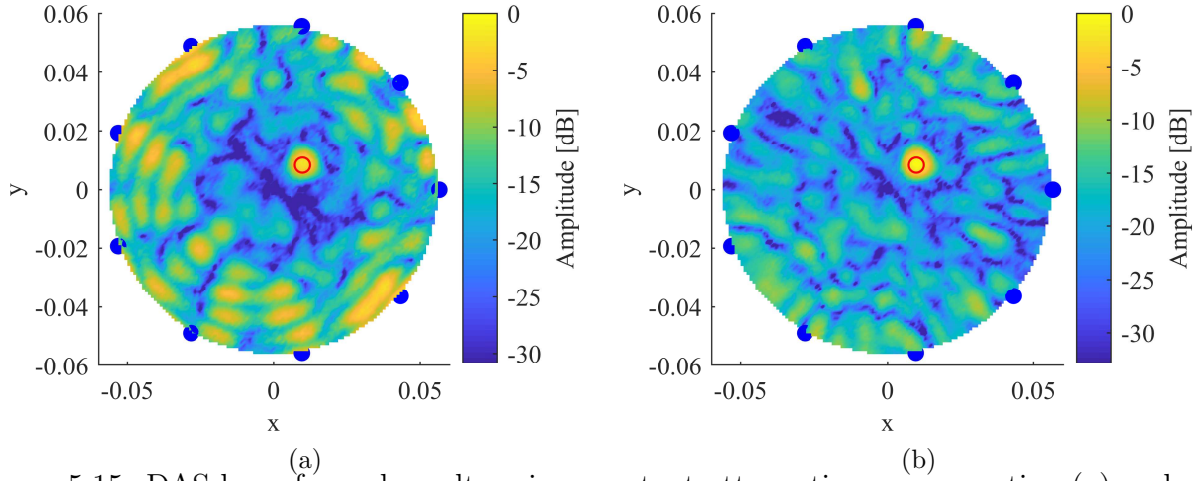


Figure 5.15: DAS beamformed results using constant attenuation compensation (a) and using no attenuation compensation (b). The red circle indicates the target location.

5.4.2 Rectangular array simulation and beampattern analysis

Beampattern analysis

The rectangular array with Tx, Rx, and virtual element positions, is shown in Fig. 5.17a. 1D Beampatterns, obtained using an analytical model, are displayed for imaging directions along the array's x-axis (azimuth = 0°) and y-axis (azimuth = 90°), in Figs. 5.17b and 5.17c, respectively. Sidelobes with amplitude over -6 dB are present, and grating lobes occur at elevation levels below 50° . The array is severely undersampled: Virtual element distance is around 2 cm, whereas the carrier wavelength in tissue is around 1.4 cm and $1/2\lambda$ spacing is required to avoid spatial aliasing.

Imaging plane rotation The 2D beampattern is displayed in Fig. 5.18a. This figure predicts that beampatterns along the x and y direction will result in the worst performance, and beampatterns along a rotated imaging plane should provide reduced spatial aliasing. To aid in understanding of imaging plane rotation, the concept is visualised in 3D in Fig. 5.20. Note that

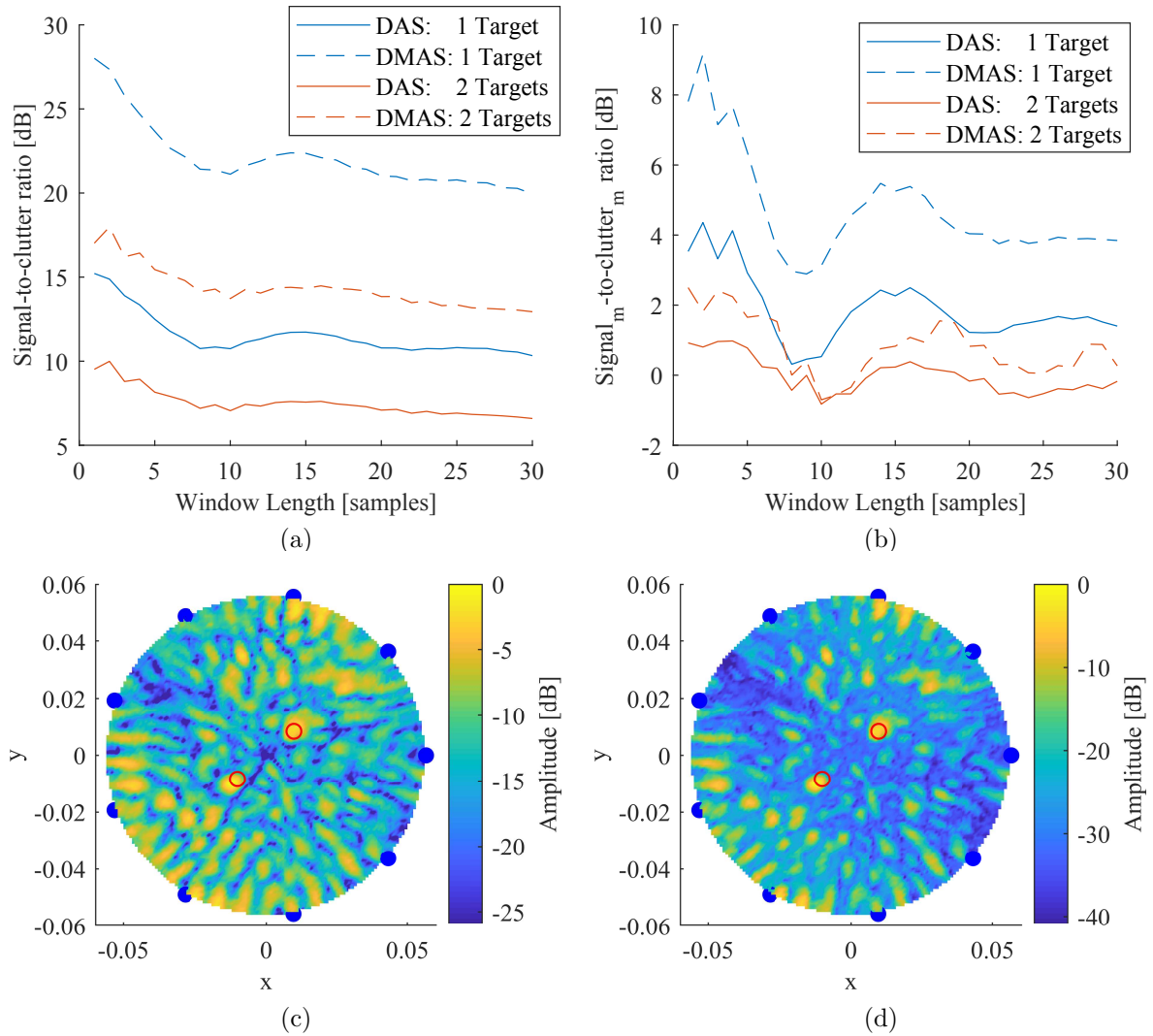


Figure 5.16: Signal-to-clutter ratio (a) and signal-maximum-to-clutter-maximum ratio (b) versus window length, for both DAS and DMAS beamforming. Examples of imaging results of the 2 Targets condition, using DAS (c) and DMAS (d). Target locations are indicated by red circles

in Fig. 5.20, the rectangular array is assumed to lie in the coronal (x-z) plane, as dictated by the standard anatomical coordinate system. The current section however, considers the array geometry to lie in the x-y plane, as is a standard assumption for 2D array analysis and conforms with definitions given in Section 5.3.3.

Beampatterns for four rotation angles are shown in Figs. 5.18b-e, with the imaging direction indicated in 5.18a. The 1D beampatterns along rotated imaging planes show significantly improved performance compared to beampatterns in Fig. 5.17. A beampattern at azimuth = 77° (the symmetrical equivalent of 113° , which would roughly align with the heart long axis), shows side lobes at an elevation of 48.3° , with 8.5 dB amplitude attenuation. Rotation angles are not dependent on medium permittivity.

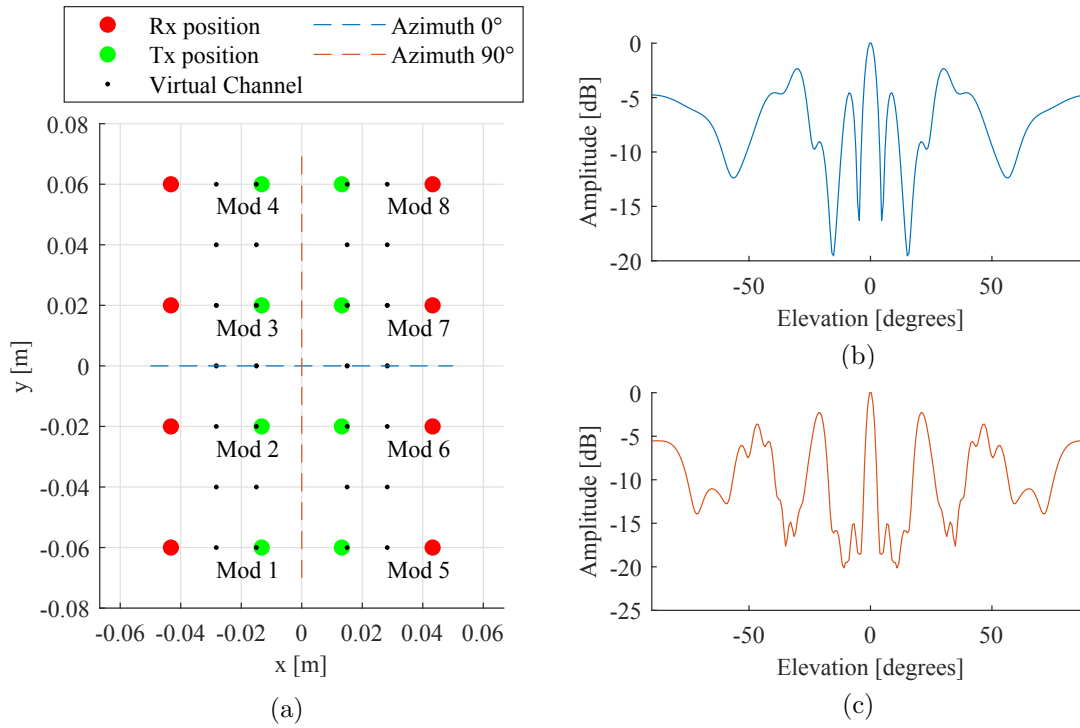


Figure 5.17: Rectangular array with Tx, Rx, and virtual element positions (a). 1D Beampatterns are displayed for imaging directions along the array's x-axis (b) and y-axis (c). Imaging directions are indicated in (a) using dashed lines.

Element weights and element pruning In the current array geometry, no improvements were found by applying element weights or element pruning. A Chebyshev window resulted in main beam broadening (as expected) and smoothing of the overall beampattern, but reduced attenuation outside the main lobe (results not shown). An attempt was made to prune the number of virtual elements, by excluding redundant channels. This did not affect the main lobe or the grating lobes, but the lower number of channels significantly reduced signal suppression from other DOA (results not shown).

2D Numerical simulation

The rectangular array and a homogeneous medium with average chest properties were simulated in 2D, both with and without targets (point reflectors). As a result of the projection of a 3D geometry onto a 2D plane, the Tx and Rx elements were co-located, Tx=Rx channels were excluded from further beamforming.

Array density As expected, the PSF of the rectangular array imaged across the longitudinal axis (azimuth = 90°) showed poor performance. For all beamforming attempts (DAS, DMAS, different propagation models), the target was not correctly identified (results not shown). Along

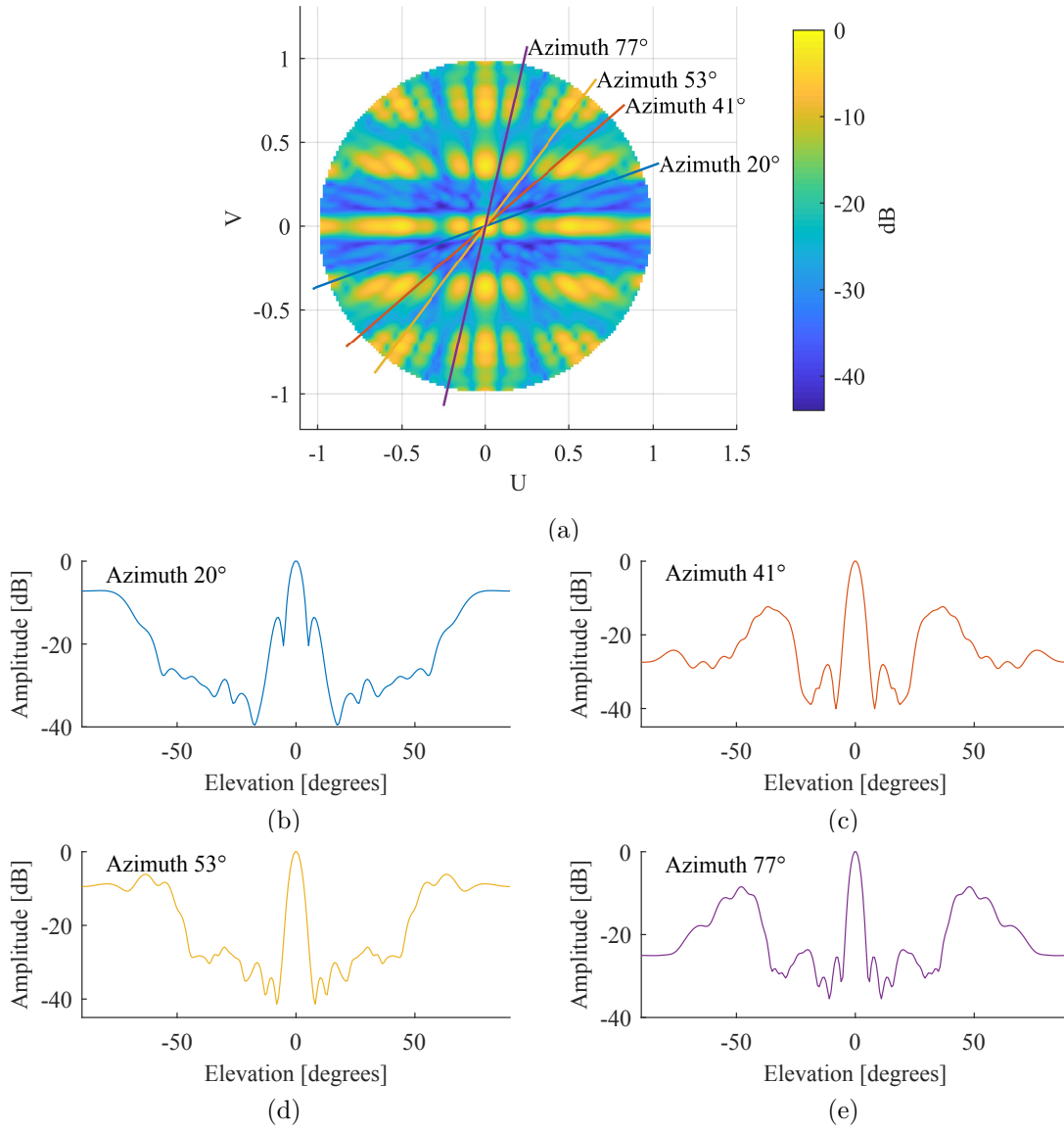


Figure 5.18: 2D Beampattern of the rectangular array (a) and four 1D beampatterns that would be obtained along a rotated imaging plane (b-e). The direction of imaging planes are indicated in (a).

the longitudinal axis, the array is severely undersampled and in a 2D projection, only 4 unique elements exist. A simulation was run with a sufficiently sampled array (sensor distance $d = 1$ cm, MIMO virtual element distance $d = 0.5$ cm), and good imaging results were obtained (Fig. 5.19a). When simulating a 2D projection of the array imaged at a rotated angle (azimuth $= 77^\circ$), the virtual element density increased which resulted in a significantly improved PSF (Fig. 5.19c). The target could correctly be identified with 24 dB SCR using DMAS and grouped average clutter rejection.

Attenuation compensation As opposed to circular array imaging, attenuation compensation was found to be highly beneficial to correct target identification in the rectangular array.

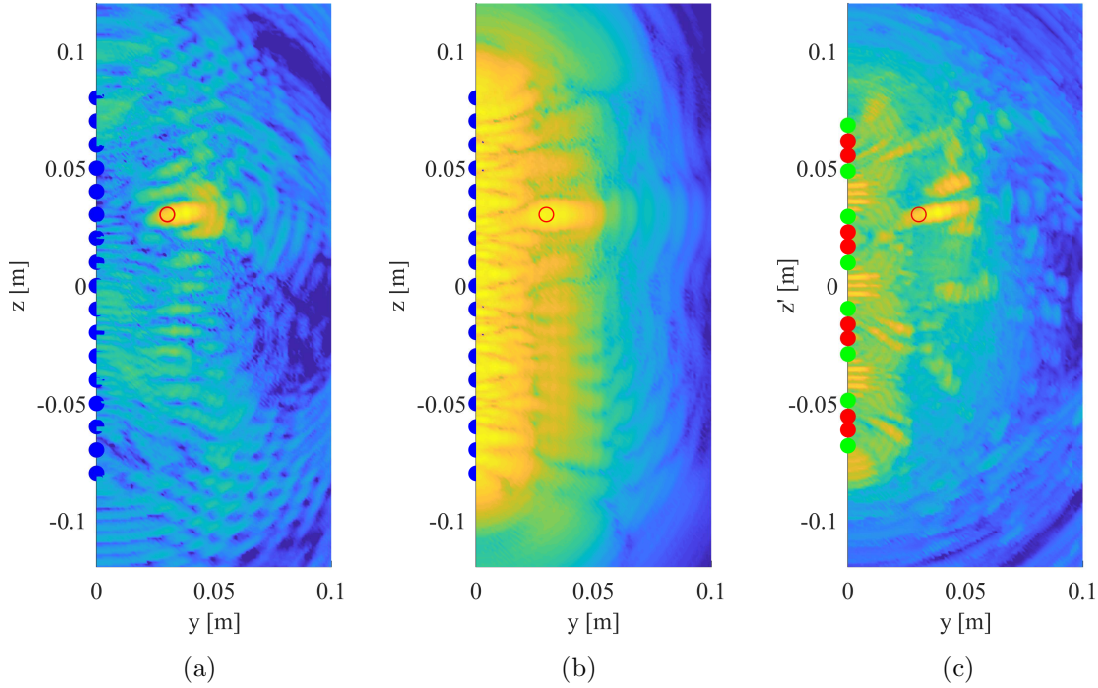


Figure 5.19: Point spread functions of rectangular arrays, as simulated in 2D. Data from sufficiently sampled arrays were DAS beamformed with (a) and without (b) attenuation compensation. Element positions of a rotated imaging plane were projected onto a 2D space (along the z' axis) to assess point spread function at different imaging planes. Results for a 77° rotation angle obtained using DMAS (c). Blue dots indicate Tx/Rx element positions. In (c), red dots indicate Tx positions, green dots indicate Rx positions.

Figs. 5.19a and 5.19b show the PSF of a sufficiently sampled array with and without attenuation compensation, respectively. Using the constant attenuation model, a flat clutter baseline was obtained. DAS beamforming with attenuation compensation (Fig. 5.19a) resulted in a 22.7 dB SCR. When DMAS was used, SCR increased to 36.8 dB.

3D Numerical simulation

Imaging plane rotation A 2D simulation of a rotated imaging plane does not account for differences in signal attenuation due to path length. A 3D simulation, with array geometry as in the experimental setup, is performed to verify improved PSF at rotated imaging plane. Imaging planes, with azimuth ϕ ranging from 1 to 90° , were defined so that they always passed through the target. To aid in understanding of imaging plane rotation, the concept is visualised in 3D in Fig. 5.20.

Array and imaging plane (at $\phi = 75^\circ$) are illustrated in Fig. 5.21a, along with the corresponding PSF (Fig. 5.21b). All PSFs were obtained using DAS beamforming with constant attenuation

compensation and grouped average clutter rejection. Performance metric SCR was found less suitable to compare PSF across ϕ : Mean voxel path length increased at reduced ϕ and attenuation compensation could not prevent high variation in clutter amplitude at remote voxels. Visual inspection of the PSF was aided by using the -6 dB contour plots (Figs. 5.21c-f). SCR versus rotation angle ϕ is shown in Fig. 5.22.

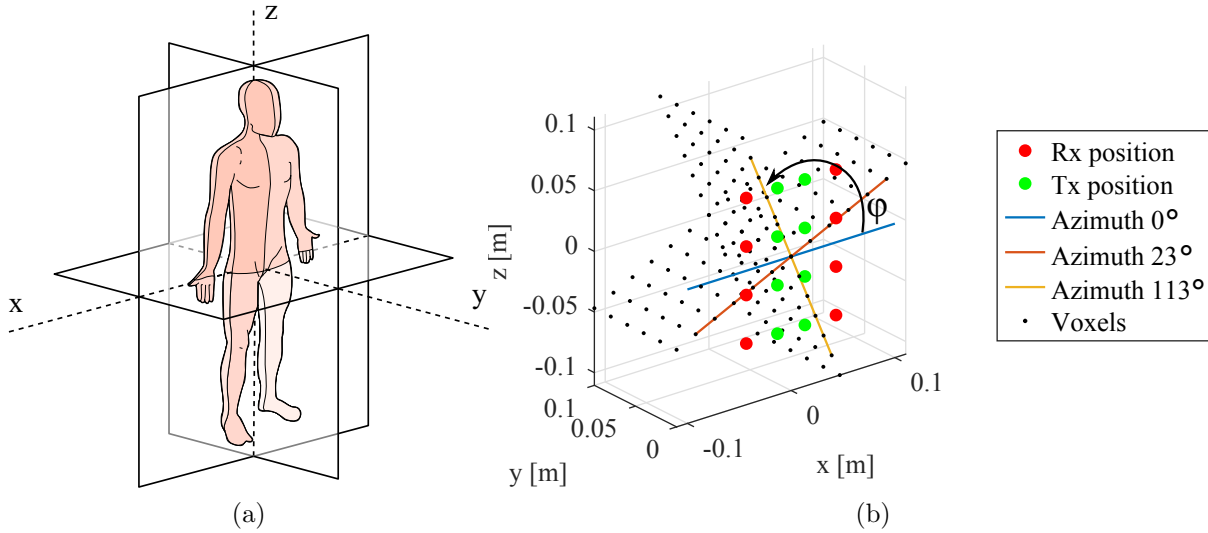


Figure 5.20: Visualisation of imaging plane rotation in 3D, in respect to the rectangular array geometry (a) and participant geometry (b). When the rectangular array is pressed on the participant chest, the imaging plane at azimuth angle (ϕ) 0° (blue line) corresponds to a transverse cross section (x-y plane in the participant geometry). When the imaging plane is rotated in counter-clockwise direction (azimuth increases), the imaging plane can be aligned with the cardiac short axis (azimuth 23° ; orange line), or long axis (azimuth 113° ; yellow line). Figure (a) adapted from [140].

5.4.3 Phantom imaging

Circular array

A semi-solid cylindrical phantom was imaged using the circular array. Due to a recording error, data from one module (M5 in Fig 5.6) was lost. A median dataset was taken across slow-time prior to beamforming. DMAS beamforming was applied to data from PG 0, 1 and 2, and the three resulting images were averaged (arithmetic mean) to obtain the final result. Using the grouped Woody average clutter rejection approach, no targets could be identified in any of the recordings. The resultant skin artefact and antenna crosstalk after clutter rejection greatly exceeded the amplitude of the signal of interest (results not shown). Successful target identification was achieved by taking a differential imaging approach. Only target 1 was correctly identified (result shown in Fig. 5.23a). SCR at target location was

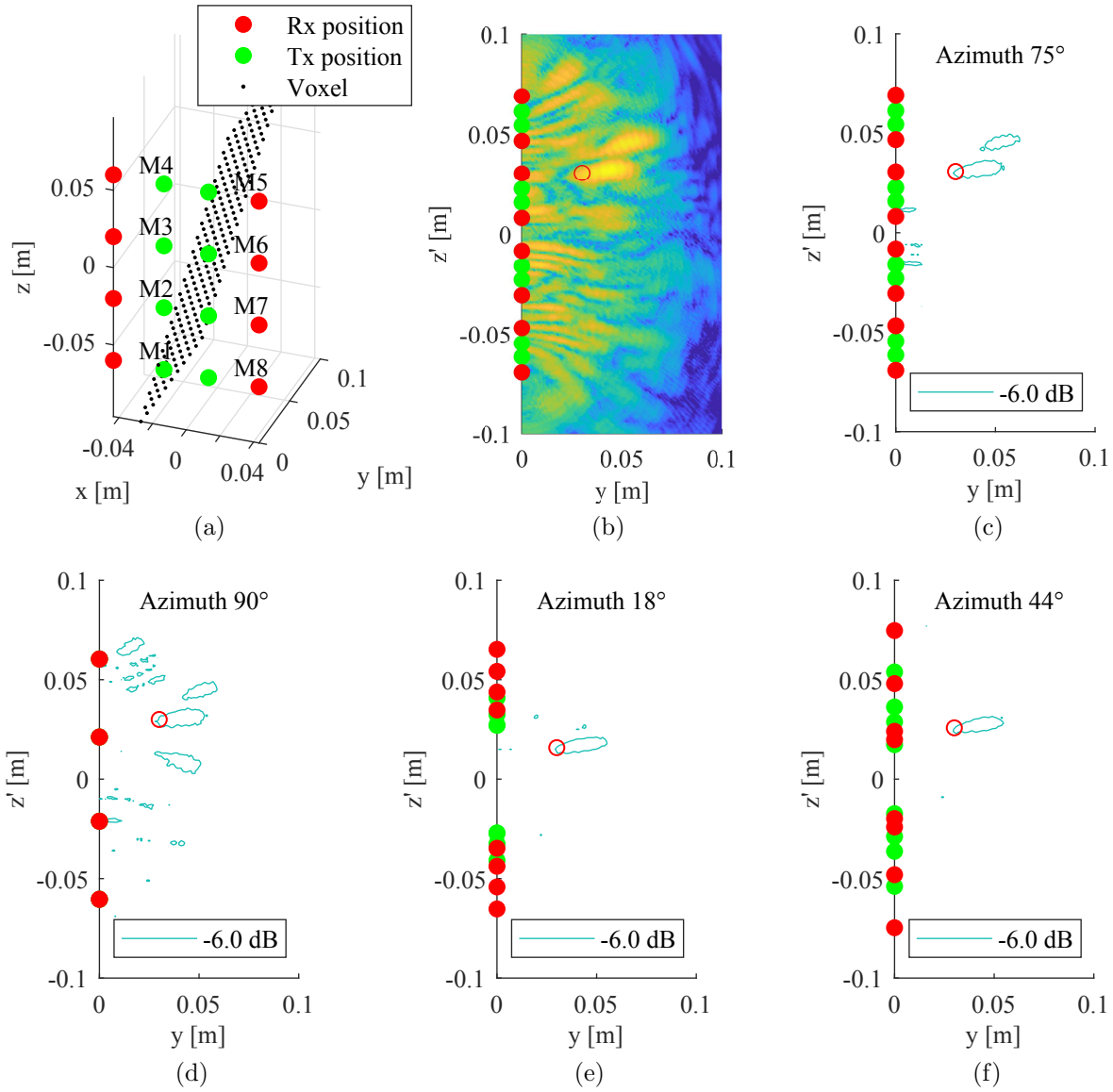


Figure 5.21: Rectangular array imaging performance across different imaging plane rotation angles ϕ . Array geometry and imaging plane at azimuth = 75° (a), along with the corresponding point spread function (b), and point spread function -6 dB contour plot (c). Contour plots for three other imaging planes are shown in (d)-(f). All point spread functions were obtained using DAS beamforming with constant attenuation compensation and grouped average clutter rejection. Red dots indicate Tx positions, green dots indicate Rx positions.

15.1 dB, SmCmR was 0.05 dB and target locating error was 0.36 mm. The identified signal maximum corresponded to the target position, but the second and third largest peaks in the obtained image were similar in amplitude and of unknown origin.

In the 2 Targets scenario, only target 1 was successfully identified, albeit at reduced SCR (Fig. 5.23b). In the 3 Targets scenario, none of the targets could be identified (results not shown). It was observed that the clutter close to modules 6 and 7 had increased in amplitude, possibly indicating increased difference between the target and empty room recording.

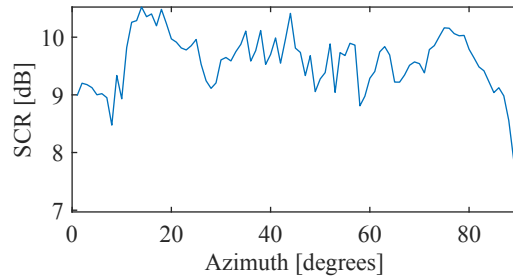


Figure 5.22: Signal-to-clutter ratio as a function of the imaging plane rotation angle ϕ .

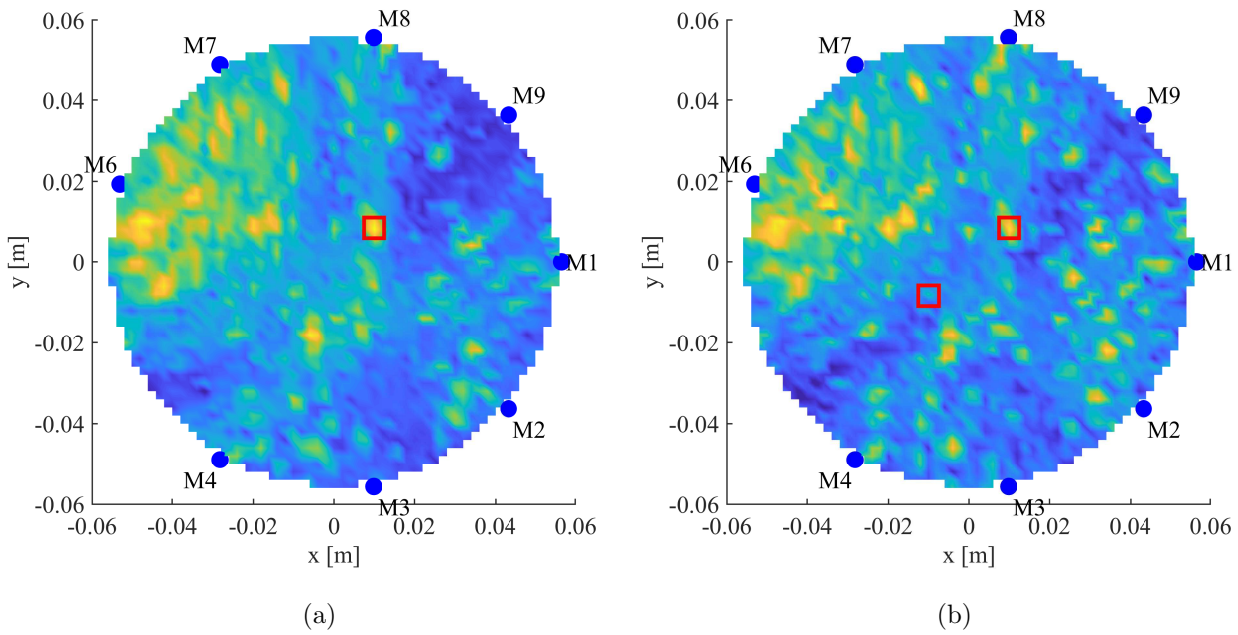


Figure 5.23: Phantom imaging results for the 1 Target condition (a) and 2 Targets condition (b). Real target location indicated by a red square. Blue dots denote Tx/Rx element positions.

To explore the problem of target identification in more detail, example signals after clutter rejection are plotted in Fig. 5.24, for 7 geometrically similar channels, with red lines indicating the focus point for the voxel containing target 1. Simulated data (Fig. 5.24a) shows the expected signal peak at the focus point, whereas real data (Fig. 5.24b, from scenario '3 Targets') shows noise across the radar frame, even after 'empty room' subtraction. An attempt was made to estimate expected signal P2P amplitude, by considering the transmitted P2P amplitude, the coupling loss (estimated in Section 3.4.4) and attenuation loss. For the first three channels (counting from the top) in Fig. 5.24b, the P2P amplitude was estimated at 0.86, 0.34, and 0.16, respectively. This signal amplitude estimates are in the orders of magnitude of measured data, but noise signal appears to be present at similar or higher amplitude.

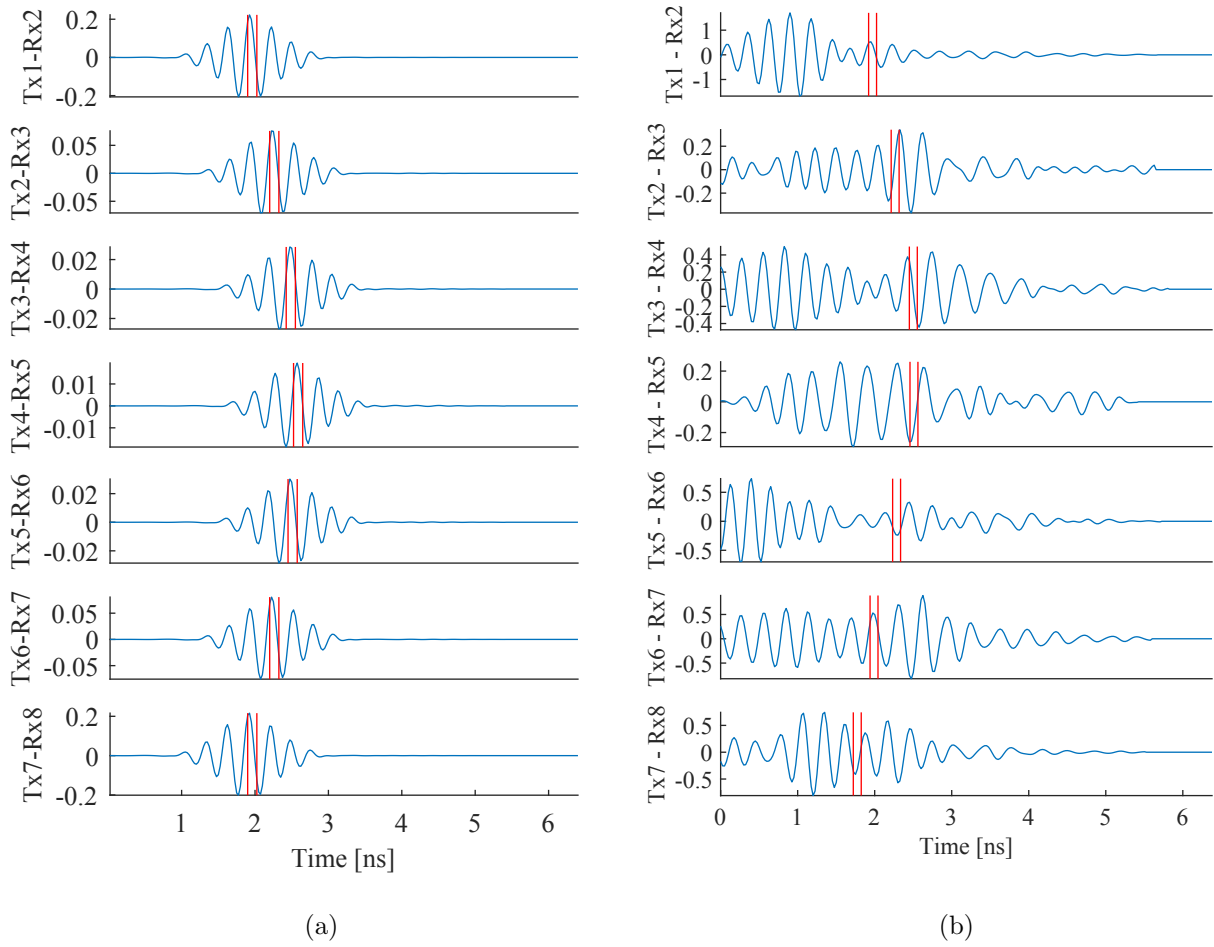


Figure 5.24: Example of phantom imaging radar data prior to beamforming, from simulations (a) and experiments (b), both after subtraction of an ‘empty room’ recording. The focal point corresponding to target 1 is indicated with red lines.

Dense semicircular array

A semi-solid cylindrical phantom was imaged using the dense semicircular array. Signal processing was identical as for the circular array phantom imaging (Section 5.4.3). Targets could not be identified for any of the recordings. No peaks in image amplitude were observed at or near the target location (results not shown). It was observed that average P2P amplitude in the pre-beamformed signals, across channels, was significantly higher (factor 2.5) than for the circular array. This was likely due to reduced clutter rejection, as no increase in signal amplitude was expected.

5.4.4 Thigh imaging

Both the dense semicircular array and the circular array were used to image the left thigh, at two proximal positions (upper and lower). Recordings were performed at PG0, both in full MIMO and in SIMO mode. DAS beamforming was used for image formation, ECG aligned averaging was applied to the obtained video data. A heatmap of signal amplitude near the heart rate frequency was constructed in an attempt to image the femoral artery. A heatmap of signal amplitude at DC was constructed with the aim of imaging static scatterers such as the femur bone.

Results from the dense semicircular array in MIMO mode are shown in Fig. 5.25. The B-mode image (Fig. 5.25a) shows a distinct peak in HR amplitude, with a location corresponding to where the femoral artery is expected to be. M-mode time series (with their voxel positions indicated by red squares), aligned to the ECG signal, are shown in Fig. 5.25c-f. Dashed lines indicate the ECG R wave. The M-mode signal at voxel 1 (peak amplitude) shows a physiologically plausible arterial pulse wave signal, whereas signals at voxels 2 and 3 (at arbitrary locations) do not show the arterial pulse. All M-mode signal y-axes were identically scaled.

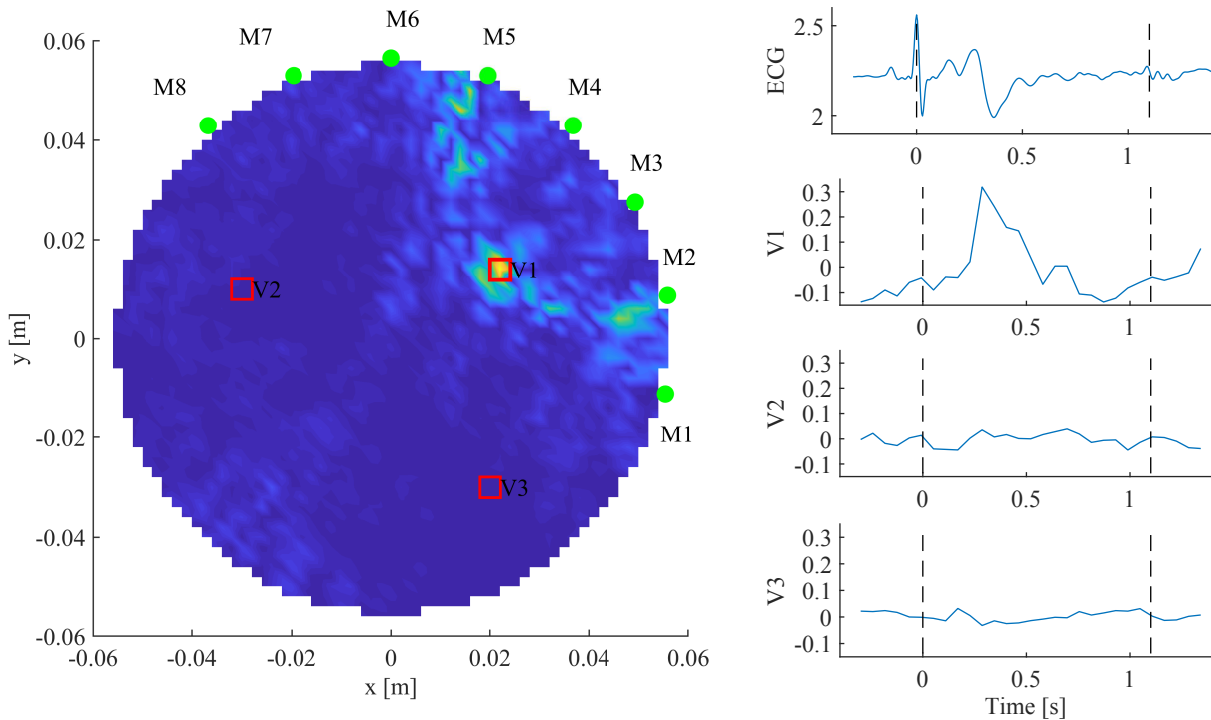


Figure 5.25: Thigh imaging result using the dense semicircular array and MIMO DAS beamforming. The B-mode image (left) shows activity at the heart rate frequency in the left lower thigh. M-mode time series are shown (right) for three voxels indicated using red squares. M-mode data are time-aligned to ECG data, dashed lines indicate the ECG R wave.

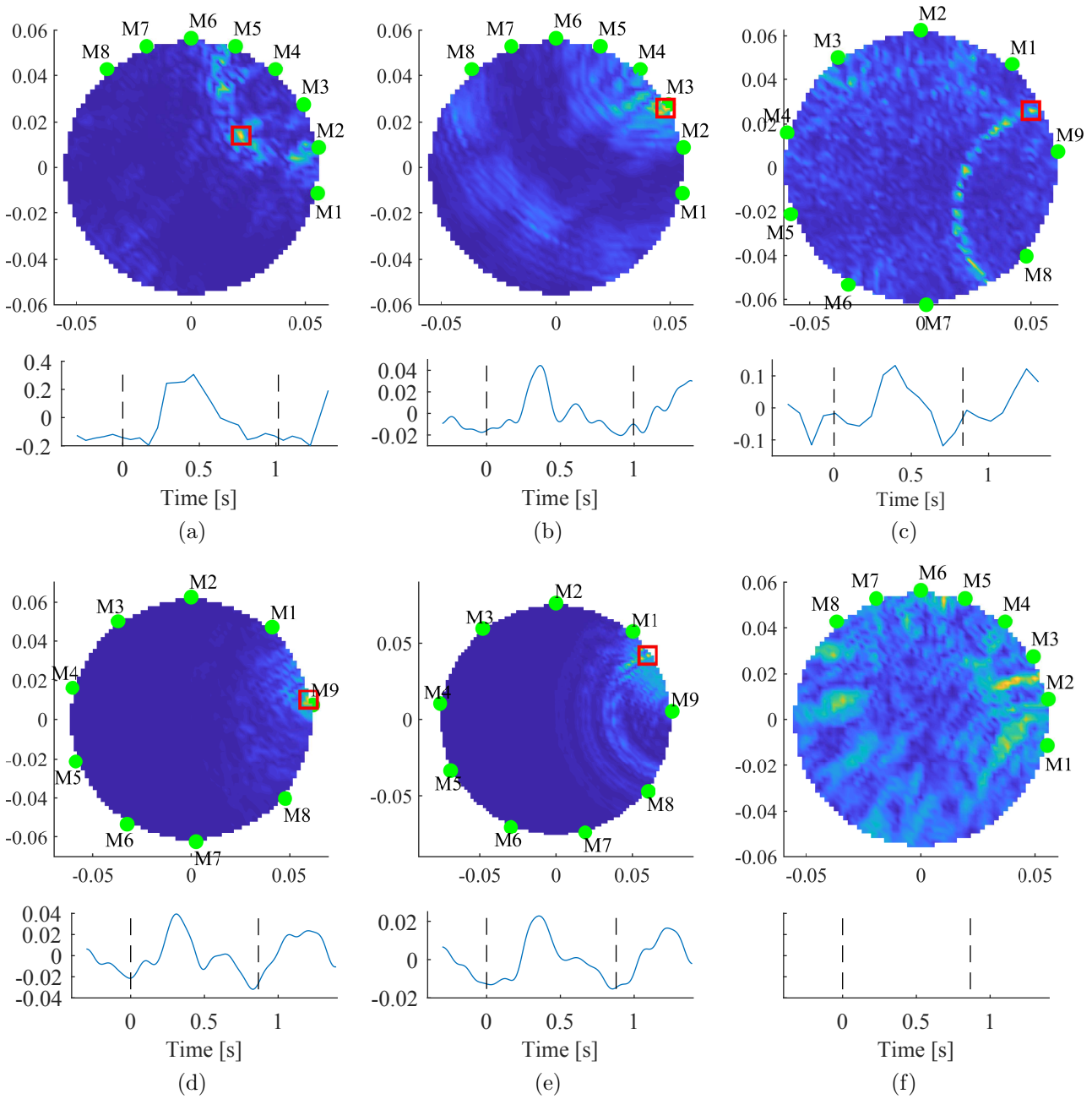


Figure 5.26: Thigh imaging results at multiple positions, using various arrays and DAS beam-forming on both SIMO and MIMO data. Both B-mode image (top) and ECG aligned M-mode time series (bottom) at red square indicators are shown. Dashed lines indicate the ECG R wave. Results given for: Dense semicircular array with MIMO data (a) and SIMO data (b), circular array with MIMO at lower thigh (c), and SIMO at lower (d) and upper thigh (e). DC imaging using dense semicircular array and MIMO data in (f).

Results from Fig. 5.25 were replicated in Fig. 5.26a. A SIMO mode recording using the dense semicircular array (Fig. 5.26b) appeared to have insufficient number of channels (factor 8 lower than for MIMO) to resolve the femoral artery location. The lower number of Tx elements did result in a higher temporal resolution with a better defined arterial pulse signal. A MIMO recording using the circular array (both on upper and lower thigh), suffered from low numbers

of channels with HR signal (lower thigh shown in Fig. 5.26c). The detected signal peak is superficial to the skin, although a second and third highest peak (similar amplitude) lie closer to hypothesised artery location. Circular array SIMO on lower (Fig. 5.26d) and upper leg (Fig. 5.26e) again suffer from a low number of channels containing the HR signal. Static imaging at DC did not result in imaging of the femur bone or any other identifiable structure (Fig. 5.26f).

5.4.5 Heart Imaging

The rectangular array was used to image the heart, along the long axis (azimuth = 113°) and short axis (azimuth = 23°). Array geometry and a subset of voxels lying on the imaging plane are shown in Fig. 5.27. Recordings were performed at PG0, in MIMO mode. DAS beamforming with attenuation compensation was used for image formation, ECG aligned averaging was applied to the obtained video data. The attenuation compensation resulted in increased wideband noise, which led to high amplitude at the HR frequency across voxels. To increase robustness of the B-mode image, SNR of the HR frequency was plotted instead of HR amplitude. ECG aligned M-mode signals are shown for all voxels indicated with a red square in the B-mode image. Dashed lines indicate the ECG R wave. Results for the long axis image in Fig. 5.28, and for the short axis image in Fig. 5.29.

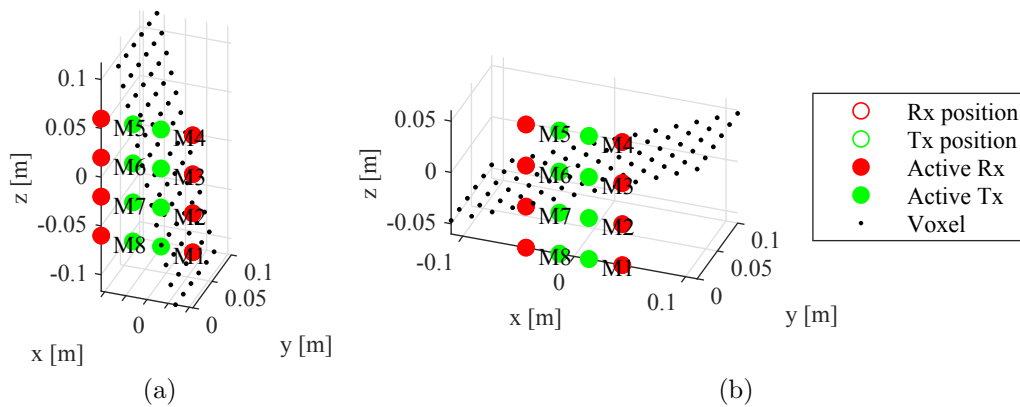


Figure 5.27: Array and image geometry for the cardiac long axis imaging plane (azimuth 113° ; a) and short axis imaging plane (azimuth 23° ; b). A subset of voxels is shown (at 2 cm spatial sampling).

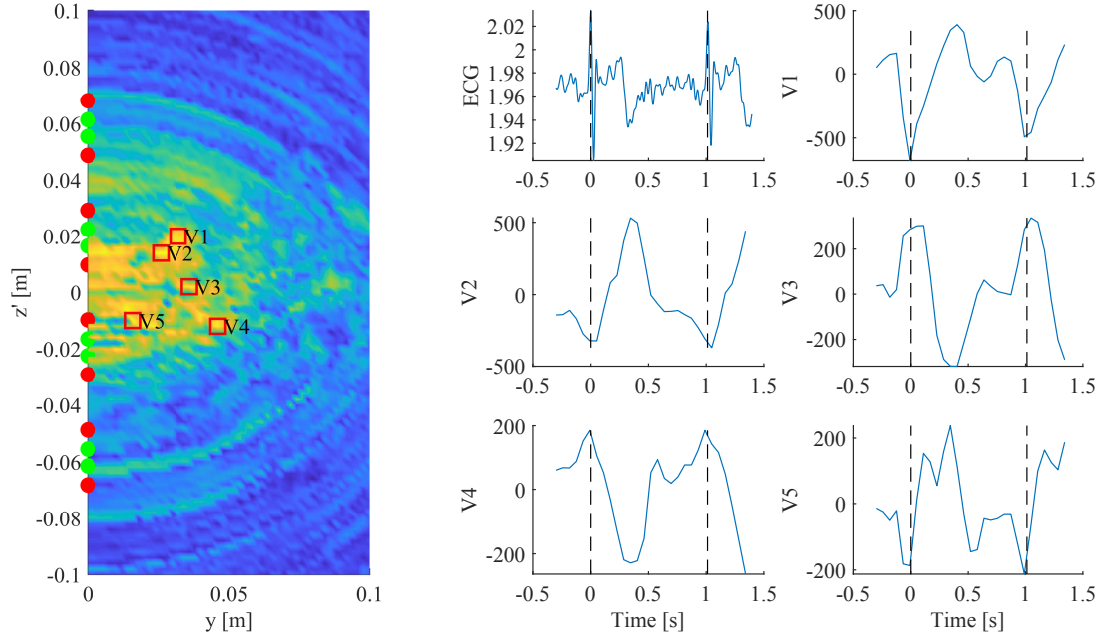


Figure 5.28: Cardiac imaging result along the long axis of the heart (Azimuth 113°) using the rectangular array and MIMO DAS beamforming. The B-mode image (left) shows SNR of heart rate activity. M-mode time series are shown (right) for five voxels indicated using red squares. M-mode data are time-aligned to ECG data, dashed lines indicate the ECG R wave.

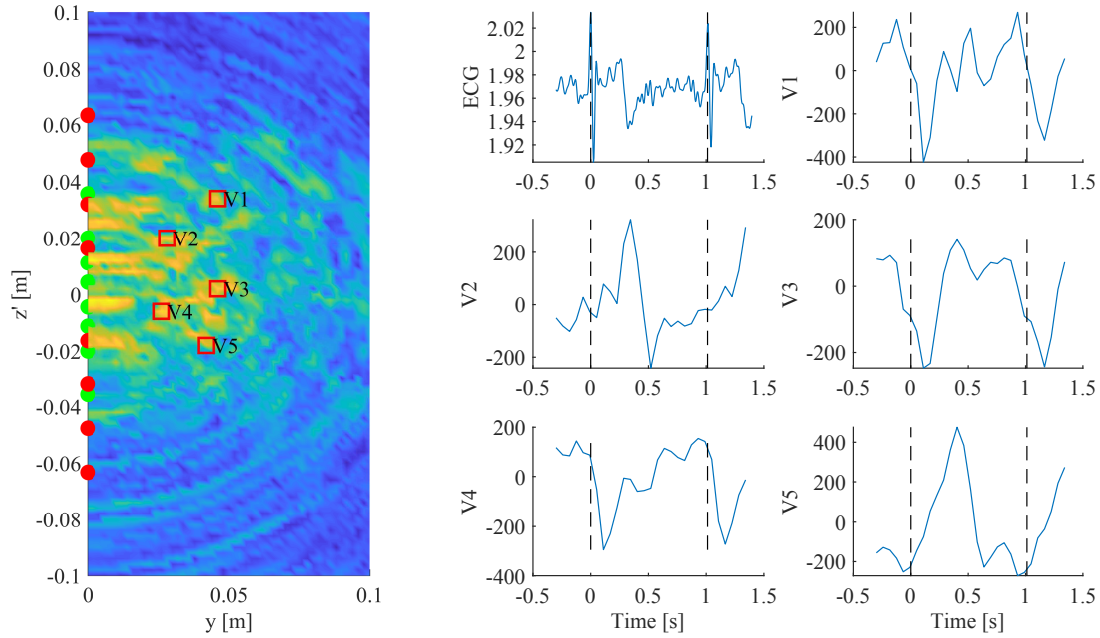


Figure 5.29: Cardiac imaging result along the short axis of the heart (Azimuth 23°) using the rectangular array and MIMO DAS beamforming. The B-mode image (left) shows SNR of heart rate activity. M-mode time series are shown (right) for five voxels indicated using red squares. M-mode data are time-aligned to ECG data, dashed lines indicate the ECG R wave.

5.5 Discussion

A system of UWB RoC devices has been used for experimental data collection and MWI of cardiovascular dynamics in a human subject. The imaging scenarios have been simulated to test various imaging algorithm parameters. Below, imaging algorithms, experimental results, and the use of UWB RoC devices for biomedical MWI will be discussed.

5.5.1 Imaging algorithm and data analysis

Clutter rejection Clutter rejection (skin artefact, backscattered signal and crosstalk) is imperative for successful static MWI. For dynamic imaging, significant clutter rejection is performed by high-pass filtering, but the skin and limbs are known to move as a result of heart contraction and cardioballistic forces. For static imaging, the best clutter rejection was achieved with differential imaging, where a no-target recording can be subtracted. In simulated data, only a small drop in performance was observed when using grouped averaging instead of differential clutter rejection. In measured data, grouped average clutter rejection proved not to be sufficient for target identification in a cylindrical phantom. Grouped Woody average clutter rejection was found to outperform simple (non-aligned) grouped average clutter rejection, but visualisation of radar data demonstrates that clutter is still the dominant signal component. In this work, grouped average clutter rejection was likely hampered by the choice of hardware, which provides far less backscattered signal absorption than conventional systems (i.e. [22, 24, 37]) and is prone to movement artefacts. The biggest gains in imaging performance for the here presented system are likely to be realised by improvements in clutter rejection. Topics for further research include the use of more advanced algorithms such PCA-based artefact removal [141], distant-based filtering using independent component analysis [38], and hardware modifications that would facilitate successful grouped averaging.

Attenuation compensation Compensation of path length dependent attenuation in theory ensures proportionate channel summation and proportionate target amplitude. Practically, attenuation of depth independent noise is a major challenge. In simulations of two different geometries, attenuation compensation was found to improve results for the rectangular array but not for the circular and semi-circular arrays. A possible explanation is that a circular array collects scattered data from all directions, therefore summed signal across channels is depth independent. In addition, the longest path length in the circular array was higher than for the rectangular array, causing more amplification. It is likely that attenuation compensation would suffer less from depth-independent noise if clutter rejection were to be improved.

Assumption of anatomy and tissue properties A correct estimate of tissue properties is required for beamforming. Considering the huge variation in anatomy between individuals, this is likely a large source of error. Individual parameter assessment as in [30] could be performed for the thigh or limb imaging case. Improvements could also be made by implementing a parameter optimization algorithm for permittivity assessment [8]. The chest is a much more complex geometry: Transmissive measurements won't be representative for the small volume of tissue between the skin and the heart, and there are no clear point scatterers for parameter optimization. In addition, the anatomy surrounding the sternum is complex and highly irregular across antennas. Future work could include experimentation with array position. Although the sternum has been found to be a good transmissive window for heart sensing, array placement beside the sternum would make the clutter space much more homogeneous. A depth-dependent permittivity model (previously proposed in [41]) could account for different volumes of lung tissue.

Beamforming algorithm In this work, the DAS and DMAS algorithms were used. DMAS was found to be highly successful at clutter rejection. With the current implementation, DMAS computation time was prohibitive for use on dynamic video data. An enormous body of research has been done on more advanced beamforming techniques such as adaptive algorithms [132] and super-resolution techniques [38]. A recommendation for future work is the development of a data-adaptive channel-quality factor for dynamic imaging.

5.5.2 Experimental results

Phantom imaging The successful identification of the first target in the cylindrical phantom is encouraging. However, the presence of peaks at similar amplitude to the target amplitude demonstrate the challenge of static imaging. Missing data from the module nearest to the second target could help explain the inability to identify this target 2. In addition, it is very likely that the insertion of the metal targets changed the overall phantom volume and the antenna-phantom interface: Differential imaging showed large differences at a subset of antennas between the 3 Target condition and No target condition. Target identification was unsuccessful for the dense semicircular array. It is likely that this is due to the gradual destruction of the phantom as targets were inserted and extracted: The No target condition was measured after the target conditions, and a crack had appeared in the phantom with air gaps where the targets had been. The estimate of P2P amplitude is encouraging: Measured signals are in the order of magnitude of estimated signal amplitude. If these estimates are correct, the gain that is provided by coherent summation should allow for resolving a target.

Thigh imaging Thigh imaging was performed as a proof of concept of RoC MWI in a relatively simple anatomical structure. The femoral artery is a major target for dynamic imaging, and static imaging could be demonstrated on the femur bone. Static imaging proved to be completely unsuccessful. The dielectric contrast of femur bone in muscle tissue is a lot lower than the contrast in case of the phantom scenario, and differential imaging could not be performed. Significant improvements in clutter rejection are required. Dynamic imaging proved successful under specific conditions. Using the dense semicircular array in MIMO mode, a clear peak was observed which corresponded to the hypothesised femoral artery position. US would provide validation of the exact artery location, but no US measurement was performed at this stage. The M-mode signal at the identified peak showed a plausible pulse wave, with clearly identifiable ejected wave, reflected wave and dicrotic notch [119]. In MIMO data, the HR signal was observed in 16 out of 64 channels, mostly between the 4 modules nearest to the artery. The use of SIMO data was expected to perform worse simply as a result of the lower number of channels. The target location could not be resolved for recordings from the circular array. This was likely due to high attenuation levels (5.2 dB/cm) and thigh diameter (15 cm), in combination with low array density.

Heart imaging Heart imaging was performed using a rectangular array, along the heart's long and short axis. Beamforming was successful to some degree: Results show that mechanical motion at the HR frequency is detected at the location where the heart is known to be. M-mode signals were periodic and peaks were found to align with ECG data, clearly indicating that signals originated from cardiac motion. Although the HR signal was also detected at very superficial positions (potentially caused by the moving skin artefact), the highest signal strength was observed at a depth of 3-4 cm, corresponding to the heart boundary. Beamforming allowed for detection of different M-mode signals, each with a unique waveforms. Interpretation of these signals at this stage is extremely challenging. Some signals clearly resemble the Wiggers schematic diagram of cardiac dynamics [142]: Both the atrial filling of the ventricle upon the ECG P wave, and ventricle contraction upon the R wave appear to be visible through transient high amplitude signals. For some voxels, a slight delay relative to the R wave is observed, which could represent the ventricular ejection delay. The intention of imaging along the cardiac long axis was to be able to distinguish between atrial and ventricular contraction. A detailed look into cardiac physiology and US M-mode data, made it clear that the heart is far too complex and motions are too large for that level of imaging to be expected. Significant improvements are needed for RoC MWI to become clinically relevant. However, this work provides the first evidence that the heart can be imaged using RoC technology.

5.5.3 The use of coherent UWB radar-on-chip for biomedical imaging

In the current work, dynamic in-body MWI of the cardiovascular system has been demonstrated, using a system of synchronised RoC devices and body-coupled antennas. After [44], this is the second report of successful dynamic MWI of the heart, and the first report of dynamic MWI of the vascular system. In addition, this work serves as the first proof of concept that imaging of the human body can be performed using a system of RoC devices.

Although MWI has been researched intensively, the use of RoC devices for this purpose is an entirely novel concept and has become a possibility only recently through advancements in CMOS radar technology. Compared to traditional radar systems, the RoC device has an extremely small form factor (recent efforts at Novelda have resulted in a radar module with dimensions of 3 by 15 mm), is very low in cost, and consumes very little power. The X2 radar module is designed to comply with international regulations on unlicensed UWB radiation (FCC part 15) and is therefore suitable for use in any medical or consumer application. Perhaps the most interesting aspect of this technology is that each RoC device has its own RF receiver, and a high number of modules can be run simultaneously. This avoids the need for complicated switching arrays, and speeds up SIMO or MIMO data acquisition by a factor that equals the number of Rx antennas.

A mainstream VNA would typically offer a dynamic range of around 120 dB, with integration. In order to achieve the UWB radar measurement speeds required for cardiovascular imaging, sweep times would have to be extremely short, leading to increased noise floor and compromised sensitivity. In [44] a sufficient frame rate was achieved, using a VNA with a price far in excess of 10,000 GBP. The imaging system has not been tested on arterial imaging, for which higher sensitivity is required. When a high number of receivers and high sensitivity is required, dynamic MWI using a VNA would not be possible.

RoC is a relatively new concept and technology is still improving. Since the start of this research project, Novelda has released the XeThru X4, which offers a number of improvements over the X2. The X4 has overall increased sensitivity, reduced jitter and a more consistent transmit pulse waveform. Due to its centre frequency at 7.29 GHz however, signal attenuation in tissue would be much higher. Losses in tissue are estimated to be as high as 16 dB/cm at 7.29 GHz, as opposed 5 dB/cm at 3.45 GHz. A future version of the X4, but at lower centre frequency would greatly benefit RoC MWI.

RoC is an exciting new technology with promising applications. In order for MWI using RoC technology to become clinically relevant, a lot of improvements will need to be made. Through this work a number of challenges related to the current system design became apparent, which

are being explored below.

Signal leakage, backscattered signal, crosstalk and the skin artefact When starting this research project in 2015, it was not anticipated that signal leakage, crosstalk, interference from backscattered signal, and the skin artefact would turn out to be one of the biggest challenges in RoC MWI. The majority of established research groups in CMI for biomedical applications have developed a static, non-portable system. Often, a coupling liquid is used not only for the purpose of increasing radiation into tissue, but also to absorb backscattered signal. The cost of increased signal absorption can simply be overcome by increasing transmitted power. Antennas are fixed in a static array, and even though cables and switch arrays are being used, none of these components are likely to move. Because of this, any system imperfection can be calibrated for. The system used in this work has the advantage of small form factor and low power. In addition, the system was designed to be modular: Any array could be constructed using the existing radar modules and control system. This, along with the chosen imaging anatomies in this work, poses the following challenges:

- Because the system is mobile and typically attached to the participant during measurements, the environment cannot be calibrated for and any participant motion will cause interference by backscattered signal reflections.
- Dielectric spacers in body-coupled antennas were designed to reduce the skin artefact, but skin-antenna interface inconsistencies (due to geometry, variation in tissue properties, or motion) hamper skin artefact reduction.
- Antennas require direct contact with the skin, therefore some adjustable fixation system must be used. This inevitably leads to inaccuracies in antenna position and orientation, which greatly hampers grouped average clutter rejection. In addition, movement of antennas during recordings could only be avoided to some extent.
- Because of the chosen imaging anatomies and desired system modularity, no coupling liquid could be used for leakage signal absorption.

To date, two research groups have developed portable microwave systems for breast cancer screening. A hand-held device employs an antenna array built into a dome, which is separated from the breast coupling dome using a coupling liquid [32]. The coupling dome ensures that the skin-device interface is constant, whereas the antenna dome ensures that none of the components in the system can move relative to one another. This configuration allows for much better calibration than was achieved in the current work. It is expected that backscattered signal and standing waves (due to high bandwidth) could form a challenge. A second group [30] has developed a wearable array in a bra for breast health monitoring. Antennas are designed to be in direct contact with the skin and no absorbing coupling liquid is used. For image formation using the bra-based system, similar challenges as encountered in this work could be expected.

Bandwidth and centre frequency A time domain UWB radar system with pulse generator provides far less flexibility than a VNA does, and is not suitable for experimentation with different bandwidths. High bandwidth leads to increased down-range resolution, which may or may not be desirable. When using circular arrays, such as in breast cancer imaging or as used in the current work, spatial resolution is equivalent to the system cross-range resolution. Cross-range resolution is dependent on pulse centre frequency and not on pulse duration. In [32], a mono pulse (extremely wide bandwidth) was used. In theory, this will give great down-range resolution. Previous research has found however that matching the system antennas to a high bandwidth is a significant challenge and standing waves occur with the use of mono pulses⁴. Increased bandwidth would be beneficial for cardiac imaging, but an anticipated challenge is extreme pulse distortion due to increased attenuation at high frequencies. Fat tissue (dominant component in breast) has very low conductivity (0.41 S/m at 3.45 GHz), whereas bone conductivity is twice as high, and muscle conductivity is as high as 2.52 S/m. When down-range resolution is less important, a lower centre frequency would be desirable for heart imaging, to reduce signal attenuation at the cost of cross-range resolution. It is likely that application specific RoC systems would be required in order to reach clinically viable products. Experimentation with a more flexible system may be required to find application specific optimal pulse specifications.

Array density In the current work, array density was limited by module dimensions. When the current modules were originally designed, it was not considered that cable leakage and movement would play a significant role, and the use of coaxial cables were to be avoided. For a future iteration, there is no reason why the module could not have the same footprint as a set of Tx and Rx antennas (2 by 5 cm). A higher array density will lead to reduced aliasing and thus better imaging performance. Increased crosstalk as a result of reduced inter-antenna distance would form a new challenge.

Imaging speed The unique advantage of using synchronised RoC devices is high imaging speed on large MIMO datasets. In the current work, using nine transmitters, a MIMO frame rate of 13.0 fps was achieved, without loss of sensitivity. Further improvements in imaging speed (or array expansion without loss of imaging speed) are possible. Frame rate on a single module is a function of signal integration and of DAC range. DAC settings determine the dynamic range of the receiver, and was set to its widest range in the current work to avoid potential signal truncation. With further reduction of signal leakage and investigation into expected signal amplitude, DAC range could be set to a fraction of the current settings, with proportionate increase in imaging frame rate.

⁴Research performed by Novelda AS, Norway

5.6 Conclusions

Based on the here presented simulations and experimental work on MWI in the human body using RoC devices, the following conclusions can be drawn:

- For the first time, it has been demonstrated that a system of synchronised RoC devices can successfully be used for dynamic MWI of the heart and the vascular system in the human body.
- Static MWI using RoC technology remains a challenge due to significant clutter sources, and further improvements to both imaging hardware and data preprocessing are required.
- The most significant challenge both in hardware and in data processing currently lies in the reduction of the skin artefact, antenna crosstalk, backscatter, and other leakage sources.
- Due to the unique opportunity of low-cost, small form factor, and portable MWI technology at high imaging speed, the use of RoC for biomedical imaging remains worth pursuing. Imaging application specific RoCs may be developed for increased sensitivity at optimal transmit pulse properties.

Conclusions and recommendation specific to the data processing and imaging algorithms are:

- A window length of 5 samples ($1/2\lambda$) was found to be a good trade off between SCR and robustness to erroneous beamforming assumptions.
- In experimental data, grouped Woody averaging for clutter rejection did not perform as well as differential imaging, but did provide a significant improvement over no clutter rejection.
- A propagation velocity model with constant dielectric properties, estimated at centre frequency was found sufficient.
- Path length dependent attenuation compensation, based on dielectric properties estimated at centre frequency, was found to increase imaging performance for a rectangular array, but reduce performance for a circular array.
- DMAS clearly outperformed DAS on SCR at the cost of increased computation time.
- A rectangular array is better utilised by rotating the imaging plane away from azimuth 0° or 90° , leading to spatial sensor dispersion in a 2D plane.
- Virtual element tapering or pruning did not improve imaging performance for the current setup.

Chapter 6

Conclusions and future work

6.1 Original contributions

This work was the first to demonstrate imaging in the human body using the novel approach of employing multiple synchronised UWB RoCs. In addition, it was the second work to demonstrate dynamic MWI. The relevance of using the novel technology of RoC lies in the fact that a MWI device could be developed at much lower cost, smaller form factor, while using higher imaging speeds. Instead of taking the more conventional approach of constructing a table-based device with an antenna array submerged in a coupling liquid, it was decided to aim for a fully portable imaging solution, which could enable future applications in ambulatory diagnostics, preventive screening and remote patient monitoring. The modular design, using body-coupled antennas and no coupling liquid, with radar transceivers connected directly to each antenna pair, resulted in numerous challenges. The original contribution of this work lies in identifying these challenges, working with hardware developing collaborators to improve and validate the system, and exploring how RoC could be utilised for in-body sensing and imaging. The personal contribution to this project can be summarised per technical chapter as follows:

Chapter 3 - System: Work with research collaborators to develop, improve, and validate the first RoC MWI system. Specifically:

- Scope out target applications and assess their feasibility:
 - Cardiovascular sensing and imaging (heart and arteries)
 - Functional neuroimaging
 - Dielectric spectroscopy for blood glucose measurement
- Work with research collaborators to develop and improve a novel imaging system

- Developing a set of system requirements
- Hardware characterisation
- Assessing the suitability of developed hardware for biomedical imaging, through experimentation
- Consider the suitability of UWB signal propagation models for imaging algorithms based on coherent summation

Chapter 4 - Sensing: Understanding the fundamentals of in-body radar sensing and its challenges, and demonstrate RoC biomedical sensing:

- Signal processing: Finding a valid metric for coherent UWB measurement of sub-wavelength in-body motion
- Experimentation: Discover the possibilities and limitations of in-body RoC measurements
- Validate the source of HR modulation in in-body sensing data
- Provide the first demonstration of in-body RoC sensing of cardiovascular dynamics

Chapter 5 - Imaging: Demonstrate RoC dynamic biomedical imaging, and investigate its challenges and opportunities:

- Algorithm optimisation using simulated data, specifically for cardiovascular dynamic imaging
- Provide the first demonstration of imaging in the human body using a system of UWB RoC devices.
- Provide the second demonstration of dynamic MWI of the heart and femoral artery

Although no specific clinical application was pursued at this point, the current work aims to cover the fundamentals of RoC sensing and imaging, and could be used as a stepping stone for a more clinical oriented project in which an application specific system could be developed based on conclusion drawn in this thesis.

6.2 Recommendations for future work

Algorithms

Throughout this work, the greatest challenge to sensing and imaging turned out to be leakage signals, composed of antenna backscatter, crosstalk and the skin artefact. The leakage signal

varied significantly between channels, which was detrimental to leakage rejection. An exploration into how the antenna-skin interface affects this signal, as well as hardware improvements and preprocessing improvement are a requirement for obtaining better dynamic imaging quality, as well as for performing imaging of static structures. Various advanced signal processing techniques have been described in the literature to tackle this problem. An additional challenge will be leakage rejection for dynamic imaging, in which case the antenna-skin interface might vary over time due to motion induced by superficial cardiovascular structures. Modelling of this time-varying artefact may be required for optimal rejection. Future work with the current imaging hardware would have to focus on preprocessing before any investigation into more advanced imaging algorithms is performed.

The imaging algorithms DAS and DMAS were used as a starting point. Most image quality gains are to be made in improving the preprocessing step. However, a wealth of advanced imaging techniques could be thrown at this problem. Apart from some of the imaging algorithms that were mentioned in this thesis, many more algorithms have been proposed by research groups working with simulated data. A first next step would be to take a data-driven, adaptive approach in DAS, by introducing a quality factor to different channels, as it was clear that not all channels were able to capture the arterial signal.

One imaging parameter that is of utmost importance to correct image formation, but was not investigated much, is the permittivity assumption. The current assumptions based on published data, anatomical assumptions, and assumptions on the signal path, could be wildly incorrect and substantially change the final result. In case of transmissive measurements (as for example thigh imaging), an attempt could be made at estimating tissue properties from collected data. For cardiac imaging, this would be less obvious as the exact path is mostly unknown. Validation of results with US measurements could provide a solution. Interesting future work would be to incorporate a more detailed anatomical model, as the assumption of homogeneity is far from reality. Likely, channel-voxel dependent effective permittivity would improve image quality.

Hardware

Apart from algorithmic improvements, future work should aim to further develop imaging hardware. The use of dielectric spacers in the body-coupled antennas proved to provide substantial reduction of losses. However, the ceramic spacer is not ideal: The weight is high, the material is challenging to process and the lack of flexibility makes for imperfect skin contact. An ideal dielectric spacer material would be pliable, easy to mould, with low conductivity but well controlled permittivity. Further work may focus on coupling materials, as the antenna skin-interface is key to reducing leakage and skin artefact. The use of gels as an additional coupling medium may be explored. Better control over antenna position and orientation will be funda-

mental to leakage reduction. In the current work, 3D printed plastic structures were used for optimal flexibility of measurements. For an application specific system design, antenna fixtures could be made much more rigid. Also improved antenna manufacturing processes (current antennas were built by hand and the dielectric spacer was challenging to cut to size) would reduce antenna variation. Elimination of sources of variation is paramount to estimating and removing the leakage signal. Finally, further improvements of module and antenna shielding and reduction of antenna cross talk would be enormously beneficial to image quality and robustness against interfering noise sources.

Radar modules in this work used the X2 chip, due to the suitability of the centre frequency. The XeThru X4 chip is an improved version of the X2 with reduced phase noise and better sensitivity, but at a higher centre frequency of 7.29 GHz. With more resources available, a radar chip could be developed specifically for MWI, combining the improvements to sensitivity of the X4, but at lower centre frequency. For intracranial imaging, a centre frequency between 1 and 3 GHz would likely be desired. For cardiac imaging, higher centre frequency (as employed by the X2) could be used to retain resolution. It must be stated that despite years of simulation studies and research with existing prototypes, there seems to be no consensus on the optimal frequency band for MWI. For further work on a cardiac imaging device, application specific simulation studies could provide more insight into this problem.

Application

In this work, imaging of the femoral artery served as a proof of concept. Although the arterial pulse waveform and PAT may serve as biomarkers for cardiovascular health, imaging of the artery alone might not be of clinical interest per se. Arterial sensing using a single RoC module, possibly integrated into a wearable, could be investigated further for clinical applications. For the scope of this work, it was decided not to delve into specific applications but rather provide a framework for potential future uses. Sensing applications of interest would be arterial sensing for continuous blood pressure estimation, arterial stiffness estimation, lung fluid assessment for chronic heart failure, assessment of intracranial pressure, and dielectric spectroscopy measurements for metabolic applications such as blood glucose levels. For applications where signal transmission is less of a challenge, the use of the XeThru X4 should be considered.

Similarly, heart imaging using RoC is an application that deserves further investigation. To study MWI as a clinical tool for assessment of cardiac function, validation of obtained signals with conventional technologies such as US would be a first step. Through collaboration with clinicians, the requirements for a minimum viable and clinically significant product could be set. Various technological improvements could be made: In the current work, the array density was limited by the size of modules. Array density could be made much higher by redesigning

the radar modules, which would add to the number of channels, thus improve signal quality. Imaging speed could be improved substantially through reducing receiver DAC range and possibly reducing signal integration. Perhaps US spatial resolution would not be within reach, but RoC has the potential to provide a more affordable and easy to use technology than US.

6.3 Final thoughts

When I walked into Tim's office to discuss PhD research project ideas about five years ago, Tim and Bassen had come up with the idea of using a new radar chip for brain imaging: A functional neuroimaging helmet that would measure the haemodynamic response as a marker for neural activity, using pulsed UWB RoC. The device would be fully portable, and make fMRI obsolete through superior temporal resolution. The term 'Braindar' was coined.

Considering the state of technology and the knowledge and expertise we had at our disposal, the goal of functional neuroimaging was perhaps somewhat ambitious for the limited time frame. We had always realised that functional neuroimaging might be out of reach, but that we would probably learn something of interest along the way. Considering stroke imaging as the low-hanging fruit was simply naive. During endless experimentation with radar sensing, on phantoms, friends and colleagues, but mostly on my own body, it became clear that this challenge was far greater than anticipated. Had we known from the start, we would probably have went with the conventional approach and ended up with a static brain scanner similar in design to what is being developed at EMTensor. Instead, we stuck to the wearable, modular concept of radar-antenna pairs which could be rearranged in a flexible manner to accommodate for any body part.

After five years of work and having gained quite a bit of knowledge and experience on radar imaging and sensing, both in-body and remotely, I am glad to say that I do think there is a viable application for RoC in in-body sensing or imaging. It isn't necessarily in functional brain imaging; It would be mad to attempt that with a non-static system, having such limited control over leakage and a lack of proper calibration opportunities. The application lies in developing accessible monitoring technologies, with cardiovascular being the most obvious target application. Healthcare patterns are shifting more towards primary care, early diagnosis of disease through preventive screening and remote monitoring. As patient data sharing and integration across providers will become more common, there will be a higher need for low cost devices which can be used to gather patient health data outside of the hospital environment. Such devices must be low in cost and portable, perhaps wearable. They must be easy to use, as primary care providers will not have access to expert knowledge on the variety of devices in their toolbox. Apart from measuring cardiac electrical activity (ECG), a modality for the

assessment of mechanical function will be indispensable. As we have demonstrated that UWB RoC technology can be used for sensing and imaging of the heart and arteries, and considering the low cost and small form factor of this technology, it appears that RoC could potentially enable accessible and easy-to-use devices for the assessment of cardiovascular function.

Appendix A

List of publications

Parts of this research have been published in a journal and conference proceedings. Below follows an overview:

- T. Lauteslager, M. Tømmer, T. Lande, T. Constandinou, “Coherent UWB Radar-on-Chip for In-Body Measurement of Cardiovascular Dynamics”, in *IEEE Transactions on Biomedical Circuits and Systems*, 2019.
- T. Lauteslager, M. Tømmer, T. Lande, T. Constandinou, “Cross-Body UWB Radar Sensing of Arterial Pulse Propagation and Ventricular Dynamics”, in *Biomedical Circuits and Systems Conference (BioCAS), 2018 IEEE*, Conference Proceedings.
- T. Lauteslager, M. Tømmer, K. Kjelgård, T. Lande, T. Constandinou, “Intracranial Heart Rate Detection using UWB radar”, in *Biomedical Circuits and Systems Conference (BioCAS), 2016 IEEE*, Conference Proceedings.
- T. Lauteslager, N. Nicolaou, T. Lande, T. Constandinou, “Functional neuroimaging using UWB impulse radar: A feasibility study”, in *Biomedical Circuits and Systems Conference (BioCAS), 2015 IEEE*, Conference Proceedings.

Other publications:

- Y. Han, T. Lauteslager, T. Lande, T. Constandinou, “UWB Radar for Non-contact Heart Rate Variability Monitoring and Mental State Classification”, in *41st Annual International Conference of the IEEE Engineering in Medicine and Biology Society (EMBC), 2019*, Conference Proceedings.

Appendix B

3D simulations

EM wave propagation in dielectric media, in 2D and 3D, was simulated using the FDTD method. Open source Python-based simulation software gprMax, originally developed for ground penetrating radar, was used [114]. gprMax does not have a graphical user interface. Instead, geometries and materials are defined through scripts and command-line interface. As Python commands are available to the user, one can dynamically update geometries between simulation iterations, which can be used to mimic moving anatomical structures.

B.1 Definition of dielectric properties

Debye model

Dispersive dielectric properties must be modelled across the bandwidth of the transmitted pulse. The different dispersion regions can effectively be modelled using the Cole-Cole model (described in Section 2.2.3). However, this model is not suitable for efficient analysis through the FDTD method. FDTD simulation software, including gprMax, usually rely on the simpler Debye model to describe a medium's dispersive dielectric properties. The Debye model is a special case of the Cole-Cole model, for which the dispersion broadening term α equals 0, and can be described both in the time and in the frequency domain. On a limited frequency bandwidth, the Debye model has been shown to be sufficient for describing the dispersive dielectric properties of biological tissues [143].

It was found that the dispersive regions in the frequency band of 100 MHz to 10 GHz (mostly gamma dispersion) could be sufficiently described using a two-pole Debye model. Complex

relative permittivity ε_r^* , as a function of angular frequency ω , was defined as:

$$\varepsilon_r^*(\omega) = \varepsilon_\infty + \sum_{k=1}^2 \frac{\Delta\varepsilon_k}{1 + j\omega\tau_k} + \frac{\sigma_i}{j\omega\varepsilon_0} \quad (\text{B.1})$$

where ε_∞ is the relative permittivity at high frequencies, $\Delta\varepsilon = \varepsilon_s - \varepsilon_\infty$, with ε_s the low frequency relative permittivity, τ is the relaxation time constant, σ_i is the static ionic conductivity, and ε_0 is the permittivity of vacuum.

Parameter fitting

Debye model parameters were generated for various biological tissues, as well as for different phantoms that were used throughout this research. Parameters were obtained by model fitting against assumed ‘true’ data. For biological tissues true data were dielectric properties from the [18] database based on [17], for phantoms these were dielectric properties measured using a VNA. Particle swarm optimisation was used to fit the two-pole Debye model to true data across frequency band from 0.1 GHz to 10 GHz. A loss function was defined to determine the error on both the real and the imaginary component of ε_r . The particle swarm function of the Matlab Optimization Toolbox was used for final parameter fitting.

B.2 Simulation

Time and space discretisation

The minimum spatial discretisation was determined by the shortest wavelength in the propagation medium. The –30 dB bandwidth upper cutoff of the X2 transmitted pulse was evaluated in the simulated medium with highest permittivity (thus shortest effective wavelength). A minimum sampling frequency of 10 samples per wavelength was set. The largest permissible step size in diagonal direction was determined, from which step size in x, y and z directions were derived. Temporal discretisation could then be determined by the spatial step and propagation velocity, according to the stability condition.

A 1D spatial discretisation of 570 nm was chosen for the 3D model and 700 nm for the 2D model.

Geometry and medium definitions

gprMax allows for definition of simple geometrical shapes such as rectangles, spheres and cylinders. Creating realistic anatomical models was outside of the scope of these simulation experiments. For most simulation experiments, targets were defined as a cylinder (in 2D) or as a sphere (3D). Geometries were evaluated in the open source data visualisation software Paraview. Medium dielectric properties were defined using two pole Debye models, with parameters obtained as described above.

Antennas

No antenna geometries were simulated: This would only be relevant to assess radiated efficiency into tissue. For most simulation experiments, DC (no target present) was simulated and subtracted, voiding the need for correct antenna properties. Tx and Rx were modelled as a voltage source.

Excitation waveform

The X2 transmitted pulse at its lowest centre frequency (PG=0) was measured, bandpass filtered to remove noise, and resampled to the temporal discretisation as defined by the step size for the relevant simulation scenario.

Scanning multiple-input and multiple-output datasets

In gprMax, multiple simultaneously receiving elements may be defined with a single transmitting element. To obtain a MIMO dataset with multiple transmitting elements, the simulation can be scripted and multiple iterations can be run subsequently, while adaptively updating Tx position.

Simulating dynamic systems

A dynamic model may be obtained by running the simulation multiple times and dynamically updating the model geometry. To simulate a pulsating artery, a cylinder was modelled with diameter d being a function of the model run. Artery diameter was modelled as a sinusoid with a frequency of 1 Hz. Two harmonics were added to create a waveform resembling the arterial pulse more closely. A single heart cycle was simulated at sampling rate of 8 Hz. Multiple heart cycles could be concatenated to closely resemble real world datasets.

Appendix C

Dielectric spectroscopy using UWB radar-on-chip: An exploration

An alternative potential application of RoC technology is that of dielectric spectroscopy. Instead of using pulse reflections to identify and range significant scatterers, the aim of dielectric spectroscopy is to estimate dielectric properties of the propagation medium. The frequency specific attenuation and propagation velocity of an RF pulse propagating through a medium with known path length, depend on the medium dielectric properties. Hence the properties could be estimated across signal bandwidth. A homogeneous medium must be assumed when using a single Tx-Rx channel. As anatomical structures are rarely homogeneous, one could only assess average or effective dielectric properties.

Two potential applications of dielectric spectroscopy are lung fluid monitoring (pulmonary oedema) and continuous blood glucose monitoring. Detection of pulmonary oedema is relevant for patients suffering from heart failure, and has been investigated previously using microwave techniques [144, 145]. A radar-based lung fluid measurement device has been commercialised by medtech company Sensible Medical (Israel). Non-invasive continuous blood glucose monitoring is highly relevant for patients suffering from diabetes and has been researched extensively. Various studies on using microwave techniques for blood glucose monitoring have been reviewed in [146]. Due to the high demand for non-invasive continuous blood glucose estimation methods and the opportunity of developing consumer friendly, wearable applications based on RoC technology, dielectric spectroscopy for blood glucose assessment was explored.

C.1 Consideration of physiology

Path length must be known for assessing dielectric properties. The only way to accomplish

this is by transmissive measurements. The upper arm was identified as a suitable measurement location: The thickness of the arm allows for recording of a clean transmitted signal, the arm is well perfused, and interstitial blood glucose measurements using commercially available monitors are typically taken from the upper arm.

A first step in assessing feasibility was predicting the range of average dielectric properties of the upper arm, and the change that can be expected as a result of increased blood sugar. When dissolving sugar in water, permittivity tends to drop and conductivity tends to rise. The underlying process is the formation of hydrogen bonds of water dipoles to the dipolar regions of sugar molecules. The hydrogen bonds reduce the freedom of water dipoles to polarise in an alternating EM field. One would expect that sugar in blood has a similar effect.

Typical blood glucose levels of an healthy individual, as well as those of people with diabetes, have been well documented. Due to the development of interstitial continuous monitors, the (temporally delayed) interstitial glucose signal is also well defined. When considering the resulting change in average permittivity through the upper arm, one encounters a major challenge: The total amount of blood and interstitial fluid in this section is not well defined. One could predict the water content of the upper arm, but different fluid compartments down the sugar metabolic pathway will have varying glucose concentrations at any given time.

Expected levels of blood glucose From various sources, the following estimates were generated: Typical blood glucose levels of a healthy person are between 70 and 100 mg/dL while fasting. When not fasting, levels will rise but should not exceed 124 mg/dL. The glucose concentration in a diabetic patients blood may vary between 30 mg/dL and 400 mg/dL. Mayo Clinic recommends emergency room treatment when the level exceeds 300 mg/dL.

Expected change in dielectric properties Dielectric properties of liquids with different levels of dissolved sugars have been quantified. Most studies consider dielectrics across a very wide range of glucose levels, with large concentration increments. Typically these studies do not correspond well to expected physiological levels. Two studies were found that presented usable data. In a first study, sucrose was dissolved in water, with a concentration range of 5 to 80 weight% (50 to 800 g/L). Published study results were digitised and a linear fit was obtained for dielectrics at 3.8 GHz. A change in ϵ' of -6 per mg/dL was observed, and a change in ϵ'' of 1.4 per mg/dL [147]. A second study dissolved glucose in water at concentrations 2.0, 10.0 and 20.0 g/L. A linear fit estimated a change in ϵ' of -0.047 per g/L, and a change in ϵ'' of 0.04 per mg/dL [148]. This same study characterised blood with two levels of glucose: 0.0 and 1.35 g/L. Surprisingly, results showed a change in ϵ' of -0.74 per mg/dL, and a change in ϵ'' of -0.22 per mg/dL [148]. That is a fifteenfold increase in $\Delta\epsilon$ compared to the water data, and a change in sign for ϵ'' . Due to some reporting ambiguity in this specific publication, it is not

clear whether reported data were accurate.

C.2 Dielectric spectroscopy experiment

Due to the incomplete and contradictory reports on blood dielectrics, and the unknown factor of the blood and interstitial fractions of the upper arm, an initial proof of concept was pursued before further theoretical work was done.

An experiment was designed to: 1) Demonstrate change in dielectric properties as a result of increasing sugar levels in water, and 2) demonstrate the ability of the X2 RoC to detect this change. A testing range of sugar concentration was chosen to be much larger than physiological range in blood.

Methods A plastic container was filled with 1.5 L of deionised water. Table sugar was added and dissolved, to obtain concentrations ranging from 0 to 30 g/L (corresponding to 3000 mg/dL, or ten times an alarmingly high level when considering physiological levels in blood). Sugar was added in increments of 1 g/L up until a level of 20 g/L. After this, 5 g/L increments were taken until a concentration of 30 g/L was reached. After each sugar addition step, water was mixed until sugar was dissolved and dielectric properties of the mixture were measured using a VNA (Agilent E8361A) and dielectric probe (Agilent 85070E). In addition, the X2 was used to record a transmissive measurement. Body-coupled antennas were wrapped in a thin, isolating rubber layer and submerged in the mixture using a Lego structure. An inter-antenna distance of approximately 5 cm was maintained.

From previous measurements it had been found that phase drift in the X2 due to heating was an issue. In order to perform recordings under consistent conditions, recordings were timed and performed at exact intervals of 3 minutes, giving the radar system time to cool down between short measurement intervals.

X2 pulse amplitude was determined for each recording, as well as phase at peak amplitude. It was assumed that the transmitted signal was the only substantial signal component, making arctangent demodulation of baseband IQ data a valid analysis method.

Results Relative permittivity ϵ' and conductivity σ (measured using VNA) are plotted as a function of sugar concentration in Fig. C.1. The observed change in dielectric properties is larger than reported in [148]. A comparison against [147] is not possible due to the different range in sugar concentrations. X2 measurements are plotted in Fig. C.2, with pulse peak

amplitude (a) and phase at peak amplitude (b). Whereas the peak amplitude signal seems to show a linear decrease, the phase signal contains unexpected jumps.

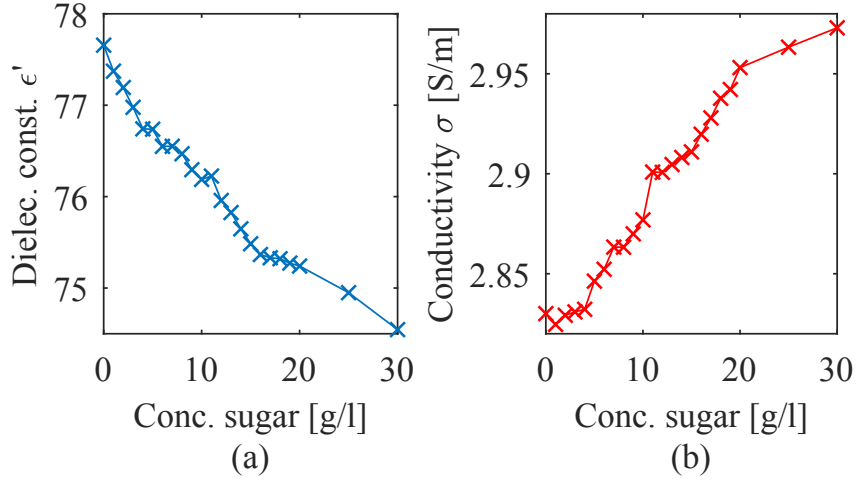


Figure C.1: Dielectric properties measured using a VNA, at 3.8 GHz, across a range of sugar in water concentrations.

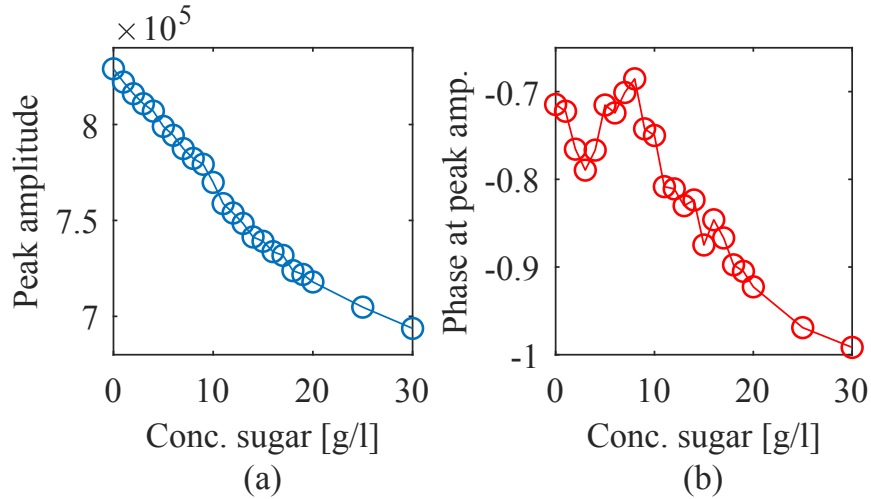


Figure C.2: Transmitted pulse peak amplitude (a) and phase at peak amplitude (b), measured using the X2, across a range of sugar in water concentrations.

Discussion and Conclusion A further analysis of the phase noise in the X2 receiver (Section 3.4.3) demonstrated that phase drift does not stabilise over time, and phase jitter cannot be eliminated without hardware modifications. The obtained experimental results should be replicated, ideally in reverse order of sugar concentration change to test sensor drift effects. Physiological sugar concentrations in blood are of a much smaller range than what was measured in the current experiment, and the volume as a fraction of average arm tissue was not even considered. For these reasons, dielectric spectroscopy using the X2 seemed out of reach

and it was not pursued as part of this research project. However, the XeThru X4 is known to have much better phase stability, and an application specific system could be build which compensates for phase noise. When taking these hardware improvements into consideration, RoC dielectric spectroscopy would make for an interesting research project.

Appendix D

Liquid phantom: Synthesis methods

Below follows the methods for synthesizing a chosen amount of liquid phantom, with dielectric properties similar to grey matter at a frequency of 4 GHz:

1. Take 63 parts (by weight) of deionised water in a large container.
2. Add 37 parts sugar. Mix thoroughly until all sugar is dissolved.
3. If heated water was used to speed up the sugar dissolving process, allow for water to cool to room temperature. Then, add 0.1 part of Virkon (DuPont, USA) to prevent bacterial and fungal growth. Stir until dissolved.
4. While stirring, add 1 part of hydroxyethyl cellulose (HEC; Merck Millipore, USA). Stir until dissolved.
5. HEC acts as a gelling agent, and will cause thickening of the mixture in the first few hours of synthesis. Stir the mixture approximately every two hours on the first day of synthesis, to avoid agglomeration of gel at the bottom of the container.
6. Stirring will trap air bubbles into the mixture, which will change the dielectric properties of the phantom. Air bubbles will escape over time. If the phantom is required immediately, a vacuum may be used to speed up the air bubble release process.
7. If kept in a closed container the shelf life of the phantom was found to exceed 3 years. However, dielectric properties do change over time as the mixture thickens and water evaporates. Permittivity can be increased at all times by adding deionised water and carefully steering the mixture.

Appendix E

Semi-solid phantom: Synthesis methods

An oil-in-gelatine dispersion was used as a semi-solid phantom. Dielectric properties are determined by the ratio of oil to gelatine components: The higher the oil fraction the lower the permittivity. To create the phantom, an aqueous gelatine solution is prepared, as well as an oil mixture. Both are heated to a temperature of 50°C and subsequently mixed together. It was found that phantoms up to 50 % oil could be created. When attempting to build phantoms with an even lower permittivity, the surfactant was found to be insufficient and the phantom oil and gelatine fractions would not fully mix.

Below follow instructions for synthesis of 400 mL of a 50 % oil phantom.

Gelatine part

1. Mix 0.2 g of p-toluic acid (powder) into 10 mL of n-propanol. Heat while stirring until the p-toluic acid has completely dissolved.
2. Mix the solution produced in (1) into 190 mL of deionised water at room temperature, in a beaker.
3. Add 34 g (dry mass) of 225 bloom gelatine to the mixture produced in (2).
4. Cover the beaker with plastic film and heat until the mixture becomes transparent (target temperature 90°C). Heat uniformly in an oven or water bath. Stir for uniformity.
5. Cool the mixture to 50°C.

Oil part:

1. Mix 100 mL safflower oil and 100 mL kerosene in a beaker.
2. Heat the mixture to 50°C.

To synthesize the semi-solid phantom:

1. Mix 200 mL of oil part with 200 mL of gelatine part in a mould (or adjust ratio for desired phantom properties).
2. Add 0.056 mL of surfactant (dish washing liquid) per mL of oil. Mix thoroughly while preventing air bubble formation.
3. When the emulsion has become white and uniform, add 0.0108 g formalin (37 % formaldehyde) per mL of the gelatine solution. Stir briefly. Gelatine setting will occur and continued mixing will lead to gelatine clots on the stirring device and finally inhomogeneity in the phantom.
4. Cool the mixture down in a bath of cold water while minimising movement of the mixture. Gelatin will set with the help of formalin. At high oil fraction mixtures, setting of gelatine should occur as quickly as possible to avoid separation of the oil and gelatine components. Allow multiple days for setting of the gelatine.

A little bit of practise and careful monitoring of temperature is required to achieve gelatine setting before component separation.

Appendix F

Safety considerations

The following considerations were made regarding EM radiation safety for participant and investigator:

- The imaging device contains multiple commercially available RoC devices. The radar chip used is the XeThru X2, produced by Novelda AS, Norway.
- The radar chip emits EM waves with frequencies between 3 and 5 GHz. EM radiation in this frequency band is non-ionising and the only known adverse health effect is the heating of tissue at very high levels of emitted power.
- The X2 chip emits EM waves that are lower in power than the noise floor. The X2 emission spectrum is in compliance with international regulations regarding emission in free space (FCC has assigned an emission mask of $-41.3 \text{ dBm MHz}^{-1}$ between 3.1 and 10.6 GHz). This means that the radar emits as much power as the average vacuum cleaner is allowed to emit unintentionally.
- Transmit gain of 8 dB is added to the radar module.
- International regulations for the exposure of humans to EM radiation exist. Exposure at the used frequencies is expressed as the SAR, which describes the potential of heating of tissue. International guidelines by the ICNIRP state that 2 W/kg (localised 10 g average) can be applied safely to humans (general public; workers exposure is higher).
- Multiple numerical simulation studies have investigated both SAR levels and tissue heating due to EM exposure, and results of similar setups were in the order of magnitude of mW/kg , with negligible temperature rises.
- The X2 transmitted power (true RMS) is measured at -10.7 dBm . With 10 dB added, only 0.53 mW is transmitted. To assess whether this falls below the safety guidelines for EM exposure, we consider the following scenario:
 - Assume that all of this power is delivered to the transmitting antenna, transmitted into the head and absorbed by a single gram of tissue. This would result in a SAR

of 0.53 mW/g. The guideline for localized SAR exposure of 2 W/kg is intended to be used as a 10 g average, resulting in an exposure of 0.053 W/kg, that is 38 times lower than the limit.

- A more realistic scenario would be to consider a larger volume of tissue in which the EM energy is absorbed. If we consider the penetration depth of EM waves at centre frequency 3.8 GHz, we find that the power of the EM field has reduced to 13.5% (an arbitrary, but frequently used value for these scenarios) after 2.4 cm of propagation into the head (estimated using the model described in Section 3.2.2). If we average the inserted power over a volume of 29.0 cm³ (the volume of a semi-sphere with radius of 2.4 cm), we find a localized SAR 109 times lower than the stated guideline.
- Both scenarios described above are pessimistic and an overestimation of the SAR levels we should actually expect: In reality it is not possible to deliver all transmitted energy to the antenna, and definitely not into the head.
- Finally, in our configuration with multiple radar modules, not more than one radar will ever transmit at a time. ICNIRP guidelines are meant to be averaged over a 6 min duration, which means that if we would be using 8 radar modules, the localized exposure averaged over time would again be a factor 8 lower.

Appendix G

Ethics

The study entitled ‘In-body sensing and imaging using ultra-wideband radar’ was approved by the Imperial College EEE Head of Department on 03/05/19 and by the Joint Research Compliance Office on 14/05/19. SETREC reference number 19IC5232.

The letter of approval is inserted on the following page.

15 May 2019

Dear Dr Timothy Constandinou

Study Title: In-body sensing and imaging using ultra-wideband radar

SETREC reference: 19IC5232

The above study was approved by your Head of Department on 03/05/19 and by the Joint Research Compliance Office on 14/05/19.

Under the Science Engineering Technology Research Ethics Committee process, a study that has been reviewed by the Joint Research Compliance Office and Head of Division/Department (or Principal), where no significant ethical issues have been identified in the protocol or ethics application, can be approved without requiring it to go to full committee.

Documents

The documents reviewed were:

- Application form (v4 12/05/19)
- Protocol (v4 12/05/19)
- Participant Information Sheet (v3 12/05/19)
- Participant Information Sheet supplement (v1 12/05/19)
- Consent Forms (v2 14/05/19)
- Risk Assessment
- JRCO Sponsorship and Insurance Request

Yours sincerely,



Ruth Nicholson,
Head of Research Governance and Integrity,
Imperial College London

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