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Author(s): Paška, Jan

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Modeling, Design, and Safety of RF-Transmitters for High Field MRI

A thesis submitted to attain the degree of DOCTOR OF SCIENCES of ETH ZURICH

presented by
JAN PAŠKA
Dipl.-Ing., Technische Universität Hamburg-Harburg
born on 8.4.1981
citizen of Czech Republic

accepted on the recommendation of
Prof. Dr. Klaas P. Prüssmann, examiner
Dr. Jürg Fröhlich, co-examiner
Prof. Dr. Daniel Erni, co-examiner

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Summary

Going to higher $B_0$ field strengths in MR imaging is promising due to the higher expected SNR. With the increase in field strength however comes a more complex behaviour of the electromagnetic (EM) field, due to the decreased RF wavelength, reaching the dimensions of the human body (3T for body imaging and 7T for head imaging). This results in non-uniform images, and even non-uniform contrast, which might lead to a misinterpretation from radiologists in the clinical routine. Furthermore, localized tissue heating might occur, therefore a more careful safety assessment becomes necessary to predict the maximal RF input power.

A more uniform transmit field is traditionally achieved by multiple RF-transmit coils with independent transmit amplifiers. The higher degrees of freedom is beneficial for the imaging performance, but complicates further the RF safety assessment. Therefore, full wave EM-modeling becomes a necessary tool for the careful safety assessment at high field strengths and for the design optimization of coil arrays.

The first part of this thesis, consisting of chapters 1 and 2, deals with the modeling, design, and safety of RF coil arrays.

In the first chapter a simulation method for a faster simulation of RF coil arrays was developed. A coil array consists of many lumped elements, such as capacitors and inductors, for tuning the coil elements, for decoupling between coils, and for matching the coils. Traditionally, the tuning, matching and decoupling in EM-simulations is done iteratively. Many iteration steps are necessary in order to find the optimal values for the lumped elements on the coil array, each one requiring a
new 3D full wave EM-simulation, leading to lengthy simulation procedures. Therefore, a new simulation method was developed that allows variable lumped elements in the 3D EM full wave simulation, and thereby reducing the number of full wave simulations to one. The actual lumped element distribution is added in a fast post-processing step, which speeds up the simulation process significantly. Furthermore, a new matching strategy is introduced, matching and simultaneously decoupling each channel with a matrix matching network, to avoid power reflection and scattering.

In the second chapter an improved stripline coil array for head imaging at 7T was designed, constructed and demonstrated. The main improvements are reduced cable currents, leading to a more safe and stable operation, and a larger coverage along the axial direction. This was achieved with a new feeding strategy. The coupling to the stripline elements was shifted from the side of the elements, as is done traditionally, to the middle of the strip. The feeding was enabled by a balanced to unbalanced network to connect the unbalanced coaxial cables to the balanced coil elements. The outer conductor of the coaxial cable could then be connected to the RF ground in the symmetry plane of the elements. An additional outer shield prevents the coil’s EM-fields, excited by the differential mode, to couple to the cable current mode. Further improvements include an increased sensitivity by finding a capacitance distribution that maximizes the sensitivity and at the same time the uniformity in the transversal plane with the simulation method developed in the first chapter. The maximal input power was also increased by using novel cooling elements to increase the power handling capability of the trimmer capacitors. The simulation of the coil array was thoroughly validated in a reference scenario with direct field measurements as well as $B_1^+$ mapping. For demonstration purposes in-vivo MR-images were acquired.

The second part of this thesis deals with the traveling wave system. The principle is introduced in chapter 3, the extension to multiple transmit channels in chapter 4, and the corresponding safety assessment and in-vivo imaging are shown in chapter 5.

In the third chapter a new way of RF transmission and reception
is introduced, called traveling wave. The $B_1$ field distribution is not linked to a particular design of an RF coil and its location – whose presence, usually close to the imaging object, changes the boundary conditions of the EM-fields – but to a waveguide mode – a wave traveling a long distance down the bore of the scanner. A $B_1$ field excited in this way is only dependent on the modes of the cross-section of the RF bore, and the coupling coefficients from the excitation antenna to these waveguide modes.

The cylindrical bore of the MR scanner can be considered as a circular waveguide. For human MR systems, the cut-off frequency, above which wave propagation inside the RF bore occurs, is close to the larmor frequency for a $B_0$ field strength of 7T. The wave is excited by an antenna at the far end of the scanner’s bore, and is used to couple to the spins, and to carry the MR signal to the antenna for reception.

In the traveling wave setup, the imaging object lies fully in the far-field of the excitation antenna, as opposed to the traditional near field coupling mechanism. The far field coupling has advantages over the near field coupling in terms of transmit field uniformity along the axial direction. The EM-fields have to have a curvature, in order to satisfy the Helmholtz equation. In the near field the curvature is dominated by the curvature in the amplitude of the EM-field, in the far field the curvature is obtained by the phase of the propagating wave. In this way, the amplitude along the axial direction remains more uniform, yielding a more uniform transmit field, and a large field of view, which makes the travelling wave concept a promising candidate for volume coils for high field strengths.

Furthermore, the traveling wave setup is beneficial for RF safety due to the large distance between the conservative electric fields from the conductors and capacitors of the excitation antenna and the human body. Patient space is also enhanced, no close fitting coils surround the patient.

In the fourth chapter, the travelling wave concept was extended to multiple channels, enabling RF shimming and parallel imaging. This was achieved by a multi-modal waveguide extension. As mentioned before, close coupling coils achieve the spatial selectivity by the lo-
cation of the coil elements. For the parallel travelling wave system, the spatial selectivity is achieved by coupling into different waveguide modes. For eight transmit channels, at least eight modes have to propagate in the RF bore. This was achieved by a dielectrically loaded waveguide extension, that was inserted into the RF bore. A partial filling with tubes filled with deionized water was found to be sufficient. The modes in the waveguide extension were excited with stubs or loops.

It was shown with $B^1_+$ mapping in a phantom, that a reasonably strong and spatially diverse transmit field distributions can be excited. This spatial diversity of the field distributions was confirmed with parallel imaging experiments, yielding a moderate noise amplification with acceleration factors of up to 3.

The results of the traveling wave setup with multiple channels obtained in an imaging phantom showed to be promising, therefore an extension to in-vivo imaging was pursued in chapter 5. Traditional EM-simulation approaches, modeling the whole RF setup in one domain was not possible due to the large size. The RF-setup consists of pieces of waveguides, therefore, the mode matching technique was found to be suitable, that allows to divide the setup into smaller subdomains at suitable reference planes, that are simulated separately. The sub-domains are then joined in a post-processing step. The finite element method was found to be more suitable than current FDTD implementations for the simulation of this setup. The last subdomain, the excitation section of the waveguide extension could not be modeled, leading to an overestimation of the power limit. The simulation approach was validated with $B^1_+$ maps in an imaging phantom, and in-vivo MR-images were acquired for demonstration purposes.
Zusammenfassung

Höhere $B_0$ Feldstärken sind vielversprechend für die Magnetresonanztomographie (MRT) aufgrund des erwarteten höheren SNR. Mit der Zunahme der Feldstärke geht jedoch ein komplexeres Verhalten des elektromagnetischen (EM) Feldes einher, aufgrund der kleineren RF-Wellenlänge, die die Masse des menschlichen Körpers erreicht (3T für Bildgebung im Körper und 7T für Bildgebung im Kopf). Dies führt zu nicht uniformen Bildern und zu nicht uniformen Kontrast, was zu einer Fehlinterpretation von Radiologen im klinischen Alltag führen könnte. Ausserdem kann eine fokussierte Erwärmung vom menschlichen Gewebe auftreten, weshalb eine sorgfältige Sicherheitsbeurteilung notwendig wird, um die maximale HF-Eingangsleistung zu bestimmen. Ein uniformeres Sendefeld wird gewöhnlich von mehreren HF-Sendespulen mit unabhängigen Sendeverstärkern erreicht. Die höhere Anzahl von Freiheitsgraden ist von Vorteil für die Bildqualität, erschwert aber die HF-Sicherheitsbeurteilung. Daher werden numerische Feldberechnungsverfahren notwendig für die sorgfältige Sicherheitsbeurteilung und für die Designoptimierung von Spulenarrays bei hohen Feldstärken.

Der erste Teil der Arbeit, bestehend aus den Kapiteln 1 und 2, beschäftigt sich mit der Modellierung, Design und Sicherheit von HF-Spulenarrays.

Im ersten Kapitel wurde ein Simulationsverfahren für die schnellere Simulation von HF-Spulenarrays entwickelt. Ein Spulenarray besteht aus vielen konzentrierten Bauelementen wie Kondensatoren und Induktoren, zum Abstimmen der Spulenelemente, zur Entkopplung zweier
Zusammenfassung

Spulen, und zur Anpassung der Spulen. Üblicherweise wird die Abstimmung, Anpassung und Entkopplung in EM-Simulationen iterativ durchgeführt. Viele Iterationsschritte sind notwendig, um die optimalen Werte für die konzentrierten Bauelemente auf dem Spulenarray zu finden, die jeweils eine neue EM-Simulation erfordern, was zu langwierigen Simulationsprozeduren führt. Daher wurde eine neue Simulationsmethode entwickelt, die variable konzentrierte Elementen in der EM-Simulation erlaubt, und somit die Anzahl der EM-Simulationen auf nur eine verringert. Die tatsächlichen konzentrierten Bauelemente werden in einem schnellen Nachbearbeitungsschritt hinzugefügt. Das beschleunigt den Simulationsprozess enorm. Desweiteren wurde eine neue Anpassungsstrategie vorgestellt, welche die Anpassung und gleichzeitige Entkopplung von jedem Kanal eines Spulenarrays mit einem Matrix-Anpassungsnetzwerk erreicht, um Reflexion und Streuung zu vermeiden.

Im zweiten Kapitel wurde ein verbessertes Streifenleitungsspulenarray für die Kopfbildgebung bei 7T entworfen, gebaut und getestet. Die wichtigsten Verbesserungen sind reduzierte Mantelströme, was zu einem sichereren und stabileren Betrieb führt und eine größere Abdeckung entlang der axialen Richtung. Dies wurde mit einer neuen Kopplungsstrategie erreicht. Die Kopplung zu der Streifenleitung der Spulenelemente wurde von der Seite, wie es herkömmlicherweise der Fall ist, zu der Mitte des Streifens hin verschoben. Die Kopplung wurde durch einen Balun ermöglicht, um das unsymmetrische Koaxialkabel mit den symmetrischen Spulenelementen zu verbinden. Der Außenleiter des Koaxialkabels wurde dann mit der HF-Masse in der Symmetrieebene der Elemente verbunden. Eine zusätzliche äussere Abschirmung verhindert, dass die EM-Felder der Spule, die durch das differentielle Signal der Koaxialleitung angeregt werden, an die Kabelmode koppeln. Eine weitere Verbesserung ist die erhöhte Empfindlichkeit, die durch eine optimale Kapazitätsverteilung, die mit dem Simulationsverfahren aus dem ersten Kapitel gefunden wurde, erreicht wurde, welche die Empfindlichkeit und gleichzeitig die Felduniformität in der transversalen Ebene maximiert. Zusätzlich wurde die maximale Eingangsleistung erhöht durch die Verwendung neuartiger

Der zweite Teil der Arbeit befasst sich mit Wanderwellen-MRT. Das Prinzip wird in Kapitel 3 eingeführt, die Erweiterung auf mehrere Sendekanäle in Kapitel 4, und die entsprechende Sicherheitsbeurteilung und in-vivo-Bildgebung werden in Kapitel 5 vorgestellt.

Im dritten Kapitel wird eine neue Art des HF-Sendens und Empfangs eingeführt, die so genannte Wanderwellen-MRT. Das $B_1$-Feld ist nicht mehr mit einem bestimmten Design einer HF-Spule und seiner Lage verknüpft, deren Anwesenheit, in der Regel in der Nähe des Imaging-Objekts, die Randbedingungen der EM-Felder ändert, sondern mit einer Wellenleiter-Mode, eine Welle, die eine weite Distanz die zylindrische Öffnung des MRT Scanners propagent. Das $B_1$ Feld, das auf diese Weise erzeugt wird, ist nur abhängig von den Moden des Querschnittes des zylindrischen Rohres und von der Kopplung der Anregungsantenne an die Wellenleitermoden.


Bei der Wanderwellen-MRT liegt das Abbildungsobjekt vollständig im Fernfeld der Erregerantenne, im Gegensatz zum traditionellen Nahfeldkopplungsmechanismus. Die Fernfeldkopplung hat Vorteile gegenüber der Nahfeldkopplung in Bezug auf Felduniformität entlang der axialen Richtung. Die EM-Felder weisen eine Krümmung auf, um die Helmholtz-Gleichung zu erfüllen. Im Nahfeld dominiert die Krümmung der Amplitude, während im Fernfeld die Krümmung der Phase der propagierenden Welle überwiegt. Dadurch wird die Amplitude in der axialen Richtung uniformer, was ein uniformeres und größeres Sendefeld ermöglicht. Das macht Wanderwellen-MRT zu einem vielver-
sprechenden Kandidaten für Volumenspulen für hohe Feldstärken.

Darüber hinaus ist der Wanderwellenaufbau von Vorteil für die HF-Sicherheit aufgrund der grossen Entfernung zwischen den konservativen elektrischen Feldern der Leiter und der Kondensatoren der Anregungsantenne und dem menschlichen Körper. Der Patient hat mehr Platz, weil ihn keine eng anliegenden Spulen umgeben.


Mit $B_1^+$-Mapping in einem Phantom wurde gezeigt, dass ein einigermassen starkes und räumlich diverses Sendefeld angeregt werden kann. Die räumliche Diversität der Feldverteilungen wurde mit parallelen Bildgebungsexperimenten bestätigt, welche eine moderate Rauschverstärkung mit Beschleunigungsfaktoren von bis zu 3 ergeben.

Die Methode wurde als besser geeignet beurteilt, als aktuelle FDTD-Implementierungen, für die Simulation dieses Setups. Der letzte Teilbereich, der Anregungsabschnitt des Verlängerungsrohres, konnte nicht modelliert werden, was zu einer Überschätzung der Sicherheitsbeurteilung führt. Der Simulationsansatz wurde mit $B_1^+$-Karten in einem Phantom validiert, und in-vivo MR-Bilder wurden zu Demonstrationszwecken aufgenommen.
Chapter 1

Introduction

1.1 MR imaging

MRI is a tomographic imaging technique, that offers not only insight into the anatomy of the human body, but also its physiology and metabolism. It is a relatively new imaging technique. Felix Bloch and Edward Purcell discovered the magnetic resonance phenomenon independently in 1946 [1,2]. Paul Lauterbur generated the first MR images in 1973 using spatially varying gradients [3], while in 1975 important ideas like phase and frequency encoding and the Fourier Transform were introduced to MRI by Richard Ernst [4]. MRI is very slow compared to other imaging modalities such as Computer Tomography. However, it offers a unique soft tissue contrast and is indispensable for the detection of tumors or joint injuries. Nowadays, there are approximately 10000 MR scanners worldwide, and approximately 75 million MRI scans are performed each year [5].

1.1.1 Basic Principles

The image formation in MRI is based on the interaction of three types of magnetic fields with matter. The first one is the main magnetic field $B_0$, a very strong, static, and homogeneous field, created by a current in a superconducting solenoid coil that is cooled by liquid Helium. The
hydrogen nuclei in the human body align with the static magnetic field. The aligned nuclei have a resonance frequency, the larmor frequency \( f_L \), that is proportional to the \( B_0 \) field:

\[
f_L = \frac{\gamma}{2\pi} B_0
\]  

(1.1)

with \( \frac{\gamma}{2\pi} = 42.58 \text{MHz/T} \), and \( \gamma \) being the gyromagnetic ratio for hydrogen.

The second magnetic field is the RF field \( B_1 \), that excites the nuclei at the larmor frequency, by an appropriate RF transmit coil. The nuclei then start to precess about the axis of the main magnetic field. After the excitation they emit RF-energy at larmor frequency that can be received with an RF receive coil. A third magnetic field, the gradient field, modulates the static magnetic field \( B_0 \) in a way such that the spatial information of the spins is encoded in the local larmor frequencies. The gradient fields operate in the acoustic range of frequencies, and are produced by three gradient coils, that generate a linear magnetic field in the three Cartesian directions. The spatial information of the spins can be retrieved by Fourier transform. The image formation through gradient encoding makes MR a slow imaging modality.

Two different components of the \( B_1 \) field are responsible for the transmission and reception of the RF energy with the spins [6]. For the transmission the mathematically positively rotating part of the transverse field \( B_1^+ \) couples to the spins. The receive sensitivity is the mathematically negatively rotating part of the transverse field \( B_1^- \). These components can be computed as:

\[
B_1^+ = \frac{B_{1,x} + i B_{1,y}}{2} \quad B_1^- = \frac{(B_{1,x} - i B_{1,y})^*}{2}
\]  

(1.2)

\( B_1^+ \) and \( B_1^- \) are only the same for a linearly polarized \( B_1 \) field, which is only possible at low \( B_0 \) strengths. At higher field strengths the \( B_1 \) field is elliptically polarized and \( B_1^+ \) and \( B_1^- \) differ [7].

The signal to noise ration (SNR) in the MR experiment is given by the sine of the flip angle by which the spin is rotated away from the main magnetic field axis divided by the Johnson noise stemming from
the coil resistance $R_{\text{coil}}$, the sample resistance $R_{\text{sample}}$, and radiation the resistance $R_{\text{rad}}$, [8].

$$\text{SNR} = \frac{SI}{\sqrt{N}},$$

with $SI \propto f^2 \sin(\gamma \tau |V \hat{B}_1^+|) \hat{B}_1^-$, \hspace{1cm} (1.3)

and $N = \sqrt{4kT \Delta f (R_{\text{coil}} + R_{\text{sample}} + R_{\text{rad}})}$.

Where $V$ is the voltage applied to the RF coil during transmission, $\tau$ is the duration of the RF pulse, $\hat{B}_1^+$ and $\hat{B}_1^+$ are the transmit field and the receive field for unit voltage, $k$ the Boltzmann constant, $T$ the temperature, and $\Delta f$ is the receiver bandwidth. Two separate coils for transmission and for reception can be used, or a single RF coil for both tasks.

### 1.1.2 MRI at High Field Strength

MRI is an intrinsically insensitive imaging modality, therefore increasing the SNR is crucial. This can be achieved by going to higher $B_0$ field strengths, as shown in Eq. (1.3) The signal is proportional to the larmor frequency squared, the noise resistance in the sample is proportional to the larmor frequency, therefore the SNR is proportional to the $B_0$ field strength. The higher frequency also comes with a shortened wavelength which is beneficial for RF-encoding in the transmit (Transmit SENSE) as well as in the receive case (parallel imaging). Currently, the highest field strength of 11.7T for whole body human imaging is under construction in Saclay, France.

However, with the increased $B_0$ field strength come also additional technical challenges. The RF wavelength comes in the order of the imaging object (at 7T $\lambda \approx 12\text{cm}$ in human tissue) leading to wave effects and standing wave patterns causing non-uniform transmit field distributions, leading to non-uniform image intensity and contrast, reducing the quality of MR images. The shorter wavelength leads to specific absorption rate (SAR) hotspots, which makes the safety assessment more difficult, as will be described in a following section.


## 1.2 RF coils for MRI

RF coil design is mainly governed by the ratio of the RF wavelength to the size of the imaging subject. At low $B_0$ field strength, the wavelength is much larger than the imaging region, the magnetic field produced by the RF coil is hardly altered by the presence of the imaging object, due to the weak coupling of the E- and H-field, and the magnetic field can be very well predicted by quasi-static approximations using Biot Savart’s law. The imaging object lies fully in the near field of the RF coil which is designed to produce a high magnetic field.

At high $B_0$ field strength, starting at 3T for body imaging and 7T for head imaging, the RF wavelength inside biological tissue comes into the range of the imaging target’s size. Far-field effects start to play a role, such as propagation and reflection of waves at dielectric boundaries. The electrical field is strongly coupled with the magnetic field. To predict the RF behaviour a quasi-static approximation is not sufficient anymore, the full wave Maxwell’s equations need to be solved. The boundary conditions of the MR scanner play an important role in the coil design. The cylindrical RF shield can be regarded as a circular waveguide for RF frequencies, for field strengths of 7T and above RF shield of human MR systems is above the cutoff frequency and propagation effects start to occur that pose an additional challenge for RF coil design, but can also be exploited.

The performance of a transmit RF coil can be measured in terms of the power sensitivity or the SAR sensitivity within a region of interest (ROI):

$$B_{1,P}^+ = \frac{\bar{B}_1^+}{\sqrt{P}} \quad B_{1,SAR}^+ = \frac{\bar{B}_1^+}{\sqrt{SAR}} \quad \text{with} \quad \bar{B}_1^+ = \frac{1}{N} \sum_{i=1}^{N} |B_1^+(r_i)|$$ (1.4)

Where the $B_1^+(r_i)$ is summed over all $N$ voxels $i$ within the ROI, and the SAR is the limiting SAR, according to [9], which will be discussed in a following section.

Another important measure is the uniformity of the transmit field in the ROI. To obtain a homogeneous image with a homogeneous contrast
across the image, and mainly to avoid dark bands in the image, because the flip angle is an integer multiple of 180°, see Eq. 1.3, the $B_1^+$ has to be uniform. One possible measure for the uniformity of the transmit field is the normalized standard deviation:

$$s_n = \sqrt{\frac{1}{N} \sum_{i=1}^{N} (|B_1^+(r_i)| - \bar{B}_1^+)^2}$$

(1.5)

1.2.1 RF Coils at low Field Strengths

RF coils at low frequencies can be divided into two classes: volume coils and surface coils. Volume coils are commonly further away from the sample and are able to excite a uniform transmit field over a large field of view (FOV), at the cost of a low efficiency and SNR. They require only one transmit amplifier. The first volume coils were solenoid coils, followed by saddle coils, and quadrature coils, such as the birdcage coil and TEM coil. Surface coils on the other hand are close to the sample. They have the advantage of a high efficiency and a good SNR, at the cost of a small FOV and an nonuniform field distribution. Common surface coil designs are the loop coil or the stripline coil.

Assembling surface coils into an array yields a large FOV while maintaining a high SNR. An optimal solution for many cases at low field strength MR imaging, is a surface coil array for reception combined with a volume coil for transmission, this yields a homogeneous transmit field and a high SNR over a large FOV. The sum of squares [10] reconstruction technique can be used to reconstruct the images in the case of array reception. The additional spatial information from the single receive channels can also be used to accelerate imaging by undersampling k-space with parallel imaging techniques [11, 12, 13].

1.2.2 RF Coils at high Field Strengths

The transmit field distribution produced by volume coils is highly inhomogeneous at high field strengths, due to the shortened wavelength, as described above. Additional degrees of freedom are needed to counteract the field non-uniformity. This is achieved by using an array of
surface coils for transmission [14]. Static shimming or dynamic shimming techniques are available to shape the transmit field.

In static shimming the phases and amplitudes of the coil array input voltages are fixed for the duration of the MR sequence. In principle, only one single transmit amplifier is necessary followed by stages of quadrature hybrids to split the amplifier output signal into equal amplitude signals for the channels of the coil array, see [14]. Phase shifters and attenuators between the quadrature hybrids and the coil array are used to adjust the phase and amplitude of the input voltages. With the knowledge of the transmit field sensitivities, using $B_1^+$ mapping [15], the magnitude least squares technique can be used to find the required pulses for a uniform transmit field distribution [16].

Parallel transmission [17, 18] is very promising for MRI at high field strengths, it allows for more degrees of freedom to control power deposition in the human while accelerating the excitation. It enables spatially selective excitation pulses [19], due to the acceleration, to excite a desired magnetization pattern. In parallel transmission not only the phases and amplitudes of the in- put voltages differ, they are also modulated in time for each channel independently. The hardware complexity increases, as it requires individual transmit chains for each channel. With increasing $B_0$ field strength also the number of required transmit channels for a uniform excitation increases. For 3T and 7T scanners up to eight transmit channels are available [16, 20], and more recently, 9.4T scanners with up to 16 transmit channels were reported [21, 22].

The most common RF transmit coil array design for high field strength is the loop coil array [22], it offers a high sensitivity. Another widely used design is the stripline array [14], this design offers a good decoupling between the channels. With the change from the near field to the far field some groups propose the use of radiative antenna elements for the use as RF coils [23, 24], inspired from antennas used for RF hyperthermia. Radiative antennas are linear dipoles that are designed to excite an EM wave in the human body rather than creating a magnetic field in the near field as is the case for conventional RF coils. Another promising approach is the combination of dipole and
loop elements in a coil array as presented in [25].

An additional challenge for high field systems is the RF power lost through radiation, due to the fact that the larmor frequency is above the cut-off frequency of the scanner’s RF bore. This effect can be exploited by using the RF-bore as a waveguide to guide the transmit and receive RF signal between the imaging sample and the excitation element. This transmission technique is called travelling wave, it offers a good field uniformity along the direction of propagation, due to the constant amplitude of travelling waves, and more patient space. The travelling wave technique is part of this thesis and will be presented in following chapters [26, 27].

1.2.3 RF coil design

RF coils and RF coil arrays play an essential role in the quality of MR images. Since the beginning of MR imaging a lot of effort was undertaken in order to improve MR images with the optimization of RF coils. Lumped element models were developed to optimize the performance of predefined RF coil structures, such as the birdcage [28]. As the wavelength effects could not be neglected anymore, transmission line models were developed by Baertlein et al [29] for the design of microstrip line TEM resonator coils. This method was extended further in [30], to include a dielectric load with an inhomogeneous cross-section, but the dielectric load remained uniform along the axial direction. The inverse design approach can be used to find surface currents on a cylindrical shell to achieve a desired $B_1$ field distribution in a FOV; the streamline function technique is then used to generate the conductor patterns of the RF coil [31].

To evaluate the performance of a receive coil array design, comparisons to the maximally achievable SNR, the ultimate intrinsic SNR (uiSNR), can be made [32]. The uiSNR can be computed in simple geometries, such as a sphere or infinitely long cylinder [33, 34]. The uiSNR in a voxel is then found based on an infinite set of modes, that describe all possible field distributions in the object that satisfy Maxwell’s equations given the boundary conditions. It is shown that higher field strengths are beneficial for parallel imaging and parallel
transmission [34], due to the increasingly diverse field distribution in the imaging object.

In [35] the EM-field in a dielectric spherical and cylindrical phantom is developed in terms of basis functions that are excited by surface current modes on a spherical or cylindrical shell. The uiSNR along with the minimally possible SAR, and the corresponding ideal current patterns are then found, based on the surface current modes. It is difficult to synthesize these ideal current patterns into a concrete coil design - they turn around the object with the larmor frequency - however, they can serve as a design guide for high field strength. An important insight gained from this analysis is that loop currents dominate the ideal current pattern at low frequencies, but as we move to higher field strengths electric dipoles become more and more necessary to achieve the uiSNR or minimum SAR [25].

All these simplified methods described above are very valuable in understanding basic principles of RF coil design, however, none of these simplified methods is able to predict the EM-fields inside an inhomogeneous load, such as the human body, which is necessary for the design of coil arrays for ultra high field human MR-systems.

1.3 Safety for RF coils

Thermal heating is the only well confirmed effect of biological tissue in response to EM-fields in the frequency range of RF fields used for MRI [36]. The deposited RF power is converted into heat due to the conductivity of the tissues. RF amplifiers can deliver up to tens of kilowatts of output power, which can lead to severe tissue damage. The CEM43 (cumulative equivalent minutes at 43°C) time is used to estimate the damage of tissue due to heat [37]. For this, the temperature distribution inside the human body has to be known, which is very difficult to obtain reliably. It depends on the outer environment, such as the temperature in the scanner room, the air flow, the RF coil and sequence used, as well as on the human object, size, internal distribution and electrical properties of tissues, vasculature, and the thermo-regulatory response. These factors are all different from indi-
vidual to individual, in addition, the thermo-regulatory response may be impaired by illnesses such as fever or by medication.

### 1.3.1 Current standards

A maximal temperature increase of 1°C was set by the International Electrotechnical Commission (IEC) for MR examinations [9]. The current approach in order not to exceed the temperature limit reliably is to set a limit on the specific absorption rate (SAR) defined as the power deposition $P_L$ within an averaging volume $V$, that has the mass $m$.

$$\text{SAR} = \frac{P_L}{m} = \frac{1}{V} \int_V E^*(r) \sigma(r) \rho(r) E(r) dr$$  \hspace{1cm} (1.6)

Where $E$ is the electrical field, $\sigma$ is the tissue conductivity, and $\rho$ is the mass density of the tissue. SAR is the source for the temperature increase, and is independent of the less known variables such as the vasculature, and the thermo-regulatory response. SAR limits are given averaged over the whole body, the head, or the extremities, and for over any 10g cube inside the human body [9].

For volume transmit coils at low field strengths the whole body SAR is the limiting factor for safety. The maximum allowed input power can be simply estimated by assuming that all power from the transmit amplifier is dissipated in the human body. The whole body SAR can then be computed as the output power, that can be monitored at the amplifier, divided by the patient mass. At higher field strength the 10g SAR is the limiting factor, and 3D EM-full wave simulations involving a human body model are necessary.

For a transmit coil array with independent transmit amplifiers the involved physics and the hardware for the safety assessment becomes more complex. The SAR distribution varies in space with the applied RF pulse, and the input voltages have to be monitored at each channel. SAR matrices have to be computed from EM-simulations for every voxel, and the SAR distribution has to be calculated for every RF pulse. For a human body millions of SAR matrices would have to be evaluated which is very time and memory consuming. The method of virtual observation points is a close upper bound to the precise SAR
distribution. It can be used to compress the amount of SAR matrices to a few SAR matrices that can be evaluated faster and memory efficient [38].

The worst case SAR is the maximal SAR for a given input power, it can be computed as the maximum of all highest eigenvalues of all SAR matrices within the human body. For the worst case SAR only the total input power of each RF pulse has to be monitored, in order to stay within safe limits. For a safety assessment based on the worst case SAR the hardware is less complex, only the total input power into the array has to be known, but it is the most conservative and hinders the image quality.

1.3.2 SAR mapping

An elegant approach for safety assessment is to derive the SAR from the electrical field and the electrical properties of the tissue derived from MR measurements. Electrical Properties Tomography [39] is one possible method, which shows promising results for low field strengths. However, it is based on symmetry assumptions of the receive and transmit sensitivities that do not hold at higher field strengths. Local Maxwell Tomography [40] does not rely on any symmetry assumptions and is a promising candidate for SAR mapping at higher field strength.

1.3.3 Temperature based approaches

The current safety limits based on SAR are rather conservative in terms of allowed input power, which limits the imaging performance at high field strengths. The tissue is merely heated in a healthy human. Tissue is damaged by temperature as mentioned above, therefore, a more direct safety assessment relying on temperature rather than on the specific absorption rate would allow a higher input power. Many groups conduct research in the direction of a temperature based safety assessment [41, 42, 43]. This might allow tighter safety limits allowing a higher input power and therefore faster sequences and better contrast.

MR thermometry is a technique to acquire relative temperature maps with MRI, by exploiting the frequency shift of the proton lar-
mor frequency with temperature [41]. This technique however is rather insensitive, furthermore, different frequency shifts were reported in literature [44], and it is time consuming especially for RF coil arrays with many channels. Local power matrices have to be acquired, therefore the measurement time scales with the number of channels squared [42, 44]. Ideally, the temperature inside the human body would be monitored during the MR scan, in-vivo MR thermometry however is challenging due to the high power requirements, long duration, and low sensitivity. Temperature measurement at single points with fluoroptic probes are faster and more sensitive, however they cannot be used to measure temperature inside a living human. They can be used to measure temperature on the skin [45], or for validation purposes in phantoms [42], or anesthetized animals [46].

The temperature can be predicted by solving Pennes’ Bioheat Transfer Equation (PBTE) [47] numerically. The PBTE relies on EM-field simulations predicting the SAR distribution, that are used as a source term for the temperature increase inside the tissue [48, 49]. Temperature simulations require not only the electrical properties of the tissue, but also the thermal properties, thermal conductivity, as well as the perfusion rate and the metabolic heat generation rate. The PBTE does not consider the increase in body core temperature due to the heating of blood, or the thermo-regulatory response of the living human. MR thermometry as well as temperature simulations can help to push the boundaries of MR imaging at high field strength in the future. However, additional research is necessary to increase speed and accuracy for a reliable and fast safety assessment based on temperature.

1.4 3D EM full wave simulations for RF-coils

For the design optimization and safety assessment of RF coil arrays at high $B_0$ field strengths the full Maxwell’s equation in 3D with a human body model, coil geometry, and the proper boundary conditions need to be solved [50]. The boundary conditions include the matching condition of the coil array channels as well as the RF shield of the MR scanner, and the shielding of the MR room itself. In most applica-
tions only the EM-fields and the RF coil’s scattering matrix at larmor frequency under the matched condition is of interest. The broadband behaviour of RF coils are only needed for special applications. Historically, with the increase in $B_0$ field strength, also the computer power increased enabling a full wave numerical solution of the EM-fields inside a human body model in a reasonable time.

The finite difference time domain method (FDTD) was the first numerical method to be used for the simulation of human body models for the safety and design of RF coils [50]. As computer memory increased other, less memory efficient, numerical methods, such as the finite element method (FEM) [51], or hybrid methods using the method of moments (MoM) combined with FDTD or FEM [52] were increasingly used. Each of these three methods have their advantages and disadvantages depending on the circumstances, and will be discussed in the following.

**1.4.1 Finite Difference Time Domain**

The finite difference time domain method (FDTD) was first introduced by Yee in 1966 [53]. It is based on a discretization of Maxwell’s Equations, either in differential form (FDTD) or in integral form (FIT), in time and space. The numerical domain is truncated by an absorbing boundary condition and discretized in a rectilinear fashion. Electrical field components are assigned to the edges, and magnetic field components to the normal of the faces of each unit cell, in between the electrical field components. Dielectric properties are specified for each of these unit cells. After an initialization of the EM-fields to zero, the numerical domain is excited by a current edge source. The excitation can be either a broadband excitation pulse, or a single frequency harmonic excitation. The electric and magnetic fields are then updated in a leapfrog manner. Each electrical field component is updated in every time step with the numerical curl of the magnetic field, and the electrical field component from the previous time step. In the following time step, the magnetic field components are updated based on the numerical curl of the electric field, and the magnetic field component from the previous time step. This updating process is repeated until a steady
state is reached. An FFT can be used to obtain the EM-field phasors at larmor frequency. For a more in-depth description of the FDTD, please refer to [54].

FDTD is an intuitive technique, easy to use and understand allowing for a natural and easy modelling of materials. This is reflected in the fact, that most available numerical body models are currently available for the FDTD method. A fine discretization of the numerical domain is possible; nowadays numerical domains with up to 100 million mesh cells can be solved, with the graphical processor unit (GPU) based accelerator technology [55]. Multiple frequencies can be solved with only one simulation run, which is useful for the design of double tuned RF coils.

On the other hand, FDTD has difficulties in simulating resonant structures with a high quality factor, such as RF coils. The energy does not leave the computational domain, leading to long computation times, resulting in a poor power balance, or even numerical instabilities, that lead to an erroneous solution. No unique permeability or permittivity is defined at material interfaces, due to the dislocated grids for the E- and the H-field. Geometrical features that do not lie along Cartesian directions cannot be modelled accurately due to the rectilinear grid, this effect is called stair-casing. A fine discretization in a part of a numerical domain leads to a large number of grid cells, unless the sub-gridding technique is used.

1.4.2 Method of Moments

The Method of Moments (MoM) is used since the 1960s to solve electromagnetic problems in the frequency domain in free space. The unknown quantity in MoM is the surface current on metallic surfaces. Homogeneous dielectric materials are modelled by the Surface Equivalent Principle by electric and magnetic equivalent surface currents enclosing the dielectric. In this way also inhomogeneous dielectric objects can be modelled by defining a magnetic and electric surface current on the dielectric boundaries. The dielectric boundaries and metallic surfaces are discretized in surface triangles. The electric and magnetic surface currents are developed in basis functions with unknown
coefficients within the surface triangles. The Electric Field Integral Equation is then written in terms of the discretized surface currents. The equation is then weighted by the basis functions to obtain a set of equations. The resulting system matrix is a full matrix that has to be inverted to solve for the unknown coefficients of the basis functions for the electric and magnetic currents. A more detailed description of the method can be found in [56].

MoM is a very powerful simulation tool for the numerical analysis of antenna structures in free space. In MRI it can be used for the simulation of RF coil arrays in the unloaded situation or loaded by simple homogeneous phantoms.

MoM is not very well suited for the simulation of large inhomogeneous dielectrics, such as the human body. The complicated geometry and the large surface of dielectric boundaries would result in a large system matrix that is not possible to solve on current computers. Therefore, it is not suited for the safety assessment of RF coil arrays at higher field strength where local SAR is the limiting factor. For the simulation of the human body hybrid methods can be used, in which MoM is used to model the RF coil and FDTD or FEM are used to calculate the EM-fields inside a human body model.

### 1.4.3 Finite Element Method

The finite element method (FEM), is a numerical method that approximates the solution to boundary value problems. It is used since the 1940s and has a wide range of applications in engineering and mathematics. For electromagnetic 3D full wave problems, the numerical domain is truncated by appropriate boundary conditions, and subdivided into smaller subdomains, called finite elements, usually of tetrahedral shape. Within each of the subdomains the electrical field is expanded in simple basis functions with unknown coefficients. The Galerkin method of weighted residuals is used to set up a system of equations by multiplying the vector wave equation for the E-field by the basis functions and integrating over the whole computational domain. The resulting system matrix, that is a square sparse matrix, has to be inverted to solve for the unknown coefficients. The magnetic field can
then be computed with Faraday’s law. FEM is very well described in [57].

The tetrahedral subdomains are well suited to model complex structures and materials. The unstructured mesh with higher order, called curvilinear, tetrahedral elements is very well suited for the accurate modelling of curved surfaces and fine structures. The simulation time for FEM frequency domain solvers is independent on the number of excitations, which is beneficial for coil arrays with a large number of channels. In most EM-simulations for MRI the EM behaviour of the coil array only at the larmor frequency is of interest, therefore a frequency sweep is not needed.

A disadvantage of FEM is the high memory requirement due to the necessary inversion of the large system matrix. Nowadays, modern PCs can handle the computational burden of simulating a human body model with a fine enough discretization.

1.4.4 Validation of Simulations

For safety assessments based on EM-simulations, the validation of the simulations is a crucial step in order to trust the numerical results. Reference scenarios that can be simply measured are used, such as cylindrical or spherical phantoms or the ASTM phantom [58] filled with a tissue simulating liquid. More recently a human head phantom with different compartments for the different tissue types in the head was developed in order to have a more realistic reference scenario [59], which becomes more important at higher field strengths.

MRI itself can be used to validate EM-field simulations of RF coils. With the $B_1^+$ mapping technique [15], the amplitude of the $B_1^+$ as well as the phase differences between the channels of an RF coil array can be measured in a phantom or in-vivo. The previously mentioned temperature mapping techniques can be used to validate SAR distributions in a phantom [42]. This technique is not ready to be used in vivo. Direct field measurements with electric and magnetic field probes inside a phantom can be used to measure all three components of the E- or H-field [60, 61, 62]. With the help of an automatic positioning system more precise and faster measurements are possible [60, 61]. Scattering
matrix measurements of RF coil arrays using a network analyser is also a valuable validation step.

1.4.5 Human Body Models

The human body is a highly complex inhomogeneous object. The creation of a body model for numerical simulations is a labour intensive and costly process, requiring manual segmentation of image based data, done by medical doctors and radiologists [8]. A first voxelized whole body model with an isotropic resolution of 5 mm was obtained in the 2000s by segmentation of anatomical images [8] of a cadaver from the Visible Male Project [63]. A numerical model of a woman with a fetus was developed in 2006 [64]. In 2010, the virtual family [65], consisting of surface based models of an adult male and female and two children, with 80 tissue types, based on high resolution MR images with a resolution of 0.5 to 2 mm became available. More recently, the virtual family grew to a virtual population of 10 human body models, including an obese male, and three body models of pregnant women were developed. A surface based human body model exists for the finite element method with a geometric accuracy of 2 mm consisting of 24 different tissue types [51].

To ensure a rigorous safety assessment, the RF coil has to be simulated with a wide variety of numerical human body models, reflecting the variety in the patient population. The power limits have to be set such that the RF coil is unconditionally safe for all numerical body models, leading to a conservative power limit. Anomalies in some patients that are not reflected by the numerical models used for the safety assessment might exceed the safety limit.

More recently, a new approach in safety assessment was introduced, in which patient specific human body models are created based on whole body MR image data. The individual human body model is then simulated prior to the actual MR exam [66,67]. This leads to a less conservative power limit, and ensuring unconditional safety. However, this approach is very time consuming.

The material properties of human tissue were measured in [68, 69, 70, 71], the measured data agreed well with literature. The measure-
ments were performed in-vitro at RF frequencies using a open ended coaxial probe connected to a network analyzer. Where possible, in-vivo data was acquired, on the skin or the tongue. Biological tissue within the same individual is inhomogeneous, it can have a variation of the dielectric properties of 5 – 10% at RF frequencies, due to physiological processes.
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Chapter 2

Multiple Port and Field Analysis

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2.1 Introduction

Full wave EM-solvers became indispensable tools for the design and safety assessment of RF-coil arrays at ultra-high field strength (> 4T) [1, 2, 3]. At ultra-high fields, the RF electric and magnetic fields are coupled and their interactions with the imaging object become more complex. The RF-wavelength reaches the range of the imaging object, resonance and standing wave effects occur, leading to an inhomogeneous RF-field [4, 5, 6]. Multiple RF coils (RF coil arrays) are used to counteract these phenomena [7, 8]. They are designed to achieve a desired field distribution inside the imaging object (e.g. high and uniform \( B_1^+ \), low SAR, low noise and a high \( B_1^- \)), and RF network behaviour (matching and low coupling between the channels).

The design process of coil arrays using EM-simulations is very time consuming. For a given scenario including an RF coil geometry with lumped elements at fixed positions and a specific load, many iterations are necessary to find the optimal values for the lumped elements in order to achieve the desired EM field, and network behaviour [9]. Each iteration involves a full wave 3D EM-simulation, and can take several hours of computation time, for complex loads, such as human models. Therefore, a new simulation method allowing for faster calculation of the EM-fields associated with a specific RF coil configuration with variable lumped elements is needed.

We developed a method that allows for variable lumped elements in combination with any full wave 3D EM-field simulation [10], this method was simultaneously published by Kozlov et al. [11], and later by Lemdiasov et al. [12]. The RF-setup is divided into a 3D EM-field domain and a network domain containing the lumped elements and ports, that are replaced by sources in the 3D full wave simulation. The corresponding EM-fields and the relation between these sources in the form of a scattering matrix are needed in the following. The actual lumped element distribution can then be added in a post-processing step to the results obtained by the 3D simulation. The resulting scattering matrix of the coil array with the actual lumped element distribution can be computed using network theory. The corresponding 3D-EM
field solution can be obtained simultaneously by superimposing the EM-fields in a way such that the boundary conditions at all sources are maintained. The solution for a given set of lumped elements is obtained within seconds, as compared to hours for a 3D EM full wave simulation. In this way, only one full wave 3D simulation is required (for time domain methods a simulation refers to N simulation runs, with N being the number of sources in the EM-domain; for frequency domain methods only one simulation run is required for N sources).

In [11] the formulation is given in terms of currents and voltages. In [12] the problem is stated in terms of scattering parameters, but no explicit expression for the EM-fields as a function of the lumped elements is given. In this chapter the relationship of the EM-field domain and the network domain is derived in terms of scattering parameters. Furthermore some useful tools for RF coil array design, such as single channel matching with two port matching networks is reviewed, and a new matching method that matches and decouples all channels of the RF-coil array is introduced. The formulas derived in this chapter can be easily implemented in e.g. Matlab or Python, to avoid the cumbersome scripting languages of commercially available 3D full wave solvers, and to implement efficient optimization procedures.

The proposed tools will be demonstrated in the design process of a transmit stripline RF coil array for head imaging at 7T. First, the ultimate transmit efficiency for the given array topology is shown, in which all lumped elements are replaced by excitation ports, resulting in 56 excitation ports, that are matched and decoupled. Furthermore, the lumped element distribution of the array will be optimized for an operation with eight transmit channels, that are driven in quadrature. Different figures of merit at the network level as well as at the EM-field level in different regions of interest (whole volume, transversal, and the sagittal slice) will be investigated.
2.2 Theory

The traditional way to incorporate lumped elements in a 3D full wave EM-simulation is shown in Fig. 2.1. They are fixed and embedded in the numerical domain, and implemented as a boundary condition for the EM-fields. The E- and the H-field have to have the ratio given by the impedance. The advantage of the traditional simulation method is that no post-processing step is necessary. The EM-fields can be interpreted right away. This is only practical, however, if the lumped elements are known already. Changing their impedance would require an additional full wave simulation. In the design process of an RF coil array the lumped element values are not known in advance, and many iterations would be necessary, due to the complex coupling behaviour within the array, in order to tune and match the coil array or achieve a desired field distribution.

Therefore, a new method allowing for variable lumped elements is necessary. For this purpose they are replaced in the numerical domain by lumped ports, see Fig. 2.2a. The values of the lumped elements are added in a post-processing step, as shown in Fig. 2.2b. The EM-fields and the scattering matrix of the original EM-simulation are transformed in a post-processing step by ensuring the boundary conditions at the lumped ports. The post-processing step is very fast, making this
method suitable for the design of RF coil arrays.

The transformation of the original scattering matrix \([S]_0\) at the reference plane 0, computed in the EM-domain through an RF-network represented by the junction scattering matrix \([S]_j\) can be found by applying the boundary conditions at the reference plane 0 (see Fig. 2.3). The resulting scattering matrix \([S]_1\) at the reference plane 1 can be computed as:

\[
[S]_1 = [S]_{j,11} + [S]_{j,12} [S]_0 ([I] - [S]_{j,22} [S]_0)^{-1} [S]_{j,21} \tag{2.1}
\]

Where \([I]\) is the identity matrix. \([S]_{j,kl}\) are block-matrices of the junction scattering matrix:

\[
[b_1] = [S]_j [a_1] \quad \text{with:} \quad [S]_j = \begin{bmatrix} [S]_{j,11} & [S]_{j,12} \\ [S]_{j,21} & [S]_{j,22} \end{bmatrix} \tag{2.2}
\]

With \(a_k\) and \(b_k\) being the forward and backward traveling wave vectors at the kth reference plane \((k = 0, 1)\), the forward direction is defined towards the junction matrix. The block-matrix \([S]_{j,11}\) has the dimension \(M \times M\) and describes the reflection and coupling between the ports at reference plane 1. The block-matrix \([S]_{j,22}\) has the dimension \(N \times N\) and describes the reflection and coupling between the ports at reference plane 0. The block-matrix \([S]_{j,12}\) has the dimension \(M \times N\) and describes the coupling between ports at the reference plane 1, to ports at the reference plane 0. The block-matrix \([S]_{j,21}\) has the dimension \(N \times M\) and describes the coupling between ports at the reference plane 0, to ports at the reference plane 1.
The EM-fields can be transformed from the reference plane 0 to the reference plane 1 according to:

\[ [E]_1 = [E]_0 \cdot ([I] - [S]_{j,22}[S]_0)^{-1} \cdot [S]_{j,21} \]  

(2.3)

Where \([E]_0 = [E_{0,1} \ldots E_{0,N}]\), and \([E]_1 = [E_{1,1} \ldots E_{1,M}]\) are vectors of electric fields at the reference planes 0 or 1. The \(i\)th component of either vector is the electrical field that is excited selectively by the forward wave with unit amplitude at the \(i\)th port, at the respective reference plane. The magnetic fields are transformed in the same way.

In the following some useful junction scattering matrices for the post-processing steps of single RF coils and arrays will be introduced and derived. Their practical relevance will be further demonstrated in the method section.

### 2.2.1 Single Channel Matching Network

After considering a general junction scattering matrix, we will now consider a concrete junction scattering matrix: a matching network for a single RF-coil. A matching network transforms any RF coil...
impedance to the characteristic impedance of the coaxial cable, to maximize the transmit efficiency (see Fig. 2.4). The matching network may be represented by a single excitation port in the 3D EM-simulations, as long as the matching network is much smaller than the RF-wavelength.

Different matching network topologies can be used (π- or L-networks) with two or more capacitors or inductors. We will not limit the matching network to a predefined topology, but define it only by the elements of its scattering matrix \([S]_M\). The synthesis of the network for an experimental design is an additional task that will not be covered here.

It will be assumed that the matching network consists of reciprocal and lossless elements, i.e. ideal capacitors and inductors. The condition for a reciprocal network is:

\[
[S] = [S]^t
\]  
(2.4)

From this equation we can easily see that \(S_{M,12} = S_{M,21}\). The condition for a lossless network is:

\[
[S]^H \cdot [S] = [I]
\]  
(2.5)

Where \([I]\) is the identity matrix, and \((.)^H\) is the conjugate complex of the transpose. Solving for the input reflection coefficient using Eq. 2.1 we can see that:

\[
\Gamma_{in} = S_{11} + \frac{\Gamma_L S_{12} S_{21}}{1 - \Gamma_L S_{22}}
\]  
(2.6)

Combing Eqs. 2.4, 2.5 and 2.6 with the condition \(\Gamma_{in} = 0\), we can
solve for the scattering parameters of the matching network:

\[
\begin{align*}
S_{M,11} &= -\Gamma_L e^{i2\phi} \\
S_{M,12} &= \sqrt{1 - |\Gamma_L|^2} e^{i\phi} \\
S_{M,21} &= \sqrt{1 - |\Gamma_L|^2} e^{i\phi} \\
S_{M,22} &= \Gamma_L^* 
\end{align*}
\] (2.7)

From Equation 2.7 we can see that the elements of the two port matching matrix are given by the reflection coefficient of the coil \(\Gamma_L\), up to an arbitrary phase \(\phi\). This arbitrary phase corresponds to a transmission line of arbitrary length at port 1, with the reference impedance of port 1.

### 2.2.2 Single Channel Array Matching

In most RF coil arrays each channel is matched individually to the characteristic impedance of the coaxial cables of 50\(\Omega\), allowing for coupling between the channels. This matching strategy is easily manufactured - for each matching network only two lumped elements are needed. This is sufficient for coil arrays with low mutual coupling (< -12dB), the power reflection will be reasonably low for any driving condition. For arrays with a high mutual coupling between the channels a more complicated matching network is needed to avoid a high reflected power. This network will have to connect the channels between each other to cancel coupling. This will be described in the following section.

The unmatched coil array can be described by an \(N\)-port scattering matrix \([S]_C\). The matching networks, that are connected to each channel, can be represented by two port scattering matrices \([S]_{M,i}\), as described in the previous section. The matching networks need to be optimized, such that the input reflection coefficients \(\Gamma_{in,i}\) of the resulting scattering matrix \([S]_{tot}\) vanish (see Fig. 2.5).

The matching networks for each individual channel cannot be computed in a single step like in the single channel case shown in the previous section, but have to be found by an iterative optimization procedure. This is due to the coupling behaviour of the coil array. Matching
one port changes the boundary condition for all other ports, and their load reflection coefficient $\Gamma_{L,i}$, and as a consequence leads to a mismatch.

A matching approach inspired by the matching of a coil array on the bench turned out to be applicable for the post-processing as well. In the experiments, each channel of the coil array is matched individually, from the first through the $N$th channel. This procedure is repeated until the reflection coefficients of every channel is below a sufficiently low value i.e. $-30\text{dB}$.

The matching procedure in the simulation is easier than the experimental matching procedure. As was shown in the previous section, for a match to the characteristic impedance the output reflection coefficient of a lossless and reciprocal matching network has the be to the complex conjugate of the reflection coefficient looking into the load, see $S_{M,22}$ in Eq. 2.7 Therefore, matching port $i$ corresponds to terminating it with the complex conjugate of the reflection coefficient looking into the port $\Gamma_{L,i}$. $\Gamma_{L,i}$ can be computed using Eq. 2.1 The reflection coefficient terminating each port is updated in this way iteratively until the sum of the differences of $\Gamma_{L,i}$ and $\Gamma_{out,i}$ is below a certain value. The corresponding matching networks can then be computed with the found $\Gamma_{out,i}$ using Eq. 2.7
Fig. 2.6: Coil array with a matching matrix and excitation sources for each channel, illustrated with scattering matrices.

2.2.3 Matrix Matching Network

To maximize the power efficiency for a given coil array, it is not enough to match the individual channels. Power loss, due to coupled power into other channels, decrease the forward power, and therefore the power efficiency. A more general network is necessary, a matrix matching network, that is connected to the coil array and prohibits not only reflection into the same channel, but also coupling into all other channels.

The matrix matching network needs to connect all channels between each other in order to be able to achieve the decoupling. In order to maintain degrees the of freedom that a $N$ port coil array offers, the number of channels must not change. Therefore the matrix matching network is a $2N$ port network, represented by the scattering matrix $[S]_M$, whose first $N$ ports are connected to the transmit amplifiers, and the last $N$ ports to the coil array, that is represented by the scattering matrix $[S]_C$ (see Fig. 2.6).

For matching and decoupling, the matrix matching network $[S]_M$ has to be designed, such that the following condition is fulfilled:

$$[S]_{tot} = [0]$$ \hspace{1cm} (2.8)

Furthermore the matrix matching network is assumed to be lossless
and reciprocal. With Eqs. 2.4, 2.5 and 2.1, we can solve for \([S]_M\):

\[
\begin{align*}
[S]_{M,22} & = [S]^H_C \\
[S]^H_{M,21} & = [I] - [S]^H_C [S]_C \\
[S]_{M,12} & = [S]^T_{M,21} \\
[S]_{M,11} & = ([S]^H_{M,12})^{-1} [S]_C [S]_{M,21}
\end{align*}
\] (2.9)

The matrices \([S]_{M,ij}\) are \(N \times N\) block matrices of the matching matrix \([S]_M\), the index 1 of the block matrices refers to the ports 1 to \(N\), the index 2 refers to the ports \(N + 1\) to \(2N\). The solution for the matching matrix is unique up to the choice of the matrix square root algorithm for the computation of \([S]_{M,21}\).

The ports of the decoupled and matched coil array cannot be assigned to the individual elements anymore, but the number of degrees of freedom is preserved. If the matrix matching network is connected with coaxial cables to the coil array, a low intrinsic coupling and a pre matching of the coil array is desirable, in order to avoid high loss on the coaxial cables due to standing waves. Variable lumped elements can be used for the matrix matching network [13], in order to adjust for variations in the coil scattering matrix \([S]_C\), due to different loading conditions. The matrix matching network is difficult to synthesize and construct, design simplifications can be achieved by assuming a circular scattering matrix of the coil array. The matrix matching is a good way to determine the ultimate performance in terms of power efficiency for a given coil array in simulations, as will be shown in the following.

### 2.2.4 Principal Component Analysis

The Principal Component Analysis (PCA) is a valuable tool to assess the maximally achievable transmit field strength of transmit coil arrays, and potentially the number of degrees of freedom for parallel transmission. The PCA will be shortly reviewed in this section.

For a transmit coil array with \(N\) channels, the transmit field strength in the voxels at positions \(r_i\) \((i = 1 \ldots M)\) in a given region of interest (ROI) can be represented by the vector \(B^+_i\), with \(B^+_i = [B^+_1(r_1) \ldots B^+_1(r_M)]^T\).
It can be computed as a function of the $N$ dimensional input voltage vector $V$:

$$B_1^+ = [\hat{B}_{1,1}^+ \ldots \hat{B}_{1,N}^+] \cdot V = [\hat{B}_1^+] \cdot V \quad (2.10)$$

Where the $\hat{B}_{1,i}^+$ are the transmit sensitivities of the individual channels, with $\hat{B}_1^+ = [\hat{B}_{1,i}^+(r_1) \ldots \hat{B}_{1,i}^+(r_M)]^T$. The field energy $|\vec{B}_1^+|^2$ in the ROI can then be computed as:

$$|\vec{B}_1^+|^2 = V^H \frac{1}{M} [\hat{B}_1^+]^H [\hat{B}_1^+] V \quad (2.11)$$

The $N \times N$ transmit field covariance matrix $[\Psi]_{Tx}$ is hermitian. By performing an Eigendecomposition on $[\Psi]_{Tx}$, we can write:

$$[\Psi]_{Tx} = [U][\Lambda][U]^H \quad (2.12)$$

With $[U]$ being a unitary matrix, and having the eigenvectors of $[\Psi]_{Tx}$ in its columns, and $[\Lambda]$ a diagonal matrix with the corresponding sorted eigenvalues on its diagonal, starting with the largest eigenvalue.

The $i$th eigenvalue is the field energy that is obtained when the input voltage vector equals the $i$th eigenvector, as can be seen in Eq. $2.12$. The $i$th eigenvector and the corresponding field distribution is the $i$th principal component. The corresponding eigenvector is also called the variance of the respective principal component. The contribution of each principal component to the overall variance in the data can easily be seen by the variance explained, that can be computed as:

$$variance \ explained = \frac{\Lambda_{i,i}}{\sum_{k=1}^{N} \Lambda_{k,k}} \times 100\% \quad (2.13)$$

With $\Lambda_{i,i}$ being the $i$th eigenvalue. Principal components that explain less than a certain threshold of the variance (e.g. 1%) in the field distribution can be neglected.
2.3 Application Example

For demonstration purposes a stripline head coil array was simulated with eight stripline elements. Each element consists of a strip above a groundplane, that have the same length of 250 mm. The strip is divided into 4 equal parts. Strip capacitors $C_{\text{strip}}$ are inserted in the two outer gaps, a matching circuit followed by an excitation port is inserted in the middle gap. The strip is connected on both ends to the groundplane through the end capacitors $C_{\text{end}}$. At both ends of the groundplanes decoupling capacitors $C_{\text{dec}}$ are placed between the elements. The eight elements are covered by an octogonal copper shield. The coil array is loaded by a cylindrical phantom with a diameter of 20 cm and a length of 30 cm, a relative permittivity of 80 and a conductivity of 0.5 S/m. Each of the lumped elements are replaced by lumped ports in the 3D full wave simulation, resulting in 56 excitation ports. The 3D full wave finite element solver HFSS version 14.0 (Ansys, PA, USA) was used for the simulation, using first order basis functions. The RF bore was included in the simulations as a cylinder. The outer shell of the cylinder was simulated as a perfect electrical conductor boundary condition and the two faces were terminated with a radiation boundary condition. The simulation setup excluding the RF bore is shown in Fig. 2.7.

2.4 Results

2.4.1 Best case performance with 56 ports

To assess the best case performance of the coil array, the ports intended for lumped elements were also used as excitation ports. The resulting 56-port RF coil array was matched and decoupled by a matrix matching network as described previously. A principal component analysis was performed to evaluate the variance and strength of the $B_{1}^{+}$ fields that are excited by the 56 ports in two regions of interest (ROIs): in the whole phantom and in the central transversal plane.

Fig. 2.8a,b shows the variance that each principal component explains in percent and the averaged $B_{1}^{+}$ in the whole phantom. We can
observe a high variance up to the 32\textsuperscript{nd} principal component, due to the variation of the ports along the circumference as well as along \(z\) direction. Fig. 2.9 shows the amplitudes of the \(B_1^+\) fields of the first four principal components in the central transversal, coronal and sagittal planes.

Fig. 2.10\textsuperscript{a,b} shows the variance that each principal component explains in percent and the averaged \(B_1^+\) in the central transversal slice. Here, only the first eight principal components contribute significantly to the overall field distribution. The 9\textsuperscript{th} to the 16\textsuperscript{th} principal components have some minor contribution. The contribution of the other principal components is negligible. This is expected due to the geometry of the coil array; the main contribution stems from the eight single stripline elements, that each have 5 ports. The ports from the same stripline element will excite a similar field distribution in the central transversal slice, this is why we can expect 8 main principal components. Minor contributions to the field variation come from the decoupling capacitors that are offset by 16.25° with respect to the ports on the stripline elements. The amplitudes of the \(B_1^+\) fields of the first 12
Fig. 2.8: Principal component analysis of the $B_1^+$ fields in the whole cylindrical phantom, (a) variance explained, (b) average value in $\mu T/\sqrt{kW}$.

Fig. 2.9: Principal component analysis of the $B_1^+$ fields in the whole cylindrical phantom. $B_1^+$ of the first four principal components in the three central orthogonal slices in $\mu T/\sqrt{kW}$. 
Multiple Port and Field Analysis

Fig. 2.10: Principal component analysis of the $B_1^+$ fields in the central transversal slice of the cylindrical phantom, (a) variance explained, (b) average value in $\mu T/\sqrt{kW}$.

The principal components are shown in Fig. 2.11. Note that some of the amplitude field distributions look similar, but they differ in phase.

2.4.2 Optimization of the 8 channel coil array

In this section we will investigate the optimal lumped element distribution for the eight channel RF coil array in quadrature drive. The position of the excitation is chosen in the center of each of the eight stripline elements. The end capacitor $C_{end}$ and the decoupling capacitor $C_{dec}$ were used to optimize the fields and network behaviour of the coil array.

To reduce the number of variables the two strip capacitors $C_{strip}$ on each of the stripline elements were short-circuited. It was found that the strip capacitor alone did not have a big influence on the field distribution, only the combined capacitance of the strip capacitor $C_{strip}$ and the end capacitor $C_{end}$.

Fig. 2.12 shows the mean $B_1^+$ and normalized standard deviation $s_n$ for quadrature drive and an input power of 1kW plotted versus the decoupling capacitor $C_{dec}$ for varying end capacitors $C_{end}$ and for different ROIs. Each channel of the array was matched with a single channel matching network to $50\Omega$, the characteristic impedance of the coaxial cables. The array shows the highest power efficiency for an end capacitor of $C_{end} = 1\, \mu F$. For this value, the power efficiency as well as the normalized standard deviation stays constant except for a decoupling
Fig. 2.11: Principal component analysis of the $B_1^+$ fields in the transversal slice of the cylindrical phantom. $B_1^+$ of the first twelve principal components in the transversal slice in $\mu T/\sqrt{kW}$. 
capacitor of $C_{dec} \approx 6 \text{pF}$, where the power efficiency as well as the normalized standard deviation show a sharp minimum. The high power sensitivity of the array for the decoupling capacitor of $C_{end} = 1 \text{pF}$ comes at the cost of a high field non-uniformity along the $z$-direction, as shown in Fig. 2.12 of about 62%. A good trade-off is a higher end capacitor of $C_{end} = 5 \text{pF}$ that shows a global maximum of power sensitivity for a decoupling capacitor of $C_{dec} \approx 6.2 \text{pF}$, and a high field uniformity in all ROIs. This end capacitor value was chosen to investigate the network behaviour in the following.

In Fig. 2.13 the power sensitivity and the normalized standard deviation in quadrature drive for an input power of 1 kW are plotted versus the decoupling capacitor in the ROIs described previously for two different matching conditions. In the first case, each channel of the array is matched with a two port matching network to the characteristic impedance of the coaxial cables, allowing the ports to couple. In the second case, the array is matched and decoupled using a matrix matching network. The end capacitor is $C_{end} = 5 \text{pF}$ for both matching conditions. The matrix matched coil array and the single channel matched coil array have the same field uniformity for all ROIs and all decoupling capacitors, indicating that the field distribution hardly changes. However, the power sensitivity is always higher for the matrix matched coil array than for the single channel matched coil array, which results from the reflected power that is injected back into the coil array through the matrix matching network.

The difference of the $B_1^+$ field distributions $\Delta(B_1^+)$ in the whole cylindrical phantom of the coil array for the two different matching strategies versus the decoupling capacitor is shown in Fig. 2.14. The difference was computed according to:

$$\Delta(B_1^+) = 2 \sqrt{\frac{\int_V |B_{1,1}^+ - B_{1,2}^+|^2 dV}{\int_V |B_{1,1}^+ + B_{1,2}^+|^2 dV}}$$  \hspace{1cm} (2.14)$$

$B_{1,1}^+$ and $B_{1,2}^+$ are the transmit fields of the coil array with single channel matching with a forward power of 1 kW, and the matrix matching with an input power of 1 kW, respectively. The difference of the
Fig. 2.12: Mean $B_1^+$ and normalized standard deviation $s_n$ in quadrature drive for an input power of 1 kW vs. the decoupling capacitor $C_{dec}$ for different end capacitors $C_{end}$ in different ROIs. In (a) the entire phantom, (b) central transversal slice, and (c) central sagittal slice.
Fig. 2.13: Mean $B_1^+$ and normalized standard deviation $s_n$ in quadrature drive for an input power of 1 kW vs. the decoupling capacitor $C_{\text{dec}}$ for single channel matching (single) and matrix matching (matrix) in different ROIs. In (a) the entire phantom, (b) central transversal slice, and (c) central sagittal slice.
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\[ \Delta (\mathbf{B}_1^+) \text{[%]} \]

\[ C_{\text{dec}} \]

Fig. 2.14: Difference in transmit field distributions for the single channel matched coil array and the matrix matched coil array versus the decoupling capacitance.

Transmit fields is non-zero, due to the different boundary conditions, that the two matching topologies present to the excitation ports of the coil array. However, the difference is vanishingly small. It reaches a maximum, where the single channel matched array has the highest coupling, which can be expected.

Figs. 2.15 and 2.16 show different properties of the single channel matched coil array and the matrix matched coil array for an end capacitor of \( C_{\text{end}} = 5 \text{pF} \) versus the decoupling capacitor \( C_{\text{dec}} \). The coil array is driven in quadrature, and the input power is set to 1 kW. Subfigure (a) shows the power budget; the line marked with \( \text{Cyl} \) is the power loss in the cylindrical phantom, the line marked with \( \text{Ref} \) is the total reflected power summed over all channels, and the line marked with \( \text{Cond} \) is the conductivity loss on the coil conductors and the RF shield. In subfigure (b) the maximal coupling of the coil array is shown. The real and imaginary part of the impedance looking into the excitation port before the matching network are shown in subfigure (c). Subfigure (d) shows the currents flowing through the different capacitors and the excitation port. \( I_e \) represents the current flowing through the end capacitor \( C_{\text{end}} \), \( I_s \) the current flowing through the port for the strip capacitor, \( I_p \) the current through the excitation port, and \( I_d \) is the current through the decoupling capacitor.

Let us first consider the single channel matched case in Fig. 2.15. For the value of the decoupling capacitor of \( C_{\text{dec}} = 6.2 \text{pF} \) we can ob-
serve a maximum of the power loss in the cylindrical phantom. The electrical field is focused in the phantom, therefore the associated magnetic field must also be focused in the phantom, explaining the maximum in transmit sensitivity in Fig. 2.13. The maximum coupling of the eight port scattering matrix has a local minimum of about $-10\, \text{dB}$. The input impedance is inductive, and can be matched with a capacitive $L$-type matching network, consisting of two trimmer capacitors. For any given decoupling capacitor the RF-current through the lumped elements on the strip is almost constant. The current through the decoupling capacitors is negligible with respect to the current through the stripline elements, except for the value of $C_{\text{dec}} = 6.2\, \text{pF}$. Here, $I_d$ has a maximum and exceeds the currents on the stripline, indicating an end-ring resonance.

The transmit properties for the matrix matched coil array are shown in Fig. 2.16. The reflected power and the maximal coupling are close to zero for all decoupling capacitors, due to the matrix matching network. Note that for the value of the decoupling capacitor of about $C_{\text{dec}} = 5.3\, \text{pF}$ the conduction loss exceeds the power loss in the phantom. This is the same as for the single channel matched array, but can not be seen easily in Fig. 2.15 due to the high reflected power for this decoupling capacitor value. The coil impedance is measured before the matching network, looking from the load, and therefore is the same as for the single channel matched array.

### 2.5 Discussion & Conclusion

A fast simulation method for the optimization of the lumped element distribution at predefined positions on a given RF coil array structure is presented in this chapter. This speeds up the RF coil design process significantly allowing extensive optimization procedures based on EM-fields ($\text{SAR, } B_1^+, B_1^-$) and network parameters (maximal coupling, input impedance) being performed in a short time. Only one 3D full wave EM-simulation has to be performed, in which the lumped elements, that have to be optimized, are replaced by excitation ports. The EM-fields as well as the network behaviour as a function of the lumped
Fig. 2.15: Transmit properties for the 8 channel RF coil array in quadrature drive and an input power of 1 kW with single channel matching, with an end capacitor of $C_{\text{end}} = 5 \, \text{pF}$ versus the decoupling capacitor $C_{\text{dec}}$, (a) power budget, (b) maximal coupling, (c) coil impedance at the excitation ports, (d) RF-currents in the ports of the coil array.

Fig. 2.16: Transmit properties for the 8 channel RF coil array in quadrature drive and an input power of 1 kW with matrix matching, with an end capacitor of $C_{\text{end}} = 5 \, \text{pF}$ versus the decoupling capacitor $C_{\text{dec}}$, (a) power budget, (b) maximal coupling, (c) coil impedance at the excitation ports, (d) RF-currents in the ports of the coil array.
element distribution on the coil array can be computed in a fast post-processing step, satisfying the boundary conditions. In this way, the iterative search of lumped elements, such as matching and decoupling, is moved from time consuming 3D simulations to fast post-processing steps. Any full wave simulation can be used for this method, but the finite element method (FEM) is favourable. The simulation time in FEM does not depend on the number of excitation ports, and it offers reasonably fine discretization of the numerical domain for human body models on modern PCs. This method is integrated in some commercial EM-simulation softwares nowadays (HFSS, CST, SEMCAD), however the implementations do not offer the freedom of an own implementation in a scripting language of choice. All results shown in this chapter are based on a single 3D full wave simulation, that lasted a day of computation time on a computer equipped with 256GB RAM, and approximately one hour of post-processing time (for 750 different lumped element distributions). To obtain the same results in the traditional way, carrying out 750 full wave simulations would last approximately 2 years.

Furthermore, two matching topologies are presented. The traditional matching strategy, where each port is matched individually with a two port matching network, allowing for mutual coupling between channels, and a new matching strategy was introduced, called matrix matching that matches and decouples the coil array with a more complex matching network, having double the number of ports as the coil array itself. This complexity however pays off with a higher transmit sensitivity because no power is reflected back to the amplifiers.

The theoretical framework developed was demonstrated on a simulated 7T eight channel RF head coil array. The 48 capacitors and eight matching networks on the RF coil array are replaced by lumped ports in the 3D full wave simulation, totalling 56 lumped ports. First, the ultimate transmit sensitivity of this array given the topology and locations of the lumped elements was computed by treating all of the 56 ports as excitation ports. The ports of this 56 channel array were matched and decoupled, by the matrix matching network. A principal component analysis on the transmit fields yielded a maximal power sensitivity
of $12 \mu T/\sqrt{\text{kW}}$ averaged over the whole phantom, and $23 \mu T/\sqrt{\text{kW}}$ averaged over the central transversal slice. 32 significant principal components were found for the whole cylindrical phantom, and eight for the transversal slice. Furthermore, the eight channel head coil array was optimized for quadrature drive. An optimal set of capacitors was found, for the single channel matched array and compared to the same array that is matched and decoupled with a matrix matching network. This optimized 8 channel array yielded a power sensitivity of $8.5/9 \mu T/\sqrt{\text{kW}}$ in the whole cylindrical phantom and $12/13 \mu T/\sqrt{\text{kW}}$ for the central transversal slice, for the single channel matched and the matrix matched array, respectively. The matrix matched array had a higher transmit sensitivity because it has no power loss due to reflection back into the coaxial cables. Renormalizing the transmit fields of the two matching strategies to the same forward power yielded almost the same transmit field distributions. This does not have to be the case for arrays with higher intrinsic couplings


[8] G Adriany, PF Van de Moortele, Fl Wiesinger, S Moeller, JP Strupp, P Andersen, C Snyder, Xi Zhang, W Chen, KP Pruessmann, P Boesiger, T Vaughan, and


Chapter 3

Symmetric Feeding Strategy for RF Coils

To be submitted:

3.1 Introduction

Increasing the static magnetic field strength in MRI promises a higher SNR, but on the other hand also leads to a decreased wavelength of the radio frequency (RF) electromagnetic (EM) fields in human tissue and a local SAR hotspots in the patient [1]. At high field strengths (> 3 T) the wavelength enters the range of the average dimensions of the human body [2]. At ultra high field strength (> 7 T) the wavelength in human tissue (∼ 14 cm) reaches the dimensions of the human head leading to a non-uniform transmit field distribution also in head imaging [3]. Additional degrees of freedom are necessary to counteract the field non-uniformity. RF transmit coil arrays with independent transmit amplifiers for static [4] or dynamic shimming [5] can be used to achieve a higher uniformity of the transmit field. Further design issues regarding RF coil arrays for ultra high fields such as coupling and radiation get even more demanding than for lower field strength systems, therefore also entering a different regime regarding coil array design. Transmit coil arrays can be divided into different topologies, e.g. loop arrays [4], stripline arrays [4, 6], radiative antenna arrays [7], or traveling wave transmit systems supporting multiple channels [8].

The stripline array is a widespread design for high field transmit arrays [4]. It is simple to manufacture and enables good decoupling between the elements [6]. The most widely used striplines are standing wave type elements of a length of about a quarter wavelength. Conventionally the stripline elements are fed at one end of the element through a capacitive matching network. However, stripline arrays suffer from a limited coverage along the axial direction, and the performance highly depends on different design issues leading to significant common mode currents on the feeding coaxial cable.

The limited coverage along the axial direction of the standing wave stripline is given by the small effective wavelength - between that of air (1 m) and human tissue (14 cm) - at 7T illustrated by MR-images in [4]. The coverage could be enhanced with a traveling wave stripline [6, 9], in which power is injected at one end of the stripline and terminated by the impedance of the transmission line at the other. This design
would avoid standing waves on the element on one hand, but is less power efficient due to the loss in the termination impedance on the other. A similar more general concept is travelling wave transmission within the bore [10], which achieves a large coverage along the axial direction, but is also less power efficient due to radiation loss. A stripline array offering both, a good coverage and a high SAR-efficiency ($B_1^+ / \sqrt{\text{SAR}}$), remains an engineering challenge.

The common mode currents of the stripline element are mainly induced by the asymmetric feeding, which is very simple regarding manufacturing, but has the disadvantage of connecting the cable shield to the element’s groundplane that is on a non-zero RF-potential at the connecting position. Furthermore, the cable is routed through a region with strong and spatially variable EM-fields produced by the differential mode of the coil itself. Therefore, the strength of the cable currents is highly dependent on the specific configuration of the cable routing. The common mode currents produce parasitic magnetic fields that disturb the wanted magnetic field distribution excited by the RF-coil, potentially causing image artefacts [11]. In addition, the cables carrying high currents on their shields pose a safety risk when routed close to the patient. RF-coils are designed for differential mode operation, and are inefficient when excited in common mode. Power leakage of the differential mode reduces the power efficiency in transmit mode, and causes a decrease of the SNR in receive mode. The power leakage of the differential mode is caused by the coupling of the differential mode to the common mode and the associated potentially increased mismatch leading to an increased reflection of the differential mode. In the array case, not only the unavoidable coupling of the differential modes, also the common modes on the cables couple to each other leading to an unstable and demanding matching of the array. In EM-field simulations the RF-coils are excited by differential mode sources, the common mode currents are not taken into consideration. In these cases safety assessments and design studies based on EM-field simulations assume a reliable suppression of the common mode [12]. If not, the contribution of the common mode has to be additionally evaluated for each possible cable routing.
Cable traps are commonly used to suppress cable currents, several of them need to be placed along the cable, especially close to the coil element. The cable traps close to the array elements however might disturb the EM-field that the coil produces when excited by differential mode currents. Cable traps might also occupy a lot of space especially for arrays featuring a large number of channels. The cable traps might also couple to each other via their stray fields, an external EM-field might couple via the cable traps to the common mode of the cable. Cable currents might also capacitively couple to neighboring cables bypassing cable traps meant to suppress them. The common mode problem of the stripline array is not solved by cable traps and remains an unsolved issue.

The shortcomings of the stripline array were addressed in this work with a modified feeding strategy, which mainly consists of feeding the array elements in the symmetry plane using phased currents. In this way, the coverage is effectively doubled by the formation of two current lobes. Cable currents are suppressed because the coaxial cable’s outer conductor is connected at a location having zero potential. In addition, the coaxial cables are routed through an outer shield, which is close to the coil element. This avoids the coupling of the common mode currents to the EM-field generated by the coil. To facilitate the central feed a balanced to unbalanced circuit (balun) is needed between the coaxial cable and the coil element. The chosen balun design consists of a $\lambda/2$ line to create the out of phase current. This type of stripline element was first introduced by Brunner et al. [13] for a head coil array, and applied in [14,15,16] for a torso array for body imaging.

An 8-channel head coil array with improved stripline elements was constructed. A good common mode rejection was shown experimentally. Single element measurements were performed to show the improved common mode rejection with respect to the traditional design. Cooling elements were used to increase the power capability of the trimmer capacitors. The decoupling capacitors of the array were numerically and experimentally optimized for high $B_{1+}$ sensitivity and low coupling. A careful safety assessment by full 3D EM-simulations was performed. The simulations were validated by direct measure-
Methods

3.2 Methods

3.2.1 Construction

Two stripline coils were constructed, that have the same geometry, but differ in their feeding strategy. The first coil is fed in the traditional way, from the side, with a capacitive matching network. The second coil is fed in the middle of the strip through a balun and a capacitive matching network. Both elements are constructed with FR-4 PCB material with a 35 μm copper foil on one side. The elements consist of a groundplane-PCB that is 250 mm long and 120 mm wide, that is fully covered with copper, and a strip-PCB that is 250 mm long and 23 mm wide. 5 mm of the copper foil were removed on the long sides of the strip-PCB, resulting in a conductor width of 13 mm. The strip was screwed with spacers on the groundplane, such that the conductors of the strip and the groundplane are 20 mm apart. The fixed capacitors used on both coils are high power non magnetic capacitors (CPX-series, Temex Ceramics, Pessac, France); they were screwed on the coil elements for higher flexibility in the design process, except for the ones used in the matching network. High voltage non-magnetic trimmer capacitors from different manufacturers were used for the matching networks on both stripline coils.

The side fed coil is shown in Fig. 3.1. One end of the strip was connected through a 2.5 pF fixed capacitor to the groundplane. On the other end a parallel trimer capacitor connects the strip and the groundplane for tuning. The groundplane and the strip were lengthened on this end by 17 mm for a capacitive matching network. A connector was placed on the extension of the groundplane. The inner conductor of the connector was connected through a second trimer capacitor in series with a 1.8 pF fixed capacitor to the strip for matching,
Fig. 3.1: Side-fed stripline element. (a) Picture (b) Schematic.

The outer conductor was soldered to the groundplane. Air dielectric trimmer capacitors were used (NMAP15HV, Voltronics Corporation, Salisbury, MD, USA).

The mid-fed coil is shown in Fig. 3.2a. The strip and the groundplane of the mid-fed coil were connected on both ends with 11 pF fixed capacitors. The strip of the mid-fed coil was cut at three positions, for capacitors and matching, dividing it into four equal parts. 8.9 pF fixed capacitors were screwed in the outer slots on the strip. The middle slot on the strip was connected to a two wire line, which routes the balanced voltage behind the groundplane to a matching board, that was soldered on the back of the groundplane, see Fig. 3.2c. The wires of the two-wire line have a diameter of 2 mm and are 9.5 mm apart. On the matching board, the two wires of the two-wire line were connected through a series circuit of a fixed and a trimmer capacitor for tuning. The additional fixed capacitor of 11 pF is necessary to reduce the voltage over the trimmer to avoid voltage breakthrough. The balanced voltage of the two wires is now transformed through a $\lambda/2$ line to an unbalanced voltage, see Fig. 3.2d. A semi-rigid cable (EZ_250_TP_M17, Hu-
ber+Suhner, Herisau, Switzerland) with a low attenuation of 0.11 dB/m at 300 MHz was used for the $\lambda/2$ line. This type of balun provides an impedance transformation-ratio of 1:4, [17]. One wire was connected through a second trimmer capacitor for matching, to the inner conductor of a coaxial cable. The outer conductor of the coaxial cable was soldered to the symmetry plane of the groundplane that has zero potential. Sapphire dielectric trimmer capacitors were used for the mid-fed coil (TG 094, Temex Ceramics, Pessac, France). Cooling elements made of aluminium nitride (Ceramdis GmbH, Elsau, Switzerland) were used to decrease the heat on the most power sensitive parts, the trimmer capacitors, and thereby increasing the maximal average input power. Aluminium nitride is an electrical isolator, but has a high thermal conductivity. Therefore one $50 \times 50 \times 1$ mm aluminium nitride plate could be used to cool both trimmers. The heat transfer between the cylindrical trimmers and the plate was increased by heat conductors made of copper that had a cylindrical cut-out on one side and were flat on the other side, increasing the area of the heat exchange. The heat conductors were glued with heat conducting glue on both ends of the two trimmers. The cooling plate was then pressed by screws on the four heat conductors, see Fig. 3.2b. The heat exchange between the plate and the heat conductors was increased by heat conducting paste. The groundplane was slotted in stripes that are 20 mm wide to reduce eddy currents at gradient frequencies, the stripes are shorted at the larmor frequency with 1 nF SMD capacitors.

An 8-channel head coil array was constructed in a modular way, consisting of 8 mid-fed stripline elements, described above, and a shield, see Fig. 3.3. The shield consists of eight shield PCB boards, that are 350 mm long and 153 mm wide, with a thickness of 3.2 mm, and have a 35 $\mu$m thick copper foil on one side. The shield boards were screwed, with the copper conductors showing inside, on two octagonal PMMA holders, forming an octagonal prism. The side length of the octagon is 153 mm to fit the shield boards such that two neighboring shield boards touch each other. The holders have a circular opening in which a cylindrical tube with 280/290 mm inner and outer diameter, and a length of 350 mm fits tightly in. The tube and the shield serve as a housing for
Fig. 3.2: Mid-fed stripline element. (a) side view (b) the back of the ground-plane including cooling element, (c) matching network (cooling element removed), (d) schematic.
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Fig. 3.3: 8-channel head coil array (inner cover removed). The decoupling capacitors for the bottom element are marked by white circles.

the array. The PCB boards of the shield were slotted in 20mm stripes, and shorted with 1nF SMD capacitors to reduce eddy currents at gradient frequencies. Neighboring shield boards were shorted at larmor frequency with 1nF ceramic capacitors for the same reason. A mid fed element was positioned centrally on each of the shield boards and screwed with spacers, such that the distance between the conductors of the shield and that of the groundplane of the element is 15mm. Neighboring stripline elements were connected to each other through decoupling capacitors $C_{dec}$ at the end of the groundplanes, marked by circles in Fig. 3.3. The coaxial cables of the stripline elements were routed through a hole in the shield boards. Each coaxial cable was screwed on its shield board at three positions for mechanical stability of the cable element connection.

The phantom used in the measurements is a cylindrical PMMA container with an inner and outer diameter of 190/200mm and a length of 300mm. It was filled with water, salt and copper sulfate. The elec-
trical properties of the liquid were measured with a dielectric probe kit (8570 E, Agilent Technologies Inc., Santa Clara, CA, USA). The relative permittivity is $\varepsilon_r = 81$ and the conductivity is $\sigma = 0.5 \text{ S/m}$ at $300 \text{ MHz}$. The $T_1$ of the liquid was measured with a series of inversion recovery sequences to be $230 \text{ ms}$.

### 3.2.2 Simulation and safety assessment

Three different setups of the coil array were simulated using the 3D finite element solver HFSS (Ansys, PA, USA) on a computer with 256 GByte RAM and 24 cores. In the first setup the coil array was in free space and loaded by the cylindrical phantom described above. In the second setup the array was positioned inside the bore of the MR scanner, and loaded by the cylindrical phantom. In the third setup the array was inside the bore of the MR-scanner, loaded by the head of a human. The conductors of the coil array were modelled as sheets with a finite boundary condition with the conductivity of copper. The ground-plane and the shield were modelled as a continuous surface, neglecting the splitting for eddy current reduction. The capacitors on the strips, the end capacitors, and the decoupling capacitors were modelled as lumped $50 \Omega$ ports. The matching network and the two wire line were modelled as a lumped $50 \Omega$ port in the middle of the strip. This resulted in 56 sources for the complete coil array. The free space condition in the first setup was achieved by a spherical radiation boundary condition. The sphere had a minimal distance of a quarter of a wavelength to the coil array. The phantom was modelled as a homogeneous cylinder with the measured electrical properties mentioned above. The bore of the scanner for the second and third simulation was modelled as a cylindrical finite conductivity boundary condition with the conductivity of copper with a diameter of 580mm and a length of 1.3 m. A radiation boundary condition was set on both ends of the cylinder, for the simulation involving the cylindrical phantom. In the setup involving the human body model [18], that was scaled by a factor of 0.9 in all three dimensions, the head end of the cylinder was terminated by a radiation boundary condition. The computational domain on the feet side had to be extended by an air cylinder to include the human body.
The radius of the air cylinder was a quarter of a wavelength larger than that of the bore; the length was such that the distance to the feet of the human body model was also a quarter of a wavelength. The surface of the air cylinder was terminated by a radiation boundary condition. The excitation frequency was set to 300MHz, and first order basis functions were used. The abortion criteria of the adaptive mesh-refinement were set to 0.001% difference in amplitude and 1° difference in phase of the 56 port scattering matrix, and 1% difference of the root mean square of the E- and H-field at 12 selected points within the coil array, of two consecutive refinement passes. The simulations converged to a stable solution with respect to the abortion criteria with 3.5 to 4 million tetrahedrons in the different simulation setups. The capacitors on the strip, the decoupling capacitors, and matching networks were added in a post-processing step [19,20,21]. The matching network was assumed to be lossless. The capacitors were modelled with a quality factor of $Q = 100$.

The power balance $PB$ of the simulation including the coil array loaded by the human was computed for quadrature excitation. The forward power into the computational domain has to equal the power lost to heat in the volume, which consists of the power loss inside the human $P_{L,\text{human}}$ and the power loss in the conductors of the coil array and the RF-bore $P_{L,\text{cond}}$, and the power leaving the computational domain through radiation $P_{\text{rad}}$, see [22]. The forward power into the computational domain is the difference of the input power $P_{\text{in}}$ and the sum of the reflected power $P_{\text{ref}}$ and the power loss inside the capacitors, that were added in a post-processing step $P_{L,\text{caps}}$.

$$PB = \frac{P_{\text{in}} - (P_{\text{ref}} + P_{L,\text{caps}} + P_{L,\text{cond}} + P_{L,\text{human}} + P_{\text{rad}})}{P_{\text{in}}}$$

For the safety assessment, the worst case specific absorption rate (SAR) for an input power of 1W was computed (as introduced by [23], and later adapted by [24]) in three different parts of the human body: head, trunk, and extremities. The worst case SAR is computed as the maximum of the highest eigenvalues of the SAR-matrices of all 10g voxels in each body part. For each body part the SAR-matrices were
computed on a equidistant grid with the spacing $d$. The spacing $d$ is one fifth of a side length of a cube of water with the mass of 10 g. The maximum allowed input power into the array was computed for the three body parts, with the according SAR-limits [25] that differ for the different body parts. The minimum of the input powers was then chosen as the limiting input power for the MR in-vivo experiments.

### 3.2.3 Bench measurements

The common mode behavior of the side-fed and mid-fed single coil elements were assessed with bench measurements. In a first experiment, the cable currents were quantified with $S_{21}$ measurements. Both elements were matched to $50\,\Omega$ at 298 MHz, and connected through a 2 m long cable to the first port of a two port Network Analyzer (NWA) (E5061C ENA, Agilent Technologies Inc.). The second port was connected to a current probe that can sense the cable current along the length of the cable. In a second experiment the influence of the differential to common mode coupling on the matching condition of the differential mode of the two single elements was evaluated with a reflection measurement. The single elements were connected through a 2 m long cable, with a floating cable trap [11], to the NWA. The elements were matched below $-20\,\text{dB}$ for one position of the cable trap. In this experiment, the reflection coefficient $S_{11}$ of the differential mode was measured as a function of the position of the floating cable trap. To evaluate the stability of the coil array in terms of differential to common mode coupling, the S-matrices of the coil array for two different setups were measured with a 16 port NWA (N5230C PNA-L, Agilent Technologies Inc.). In both cases the array was loaded by the cylindrical phantom mentioned above. Each channel of the coil array was matched for the first setup; here the coaxial cables were connected directly to the NWA. For the second setup the outer conductors of the coaxial cables were shorted and then connected to the NWA.

To ensure a linear operation, a high voltage test was performed on each channel of the coil array. A 1 kW pulse with a length of 10 $\mu$s, a duty cycle of 1%, and a carrier frequency of 298 MHz was sent into a single channel of the coil array, while the other channels were termi-
nated with 50Ω. This was repeated for each channel. The array was
unloaded and matched, here we expected the highest voltage drop over
the capacitors. The magnetic field produced by the array was measured
with a pickup coil, and demodulated to the baseband.

To assess the maximum input power of the mid-fed stripline ele-
ment for non-in-vivo use, and the effect of the cooling elements, we
measured the temperature on the most heat sensitive parts of the mid-
fed element - the trimmer capacitors - with and without the cooling el-
ements. The mid-fed stripline element was excited with a block-pulse
signal with a 298MHz carrier frequency, with varying duty cycle, and
thereby varying input power. The temperature on the trimmers was
measured with a fiber optic temperature measurement sensors (Lux-
tron 790, LumaSense Technologies Inc., Santa Clara, CA, USA), that
were attached with heat conducting paste to the trimmers. The average
power was increased after the temperature on the trimmer capacitors
saturated. The stripline element was loaded with the cylindrical phan-
ton.

3.2.4 Measurements - DASY

The E- and H-fields of the coil array loaded by the cylindrical phantom
were measured with an automated measurement robot with miniatur-
ized field probes (DASY52, Speag, Zurich, Switzerland). The array
was loaded by the cylindrical phantom described above and matched
at all channels. One channel of the array was excited with a continuous
wave signal at 300MHz, all other channels were terminated with 50Ω
resistors. The input power was set to 1 W, and was monitored through-
out the measurement. The root mean square of the E- and H-fields was
measured in the three central orthogonal planes on a 5 mm grid inside
the cylindrical phantom. The sagittal plane is parallel and the coronal
plane is perpendicular to the excited coil element.

3.2.5 Imaging Experiments

The MR images were acquired with a 7T Philips Achieva scanner with
a MultiX system with 8 independent transmit/receive channels (Philips
Symmetric Feeding Strategy for RF Coils

Healthcare, Cleveland, OH, USA). Each channel of the coil array was connected through a home-built transmit receive switch to the transmit and receive chains of the MR scanner. The coil array was used in transceive mode for all imaging experiments. The amplitude and phase of each of the eight transmit amplifiers (7T1000M-8C, CPC, Hauppauge, NY, USA) can be controlled separately. The transmit amplifiers have a maximal pulse power of 1 kW and a maximal duty cycle of 10%, resulting in a maximal output power of 100 W. For the in-vivo experiments the repetition times ($T_R$) were used to reduce the input power to stay within SAR limits.

For the $B_1^+$ mapping a multiple flip-angle method was used [26]. First, the single channel $B_1^+$ maps in the central transversal slice of the cylindrical phantom were acquired with the interferometric method [26]. In a second experiment, the coil array was driven in quadrature and $B_1^+$ maps were acquired in all three central orthogonal slices (transversal, sagittal, and coronal). The duration of the non-selective block prepulse was 3 ms, whose flip angle was increased from zero to the maximal flip angle of 600°, in $\Delta \alpha = 42.75^\circ$ steps. The slice excitation flip angle was 16°. A 2D gradient echo sequence was used for image acquisition. The field of view (FOV) was $230 \times 230$ mm for the transversal slices and $320 \times 320$ mm for the sagittal and coronal slices, with a resolution of $64 \times 64$, and $96 \times 96$, respectively. The repetition time was set to $T_R = 300$ ms. In vivo $B_1^+$ maps were acquired in a third experiment in three orthogonal slices with quadrature excitation, such that the individual $B_1^+$-fields add constructively in the center of the transversal slice. All parameters of the $B_1^+$ sequence were the same as described above for the cylindrical phantom, except the FOV was $230 \times 230$ mm and the image resolution was $64 \times 64$ for all slices. The $T_R$ had to be adjusted according to the SAR limits. To normalize the $B_1^+$ maps to the input power, the power going into each of the eight channels of the coil array was measured with a directional coupler and a spectrum analyzer (FSL, Rohde & Schwarz, Munich, Germany).

To demonstrate the imaging performance of the coil array in-vivo, three different imaging sequences were acquired in quadrature excitation - a refocused gradient echo, an proton density weighted se-
sequence, and a $T_1$ weighted sequence. The FOV in all images was set to $230 \times 230$ mm. The refocused gradient echo sequence had a resolution of $0.6 \times 0.6$ mm in plane and a slice thickness of 4 mm, a $TR$ of 200 ms, and an echo time $TE$ of 3.8 ms. The flip angle $\alpha$ was 10°, the number of averages was 2, and the total scan time was 2 : 40 min. The proton density weighted imaging sequence was realized as a ultrafast gradient echo sequence without an inversion pulse. The sequence had a resolution of $0.6 \times 0.6$ mm and a slice thickness of 3 mm, a $TR/TE$ of 9.5/4.6 ms, a flip angle of 7°, and number of averages of 20. The total scan time was 1 : 15 min. The $T_1$ weighted sequence was realized as a ultrafast gradient echo sequence with a resolution of $0.6 \times 0.6$ mm, a slice thickness of 3 mm, the adiabatic inversion pulse had a delay of 1200 ms, and was repeated every 4000 ms. The $TR/TE$ was 8/3.9 ms, the flip angle was 7°, the number of averages was 30, and the total scan time was 3 : 58 min.

### 3.3 Results

#### 3.3.1 Single Elements

The common mode currents along the cable for the side-fed and mid-fed elements were measured. The maximal differential to common mode coupling was $-15$ dB for the side-fed stripline element, and $-28$ dB for the mid-fed element. Both maxima occurred close to the coil element. The highest reflection of the differential mode as a function of position of the floating cable trap was measured to be $-9$ dB for the side-fed element and $-23$ dB for the mid-fed element.

The saturated temperature on the parallel trimmer capacitor on a single mid-fed stripline coil as a function of input power is shown in Fig. 3.4. The series trimmer capacitor showed a lower increase in temperature. The temperature rise is almost linear. The temperature increase in saturation per input power is $1.2 \text{ K/W}$ without and $0.6 \text{ K/W}$ with the cooling element. The maximum temperature increase of the cooled trimmer capacitors for an input power of 100 W is 60 K, which is below the maximum temperature of the trimmer capacitor of $125^\circ \text{C}$.
3.3.2 Array

To find the optimal value for the decoupling capacitor $C_{\text{dec}}$, the coil array was evaluated as a function of the decoupling capacitor in terms of transmit performance. The chosen performance criteria are the transmit field strength, the field uniformity, and the maximal coupling of the array. The simulation setup is the array inside the scanner’s bore loaded by the cylindrical phantom. The decoupling capacitor $C_{\text{dec}}$ was varied from 0.1 pF to 10 pF. Each channel of the array was matched for each value of the decoupling capacitor. Fig. 3.5a shows the mean $B_1^+$ in quadrature drive in the central transversal slice for a total input power of 1 kW. Fig. 3.5b shows the normalized standard deviation $s_n$ in percent of the $B_1^+$ field in quadrature drive in the central transversal plane. The maximal coupling in $dB$ is shown in Fig. 3.5c. The coil array has the best transmit performance for a decoupling capacitor of $C_{\text{dec}} = 6.2$ pF. For this value the $B_1^+$ reaches a maximum, at the same time the $B_1^+$ non-uniformity reaches a minimum and the maximal coupling of the coil array has a local minimum that is reasonably low. This value was used for the construction of the coil array.

The matching stability at the larmor frequency of 298 MHz with respect to common mode currents is shown in Fig. 3.6. $S_{11}$ and $S_{12}$ are plotted versus frequency in the upper and lower plot, respectively.
Fig. 3.5: Simulated array performance as a function of the decoupling capacitor $C_{\text{dec}}$ loaded by the cylindrical phantom. (a,b) mean $B_{1}^{+}$ and normalized standard deviation in the central transversal slice in quadrature drive, and (c) maximal coupling between channels. The dashed line shows the chosen decoupling capacitor of $6.2 \, \text{pF}$. 
The coil array passed the high-voltage test. A stable and rectangular pulse-shape of the magnetic field for each channel indicated a regular operation of the array. The measured S-matrix of the coil-array loaded by a human head is shown in Table 3.1. The highest coupling occurs between neighboring coil elements.

### 3.3.3 Safety Assessment

The root mean square of the H- and E-fields produced by the coil array inside the cylindrical phantom when one coil is excited with 1W at 300MHz is shown in Fig. 3.7a and Fig. 3.7b respectively. The upper rows show the fields measured with the DASY measurement system, the lower rows show the simulated fields of the coil array loaded by the cylindrical phantom in free space.

The $B_1^+$ maps of the coil array loaded by the cylindrical phantom with $1kW$ input power are shown in Fig. 3.8. The left columns show the measured maps, the right columns show the simulated maps of the simulation including the cylindrical phantom inside the bore of the
Table 3.1: Measured S-matrix in dB, coil array loaded by a human head.

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Fig. 3.7: Comparison of the measured (left column) and simulated (right column) E- and H-fields of the eight channel head coil array, in the transversal, coronal, and sagittal plane (from top to bottom) inside the cylindrical phantom, single channel excitation. Measurements were performed with E- and H-field probes on a 5 × 5 mm grid. Input power is 1 W. Root mean square of the (a) H-field in [A/m], and (b) E-field in [V/m].
Fig. 3.8: Measured and simulated $B_1^+$ in $\mu T/\sqrt{\text{kW}}$, of the eight channel head coil array loaded with the cylindrical phantom. (a) transversal plane single channel excitation, (b) all three orthogonal slices, quadrature excitation.

<table>
<thead>
<tr>
<th>$P_{\text{in}}$</th>
<th>$P_{\text{ref}}$</th>
<th>$P_{\text{L,caps}}$</th>
<th>$P_{\text{L,human}}$</th>
<th>$P_{\text{L,cond}}$</th>
<th>$P_{\text{rad}}$</th>
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<td>0.47</td>
<td>0.06</td>
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<td>0.003%</td>
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Table 3.2: Power budget in Watts and power balance $PB$ in per-cent for quadrature drive.

The power balance in the simulation involving the human body model is 0.003% for quadrature drive. The different loss mechanisms are listed in Table 3.2. The limiting worst case SAR is 1.1 W/kg for 1 W of input power, and is located in the eyes. The according SAR-limit in the first level controlled mode is 20 W/kg averaged over 6 minutes in the head [25], and therefore the maximum total input power into the array averaged over 6 minutes is 18.2 W. The maximal input power was not exceeded for the following in-vivo MR experiments.

Fig. 3.9a shows the $B_1^+$-maps for quadrature drive for three orthogonal slices for an input power of 1 kW. The left column shows the measured and the right column shows the simulated $B_1^+$-maps. To
Fig. 3.9: In-vivo simulation and measurements for quadrature excitation in three orthogonal planes. (a) Left row shows the measured, right row shows the simulated $B_1^+$ maps in $\mu T/\sqrt{kW}$, (b) SAR-map in W/kg for an input power of 1 W.

stay within safety limits the $TR$ of the mapping sequence was set to 500 ms. The simulated SAR-maps for quadrature drive for the three central orthogonal slices for an input power of 1 W are shown in Fig. 3.9b.

3.3.4 MR Imaging experiments

The in vivo images without intensity correction are shown in Fig. 3.10. Fig. 3.10a shows the refocused gradient echo images in all three orthogonal slices. The transversal slice acquired with the proton density weighted and $T_1$ weighted sequence are shown in Figs. 3.10b and c respectively.
Fig. 3.10: In vivo images, FOVs 230 × 230 mm: (a) Refocused gradient echo, resolution (R)=0.6 × 0.6 mm, slice thickness (ST)=4 mm, TR/TE = 200/3.8 ms, flip angle (α)=10°, number of averages (NSA)=2, scan time (T)=2 : 40 min, in all three slices. (b) Ultrafast gradient echo without inversion pulse, R = 0.6 × 0.6 mm, ST = 3 mm, TR/TE = 9.5/4.6 ms, α = 7°, NSA = 20, and T = 1 : 15 min. (c) Ultrafast gradient echo, the adiabatic inversion pulse had a delay of 1200 ms, and was repeated every 4000 ms, R = 0.6 × 0.6 mm, ST = 3 mm, TR/TE = 8/3.9 ms, α = 7°, NSA = 30, and T = 3 : 58 min.
3.4 Discussion & Conclusion

An 8 channel stripline head coil array with several design-improvements is proposed in this work. The most important of which is the reduction of cable currents, and a better coverage along the $z$-direction. Further improvements include an increased power efficiency, and an increased allowed input power.

The common mode currents of the stripline array were reduced by a new feeding strategy of the single elements and a smart cable routing. This improves the patient’s safety, the stability of the matching, and the repeatability of the measurements. The single stripline elements were fed in the potential-free symmetry plane through a balun, the coaxial cables were then routed through an additional shield, to prevent the EM-field, excited by the differential mode, to couple to the coaxial cable. The improved common mode rejection was shown experimentally with bench measurements in the single element, as well as the array case. In the single element case, the common mode behaviour of the new stripline element was compared to the traditional feeding strategy from the side. The single mid-fed element had a significantly better common mode rejection in comparison to the traditional feeding strategy by 13 dB. This manifested itself in a stable matching of the differential mode of the mid-fed element, whereas the side-fed element is highly unstable with respect to matching, when the common mode was manipulated by moving a floating cable trap on the coaxial cable. In the array consisting of mid-fed elements a stable matching of the differential modes was shown. The matching and coupling did not change, when manipulating with the common modes by shorting the coaxial cable’s outer conductors.

We have a high confidence in the safety assessment, due to the excellent agreement of measured and simulated results obtained in phantom experiments. The in-vivo validation shows also a reasonable agreement, taking into account the difference of the human body model, and more difficult positioning as compared to the phantom. The agreement is partly due to the aforementioned common mode suppression in the experimental setup, in the simulations only the differen-
tial mode is considered, and the precise construction of the coil array, based on PCB boards.

A large coverage along the axial direction was achieved with the modified feeding strategy. The single elements were fed in the middle with an out-of-phase current enabling to form a current lobe on both arms, producing a more homogeneous field distribution. This was predicted by EM-simulations, confirmed by measurements, and imaging experiments, as can be seen in the coronal and sagittal slices of the phantom experiments and in-vivo experiments. The RF-shimming capability of the array was then demonstrated in in-vivo imaging experiments. Homogeneous RF-shims are possible in all three orthogonal slices, allowing images with a large FOV with good contrast. The whole brain and cerebelum, brainstem, medulla, and even the upper part of the spinal cord are covered.

An in-vivo $B_1^+$-sensitivity of $11 \mu T/\sqrt{kW}$ was measured, comparable to previously reported sensitivities in stripline arrays [4, 27]. The value of the decoupling capacitor $C_{dec}$ on the coil array was optimized for a high $B_1^+$, a uniform field, and low mutual coupling between coil elements using the field superposition method [19, 20]. The current distribution of the optimized coil array resembles that of a degenerate birdcage coil, due to high currents through the decoupling capacitors $C_{dec}$.

Cooling elements were used to double the maximally allowed average input power of the array. The most sensitive parts of the coil array with regard to input power, the trimmer capacitors, were cooled. The special material properties of the cooling element made it possible to cool both trimmers on a single element with one cooling element. The used material is aluminium-nitride which is an electric isolator, but heat conductive. The increase in allowed input power is not beneficial for in-vivo imaging, here, the limiting factor for the input power is the local SAR in the human body [25]. For phantom imaging however, the allowed input power is increased, allowing a higher duty cycle range, and therefore enabling faster testing of the coil array. Due to reduced thermal stress, the life time of the trimmer capacitors potentially increases.
3.5 Acknowledgements

The authors would like to thank Beyhan Kochali from the Speag company for the measurements of the E- and H-fields, Hans-Rudolf Benedikter and Thomas Schmid for their help and advise with the RF-measurements, Anke Henning for help with the $B_1^+$ mapping, Stephen Wheeler for kind and patient manufacturing, and Niklaus Zölch for the $T_1$ measurement.
Bibliography


Chapter 4

Travelling Wave MR

This work was published in:

4.1 Basic Conception and Demonstration

Uniform spatial coverage in NMR and MRI is traditionally achieved by tailoring the reactive near field of resonant Faraday probes, such as a birdcage resonator [1, 2, 3, 4]. This approach is valid when the RF wavelength at the Larmor frequency is substantially larger than the target volume, which does not hold for modern, wide-bore, high-field systems. At the currently highest field strength that is used for human studies, 9.4 tesla [5, 6], the resonance frequency of hydrogen nuclei reaches 400 MHz, corresponding to a wavelength in tissue on the order of 10 cm. At such short wavelengths head or body resonators form standing-wave field patterns, which degrade MRI results by causing regional signal losses and perturbing the contrast between different types of tissue.

The non-uniformity of standing waves is due to the underlying electrodynamics, which require that the magnetic field exhibit curvature according to its frequency and the ambient material. Standing waves fulfill this condition by spatial variation of the field magnitude (Fig. 4.1a). However the required field curvature can also be translated, partly or wholly, into phase variation. Causing the underlying field pattern to propagate through space such phase variation reduces the variation of the field magnitude. Notably, the limiting case of a plane wave exhibits perfectly uniform magnitude at any wavelength. In addition, travelling RF waves offer a natural means of exciting and detecting NMR across large distances (Fig. 4.1b).

Despite these attractive features travelling-wave NMR has not been explored so far. In traditional cylindrical setups the formation of travelling waves at the NMR frequency is suppressed by structures surrounding the sample, such as gradient coils, cryostats, and RF screens. Their conductive surfaces admit axially travelling waves only beyond some cut-off frequency which is roughly reciprocal to the bore width. Therefore travelling-wave NMR requires a high-field magnet that also has a wide bore in order to bring the cut-off frequency below the NMR frequency.

To fulfill this requirement we used a cylindrical, superconducting
7.0-tesla magnet with a 58 cm-diameter bore lined with an RF screen (Fig. 4.1c). The screen was made from a stainless steel mesh designed to provide high conductance at RF frequencies while blocking audio-frequency eddy currents induced by the surrounding gradient coils. When enclosing only air the bore has a cut-off frequency of 303 MHz, which is still just above the proton Larmor frequency of 298 MHz. However, the frequency limit is reduced when dielectric material is brought into the bore. Even small amounts of dielectric loading enable the formation of axially travelling waves at the NMR frequency, effectively using the RF screen as a waveguide. A human body in particular - containing large amounts of water, which has a high permittivity - reduces the cut-off frequency sufficiently to clearly enter the travelling-wave regime.

NMR via such travelling waves requires a new type of probe. Instead of the reactive near field of the sample a travelling-wave probe
must couple to the propagating modes of the waveguide. To do so it no longer needs to be close to the sample but can be placed anywhere along the bore. Requirements of this sort are well known in microwave engineering and can be addressed by a range of technical solutions. For the present work the NMR probe was implemented in the form of a circularly polarized patch antenna (Fig. 4.1d).

Using this setup the principle of travelling-wave NMR was first demonstrated by spectroscopy of an aqueous 10% ethanol solution. Proton NMR was excited and detected by the patch antenna, which was initially mounted at the end of the RF screen, 70 cm from the sample. The resulting spectrum (Fig. 4.2a) shows the expected dominant water peak as well as the methyl and methylene resonances of the ethanol molecule. The experiment was then repeated with gradually increasing antenna distances. As the magnified methyl triplets in Fig. 4.2b show, consistent spectrum quality was obtained with the probe placed well outside the magnet and a well-resolved spectrum was still detected at a distance of 2.6 m from the sample. The evident loss of sensitivity at large distance reflects the expected decrease in coupling between the antenna and the modes of the bore. Higher sensitivity at large distances would be achieved with an antenna of greater directivity or a longer waveguide.

The spatial uniformity of the travelling waves was studied in an extended sample of 50 cm in length, formed by two adjacent bottles filled with mineral oil. On its own, this arrangement did not yield completely uniform RF coverage, as illustrated by the imaging results shown in Fig. 4.3a. The residual non-uniformity indicates the presence of a standing RF wave superimposed on the intended travelling component. It is caused mainly by slight reflections at the transitions between the bottles and the empty sections of the bore, which entail changes in wave impedance. The reflections can be mitigated by wave-impedance matching and additional loading. To demonstrate this two further bottles were added at the far end of the sample, the second one containing a conductive water solution to act as a termination dissipating incident wave energy. This modification indeed rendered the MRI results substantially more uniform, indicating the presence of almost purely
Fig. 4.2: Demonstration of travelling-wave NMR in an aqueous 10% ethanol solution. 
a) NMR spectrum obtained at an antenna distance of 70 cm, showing water (1), methylene (2), and methyl (3) resonances. 
b) Details of the methyl triplet as observed with increasing distance.

travelling RF waves (Fig.4.3b).

For reasons of safety, initial in-vivo experiments targeted only a volunteer’s lower extremities ensuring that the chest and head remained outside the waveguide. The antenna was placed at the opposite end of the bore, 70 cm from the ankle. The resulting magnetic-resonance image (Fig. 4.4a) shows the right lower leg with good uniformity over a large volume. The field of view of 50 cm is the maximum possible with the gradient system used and is not limited by the travelling-wave concept. Wave-impedance matching was not necessary in this case because the leg per se forms a sufficiently smooth, tapered impedance transition. For comparison the same imaging procedure was repeated with a commercially available birdcage resonator optimized for head MRI at 7 tesla. The result thus obtained (Fig.4.4b) exhibits smaller coverage, reflecting inherent limitations of resonant probes. The standing-wave nature of its rung currents limits the feasible length of the birdcage probe, which is 17 cm for this model. On longer rungs the RF current would become even more non-uniform [7], causing a similar
axial sensitivity drop-off and rendering the resonator unstable [8].

![Fig. 4.4: In-vivo results: a) Travelling-wave MRI of a human lower leg in-vivo. b) Identical scan performed with a traditional resonant probe.](image)

While it was possible to cover the lower legs uniformly, significant RF attenuation is expected to occur along the full length of a human body. Simulations of the setup in Fig. 4.1c, assuming an adult male subject, indicate that in total the body absorbs approximately 90% of the RF power coupled into the waveguide and only the remaining 10% are radiated off its far end. Less attenuation is expected for shorter or slimmer subjects or when using an even wider bore. Based on the same simulations the coupling efficiency of the patch antenna was estimated at 80%, a value that can certainly be improved by optimizing the strategy of driving the waveguide modes.

Besides coverage and uniformity the travelling-wave concept will also affect the sensitivity and RF power efficiency of NMR experiments. According to the reciprocity of NMR signals [9] the sensitivity of an RF probe is closely related to its efficiency, i.e., its yield of circularly polarized RF magnetic field at reference input power. Res-
Resonant near-field probes achieve high efficiency by concentrating RF energy and dissipation in the sample and the probe itself. By contrast, the travelling-wave approach relies on RF energy flowing through the setup, requiring that part of it be absorbed beyond the target volume. This can be done by a dedicated absorber device or diffusely outside the waveguide as was the case in our initial experiments. With either solution the necessary absorber losses will take some toll in terms of efficiency and sensitivity, constituting a drawback of travelling-wave probes compared with resonators.

Reduced probe efficiency is the lesser concern because it can be addressed by using higher driving power. In-vivo MRI is usually not limited by technical RF power constraints but rather by sample heating, to which the absorber losses do not contribute.

The corresponding sensitivity loss, caused by thermal noise originating from any material that absorbs RF power during transmission, is more limiting. One potential way of avoiding this loss is to use a cryogenically cooled absorber structure. For MRI applications it is also conceivable to combine travelling-wave excitation with local detection by an array of detunable surface resonators [3]. Such a hybrid approach will reconcile the improved coverage and safety advantage of travelling-wave excitation with the sensitivity benefit of close-range array detection.

With respect to net sensitivity another potential concern is the phase delay that results from signal propagation to and from the resonant nuclei. For large samples it will give rise to significant phase differences.

![Fig. 4.3: Example of wave-impedance matching in travelling-wave MRI: a) Non-uniform coverage of two phantom bottles is caused by residual standing RF waves, which can be suppressed by b) wave-impedance matching and dissipation in a termination load.](image-url)
between signals travelling different distances. This effect is illustrated in Fig. 4.5a, showing travelling-wave MRI of a long, water-filled cylinder with antennas placed at both ends of the waveguide. Using the same antenna for transmission and reception ($1 \rightarrow 1$ or $2 \rightarrow 2$) results in a linear distribution of the image phase. MRI will not be hampered by such phase variation as long as it is resolved by the imaging process. However spectroscopic experiments could suffer from delay-related dephasing. To address this problem the transmit and receive antennas should generally be designed and positioned such that the total phase delay is the same across the volume of interest. In the previously mentioned experiment this situation was accomplished by using either of the two antennas for transmission and the opposite one for reception. As shown in Fig. 4.5a the linear phase patterns nearly vanished in these configurations ($1 \rightarrow 2$ or $2 \rightarrow 1$). Alternatively, after travelling-wave excitation from either side a gradient encoding blip could be applied to compensate for variable phase delay, effectively refocusing the spin radiation for reception by the same antenna as used for RF transmission. Since the signal received in a travelling wave setup is directly linked to the radiation of electromagnetic power, this experiment shows that the radiation from spin ensembles can be steered by magnetic field gradients [10, 11]. Furthermore, the distinct phase gradient modulated on the NMR signal dependent on the direction of the signal propagation delivers as such the possibility for spatial signal localization. In the case ideal linear phase gradients can be established for each detector, the signal received by each antenna corresponds to the signal acquired after a phase encoding gradient moment dependent on the direction of propagation. This means that in a spin warp acquisition each antenna can acquire a different line in k-space simultaneously to anther antenna if the phase encoding is chosen in the direction of propagation. Considering an accelerated SENSE scan [12] the encoding provided by the RF is ideal corresponding to a $g$-factor of 1.

Entering the far-field realm, travelling-wave NMR prompts analogies with electron spin resonance [13] and nonlinear optics [14]. Adopting principles and devices from these fields may enable the study of phenomena analogous to, e.g., photon echoes [15], four-wave mix-
ing [16], and self-induced transparency [17]. Potentially useful analogies can also be drawn with the large variety of more widespread technologies that rely on travelling-wave phenomena, including manifold imaging modalities, radar, and telecommunication. The above discussed spatial localization by the propagation phase delay can for instance directly be related to holographic techniques. In the present work, the established concepts of waveguides and antennas have already proven useful.

Clearly, by adopting propagating waves NMR also incurs known complications of this realm, such as material-dependent diffraction and attenuation. These effects will likely be encountered, e.g., in MRI of the human torso, exhibiting pronounced dielectric interfaces such as the shoulders. Figure 4.5b shows an initial experimental illustration of dielectric perturbations in a phantom of similar size and shape as a human torso. One way of countering these effects is wave-impedance matching, which has been demonstrated here in a basic form. Analogously to index matching in microwave and optical technology, wave-impedance matching can be achieved using a wide range of distributions of dielectric, magnetically permeable, or conductive material and thus offers great inherent freedom for tailoring resulting RF field distributions.

By improving the extent and uniformity of spatial coverage travelling-wave MRI promises to facilitate the exploration of the highest currently available field strengths for human studies. A further promising area of application is high-field screening. The ability to perform spatially resolved NMR in larger volumes may simplify studies of large numbers of small animals [18] or inanimate samples in parallel.

Finally, introducing a significant distance between the sample and the NMR probe has multiple beneficial side-effects. In such a configuration the probe is not loaded by losses in the sample, which simplifies impedance matching and renders the probe performance substantially more robust than that of near-field high-frequency probes. This situation also simplifies safety considerations in human studies by avoiding exposure to strong short-range electric fields emanating from the probe. Furthermore, placing the probe far away frees up space in the
Fig. 4.5: Travelling-wave MRI of very large samples. a) Two antennas, one at each end of the waveguide, were used to image a long water cylinder. Transmission and reception with the same antenna (1 → 1, 2 → 2) give rise to linear phase delays, as shown in graphs of the image phase along the phantom. Constant net phase delay is achieved by transmitting with either of the antennas and receiving with the opposite one (1 → 2, 2 → 1). b) Image of a human torso phantom. Dielectric boundaries introduce discontinuities of the wave impedance, causing field perturbations that diminish towards the lower regions of the phantom.

center of costly high-field magnets [19], may improve the comfort of human subjects and facilitates bringing in alternative equipment, such as stimulation devices for studies of brain function.

4.2 Methods

All experiments were performed in a cylindrical superconducting 7.0-Tesla magnet (Philips Healthcare, Cleveland, Ohio), equipped with a three-axis gradient system and lined with an RF screen made from a stainless-steel mesh (diameter = 58 cm, length = 135 cm). NMR and MRI data acquisition was performed with an integrated console and spectrometer (Achieva, Philips Healthcare). Throughout, the RF transmitter power was limited to the same values as for human head ex-
posure with a local resonator, i.e., 10 W average forward power and 1.4 kW peak power.

The custom-designed patch antenna is shown in Fig. 4.1d. Its front plane is a 350-mm diameter disk made of copper sheet glued onto a 25 mm-thick acrylic (PMMA) former. The backplane is formed by a square copper sheet mounted on 5 mm acrylic. Acrylic yields negligible proton NMR signal at the echo times used in this work. A variable gap between the two planes was used to tune the antenna to the proton NMR frequency of 298 MHz. The antenna was driven at two points in the front plane forming a right angle with the center of the disk. The radial positions of the feed points were chosen such as to match the impedance of the 50 Ω feed lines. To produce circular polarization in the transmission mode the two ports were fed through a 90° hybrid splitter. In receive operation the two channels were connected to independent receive lines of the spectrometer and combined digitally. Switching between transmit and receive operation was performed by TTL-controlled diode switches.

The spectra shown in Fig. 4.2 were obtained from a 10% solution of ethanol in water, contained in a 1-litre glass beaker. To support the formation of axially travelling waves bottles of distilled water were added on either side of the beaker as dielectric loads. NMR spectra were acquired with a localized STEAM sequence (stimulated echo acquisition mode) with an echo time of 11 ms and a repetition time of 3 s. Low-flip-angle RF pulses were used such that the selected volume of 9 cm$^3$ changed negligibly with the RF amplitude. The acquisition bandwidth was 2 kHz and 16 phase cycles were performed to cancel spurious echoes. When changing the position of the RF probe all console and spectrometer settings were kept constant.

The data shown in Fig. 4.3 were obtained from bottles filled with 3 litres Marcol 82 mineral oil, using a gradient-echo sequence with an echo time of 2.9 ms, a repetition time of 42 ms, and a bandwidth per pixel of 338 Hz. A very small flip angle was chosen to ensure that the image intensity depends linearly on the transmit and receive sensitivities of the probe.

The in-vivo images of a volunteer’s right lower leg (Fig. 4.4) were
acquired with a 3D gradient-echo sequence, yielding an isotropic resolution of 1 mm in 6 minutes. The echo time was 3.1 ms, the repetition time was 12 ms, and the bandwidth per pixel was 376 Hz. All sequence parameters were kept the same for the two probes, except for the transmit power, which was reduced to 25 % for the resonator to obtain similar flip angles in the center. The axial length of the field of view of 500 mm was again limited by the gradient system. The resonant probe was a commercially available birdcage resonator with a diameter of 30 cm and a length of 17 cm (Nova Medical Inc., Wilmington, MA). Both legs resided within the resonator to keep the volunteer in the same position.

Electromagnetic simulations of the in-vivo setup were performed using the finite difference time domain (FDTD) technique (SEMCAD®, Schmid&Partner Engineering AG). The cylindrical RF screen was modeled as a perfect electrical conductor and an anatomical model of an adult male [20], featuring in-vivo dielectric tissue properties, was positioned at its center. The patch antenna was placed equally as in the in-vivo experiment, i.e., at a distance of 70 cm from the ankle. The flow of RF energy through the waveguide was assessed by integrating the Poynting vector over transverse control planes. A first control plane between the antenna and the human served to determine the power coupled into the waveguide. A second one in the loss-free region close to the end of the waveguide was used to calculate the power that is radiated off the far end of the waveguide. The power difference was validated by volume integration of ohmic losses over the entire body.

In the experiments reported in Fig. 4.5a, an RF-spoiled 2D gradient-echo sequence was used to image a 2 m long, water-filled cylinder of 7 cm in diameter. Two antennas were used simultaneously, one at each end of the bore, at a distance of 10 cm from the respective end of the waveguide to minimize reflections. In these experiments only a single port of each antenna was used for RF transmission and signal reception to avoid any phase perturbation by imperfect port combination. In two successive sets of experiments either of the antennas was used for transmission while receiving image signals with both. All scans were performed twice with an echo time difference of 0.5 ms, permitting
to estimate and remove any phase contributions from local frequency offsets.

The remaining image phase results exclusively from RF propagation, including unknown constant offsets of various sources, such as cabling and spectrometer calibrations. To eliminate these arbitrary offsets each phase plot was offset corrected to null the phase in the center ($z = 0$). The nominal field of view was 50 cm. In order to avoid phase distortions by potential gradient delays, the readout direction was orthogonal to the cylinder axis and no correction for gradient non-linearity was applied. The range of meaningful phase values was thereby reduced to approximately 35 cm.

Figure 4.5b shows a torso phantom made of PMMA as proposed by the American Society for Testing and Materials (ASTM), filled with gelled water ($\varepsilon_r = 80, \sigma = 0.47 S/m$). Imaging of this phantom was performed by a low-flip-angle, multiple-slice, RF-spoiled 2D gradient-echo sequence with a repetition time of 50 ms per slice, yielding an effective resolution of 1 mm in-plane and a slice thickness of 4 mm.
Bibliography


Chapter 5

Travelling-Wave RF Shim-ming and Parallel MRI

This work was published in:

5.1 Introduction

One of the most interesting and challenging aspects of high-field magnetic resonance imaging (MRI) is the complex behavior of the radiofrequency (RF) fields that are involved. At the length scale of the human body, operating frequencies of 300 MHz and beyond [1, 2, 3, 4] give rise to pronounced RF wave behavior, including interference, diffraction, propagation, and resonance, as well as variable interaction with different tissues and materials [5, 6, 7, 8, 9, 10].

For signal reception, more complex RF behavior can be beneficial because it usually entails greater diversity of possible sensitivity patterns. In conjunction with array detection, enhanced spatial RF diversity at high field has been shown to translate into higher intrinsic sensitivity [11, 12] and more effective parallel imaging [12, 13, 14]. For RF transmission, however, wave phenomena are primarily a challenge because they hamper uniform slice and volume excitation. In particular, birdcage resonators, which are commonly used for transmission in clinical MR systems and which yield highly uniform RF fields in near-field conditions, are prone to interference effects such as central brightening [6, 15, 16] in the wave regime. They are also subject to size limits related to the standing wave nature of their reactive near fields [17, 18].

In response to these issues, high-field MRI increasingly relies on transmission concepts that address and even exploit RF wave behavior. The most widely pursued approach of this kind is multiple-channel transmission, which permits tailoring the superposition of multiple RF field components for individual imaging situations (RF shimming [19, 20, 21, 22]). In analogy to parallel imaging, the spatial diversity of transmit fields can also be used to accelerate the formation of arbitrary excitation patterns with time-varying gradients (Transmit SENSE [23, 24, 25]).

Another approach is to make deliberate use of wave propagation. Propagating RF fields translate part of the inherent field curvature, which is required by electrodynamics, into spatial phase variation. Therefore they offer greater magnitude uniformity than standing waves, in
which the same curvature is realized through magnitude variation alone. In addition, wave propagation along a sufficiently wide magnet bore permits MR excitation and detection across large distances, effectively using the bore as a waveguide. These benefits have recently been demonstrated with a patch antenna as a transmit-receive probe [26,27]. Traveling waves along the magnet axis can also be excited and detected within the bore, using a small local probe [28]. Their transmission into the sample is subject to diffraction at material interfaces, which can be mitigated by matching with dielectric materials [26, 29]. Traveling waves are also subject to attenuation along conductive loads. To counter this problem and to optimize RF energy delivery to a target region, it has been proposed to complement the waveguide with a coaxial shield [30]. Furthermore, the benefits of traveling-wave excitation can be combined with those of local detection by combining antenna transmission with a close coupling receive array [31, 32].

A downside of current traveling-wave implementations is their lack of spatial diversity in the transmit fields and receive properties of remote probes. As a consequence, they do not support multiple-channel transmission and also do not exploit the high-field potential of parallel imaging. A first example of multiple-channel traveling-wave imaging was arguably reported in Ref. [26], using two transmit-receive antennas at opposite ends of the waveguide and studying differences in spatial characteristics related to propagation delays. However, affording only minimal, twofold field diversity, such a setup will offer only very limited RF shimming and parallel imaging capability.

The aim of this work is to overcome this deficit and explore the feasibility of traveling-wave MRI with a larger number of remote probes. With close-coupling arrays, spatial diversity of RF characteristics is achieved automatically when mounting the array elements in different positions. This is not sufficient, however, for a traveling-wave setup, in which remote waveguide ports act as effective array elements. In such a setup, spatial diversity of RF fields within the sample must be relayed to the port level by corresponding diversity of propagating fields between the ports and the sample. In a significant length of waveguide, the choice of suitable fields for this purpose is essentially given by the
Waveguides propagating modes, which are characterized by individual cut-off frequencies. Therefore, one prerequisite for the efficient use of multiple channels is to bring the cut-off frequencies of sufficiently many waveguide modes below the respective Larmor frequency. In this work, this is achieved by dielectric loading. Based on the resulting mode diversity, an effective transmit-receive array is formed by stub and loop ports at the end of the bore. The performance of the traveling-wave array is studied in examples of RF shimming and parallel imaging, including an investigation into the role of the amount of dielectric loading.

5.2 Theory

Most MRI magnets have cylindrical bores lined with an RF shield, effectively forming a circular waveguide. The oscillating electromagnetic fields that such a waveguide supports are found by solving the corresponding homogeneous Helmholtz equation, yielding the so-called waveguide modes [33, 34]. Each mode is characterized by transverse electric and magnetic field patterns $E_T(x, y)$, $B_T(x, y)$ and a cut-off frequency $\omega_c$, which is the minimum oscillation frequency required to permit the mode to propagate. For a given oscillation frequency $\omega$, the cut-off frequency determines the axial wave number

$$k_z(\omega) = \sqrt{\mu \varepsilon} \sqrt{\omega^2 - \omega_c^2}. \quad (5.1)$$

with $\mu$, $\varepsilon$ denoting the permeability and permittivity of the waveguide medium. Together with the mode patterns, the wave number defines the global field phasors

$$E(r) = E_T(x, y) \cdot e^{ik_z(\omega)z}$$
$$B(r) = B_T(x, y) \cdot e^{ik_z(\omega)z} \quad (5.2)$$

which reflect wave propagation if $\omega \geq \omega_c$ and exponential axial decay otherwise.
The waveguide modes can be divided into two disjoint sets according to their axial field components. In transverse magnetic (TM) modes, the axial magnetic field is zero, \((B_T)_z \equiv 0\). Their exclusively TM field suits them particularly for MR excitation. However, the TM modes with the lowest cut-off frequencies exhibit a zero of the magnetic field at the center. Transverse electric (TE) modes have no axial electric field, \((E_T)_z \equiv 0\). Their magnetic field is not purely transverse, but its transverse components increase with the wave number \(k_z\). Both types of modes are commonly labelled by two numbers, indicating circumferential and radial field variation. For instance, for an empty circular waveguide, the modes with the lowest cut-off frequency are two degenerate TE\(_{11}\) modes, which exhibit one cycle each of circumferential and radial variation. Two modes are termed degenerate if they have the same cut-off frequency but linearly independent field patterns.

The cut-off frequencies are inversely proportional to the waveguide diameter. In whole-body MRI systems with typical bore diameters of about 60 cm, the cut-off frequency of the TE\(_{11}\) modes is 300 MHz, followed by TM\(_{01}\) only at about 400 MHz. Therefore, traveling-wave experiments so far have largely relied on TE\(_{11}\), along with minor contributions from nonpropagating fields, which can mediate residual coupling across finite distances. In most implementations, the modes were excited by a single patch antenna, which impresses an approximately fixed current distribution at one end of the waveguide, creating one fixed mode superposition.

To extend this approach toward multiple-channel operation, a larger number of propagating modes must be made available by lowering their cut-off frequencies. According to microwave theory this can be done in a number of ways, involving dielectric filling [10, 35], conductive structures [36], or metamaterials [37]. The first approach is attractive because it can rely on simple structures and materials with low losses at the frequencies targeted here. On this basis, dielectric filling was adopted for this work, similar to Ref. [9], in which it was previously used to modify the resonant modes of a magnet bore. A solid filling of the waveguide with an isotropic material of relative permittivity \(\varepsilon_r\) would lower the cut-off frequency of all modes according
A solid filling with $\varepsilon_r \approx 5$ would be sufficient to support eight propagating modes in a circular waveguide of 60 cm in diameter. However, such a filling would either be expensive or very heavy and difficult to mount and handle. Furthermore, it would affect TE modes and TM modes equally although the latter might be expected to be more useful for MR excitation and detection. For these reasons, fractional filling with highly dielectric inserts was preferred. TM modes can be targeted selectively by using slim axial inserts, which hardly affect TE modes due to their lack of axial electric fields. The resulting modal structure depends on the number, shape, and placement of the inserts and cannot be readily deduced from the mode structure of the empty waveguide. However, if the structural features of the filling are small compared to the RF wavelength, it can be approximated as a solid filling with an effective permittivity for each mode:

$$\varepsilon_{\text{eff}} = \left( \frac{\omega_c^{\text{empty}}}{\omega_c^{\text{filled}}} \right)^2.$$

(5.4)

The effective permittivity can be determined directly by the Maxwell-Garnett formula [38, 39], or numerically by electromagnetic modeling.

Given sufficient propagating modes, their spatial diversity can be tapped by a suitable set of ports at the end of the waveguide. To preserve diversity, the ports must be disposed such that they couple to linearly independent mode superpositions in the feed section. At the interface between the dielectric filling and the sample, the propagating modes couple into the latter, translating given mode combinations into net field patterns. This coupling is expected to depend strongly on the width of the gap between the filled section and the sample since the otherwise propagating modes extend only evanescently into this empty part of the bore. Significant gaps will limit the transfer of RF power, which will instead be reflected back to the port level. Moreover, evanescent coupling also reduces the available diversity of field
patterns in the sample because the remaining power transfer will be dominated by those modes with the lowest cut-off frequencies. For effective multiple-channel operation, the sample should therefore be placed closely to the end of the filled section of the waveguide. The ports will then span a set of independent field patterns in the sample, thus offering the functionality of a transmit-receive array (Fig. 5.1).

5.3 Methods

5.3.1 Waveguide Extension

The experimental setup was based on a 7T Philips Achieva system (Philips Healthcare, Cleveland, OH) with a circular RF screen of 58 cm in diameter. For truly remote operation in clear waveguide conditions, the systems bore was elongated by a waveguide extension of 2 m in length (Fig. 5.2). It was made from a piece of polyethylene pipe with an outer diameter of 56 cm whose outside was lined with a brass mesh based on 0.25 mm wire, which was chosen such as to screen RF currents while hardly admitting any eddy currents at gradient switching frequencies. For sliding the extension into the MR system, it was equipped with two parallel grooves such as to fit onto rails present in the original bore. To safely maintain a closed conductive surface around the pipe, the grooves were covered with 0.5 mm sheet copper mounted with nylon screws. At its remote end, the waveguide was equipped with an end-cap based on a polyvinylchloride disk, which was also covered with the aforementioned brass mesh. The end-cap was mounted with nylon screws and contacted to the waveguide by copper sheets soldered onto both parts. The waveguide was thus terminated with a short circuit at its remote end while it remained open at the opposite end. The entire construction was supported by a wooden frame equipped with wheels to facilitate insertion into the bore.

Dielectric filling was realized by cylindrical PMMA tubes (length = 2 m, inner/outer diameter = 34/40 mm), filled with distilled water ($\varepsilon_r = 81$, $\sigma = 0.0001$ S/m). A total of 52 tubes were arranged on a transverse Cartesian grid with a pitch of 58 mm, supported by two
Fig. 5.1: Multiple-channel traveling-wave MRI. a: The sample to be imaged is placed in a waveguide for RF transmission and reception. In this work, the screened magnet bore was used as the waveguide, elongated by a cylindrical extension (Fig. 5.2). Multiple propagating electromagnetic modes in the waveguide are required to relay diverse field patterns in the sample to remote ports. b: To obtain linearly independent mode combinations, eight ports were disposed such as to target eight individual modes. For preferential coupling to these modes, seven stubs and one loop were placed to couple maximally to the respective modes electric or magnetic field.
Fig. 5.2: A waveguide extension was filled with up to 52 water-filled PMMA tubes for dielectric loading. a: Proximal end of the extension, which is inserted into the bore. b: Distal end with an endcap containing six waveguide ports (five stubs and one loop). Another two ports were implemented as radial stubs in the waveguide wall. The modes targeted by the numbered ports are listed in Table 5.1. c: Schematic of the electrical construction of each feed structure. d: Side view of the waveguide extension, showing its full length of 2 m. e: Prospective placement of a subject, 5-10 cm from the end of the extension. The phantoms used in this work were positioned at the same distance.
circular pressboard plates with suitable holes. The waveguide modes of the extension, with and without filling, were studied by 2D finite-element eigenmode analysis using COMSOL (COMSOL AB, Stockholm, Sweden). The results of these calculations are summarized in Table 5.1 which lists all modes that propagate in the presence of the filling at the targeted Larmor frequency of 298 MHz. The table includes the corresponding cut-off frequencies with and without filling, as well as the corresponding effective permittivities according to Eq. 3. These results illustrate that, as intended, the cut-off frequencies of TM modes are strongly reduced, with an effective permittivity of $\varepsilon_{\text{eff}} = 23.5$, which roughly corresponds to the volume average of $\varepsilon_r$. The dielectric effects on the TE modes are relatively small, with an effective permittivity of $\varepsilon_{\text{eff}} = 1.6$, enabling only TE$_{11}$ to propagate. Overall, including the multiplicities of degenerate modes, the dielectric filling makes 17 propagating modes available.

5.3.2 Multiple-Channel Transmission and Detection

To utilize these modes for multiple-channel operation, eight ports were introduced at the end of the waveguide. As mentioned earlier, the overriding goal in choosing such ports is to form net transmit fields and receive sensitivities of high spatial diversity in the sample. For a given type of sample and dielectric filling, the port design could be rigorously optimized. However, in the present initial implementation a simpler but still effective strategy was pursued. Relying on the intrinsic orthogonality of the modes, each port was disposed such as to maximize its coupling to one associated mode as described later. In this fashion, coupling to other modes is not explicitly suppressed but was found to be sufficiently small to ensure substantial linear independence. The modes targeted were the 6 TM modes of lowest cutoff frequency (TM$_{01}$, TM$_{02}$, and TM$_{11}$, TM$_{21}$ with two degenerate modes each) and the two degenerate TE$_{11}$ modes (Fig. 5.1b). In the following, the associated ports will be referred to by the mode names.

The port positions were chosen on the basis of the simulated modal field distributions. Except for TM$_{01}$, all ports were realized with copper stubs with a diameter of 7 mm, positioned such as to coincide with
a strong, collinear electric field in the targeted mode pattern. For the TM\(_{02}\), TM\(_{11}\), and TM\(_{21}\) ports, suitable positions for axial stubs were found on the end-cap (Figs. 5.1b and 5.2). To span the degenerate TM\(_{11}\) and TM\(_{21}\) modes, the stub positions were chosen such as to form 90° and 45° angles, respectively, with the center of the end-cap. To couple to the TE\(_{11}\) modes, two radial stubs were placed such as to form a 90° azimuthal angle, at a distance of one quarter of the effective wavelength from the end-cap (Figs. 5.1b and 5.2). The TM\(_{01}\) port was realized with a loop made from hand-formable coaxial cable (SU-COFORM_47_CU, Huber&Suhner AG, Pfaeffikon, Switzerland). The loop was placed on the end-cap such as to couple to the circumferential magnetic field of the TM\(_{01}\) mode at a radius where that of the TM\(_{02}\) mode is zero (Figs. 5.1b and 5.2).

All ports were mounted with 7/16 series panel connectors (Huber&Suhner AG), including small holes in the waveguide for the center conductors. Inside the waveguide, the stubs were screwed onto the respective center conductors. For matching, the length of the stubs was adjusted. The loop conductor was soldered to its panel connector and matched by changing its circumference. All ports were individually matched in the presence of the dielectric filling as well as the first phantom described later.

Each port was equipped with a custom-made, low voltage transmit-receive switch with an integrated preamplifier (MGA53453, Agilent, Santa Clara), yielding an isolation of more than 60 dB in transmission and a gain of about 18.5 dB in reception. The insertion loss in transmission mode was below 0.3 dB and the noise figure of the entire unit was 1.5 dB. Via the switches, the eight ports were connected to an eight-channel parallel transmit system (MultiX, Philips Healthcare) and to the regular Achieva spectrometer. For this study, the peak power of each amplifier and the total average power over all channels were limited to 500 W and 100 W, respectively.

5.3.3 Imaging Experiments

For all experiments, the waveguide extension was inserted into the magnet bore. For most scans, it was positioned such that the dielectric
Table 5.1: Propagating Modes in the Waveguide Extension. For all of these modes, the cutoff frequency in the empty waveguide was brought below the Larmor frequency of 298 MHz by full dielectric filling with 52 inserts, resulting in associated effective permittivities and axial wavelengths. Eight of these modes were targeted by a port each at the end of the waveguide (2nd column).

inserts reached up to a distance of 35 cm from the systems isocenter. Only for one part of the study, aiming to image the field distribution in the inserts, the extension was pushed in somewhat further. Two cylindrical phantoms were employed, one filled with a liquid mimicking in-vivo dielectric properties (diameter = 20 cm, length = 30 cm, \( \varepsilon_r = 56, \sigma = 0.78 \text{ S/m} \)) and one filled with saline (diameter = 20 cm, length = 34 cm, \( \varepsilon_r = 81, \sigma = 0.84 \text{ S/m} \)). Additionally, a watermelon (diameter ≈ 20 cm) was imaged as an example of a structured sample. The phantoms and the watermelon were always placed around the isocenter leaving a gap of 5 to 10 cm to the dielectric filling of the feed section.

To estimate transmit fields and receive sensitivities of the traveling-wave array, a low-flip-angle 2D gradient-echo sequence was used (transverse, resolution = 2 mm, slice thickness = 5 mm, TR = 100 ms). The long repetition time was chosen to limit saturation effects. The field of view was 300 mm for the phantoms and 450 mm for imaging the dielectric inserts. For each slice of interest, this scan was performed eight times, once with each transmit channel for excitation, using all eight channels for data acquisition throughout. Based on the resulting 64 reference images, maps of the relative transmit and receive characteristics were obtained by singular-value decomposition on a pixel-by-pixel ba-
sis [40]. The same scan as mentioned earlier, yet without excitation, was used to determine the noise covariance of the array [41]. Relative sensitivity maps and noise data of this kind, obtained from the first phantom, were used to study parallel-imaging performance in terms of the g-factor [42]. Relative transmit fields served to calculate RF shims, using magnitude-least-squares fitting of given target distributions [43].

To determine the transmit efficiency of the described setup, absolute $B_1^+$ mapping was performed in the saline phantom. The first phantom was not suitable for this purpose because it exhibited excessive chemical-shift induced peak splitting. $B_1^+$ mapping relied on the method described in Ref. [44], which requires 2D gradient-echo imaging with a variable nonselective pre-pulse as well as off-resonance mapping. The former sequence was performed with the following parameters: matrix = 64 × 64, field of view = 250 mm, TR = 200 ms, and pre-pulses of 10 ms. Corresponding off-resonance maps were obtained by regular gradient-echo imaging with two echo times differing by 0.5 ms. Based on the resulting data, $B_1^+$ maps were calculated by nonlinear fitting as described in Ref. [44]. Occasional outlier pixels due to partial volume effects were eliminated by a short-range median filter.

$B_1^+$ maps were obtained for three transverse slices at different axial positions, using individual RF shims. For two of these slices, the RF shims were designed to approximate uniform excitation. For the slice closest to the feeding system, the shim was designed to maximize $B_1^+$ at one single position using the phase-based approach described in Ref. [45]. Additionally, $B_1^+$ maps of each individual transmit channel were obtained, using the method described earlier together with the interferometric approach detailed in Refs. [44, 46, 47, 48]. In all $B_1^+$ mapping experiments, the actual power delivered to the individual ports was determined with a spectrum analyzer (FSL 9 kHz-3 GHz, Rohde & Schwarz, Munich, Germany) via a -20 dB directional coupler. The resulting power measurements were corrected for cable losses, which were measured separately with a network analyzer (Agilent ENA E5071C).

For imaging of the watermelon, the first phantom was additionally placed in the bore to mitigate axial standing waves in the sample. The
phantom was positioned adjacent to the melon in the direction of transmit field propagation, leaving a gap of 10 cm. In this fashion, wave propagation past the sample was facilitated and eventual reflections were attenuated by losses in the phantom, similar to a setup described in Ref. [26]. After the acquisition of relative transmit and receive sensitivity maps, a uniform RF shim was calculated for a central transverse slice through the watermelon. Using this RF shim setting, a higher-resolution image (0.7 mm in-plane, 5 mm slice-thickness) of the same slice was acquired with TR = 16 ms. Based on this full-Fourier scan, parallel imaging with varying acceleration was mimicked by image-domain aliasing and subsequent SENSE reconstruction.

A final set of experiments aimed to study how the performance of parallel imaging depends on the number and structure of available propagating modes and thus on the degree of dielectric filling. To this end, the number of dielectric inserts was successively reduced from 52 to 12. Due to consequent impedance changes, the ports needed to be re-matched for each filling configuration. To avoid geometric changes inside the waveguide, this was done by basic L-matching networks consisting of a shunt capacitor and a piece of coaxial cable for phase adjustment. The resulting scattering matrices were measured using the network analyzer. To assess parallel imaging performance, sensitivity maps of a central transverse slice through the first phantom were again acquired along with noise data to permit calculating g-factors. The number of available propagating modes was equally re-determined for each filling configuration, using finite-element modeling.

5.4 Results

5.4.1 Matching

The initial geometric matching resulted in stub lengths between 165 mm and 230 mm and a circumference of 360 mm for the loop exciter. The remaining reflections amounted to between -33 dB and -6 dB at an average of -11 dB. The maximum port-to-port coupling was -9.1 dB at an average of -14.5 dB.
5.4.2 Parallel Transmission

Figure 5.3 shows maps of the eight relative transmit characteristics in the dielectric inserts (a) and in the first phantom (b). As expected, due to nonexclusive feeding of the modes and due to the nonuniformity of the dielectric filling, the field patterns observed in the inserts do not match individual mode patterns of a uniformly filled waveguide. However, they do exhibit the desired spatial diversity, which is even more evident in the phantom. Figure 5.3c shows the result of RF shimming in the same phantom, aiming to focus on the indicated circle. It confirms the fundamental feasibility of RF shimming with the described traveling-wave setup. It also shows close correspondence between the calculated linear superposition of images obtained with individual-port excitation (planned) and the image obtained with the actual RF shim (measured).

Figure 5.4 shows the absolute $B_1^+$ maps measured in the saline phantom. The highest efficiency of 6.9 $\mu T/\sqrt{kW}$ was reached in the focus of the point shim in the most proximal slice (Fig. 5.4a). It corresponds to a nutation frequency of just over 400 Hz at the maximum total power of 1.9 kW that was used during $B_1^+$ mapping of this shim. This is also illustrated by the inset plot of the signal levels observed at this position as the amplitude of the 10 ms pre-pulse was varied. In the other two slices, uniform shims yielded mean efficiencies of 4-5 $\mu T/\sqrt{kW}$ (Fig. 5.4b,c). The $B_1^+$ maps of the individual channels (Fig. 5.4d) again illustrate the spatial field diversity that is afforded by multiple propagating modes, with a maximum single-channel efficiency of about 4 $\mu T/\sqrt{kW}$. The fact that only channel 6 had a significantly weaker excitation through the slice while maintaining field diversity among the channels shows that most of the modes are able to couple with substantial strength into the sample over the gap between the dielectric filling and the dielectric load of the phantom itself.

5.4.3 Parallel Imaging

Figure 5.5 shows maps of relative receive sensitivities in the inserts and phantom, corresponding to the transmit maps shown in Fig. 5.3.
Fig. 5.3: Multiple-channel transmission. a: Relative transmit fields in a transverse slice through the dielectric inserts. Each of these images reflects transmission through the single port with the number indicated (cf. Fig. 5.2). b: Analogous display for resulting relative transmit fields in a cylindrical phantom. c: Example of RF shimming, focusing on the dashed circle. The image on the right was obtained with the RF shim, closely matching that on the left, which shows the magnitude of the equivalent weighted sum of measured transmit fields. All plots show relative values between 0 and 1 according to the given color bar.
Fig. 5.4: Mapping of transmit efficiency based on absolute $B_1^+$ maps in three evenly spaced transverse slices. a: Efficiency of an RF shim focused on the position indicated by the cross in the proximal slice. The adjacent plot shows the signal variation at the targeted position caused by the variable prepulse in the $B_1^+$ mapping sequence. b, c: Efficiencies of RF shims targeting uniform excitation in the central and distal slices. d: Efficiencies of the eight transmit channels driven individually (proximal slice).
Fig. 5.5: Relative receive sensitivities of the eight ports as observed in transverse slices through (a) the filled section of the waveguide and (b) a cylindrical phantom. c: Maps of the resulting g-factors for Cartesian parallel imaging with varying reduction factors and fold-over directions.

Similar to the transmit characteristics, these maps do not reflect any direct correspondence with mode patterns of an ideal waveguide but do illustrate the spatial diversity that is afforded by the underlying propagating modes. As Fig. 5.5c shows, the achieved degree of spatial diversity is sufficient for parallel imaging with significant acceleration factors. Twofold undersampling in either dimension is possible at rather minor g-factor penalty and for threefold undersampling in either dimension the maximum g-factor remains below 3.

Parallel imaging of the watermelon is shown in Fig. 5.6. The receive sensitivities in this sample (Fig. 5.6a) are again highly structured and diverse. Uniform RF shimming and SENSE combination of the received data yielded Fig. 5.6b, which exhibits remarkable uniformity of depiction. Note that high image intensity in the center reflects genuine image contrast unrelated to RF nonuniformity, which would cause smoother intensity variations. Figure 5.6c,d illustrates the feasibility of
accelerated imaging in this structured sample, using undersampling by factors of 2.5 and 4.

Figure 5.7 shows the dependence of the parallel imaging performance on the number of modes supported by the waveguide. Reducing the dielectric filling from 52 to 36 and 24 inserts decreased the number of propagating modes from 17 to 13 and 10, respectively. With only 12 inserts, 5 modes are still available. In agreement with theory, the empty feed waveguide, being slightly smaller in diameter than the bore of the MR system, was found to support no propagation at all at 298 MHz. This numerical result was illustrated by the fact that none of the ports could be matched in this case and no quantifiable MR signal could be acquired. In the bottom half of Fig. 5.7, the parallel-imaging performance is analyzed for the four filling configurations. Remarkably, the feasible range of acceleration factors changes only moderately as the mode diversity decreases. For acceleration up to threefold, the most scarce dielectric filling with only 12 inserts even yielded the lowest mean g-factors and also some of the lowest g-factor maxima. Only for high acceleration factors, the lack of field diversity appears to become relevant, causing a steeper increase of the g-factor penalty than observed with a larger number of available modes.

Figure 5.8 finally, shows the noise correlation and scattering matrices of the four setups with different degrees of dielectric filling. One clear observation in these data is a tendency of noise correlation to increase as the number of available modes decreases. This trend directly reflects the decrease in spatial diversity of electric field patterns in the lossy sample, resulting in more shared noise at the port level.

5.5 Discussion

This study demonstrates that multiple-channel MRI can be performed with a remote probe array, relying on traveling-wave excitation and detection. To make effective use of multiple channels, the traveling-wave approach requires multiple propagating modes that relay spatial diversity of RF fields from the sample to the waveguide ports. Dielectric loading of the bore is an effective means of increasing the number and
Fig. 5.6: RF shimming and parallel imaging in a watermelon. a: Relative receive sensitivities of ports 17 in a central transverse slice. b: Array image obtained with a uniform RF shim and full Fourier encoding. Note that the high image intensity in the center reflects genuine image contrast unrelated to RF nonuniformity. c, d: SENSE reconstructions from data obtained by 2.5- and 4.0-fold aliasing of the full-Fourier data.
Fig. 5.7: Dependence of parallel imaging performance on dielectric filling of the waveguide. Top: the dielectric load was gradually reduced from 52 to 12 inserts, reducing the number of available propagating modes from 17 to 5. Bottom: resulting g-factor behavior for Cartesian parallel imaging in a central transverse slice through a cylindrical phantom. Different colors correspond to different filling configurations as indicated earlier.
Fig. 5.8: Noise correlation and S-matrices (magnitude shown) as obtained with the different filling configurations shown in Figure 5.7.
In addition to being of certain conceptual interest, traveling-wave arrays offer several features of potential practical relevance. Compared to close-coupling arrays, one interesting aspect is that all lumped elements and other potential high-voltage points are removed to the port level at the end of the waveguide. The remote approach thus eliminates a potential source of electrical hazard, e.g., by the formation of arcs in transmit operation. It also naturally avoids power deposition in the sample via conservative electric fields that emanate from capacitors or other components that concentrate electric field. The waveguide ports described in this work require matching like close-coupling arrays, but they do not need to be resonated with typically costly variable high-voltage components. Moreover, according to initial experience, the remote array exhibits relatively small sensitivity to load variations and performs robustly without re-matching for different samples.

With respect to the comfort of human subjects and potential claustrophobia, dielectric filling is arguably a drawback. It was found, however, that rather scarce filling with only 12 tubes, occupying $\sim 5\%$ of the bore cross section, was sufficient for effective multiple-channel operation. At such small density, the transparent tubes have a limited effect on the perception of space in the bore.

As previously discussed in Ref. [26], one general drawback of traveling-wave probes is a certain toll in transmit efficiency and receive sensitivity due to deliberate radiation losses. The transmission and detection performance of a traveling-wave array is additionally limited by the fact that it encompasses the entire sample with all lossy material therein. In this respect, its performance is best compared with body coils and body arrays [19, 49]. Against this background, the observed $B_1^+$ efficiencies of 4-7 $\mu T/\sqrt{\text{kW}}$ in the saline phantom are an encouraging preliminary result. Notwithstanding, local receivers clearly offer higher detection sensitivity. Therefore, remote array transmission with local array reception is an attractive combination, reconciling the advantages of both approaches in terms of safety, matching robustness and sensitivity.

In addition to basic efficiency and sensitivity, the performance of
the remote array depends crucially on the spatial diversity of its RF fields in the sample. Relying on between 5 and 17 different modes coupled to eight ports, RF shimming and parallel imaging with up to fourfold acceleration have been successfully performed. Heuristic placement of the ports proved sufficient to couple to linearly independent mode superpositions. Nevertheless, some improvement in array performance can likely be accomplished by optimizing the port design for spatial field diversity in the sample. The benefit of such optimization will generally be larger for more propagating modes and for fewer ports.

One straightforward means of supporting a larger number of modes is even stronger dielectric loading. However, an ever higher refractive index in the filled part of the waveguide will eventually result in excessive reflection at the transition to the imaging volume. This limit was not yet reached in this work, as shown by limited scattering at the port level. Another conceptually simple approach for doubling the number of available modes is the inclusion of ports at both ends of the waveguide. As described in Ref. [26], such bi-directional transmission leverages propagation delay to introduce spatial diversity between modes that differ only in the direction of propagation. Even with ports only at one end, some spatial diversity along the waveguide axis is available because non-degenerate modes exhibit different axial wave numbers $k_z$ and thus different wavelengths. As a consequence, limited axial RF encoding is possible even with a single-sided traveling-wave system. On the downside, differences in axial wavelength also mean that RF shims that involve non-degenerate modes must necessarily vary in magnitude along the axis. However, at the long axial wavelengths used currently, such mode interference effects are smaller than those of diffraction, reflection and attenuation by typical imaging targets.

Dielectric loading with axial inserts was used to lower the cut-off frequencies particularly of TM modes. With this approach, high and approximately equal effective permittivity $\varepsilon_{eff,TM} \approx 23.5$ was achieved for all TM modes, along with a low value $\varepsilon_{eff,TE} \approx 1.6$ for the TE modes (Table 5.1). The former value corresponds well to the average volume permittivity in the waveguide ($\varepsilon_r = 23$), while the latter is well
approximated by the Maxwell- Garnett formula [38, 39], which yields $\varepsilon_r = 1.7$. Such mesoscopic analysis permits estimating the mode structure of a filled waveguide without actual field modeling in the structured material. Particularly, the Maxwell-Garnett formula predicts that the transverse effective permittivity of the described axial filling will remain rather low up to very high filling fractions due to a shielding effect of the polarized inclusions. Interestingly, the inverse of this effect could be used to achieve much higher effective transverse permittivity with a small volume fraction of dielectric material. A bulk insert consisting of air-filled tubes surrounded by water would yield an effective transverse permittivity close to the volume average of $\varepsilon_r$. Such a setup would hence be expected to lower the cut-off frequencies of TE modes equally effectively as those of TM modes.

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Chapter 6

Safety and In-vivo imaging for a Travelling Wave Multiple Channel Setup

To be submitted:

6.1 Introduction

High field MRI (> 7T) is promising due to a higher expected SNR. It also poses challenges; The RF-wavelength is in the range of the imaging sample, leading to a pronounced wave behaviour [1]. The skin effect hampers the EM-field to penetrate into deep tissue, and multiple reflections at dielectric boundaries and conductors lead to standing waves. All these effects lead to an non-uniform EM-field which is beneficial for parallel imaging techniques [2, 3]. Another effect that arises at high field strength human scanner’s is the propagation of waveguide modes in the scanner’s bore [4].

The RF-transmit field is hit by these effects [5]. The non-uniform H-field results in an non-uniform transmit field, causing non-uniform images or even signal nulls. The non-uniform E-field leads to local SAR hotspots, that can lead to tissue damage. Field enhancement close to the coil elements can be observed, and lower field strength in deep tissue. This can be overcome by radiative antenna elements [6], whose Poynting vector is directed towards the tissue. The propagation of waveguide modes allows RF-power to radiate away from the imaging sample lowering the power efficiency of RF-coils. More degrees of freedom are necessary to mitigate these effects.

Multiple RF coils can be used [7, 8] that offer the opportunity to steer the EM-field, with a coil array connected to a butler matrix or to independent transmit amplifiers [9]. The pulses can be driven in a fixed setting of amplitudes and phases of the transmit elements (static shimming) [7], or in a time dependent fashion (dynamic shimming) [10]. This comes at the expense of hardware complexity. Tuning and matching of close coupling coil arrays is a time consuming issue. This also applies to the numerical simulations needed to achieve a reliable safety assessment.

The propagating waveguide modes at high field MRI can be exploited for RF transmission from the distance. The concept of travelling waves offers several advantages compared to close coupling arrays, such as a larger FOV, increased field uniformity along the axial direction, more patient space, and easier matching and tuning of the
transmit elements due to far field coupling. The lower SNR of travelling wave [11] can be overcome by local receive coils [12]. Several excitation mechanisms were reported to couple into travelling waves. In [13] a patch antenna was used to excite the travelling waves in the scanner’s bore, in [14, 15, 16] the bore is modified to excite coaxial waveguide modes, in [17] a loop coil was used to couple into travelling waves. These implementations are easy to build, but they have, due to their single quadrature channel, limited RF-shimming capabilities. This can be overcome by using dielectric pads [18, 19, 20], lining the inner surface with a adaptive meta-material [21], dielectric lining of the bore [22], or using absorbing layers [23] to match the sample to the travelling wave to improve image homogeneity, increase SNR, and power efficiency.

A new approach is to combine travelling wave with multiple channel transmission. This can be achieved by coupling into higher order modes [24]. At 9.4T human systems with a wide bore 2 modes are propagating [25]. At a 7T human scanner, a single mode is propagating, but the number of propagating modes can be increased by dielectric loading of the scanner’s bore [26, 27].

In [26], a multimodal waveguide extension (WGE) is used to transport spatially diverse field distributions to the imaging area, first experimental results in phantoms exist. Homogeneous as well as local RF-shims were demonstrated.

The combination of travelling wave with multiple channels could be beneficial for in-vivo imaging. It combines the large FOV, homogeneity, matching stability, and patient comfort of travelling wave and the degrees of freedom of parallel transmission.

The necessary step for in-vivo imaging is a careful and reliable safety assessment [28], accounting for the uncertainties particular to this RF-system. These uncertainties include the exact positioning of the excitation elements, and the coupling of the coaxial cable modes into the modes of the WGE. 3D EM-simulations using FEM or FDTD are not well suited because of the large required amount of memory, due to the size of the setup, and the complexity of the WGE. A simulation method taking advantage of the waveguide structure of the TWM
would be a better choice.

The mode matching technique [29] is widely used for EM-modelling of waveguide structures. The concept of dividing the RF-setup into an EM-domain and RF-domain for the simulation of RF-coil arrays is well known in the MR community [30, 31, 32]. The computational domain is divided into sub-domains with uniform cross-section (waveguides) and discontinuities. The EM fields in uniform waveguide sections are decomposed into waveguide modes, that are eigenvectors of the 2D cross-section of the waveguide, the field evolution along the axial direction is described by the propagation constant, the corresponding eigenvalue. The reflection, conversion and transmission into modes at discontinuities can be described by scattering matrices. The discontinuities can be modelled using 3D EM-simulations. The sub-domains can then be combined in a post-processing step, satisfying the boundary conditions at the interfaces. In this way the EM-simulation becomes more flexible, parts of the system can be easily exchanged. The EM-simulation becomes faster, due to the smaller sub-domains, and more accurate, due to the finer meshing of the smaller sub-domains.

The excitation section, where the coaxial cable ports couple to the propagating modes of the WGE, is difficult to model. The most important aspect for worst case estimates for safety is the WGE, which acts as a filter, due to its length, and limits the amount of possible field distributions in the imaging object, due to the finite number of propagating modes. Therefore, the excitation section can be neglected in this work. The safety assessment is then based on modes of the WGE, rather than on the coaxial cable ports. The space of field distributions that is spanned by the coaxial cable ports is included in the space of field distributions that is spanned by the WGE modes. Therefore the worst case SAR [33] of the fields spanned by the WGE modes will be larger or equal than the worst case SAR of the fields that are spanned by the coaxial cable ports, yielding an upper boundary of the experimentally achievable worst case SAR.

The RF-simulations were validated in a reference scenario. $B_1^+$ maps in a cylindrical head sized phantom were acquired, and compared to the simulated RF-fields. A worst case SAR estimate was computed
within the given uncertainties for a given input power. First in-vivo images were acquired with the TWM in a safe mode of operation, with static RF-shimming using eight independent transmit channels. The TWM was used in transceive mode and in transmit mode only, using a head coil array for reception, showing a homogeneous transmit profile.

6.2 Methods

6.2.1 Experimental RF-setup

The TWM setup is described in reference [26] (or chapter [5]) in detail. It consists of a PE tube, the so called waveguide extension (WGE), that has a length of 2 m and a diameter of 0.56 m. The WGE is covered with a conducting brass mesh and closed on one side with a conducting plate, which is connected electrically to the brass mesh cover of the WGE, forming a short-circuit. The opposite side is left open. The WGE is filled with 52 PMMA rods with an inner/outer diameter of 34/40 mm, that are each filled with distilled water. Distilled water has a relative permittivity of $\varepsilon_r = 80 - j1.2$ at 300 MHz. The WGE is excited on the closed side by conducting rods and a loop, that are inserted into the WGE, and connected to the inner conductor of coaxial connectors. The coaxial connectors are screwed on the outer shell of the WGE and the conducting plate, the outer conductor is electrically connected to the conducting surface of the WGE. Six excitation elements are mounted on the backplane, two are mounted on the outer shell of the waveguide, at a distance of 0.5 m from the backplane. The last 1.5 m of the WGE is a piece of uniform waveguide.

The WGE is inserted into the scanner’s bore, open end first. The bore is loaded by the imaging sample. The imaging sample is positioned in the isocenter of the scanner, and the distance between WGE and the top of the sample is 0.1 m for the imaging experiments, when the TWM is used in transceive mode (for the in-vivo and the validation setup). For the case, when the TWM is used in transmit mode only, and a local receive coil is used for reception, the distance of the WGE to the top of the human head is 0.26 m.
6.2.2 EM-modelling

The TWM described above is modelled by the mode matching technique. The RF-setup is divided into five separate regions (see Fig. 6.1), that are sorted from the load to the source. The first part is the scanner’s bore loaded by an imaging object. This is followed by the second part, the waveguide junction, consisting of a 0.1 m long empty waveguide of the bore, which can be considered as a circular waveguide filled with air, the step discontinuity, and a 0.22 m long piece of the WGE. The third part is a 1.28 m long piece of the WGE, which is a longitudinally uniform piece of multimodal waveguide. The fourth part is the 0.5 m long excitation section of the cables, connectors, transmit amplifiers, circulators TR-switches, and matching circuits. As mentioned above the excitation section of the WGE is not included in the numerical model, therefore the analysis includes only the first three parts of the TWM, the bore loaded by the human, the waveguide discontinuity, and the WGE.

The TWM consists of two waveguides; the WGE, and the circular waveguide of the scanner’s bore. In the mode matching analysis the EM-field in a waveguide of finite length is described analytically as the superposition of an infinite number of modes, obtained by the eigenvalue analysis performed on the cross-section of the waveguide [34]. Since it is not possible to obtain an infinite number of modes numerically, a finite number of dominant modes $M$ can be used to describe the EM-fields sufficiently. In the eigenvalue analysis of waveguides, the $k$th mode is described by the transversal field distributions $e_{t,k}$ and $h_{t,k}$ in the cross-section and a corresponding propagation constant $\gamma_k$ that describes the propagation along the axial direction of the waveguide. The modes can travel in the forward and backward direction, and their amplitudes and phases are described by the complex forward and backward coefficients, $a_k$ and $b_k$ respectively, defined at the discontinuities of the waveguides. The EM-fields are then given by:

$$E_t = \sum_{i}^{M} (a_i e^{-\gamma_k z} + b_i e^{\gamma_k z})e_{t,i} \quad H_t = \sum_{i}^{M} (a_i e^{-\gamma_k z} - b_i e^{\gamma_k z})h_{t,i} \quad (6.1)$$
Fig. 6.1: Sketch of the travelling wave multiple channel setup, and the decomposition into sub-domains.
The propagation constant is in general complex. The mode is called propagating, if the imaginary part is larger than the real part, and evanescent, if the real part is larger than the imaginary part. In a lossless waveguide the propagation constant can be purely real or purely imaginary, in a lossy waveguide, the propagation constant is always complex. As mentioned previously, far away from the reactive region of a discontinuity in the waveguide only a finite number of modes contribute to the field distribution, because higher order modes (i.e. modes, whose propagation constant has a large real part) decay very fast. The modes are orthogonal to each other according to [34]:

$$\int_A e_t^{i} \times h_t^{j} \, dA = \begin{cases} 1, & \text{for } i=j \\ 0, & \text{for } i \neq j \end{cases} \quad (6.2)$$

where the transversal modal fields are normalized. The integration area $A$ is the cross-section of the waveguide. This orthogonality relation holds also for inhomogeneously filled, and lossy waveguides.

As described above, the RF-system is divided into subregions, at suitable reference planes, indicated in Fig. 6.1. These reference planes are cross-sections of waveguides, that are placed sufficiently far away from waveguide junctions, which makes it possible to describe the power flow by the superposition of a limited number of dominant modes. In the present case we have only one waveguide junction, which is the step discontinuity connecting the WGE and the circular waveguide of the bore.

The RF-behaviour of the first two subregions, the WGE and the waveguide discontinuity can be fully described by scattering junction matrices $[S_j]$, the subscript $j$ stands for junction:

$$\begin{bmatrix} b^{(i)} \\ b^{(i-1)} \end{bmatrix} = [S_j^{(i)}] \begin{bmatrix} a^{(i)} \\ a^{(i-1)} \end{bmatrix} \quad \text{with:} \quad [S_j^{(i)}] = \begin{bmatrix} [S_{j,11}^{(i)}] & [S_{j,12}^{(i)}] \\ [S_{j,21}^{(i)}] & [S_{j,22}^{(i)}] \end{bmatrix} \quad (6.3)$$

The entries of these scattering matrices correspond to modes at the corresponding reference planes. The forward direction is defined towards the junction and the backward direction away from the junction,
as illustrated in Fig. 6.2. The junction matrix can be decomposed into block-matrices. The diagonal block-matrices describe the reflection and conversion of forward travelling modes in one waveguide into backward travelling modes of the same waveguide. The off-diagonal block-matrices describe the transmission of forward travelling modes of one waveguide into backward travelling modes in the other waveguide. The superscripts of the scattering matrix, and sub-matrices stand for the $i$th subregion, the superscripts of the forward and backward wave coefficients stand for the reference planes. Note that the reference impedance of the ports can be complex for the modes of lossy waveguides and evanescent modes, unlike conventional scattering matrices whose ports have a purely real reference impedance. For complex reference impedances the amplitude of the reflection coefficient may be greater than one, even for passive devices [35].

As discussed above, a finite number of modes are used to describe the fields in corresponding waveguides. The criterion for including a mode in the analysis is defined by the amplitude of the forward travelling coefficient at the discontinuities, at $z = -0.1 \text{m}$ for the modes of the WGE, and $z = 0 \text{m}$ for the modes of the circular waveguide, when excited by forward modes in the WGE with unit amplitude at $z = -0.32 \text{m}$.

Let us consider the first part of the considered RF-setup, the WGE. A mode travelling through a uniform piece of waveguide will be altered
only in phase and in amplitude by the propagation constant. There is no conversion into other modes or reflection due to geometrical changes of changes in material properties, therefore the only non-zero elements of the junction matrix \( S_j^{(2)} \) are the diagonal elements of the off-diagonal sub-matrices. The set of modes at both reference planes is the same. The first reference plane is at \( z = -1.6 \) m and the second is at \( z = -0.32 \) m. The eigenfields and corresponding propagation constants are computed by a 2D eigenvalue analysis performed by the commercial finite element (FEM) based simulation platform COMSOL. In the eigenvalue analysis, 7000 quadratic elements are used in order to obtain a sufficient amount of candidate modes for the mode matching analysis. The conducting boundary is modelled as a perfect electrical conductor (PEC). The water as well as the PMMA tubes are modelled as lossy dielectrics. Only the set of modes \( N_1 \) is considered, whose forward wave coefficients have a coupling of more than \(-30\) dB at \( z = -0.1 \) m, when excited at \( z = -1.6 \) m. The modes in the set \( N_1 \) are sorted in ascending order with respect to the real part of their propagation constant.

Now, let us consider the second part, the junction between the WGE and the circular waveguide of the empty bore. The junction scattering matrix for this subregion was computed in a 3D full wave simulation in two steps. In the first step the WGE was excited with the previously determined set of modes \( N_1 \). The mode coefficients for mode \( i \) and the \( j \)th excitation \( A_{i,j} \) were then computed in both waveguides from the EM-field, using the orthogonality relation from Eq. 6.2. The modal electrical field in Eq. 6.2 was replaced by the transversal component of the electrical field that is excited by the \( j \)th mode \( E_{t,j} \).

\[
A_{i,j}(z_k) = \int_{A} E_{t,j}(z_k) \times h_{t,i} \, dA \tag{6.4}
\]

Computation of the mode coefficient at two or more planes allows us to compute the forward and backward coefficients of the mode, \( a_{i,j} \) and \( b_{i,j} \) respectively, using a linear least squares fit, with the knowledge
of the propagation constant $\gamma_i$:

$$
\begin{bmatrix}
A_{i,j}(z_1) \\
\vdots \\
A_{i,j}(z_Z)
\end{bmatrix}
= a_{i,j} 
\begin{bmatrix}
e^{-\gamma_i z_1} \\
\vdots \\
e^{-\gamma_i z_Z}
\end{bmatrix}
+ b_{i,j} 
\begin{bmatrix}
e^{+\gamma_i z_1} \\
\vdots \\
e^{+\gamma_i z_Z}
\end{bmatrix}
$$

(6.5)

In this way the significant modes in the second waveguide are found. They form the set of modes $N_2$. In the second step the waveguide junction was excited from the other side with the set of modes $N_2$, and again, the forward and backward coefficients on both sides are computed for each simulation. Arranging the forward and backward wave coefficients in matrices $[a]$ and $[b]$, whose rows represent the different modes, and the columns the different excitations, allows us to compute the scattering matrix:

$$
[b] = [S][a] \quad \Rightarrow \quad [S] = [b][a]^\dagger
$$

(6.6)

(\cdot)^\dagger$ is the pseudo-inverse.

The waveguide junction was analyzed in 3D by COMSOL. The numerical domain was discretized using $200k$ quadratic prism elements. A 0.3 m long part of the WGE was modelled followed by a 0.13 m long part of the empty bore. As mentioned above, in a first step the numerical domain was excited through the WGE with the set of modes $N_1$, and the empty bore was terminated with a scattering boundary condition. The forward and backward coefficients and field errors with respect to the modal fields were computed in three equidistant planes in both waveguides on a 2mm rectangular grid. In the WGE with respect to the set of modes $N_1$ from $z = -0.32$ m to $z = -0.24$ m, and in the empty bore with respect to the first 100 circular waveguide modes from $z = 0$ to 0.02 m. For the bore that is modelled as an empty circular waveguide, analytical expressions for the modes are available, see [34]. The modes in the empty waveguide, that have a backward coefficient greater than $-30$ dB, when the WGE is excited at $z = -0.32$ m by the set of modes $N_1$ with unit amplitude, form the set of modes $N_2$. In the second step the excitation and scattering boundary condition were exchanged, the empty bore was excited with the set of modes $N_2$. 
Also here, the forward and backward coefficients and field errors with respect to the modal fields were computed in the same way in both waveguides. The coefficients were computed in the WGE with respect to the first 100 waveguide modes, and in the empty bore with respect to the set of modes \( N_2 \). There is contribution of modes in the WGE and in the empty bore at both reference planes that are included in the first 100 modes but not in the set of modes \( N_1 \) or \( N_2 \) respectively. But they do not contribute to the field at the discontinuity significantly, due to a fast decay, and therefore were neglected. The forward and backward coefficients from the two simulations were used to compute the junction matrix \( [S_j^{(1)}] \), that describes the coupling of the set of modes \( N_1 \) from the reference plane \( z = -0.32 \) m in the WGE to the set of modes \( N_2 \) at the reference plane at \( z = 0 \) m in the empty bore.

The last subregion, the bore loaded by the human was also simulated as a 3D full wave model. This subregion can be described by the termination scattering matrix \( [S_t^{(0)}] \) and corresponding EM-fields in the imaging object. The termination matrix relates the forward and backward modes at one reference plane only, \( b^{(i)} = [S_t^{(0)}]a^{(i)} \). It is computed the same way as the junction matrix of the waveguide junction, Eqs. 6.4, 6.5, 6.6.

For the transformation of the EM-fields to different reference planes through junction matrices, we need the fields that are excited by the forward travelling mode with unit amplitude selectively \( F' \). In the simulation however the excitation of modes is not selective, in general, and therefore we need to transform the simulated fields \( F \). EM-fields that are excited by a waveguide, can be considered as a superposition of the fields \( F' \), weighted by the corresponding forward coefficient at the given reference plane. For a set of simulations and modes we can then write:

\[
[F] = [F'][a] \quad \Rightarrow \quad [F'] = [F][a]^\dagger \quad (6.7)
\]

The rows of the matrices \( [F] \) and \( [F'] \) correspond to the different positions. The columns of \( [F] \) correspond to the fields from the different simulations, and the columns of \( [F'] \) correspond to the fields that are excited with the different modes. The matrix \( [a] \) is computed with
This subregion is simulated using the 3D FEM solver HFSS, with an adaptive mesh-refinement and mixed order basis functions, on a 256 GByte RAM computer. The numerical model consist of a 0.01 m long excitation cylinder, that has the same diameter as the bore of 0.592 m and has anisotropic permittivity, mimicking approximately the macroscopic anisotropy of the WGE, to excite the set of modes \( N_2 \). The excitation cylinder is followed by a piece of empty bore. The human model is placed at a distance of 0.1 m after the excitation cylinder, head first. The simulation is excited by waveguide modes of the excitation cylinder at \( z = -0.11 \text{m} \). The numerical domain is terminated by a cylindrical radiation boundary, coaxially with the scanner’s bore. The radius of the cylindrical radiation boundary is a quarter wavelength larger than the scanner’s bore, and it extends from the end of the bore at \( z = 0.785 \text{m} \) to a quarter of a wavelength distance from the human’s feet. The cylindrical shell of the excitation cylinder and that of the scanner’s bore are terminated with a PEC boundary condition. The forward and backward mode coefficients of the mode-set \( N_2 \), the field errors, at three equidistant planes from \( z = -0.08 \text{m} \cdots -0.06 \text{m} \) on a 2mm rectangular grid, and the termination matrix \([S'_t(0)]\), at the reference plane \( z = 0 \text{m} \) are computed with Eqs. [6.4-6.6]. The fields are transformed, as if they were excited by the set of waveguide modes \( N_2 \) selectively with unit amplitude at the reference plane \( z = 0 \text{m} \), according to Eq. [6.7]

The following subregions of the RF-system can be added by transforming the termination matrix and the EM-fields through the corresponding junction scattering matrices of the subregions. The boundary conditions at the reference planes have to be satisfied, by making sure that the ports belong to the same modes on both sides of the reference plane.

Knowing the set of selectively excited fields \([F'_i]\) at a given reference plane \( i-1 \), the EM-fields can be transformed to a second reference plane \( i \), even to another waveguide through a waveguide junction, if the relation of the forward travelling modes from the first to the sec-
ond reference plane is known:

\[
[a^{(i-1)}] = ([I] - [S_{j,22}^{(i)}][S_{t}^{(i-1)}])^{-1}[S_{j,21}^{(i)}][a^{(i)}] \quad (6.8)
\]

Here the matrices \([a^{(i-1)}]\) and \([a^{(i)}]\) are the forward coefficient wave matrices at the first and second reference planes respectively, and matrix \([I]\) is the identity matrix.

Now, the EM-fields that are excited by the modes in the second reference plane selectively can be computed with Eq. \[6.7\]. The corresponding termination matrix at the second reference plane is computed as:

\[
[S_{t}^{(i)}] = [S_{j,11}^{(i)}] + [S_{j,12}^{(i)}][S_{t}^{(i-1)}]([I] - [S_{j,22}^{(i)}][S_{t}^{(i-1)}])^{-1}[S_{j,21}^{(i)}] \quad (6.9)
\]

By the termination and junction matrices introduced in the technique described in this chapter, all the subregions can be linked and as a result, the RF-system can be analyzed as a whole, as outlined in Fig. \[6.1\].

### 6.2.3 Validation

For the validation, \(B_{1}^{+}\) maps were measured and simulated in a cylindrical phantom. The cylindrical phantom had a diameter of 20 cm, and a length of 34 cm, see [26] (or chapter [5]). The cylinder is filled with a tissue simulating liquid with the electrical properties of \(\varepsilon_r = 80\), \(\sigma = 0.84\) S/m. The cylinder is placed coaxially in the bore, starting at \(z = 0\) m. The \(B_{1}^{+}\) maps were acquired in a transversal slice. In the simulation the modelling of all subsystems remain the same, as described above, except for the 3D simulation. There, the human is replaced by the cylindrical phantom. The EM-fields were transformed, as if they were excited by the set of modes \(N_1\) at the reference plane at \(z = -1.6\) m selectively. A magnitude least squares fit (MLS), [36] was used to fit the set of simulated fields to the amplitudes of the measured \(B_{1}^{+}\) maps. The periphery of the phantom in the measured \(B_{1}^{+}\) maps shows artifacts, due to a local off resonance induced by the phantom
container susceptibility. Therefore only the inner region of the phantom, up to an inner radius of 8 cm was considered in the fitting.

6.2.4 Safety

For the safety assessment, the worst case SAR [33], normalized to the forward power at the reference plane $z = -1.6$ m, is evaluated. The power flow through a cross-section of the waveguide can be computed with Poynting’s theorem. Writing the EM-field in terms of modes weighted by their forward and backward travelling coefficients, we can find a matrix equation in quadratic form that describes the power flow. The vector of the quadratic form is constructed from the forward and backward coefficients and the elements of the power matrix $[P]$ consist of the inter-modal power flow $p_{ij}$ through the cross-section, [35]. Knowing the relation of the forward and backward wave coefficients, through the termination matrix, we can replace the backward wave coefficient, and find an expression for the power flow in quadratic form, that is only dependent on the forward wave coefficient.

\[
P = \left( \begin{array}{c} a \\ b \end{array} \right)^H [P] \left( \begin{array}{c} a \\ b \end{array} \right) = \left( \begin{array}{c} a \\ [S_t^{(2)}] a \end{array} \right)^H [P] \left( \begin{array}{c} a \\ [S_t^{(2)}] a \end{array} \right)
\]

\[
= a^H \left( \begin{array}{c} I \\ [S_t^{(2)}] \end{array} \right)^H [P] \left( \begin{array}{c} I \\ [S_t^{(2)}] \end{array} \right) a = a^H [P_a] a
\]

The elements of the power matrix $[P]$, $p_{ij}$ are computed as:

\[
p_{ij} = \int e_{t,i} \times h_{t,j}^* dA
\]

In each averaging volume the SAR matrix is computed and transformed to a SAR matrix that is normalized to a forward power of unity at the reference plane $z = -1.6$ m. The highest eigenvalue of the transformed SAR matrix is the highest possible SAR, that is achievable in this averaging volume for a forward power of unity. The whole body, the head SAR, and the 10 g local SAR were evaluated. The 10 g local
SAR was evaluated for every voxel in the human body on a grid of $d/5$, where $d$ is the side length of a water cube with the mass of 10 g. For the whole body SAR the averaging volume is the human body, and for the head SAR the human head. The maximum of all highest eigenvalues, in the different body parts (head, trunk and the extremities, whole body, and whole head) is found, the limiting SAR value according to [37] is used to compute the maximum input power to the TWM-system in the imaging experiment.

### 6.2.5 In-Vivo Imaging

Spoiled gradient echo images of the transversal, sagittal and coronal planes were acquired on a 7T Philips Achieva scanner with a MultiX system with 8 independent transmit/receive channels (Philips Healthcare, Cleveland, OH). In lack of a power monitoring system, the average forward power delivered to the exciting elements was limited by the RF duty cycle of the sequence under the assumption that the RF pulse reached the full peak power for the duration of the excitation pulse. In the first experimental setup, the TWM was used in transceive mode, in the second experimental setup the TWM was used in transmit mode only and a 16 channel array coil, Nova Medical (Wilmington, MA, USA) was used for reception.

### 6.3 Results

The number of modes in the 1.5 m long WGE, whose coupling of the forward wave is more than $-30\text{dB}$ is 19. These modes form the set of modes $N_1$. 16 modes in the empty bore have a forward wave coefficient greater than $-30\text{dB}$ at $z = 0\text{m}$ when the WGE is excited by the set of modes $N_1$ at the reference plane $z = -0.32\text{m}$ with unit amplitude. These circular waveguide modes form the set of modes $N_2$. Two of those modes are propagating, the two degenerate TE$_{11}$ modes, all other modes are evanescent. In the 3D simulation modelling the step discontinuity the mean and maximal errors of the E-field are 2.3% and 8.5%, and of the H-field 2.3% and 12.2%, with respect to the modal
fields, in the reference planes $z = -0.32\text{m}$ and $z = 0\text{m}$, when excited through the WGE and the empty bore.

The field errors are computed as:

$$
\varepsilon = 2 \sqrt{\frac{\int_S |F - F'|^2 dA}{\int_S |F + F'|^2 dA}}
$$

(6.11)

Where $F$ can be the electric field or the magnetic field.

The EM-field in the 3D-simulation including the human converges with respect to the number of tetrahedra. The number of tetrahedra at the final adaptive step was $1.4\text{Mio}$. The expansion of the fields in the three planes from $z = -0.08 \cdots -0.06\text{m}$, with respect to the modal fields yielded a mean and maximal error of the E-fields of $4.4\%$ and $5.7\%$ respectively, and a mean and maximal error of the H-fields of $4.9\%$ and $7.1\%$ respectively, with respect to the simulated fields, when the modes are excited selectively. The error of the power balance was also computed for the simulated fields; the mean value is $5\%$ and the maximal value is $14.8\%$.

The EM-field in the 3D-simulation including the cylinder also converges with respect to the number of tetrahedra. The number of tetrahedra at the final adaptive step was $1.4\text{Mio}$. The expansion of the fields in the three planes from $z = -0.08 \cdots -0.06\text{m}$, with respect to the modal fields yielded a mean and maximal error of the E-fields of $4.4\%$ and $5.7\%$ respectively, and a mean and maximal error of the H-fields of $4.9\%$ and $7.1\%$ respectively, with respect to the simulated fields, when the modes are excited selectively. The power balance was also computed for the set of selectively excited fields; the mean value is $0.2\%$ and the maximal value $0.5\%$.

For the validation setup as well as the in-vivo setup the simulations were joined in a post-processing step, the termination matrix and the fields were transformed to the reference plane at $z = -1.6\text{m}$. The resulting termination scattering matrix $[S^{(2)}_t]$ for the in-vivo case is shown in Fig. 6.3. The corresponding EM-fields in the phantom and the human, that are excited by forward waves with unit amplitude by the set of modes $N_1$ at the reference plane $z = -1.6\text{m}$ of the WGE selectively are used for the following validation and safety assessment.
The validation of the simulation is shown in Fig. 6.4. The magnitude of the $B_{1}^{+}$ are shown, the right column shows the measured field, when excited by the coaxial cable ports with 1 kW input power, the left column shows the fitted simulated fields with 1 kW input power at the reference plane $z = -1.6 \text{ m}$.

The limiting worst case SAR was found to be 0.32 W/kg and is found in the neck yielding a power limit of 31 W, in normal operating mode according to [37].

This power limit allowed a sequence timing of 0.5 ms for the main excitation lobe and a minimal TR of 150 ms with maximum duration of 6 minutes. Examples of RF shimmed images of several planes are shown in Fig. 6.5 and 6.6. For the three images in Fig. 6.5 the travelling wave system was used in transceive mode. For the image in Fig. 6.6 the travelling wave system was used in transmit mode only and a 16 channel array coil, Nova Medical (Wilmington, MA, USA) was used for reception.

*Fig. 6.3: Termination S-matrix $[S^{(2)}_t]$ for the in-vivo case, absolute value in linear scale.*
Fig. 6.4: Validation of the simulation, $|B_1^+|$ in $\mu T$. Right column, measured data normalized to 1kW input power. Right column simulated data, normalized to 1kW input power at the reference plane $z = -1.3 \text{ m}$. 

Results
Fig. 6.5: In-vivo MR images of three slices, with the travelling wave multiple channel system in transceive mode.
6.4 Discussion & Conclusion

First in-vivo images using the TWM-setup are presented in this work. RF-shimming capabilities of the TWM were demonstrated allowing in-vivo images with a homogeneous transmit profile and a large FOV. Coverage of the whole brain was achieved including the cerebellum and the brainstem, unlike close coupling head coil arrays. An increase in SNR was demonstrated by using a local head coil array for reception.

For a safe operation of the TWM a rigorous, robust, and efficient modelling technique based on the mode matching method is introduced. The RF-setup is divided into smaller sub-domains that are added in a post-processing step. The sub-domains can be represented by an N-port network, whose ports are modes of the involved waveguides. The scattering matrices describing the N-ports can have unusual properties, as compared to conventional scattering matrices. The entries can have a magnitude bigger than one, even for a passive N-port, due to the evanescent nature of the involved modes. Errors with respect to physically behaving fields are given, assessing the quality of the EM-simulations.

The simulation was validated in a reference scenario, with a head sized phantom filled with saline. The fields in the simulation overestimate the measurements by a factor of at least 10%, yielding a con-
servative safety assessment. The overestimation can be explained by additional losses, in the excitation section, the coaxial cables, the TR-switches, and standing waves in the WGE, that are not included in the simulation.

A conventional close coupling coil array outperforms the TWM in terms of power efficiency, due to a higher radiation loss and losses in the WGE. On the other hand the SAR efficiency (the ratio of the average $B_1^+$ over the entire brain and the maximal local SAR) are comparable, $0.46 \mu T/\sqrt{W/kg}$ for the TWM and $0.71 \mu T/\sqrt{W/kg}$ for a dedicated head coil array [38]. The TWM exhibits a more homogeneous transmit profile. More RF-power or cooling the water stubs in the WGE, and the walls of the scanner room would solve this issue.

The excitation section was not modelled in this paper, which saves modelling steps and yields a more conservative safety margin. The safety is based on the worst case SAR, which is inherently independent on the driving voltage vector. It depends only on the total input power. By neglecting the excitation section in the worst case SAR analysis, the derived power limit is independent of the positioning and number of the excitation elements in the excitation section. The exact positioning of the excitation elements is one of the main uncertainties of the TWM and is a critical factor for the coupling of the coaxial cable modes into the modes of the WGE. A further uncertainty is the inter-subject variability [39], which was not assessed in this work. Without the inclusion of the excitation section, the imaging performance cannot be assessed.

This modelling of the excitation section presents the remaining challenge of this work. If that is solved, it can be easily included in terms of an additional junction scattering matrix to the presented framework. Following matching networks and other RF-circuitry can be added in the same way. This would allow us to predict the imaging performance, and a less conservative power limit would allow us to reduce the safety margin, and have more power available for the imaging experiment. Measuring the pulse in the imaging experiment would allow us to predict the local SAR exactly and have even more power available for the imaging sequence. But the safety would be valid only
for a fixed positioning of the excitation elements and a fixed voltage vector applied to the coaxial cable ports.
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List of Publications

Journal Papers


Conference Proceedings


Curriculum Vitae

Personal Information

Name Jan Paška
Date of Birth 8th April 1981
Place of Birth Zlín, Czechoslovakia

Education

2014 PhD, ETH Zurich, Supervisor: Klaas Prüssmann
2007 Master of Science in Electrical Engineering, Technical University of Hamburg-Harburg, Germany, Specialized in RF Engineering.
Master Thesis: Transceive Hybrid Coil Array for High Field MRI
Student Project: Double Tuned Birdcage for Flourine and Proton MR imaging

2004-2005 Student Exchange, The University of California, Berkeley, USA
2001 Secondary School, Technisches Gymnasium an der G16, Hamburg, Germany, Specialized in Fine Mechanics
Work Experience

2007–2014  Research Associate, ETH Zurich
2006–2007  Research Assistant, Philips Research Hamburg, Germany
2003–2004  Research Assistant, Institute of Telecommunications at the TUHH, Hamburg, Germany

Skills and Interests

Languages
German, English, Czech

Software
EM-solvers: HFSS, SEMCAD, COMSOL, CST, FEKO
Circuit Design: ADS, Altium
Programming: Matlab, Python
Editing: Word, Powerpoint, Latex, Vim
Operating Systems: Linux, Windows

Machining
Milling, Drilling, Turning, Technical Drawing

Interests
Chess, Cooking, Dancing
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