

NOVEL TECHNIQUES AND INNOVATIVE DESIGNS FOR THE RF CHAIN OF MAGNETIC RESONANCE IMAGING SCANNERS

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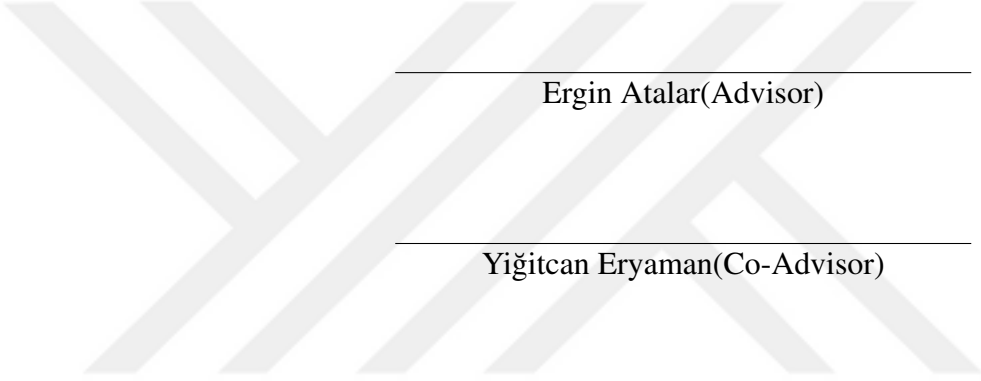
By
Alireza Sadeghi Tarakameh
July 2020

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We certify that we have read this dissertation and that in our opinion it is fully adequate,
in scope and in quality, as a dissertation for the degree of Doctor of Philosophy.



Ergin Atalar(Advisor)

Yiğitcan Eryaman(Co-Advisor)

Vakur Behçet Ertürk

Emine Ülkü Sarıtaş Çukur

Özlem Aydın Çivi

Özgür Salih Ergül

Approved for the Graduate School of Engineering and Science:

Ezhan Karaşan
Director of the Graduate School

ABSTRACT

NOVEL TECHNIQUES AND INNOVATIVE DESIGNS FOR THE RF CHAIN OF MAGNETIC RESONANCE IMAGING SCANNERS

Alireza Sadeghi Tarakameh

Ph.D. in Electrical and Electronics Engineering

Advisor: Ergin Atalar

Co-Advisor: Yiğitcan Eryaman

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In this dissertation, novel techniques and innovative designs are proposed to serve one of the main goals of MRI, which is achieving the best image quality by improving signal-to-noise ratio (SNR) while confined to the patient's safety restrictions determined by specific absorption rate (SAR) limits. This dissertation comprises five different contributions to the field. First, the co-simulation method is accelerated using the equivalent circuit model of the intended coil and introduced as a compact novel technique for the fast design of transmit array (TxArray) coils. It is shown that the existing co-simulation method can be accelerated by more than 20-fold for different types of TxArray coils. In another study, a simple strategy is introduced for optimizing the topography of transmit elements to reduce the local SAR at ultra-high field (UHF, $B_0 \geq 7$ T) MRI. It is shown that the peak local 10g-averaged SAR can be reduced more than 20% for different types of transmit elements at two different field strengths (i.e., 7 T and 10.5 T). As another contribution, confining to the safety limits, the first in vivo human head images at 10.5 T is acquired. To ensure the safety of subjects, the EM simulation model of the radiofrequency (RF) coil is validated using an expanded validation workflow, then, safe RF power levels were calculated. In another study, a novel short dipole-like element with improved SAR performance, non-uniform dielectric substrate (NODES) antenna, is introduced which provides an opportunity to design highly-dense TxArray coils for UHF-MRI. Eventually, a nine-channel transmit/receive coil is built and human cadaver spine images are acquired at 10.5 T. In the final study, a new parameter, ultimate intrinsic SAR efficiency (UISARE), is introduced and calculated as the upper bound for SAR performance of the transmit coils at different field strengths. Overall, this dissertation proposes novel techniques and innovative structures for designing the TxArray coils, which mainly contribute to enhancing image quality and patient safety by improving SNR and reducing SAR.

Keywords: Transmit array, Co-simulation, Ultra-high field, RF safety.

ÖZET

MANYETİK REZONANS GÖRÜNTÜLEME CİHAZLARININ RF TARAFI İÇİN YENİ TEKNİKLER VE YARATICI TASARIMLAR

Alireza Sadeghi Tarakameh

Elektrik ve Elektronik Mühendisliği, Doktora

Tez Danışmanı: Ergin Atalar

İkinci Tez Danışmanı: Yiğitcan Eryaman

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Bu tez’de manyetik rezonans görüntüleme (MRG)’nin en önemli amaçlarından biri olan en kaliteli görüntüyü sinyal-gürültü oranı (SGO)’nı artırarak ve özel soğrulma oranı (ÖSO) ile belirlenen hasta güvenliği kısıtlamalarına uyarak elede etmeye hizmet etmek için yeni teknikler ve yaratıcı tasarımlar önerilmektedir. Bu doktora tezi alana beş farklı katkı sunmaktadır. İlk olarak, eşzamanlı simülasyon yöntemi istenilen sargının eşdeğer devre modelini kullanılarak hızlandırılmış ve hızlı sargı dizisi tasarımı için yeni teknik olarak tanıtılmıştır. Varan olan eşzamanlı simülasyon yöntemi farklı sargı dizileri için 20 katdan fazla hızlanılabilirliği gösterilmektedir. Başka bir çalışmada aşırı yüksek alan (AYA, $B_0 \geq 7T$) MRG’de bölgesel ÖSO’yü azaltmak amaçlı verici elemanların yapısını optimize etmek için yeni bir strateji önerilmektedir. İki farklı alan gücünde (7T ve 10.5T) farklı verici elemanlar için bölgesel ÖSO’nun %20 azalabildiği gösterilmektedir. Bir başka katkı olarak hasta güvenliği kısıtlamalarına uyarak ilk canlı insan kafa görüntüsü 10.5T’da alınmıştır. İnsan deneklerinin güvenliğini sağlamak amaçlı bir genişletilmiş doğrulma iş-akışı kullanılarak radyofrekans (RF) sargısının elektromanyetik (EM) modeli doğrulanmış ve güvenli RF güc seviyesi hesaplanmıştır. Bir başka çalışmada ÖSO performansı iyileştirilmiş NODES anten isimli yeni bir dipol benzeri eleman önerilmiştir. Bahsi geçen eleman AYA-MRG’de çok sayıda elemana sahip olan verici sargı dizisi tasarlamak için bir fırsat sunmaktadır. NODES elemanlarını kullanarak dokuz kanallı bir alıcı/verici sargı inşa edilmiş ve insan cadavrasının omurga görüntüsü 10.5’da çekilmiştir. Son çalışmada verici sargıların ÖSO performansının üst sınırını belirlemek için nihai içsel ÖSO verimliliği (NİÖSOV) yeni bir parametre olarak tanımlanmış ve hesaplanmıştır. Genel olarak bu doktora tezi verici sargıların tasarımı için yeni teknikler ve yaratıcı yapılar sunmaktadır, bu da esasen görüntü kalitesi ve hasta güvenliğini SGO’yu artırarak ve ÖSO’yü azaltarak iyileştirmeye katkı sağlamaktadır.

Anahtar sözcükler: Verici sargı dizisi, Eşzamanlı simülasyon, Aşırı yüksek alan, RF güvenliği.



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Chapter 1

Introduction

Magnetic resonance imaging (MRI) is one of the safest widely-used medical imaging techniques. This nonradiative imaging technique provides an excellent soft-tissue contrast, which makes it a good candidate for most of the medical imaging applications. However, there are still some technical demands and safety issues that need further investigation. In MRI, the goal is to achieve the best image quality in the shortest time, confining to the patient's safety restrictions.

One of the main safety concerns in MRI is the radiofrequency (RF) power absorption in the subject that causes heating in tissues, which must be kept under a safety limit [1]. However, restricting the specific absorption rate (SAR) may increase imaging duration and also decrease the excitation homogeneity [2–4] which causes degradation in the image quality. This issue becomes more severe at higher field strength.

Ultra-high field (UHF, $B_0 \geq 7$ T) MRI is one of the main candidates to increase the image quality due to its numerous benefits, such as a significant increase in signal-to-noise ratio (SNR) [5–10] and improved susceptibility contrast [11, 12]. However, with an increasing static magnetic field (B_0), EM wavelengths decrease resulting in destructive and constructive interferences of RF magnetic fields, causing degradation in overall image quality. In addition, constructive interference of the RF electric field

can lead to elevated local SAR hot spots, which are an important patient safety concerns at UHF. Nevertheless, it is shown that multi-channel TxArray with improved RF performance, besides other technologies [13, 14], can be used to alleviate both issues when optimal excitation strategies are utilized [4, 15, 16].

TxArray technology [17, 18] provides higher controllability on the RF electromagnetic (EM) fields and thereby on excitation quality and SAR limits. However, it is shown that there are trade-offs between excitation accuracy, local and global SAR, and maximum and average power for parallel transmit pulses [3]. Furthermore, TxArray technology, besides the other approaches [19–35], presents possible solutions for the safety of patients with metallic implants [36–38]. However, the commonly used method [39] for the design of multi-channel TxArray coils is a time-consuming procedure that consists of a full EM simulation of the coil and an optimization process over the lumped elements of the coil. This issue can be addressed by exploiting fast EM solvers [40–45] and/or accelerating the optimization process.

On the other hand, TxArray coils composed of new transmit elements with improved SAR performance may be beneficial for mitigating the SAR issue in UHF-MRI. In the literature, various types of transmitting elements such as loop coils [46–51], transmission lines [10, 15, 52–54], and dipole-like structures [55–60] are investigated for UHF-MRI. Thalhammer et al. [46] used TxArray of loop coils for cardiac imaging at 7 T. Adriany et al. [15] employed an array of the transmission lines for head imaging at 7 T. Vaughan et al. [53] built a transverse electromagnetic (TEM) coil for whole-body imaging at 7 T. Raaijmaker et al. [55, 56] introduced single-side adapted and fractionated dipole as new transmitter elements for body imaging at 7 T. Steensma et al. [57, 58] improved the SAR performance of the dipoles at 7 T and 10.5 T by introducing the snake antenna. Vaughan et al. [10] utilized a TEM coil for the head imaging at 9.4 T. Shajan et al. [48] performed head imaging at 9.4 T using the array of loop elements. Erturk et al. [59] proposed to use an array of fractionated dipoles for torso imaging at 10.5 T. Despite all previously proposed designs, improving the performance of RF coil arrays is still required for realizing the full potential of UHF-MRI. Considerable improvement may be achieved by designing and investigating novel and

innovative coils [51, 61–68]. However, determining an upper limit for such improvements is extremely solicited. Such an upper limit will be a proper criterion for evaluating the performance of designed coils.

Correspondingly, this dissertation introduces a novel technique for the design and implementation of TxArray coils. Furthermore, some innovative designs are presented for reducing the local SAR and increasing the image quality at UHF-MRI. Finally, an upper bound for the RF excitations at different field strengths is calculated while the local SAR is restricted to the safety limits.

In Chapter 2, we propose a technique using the equivalent circuit model of the intended coil in order to accelerate the co-simulation method [39] for designing the TxArray coils. In Chapter 3, we introduce a simple strategy (i.e., putting a bump underneath the discontinuities along the coil) to reduce the local SAR at UHF. In Chapter 4, using the proposed design in Chapter 3 (bumped fractionated dipole), we acquired the first human head MR image at 10.5 T. In Chapter 5, we optimized a dipole design to achieve the maximum SAR performance. Consequently, we introduced a new transmit element (i.e., NODES antenna) and acquired cadaver spine images at 10.5 T. In Chapter 6, we introduced a new concept, ultimate intrinsic SAR efficiency (UISARE), and calculated the upper bound for local SAR performance of a transmitter. These novel techniques and innovative structures for designing the TxArray coils mainly contribute to enhancing image quality and patient safety by improving SNR and reducing SAR.

Chapter 2

Accelerating the Co-Simulation Method for the Design of Transmit Array Coils

Preface

The content of this chapter was presented in part at the 26th Annual Scientific Meeting of International Society of Magnetic Resonance in Medicine (ISMRM) [69] and it was published in Magnetic Resonance Materials in Physics, Biology and Medicine [70]. The text and the figures of this chapter are based on the journal publication [70]. Ehsan Kazemivalipour, Umut Gündoğdu, Serhat Erdoğan, Ergin Atalar contributed to this study. Ergin Atalar was involved in study conception and design, Ehsan Kazemivalipour contributed to the EM simulations and design. Umut Gündoğdu advised and developed methods for the data acquisition (i.e., GRE images and B_1^+ -maps). Serhat Erdoğan developed methods to analyze the data and was involved in seeding the initial values for the optimization problem.

2.1 Introduction

In this chapter, we introduce a fast method to find the optimum capacitor values for a transmit array (TxArray) coil.

There exist many benefits of using radiofrequency (RF) transmit array coils in magnetic resonance imaging (MRI). One common usage of this technology is for RF-shimming [71–74]. For the static magnetic field strength of 3 T or higher, the size of the human body becomes comparable to the wavelength at the Larmor frequency, and homogenous excitation cannot be achieved with the conventional birdcage body coil. Moreover, the B_1^+ field is a function of patient position and size. An array of transmitters can be used to overcome these difficulties. Furthermore, in many RF-shimming applications, the specific absorption rate (SAR) becomes a limiting constraint that can be taken under control by using TxArray coils [2, 3, 75–81]. Another possible use of the TxArray systems is to reduce the duration of the RF pulses designed for complex excitation patterns [17, 18]. As an example, the spatially selective excitation is used for limiting the size of the signal-contributing volume based on the region-of-interest (ROI), which is advantageous for fast data acquisition [82, 83]. Another potential use for the TxArray systems is to achieve implant-friendly MRI scanning [19, 20, 22, 23, 36].

TxArray coils are studied in many forms, including decoupled strip-lines [15, 84], decoupled surface coils [85, 86], dipole-like structures [55–57, 59, 60, 87, 88] and degenerate birdcage coils (DBC) [69, 89–92]. Despite the advantages of using TxArray coils for MRI, the design and manufacturing of such coils is a significant challenge due to the mutual coupling between elements of the array. One of the practical techniques described in the literature is constructing and tuning only one of the elements, initially, then assembling the remaining elements of the array in successive steps, followed by iteratively tuning and decoupling all elements at each step [89]. Unfortunately, this technique is very time-consuming, making it infeasible for a large number of elements.

The co-simulation method, which was proposed by Kozlov et al. [39], significantly reduced the difficulties of RF coil design methods. In this method, an electromagnetic

(EM) simulation is performed by replacing all discrete elements (capacitors and inductors) with excitation ports. The calculated scattering matrix (S matrix) is exported to a circuit simulator in order to optimize the values of these elements. Using this method, the values necessary for tuning and decoupling can be accurately obtained. Additionally, it makes large problems (of many independent discrete elements) feasible to solve. However, in the optimization process of this method, the initial guess plays a significant role. In the literature [93, 94], many randomly chosen initial guesses were established to avoid converging to a local minimum, which makes the optimization process time-consuming. On the other hand, there is no guarantee that the optimization will converge to the global minimum using the randomly chosen initial guesses. A pre-calculated initial guess was considered in an earlier study [2], but this strategy was not fully analyzed and did not yield the optimum solution in all cases that were tested.

In this chapter, we use analytic calculations and finite element method (FEM)–based simulations to obtain all inductive and resistive parameters necessary for the organization of the equivalent circuit model. Then, the circuit model is utilized in determining approximate values for the capacitors used in the design. These approximate values of capacitors are used as the initial values for the optimization process of the co-simulation method. To prove the effectiveness of the method, six different TxArray coils with two different loading schemes were investigated. The results of the proposed method are compared in terms of accuracy and speed to the results of the conventional co-simulation method with many randomly chosen initial guesses.

2.2 Theory and Methods

Here, the proposed method will be explained in detail for an N-channel shielded DBC shown in Figure 2.1, while, the method can be easily extended for the other types of coils as well. In this design, there are three independent capacitor values: C_t is the tuning capacitor, C_d is the decoupling capacitor, and C_m is the matching capacitor when the load is a cylindrical uniform phantom.

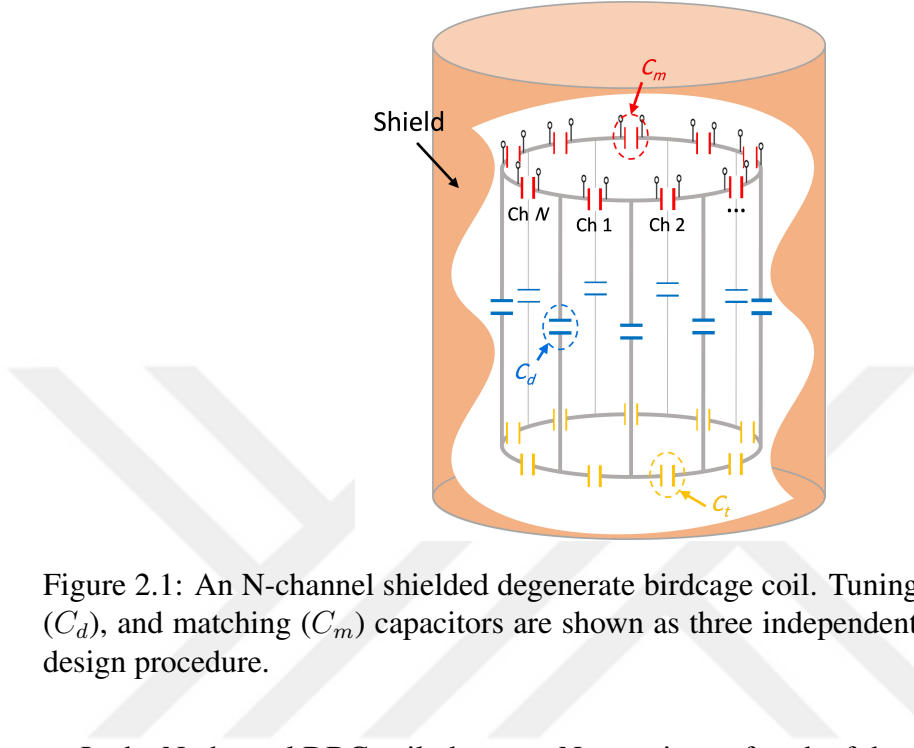


Figure 2.1: An N-channel shielded degenerate birdcage coil. Tuning (C_t), decoupling (C_d), and matching (C_m) capacitors are shown as three independent capacitors in the design procedure.

In the N-channel DBC coil, there are N capacitors of each of the three types. Some capacitors, such as C_t , can be distributed. The following theory aims to find an approximate value for each of these capacitors and use them as initial guesses in the optimization process of the co-simulation technique. In the other designs or when the load is not circularly symmetric as in the human head/body model, the numbers of independent capacitor values can be significantly higher.

The flowchart of the proposed method is shown in Fig. 2.2. In this method, we obtain a coarse estimate for the decoupling and tuning capacitor values, based on the ideal decoupling between adjacent channels and assuming no loss in the system. Later, these values are used as the initial guesses for the optimization of a simplified circuit model of the array coil. The results of this intermediate optimization process are called the fine estimates. We used these as the initial guesses for the co-simulation method (see Fig. 2.2). These steps will be described below in detail.

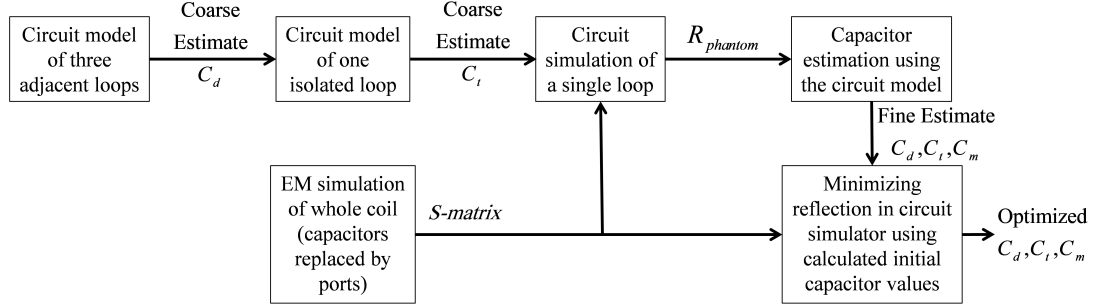


Figure 2.2: The workflow of the proposed method.

2.2.1 Equivalent Circuit Model

First, we start with the equivalent circuit model of the coil. By modeling each wire or strip as an inductor—an assumption that is only valid if the wavelength is much larger than the strip segments—a shielded DBC can be modeled as an equivalent circuit that consists of self- and mutual-inductances. Assuming an infinitely long shield, the shield acts as an electromagnetic mirror. Therefore, the shielded DBC can be analyzed using the mirror currents. Finding the position of these mirror currents in a cylindrical coordinate system is a well-known procedure [95]. We designated the current segments on the coil 1 to 3, and the mirror currents 4 to 6, as shown in Fig. 2.3a. Let $L_{j,p,k,q}$ represent the self- or mutual-inductance of the p^{th} segment in the j^{th} loop with the q^{th} segment in the k^{th} loop. Calculation of each of the inductance values can be found in the existing literature [96, 97]. Kirchhoff’s voltage law for the j^{th} loop can be written in the form of a matrix equation as follows [98]:

$$\bar{\bar{K}} \cdot \bar{I} = \frac{1}{\omega^2} \bar{\bar{H}} \cdot \bar{I} \quad (2.1)$$

where $\bar{\bar{I}}$ is a vector of mesh currents, as shown in Fig. 2.3b. This vector does not include the mirror currents since, by their definition, their values are identical to those of flow on the coil conductors. Therefore, the elements of matrices $\bar{\bar{K}}$ and $\bar{\bar{H}}$ can be

written as [98]:

$$K_{j,k} = 2(L_{j,1,k,1} - L_{j,1,k,3} - L_{j,1,k,4} + L_{j,1,k,6}) + \sum_{m=0}^1 \sum_{n=0}^1 (-1)^{m+n} [L_{j+m,2,k+n,2} - L_{j+m,2,k+n,5}] \quad (2.2)$$

$$H_{j,k} = 2 \left(\frac{1}{C_t} + \frac{1}{C_d} \right) \delta_{j,k} - \left(\frac{\delta_{j,k+1}}{C_d} + \frac{\delta_{j,k-1}}{C_d} \right) \quad (2.3)$$

where $\delta_{j,k}$ is the Kronecker delta defined as:

$$\delta_{j,k} = \begin{cases} 1 & j = k \\ 0 & j \neq k \end{cases} \quad (2.4)$$

2.2.2 Circuit Model of Three Adjacent Loops

In the design of a DBC, decoupling between the channels, tuning to the desired frequency, and impedance matching of the input ports can be considered the essential parameters.

To find the decoupling capacitor value, we ignored the power loss of the phantom and assumed that the coupling between nonadjacent loops (channels) is negligible. Consequently, the coupling between adjacent loops with a common rung was canceled using an appropriate capacitor on the common rung. Fig. 2.3b shows the equivalent circuit model of three adjacent loops, such that the loop in the middle (j^{th} loop) is supposed to be decoupled from the $(j-1)^{th}$ and $(j+1)^{th}$ loops. Therefore, one can obtain a coarse estimate for the decoupling capacitance as shown below.

$$C_d = -\frac{1}{\omega^2 K_{j,j-1}} \quad (2.5)$$

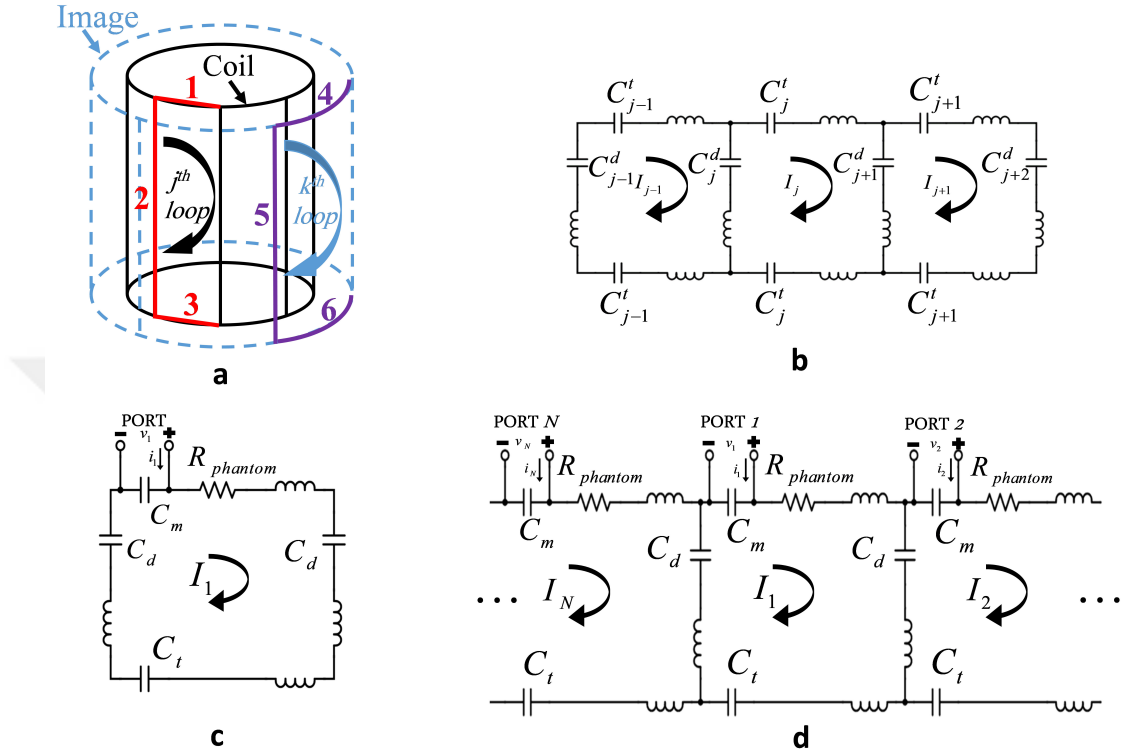


Figure 2.3: (a) All combinations of self- and mutual-inductances in the DBC including the image of the coil corresponding to the shield. (b) The circuit model of three adjacent loops in a DBC was investigated for decoupling and tuning purposes. (c) Circuit model of a single loop, which was used to determine the loading effect on the coil. (d) The final equivalent circuit model of the DBC including the loading effect of the phantom.

2.2.3 Circuit Model of One Isolated Loop

To find the tuning capacitor values, we assume perfect decoupling between all channels, no loss (as before), and that the ports are disconnected. Under these assumptions, the array structure can be treated as N separated single loop coils. The Kirchhoff's voltage law equation for this single loop can simply be written as:

$$j\omega \left[K_{j,j} - \frac{2}{\omega^2} \left(\frac{1}{C_t} + \frac{1}{C_d} \right) \right] I_j = 0 \quad (2.6)$$

Substituting the value of C_d from Eq. 2.5 into Eq. 2.6 and solving the equation for C_t provides a coarse estimate for the tuning capacitance as follows:

$$C_t = \frac{2}{\omega^2} \left(\frac{1}{K_{j,j} + K_{j,j-1} + K_{j+1,j}} \right) \quad (2.7)$$

2.2.4 Electromagnetic (EM) Simulations of the Coil

Similar to the original co-simulation method [39], all of the lumped elements on the coil that are used for tuning, decoupling, and matching purposes were replaced by $50 \, \Omega$ ports in an EM simulation environment. The iterative meshing process of the EM simulation was stopped if the greatest difference between the S-parameters of two successive iterations was less than a pre-determined value. Otherwise, the simulator proceeded with finer meshes [99]. Eventually, the S matrix extracted from the EM simulator was imported to a circuit simulator for performing a single-loop circuit simulation as well as determining an optimum set of capacitors.

2.2.5 Circuit Simulation of a Single Loop

To obtain a finer estimate of the tuning and decoupling capacitance values and determine an estimate for the matching capacitance, the loss (loading effect) is introduced. To predict the loss, a circuit simulation of a single loop was performed by assigning zero-capacitance (i.e., large impedance) to lumped elements of all the loops excluding the intended loop. The values obtained for C_d and C_t in Eq. 2.5 and Eq. 2.7, respectively, were used in the single-loop simulation. Based on the circuit model in Fig. 2.3c, $R_{phantom}$ was formulated as follows:

$$R_{phantom} = \text{real} \left\{ \frac{j}{\omega C_m (j\omega C_m Z_{simulation} - 1)} \right\} \quad (2.8)$$

where $Z_{simulation}$ is the impedance at the port and is determined by the EM simulator.

2.2.6 Circuit Model of All N Loops

Once $R_{phantom}$ is determined, the general equivalent circuit model for the DBC, including the matching capacitor and loading effect of the phantom (Fig. 2.3d), can be utilized for design purposes. The impedance matrix corresponding to the circuit model in Fig. 2.3d can be formulated as follows:

$$\bar{\bar{Z}} = -\frac{j}{\omega C_m} \left[\bar{\bar{I}} + \left(j\omega \bar{\bar{K}} - \frac{j}{\omega} \bar{\bar{H}} + R_{phantom} \bar{\bar{I}} \right)^{-1} \left(\frac{j}{\omega C_m} \bar{\bar{I}} \right) \right] \quad (2.9)$$

where $\bar{\bar{I}}$ is the identity matrix. Then, the scattering matrix of an N-port network, with the corresponding impedance matrix $\bar{\bar{Z}}$, can be determined as follows [47]:

$$\bar{\bar{S}} = \left(\bar{\bar{Z}} + Z_0 \bar{\bar{I}} \right)^{-1} \cdot \left(\bar{\bar{Z}} - Z_0 \bar{\bar{I}} \right) \quad (2.10)$$

where Z_0 represents the characteristic impedance of transmission lines (50 Ω).

Once the S parameters are calculated, the cost function can be evaluated as the total reflected power from the ports while one of the ports is stimulated by unit power:

$$Cost = \sum_{n=1}^N |S_{n1}|^2 \quad (2.11)$$

We used the steepest-descent method [100,101] to minimize the cost function. If the cost difference between two consecutive iterations is less than a predetermined value, the algorithm is terminated. The final capacitor values become the initial values for the co-simulation method.

After performing the EM simulation, the resultant S matrix was exported to the circuit simulator. In the circuit simulator, each port was connected to its corresponding capacitor. A fine estimate for the capacitors was calculated using the equivalent circuit model introduced above. Utilizing the optimization tool of the circuit simulator (which uses the gradient optimization algorithm) with the minimum reflected power constraints and finely estimated capacitor values as the initial guesses, the proper capacitor values were obtained.

2.2.7 Simulations and Experiments

To verify the proposed method, six different TxArray coils with two different loading schemes are designed (See Fig. 2.4) and simulated. One of these designs is constructed and tested with an MRI experiment.

2.2.8 Comparison of the Proposed Method With the Original Co-Simulation Method

An eight-channel head DBC (Coil 1 in Fig. 2.4), a 12-channel head DBC (Coil 2 in Fig. 2.4), a 16-channel cylindrical body DBC (Coil 3 in Fig. 2.4), a 16-channel dual-row head coil (Coil 4 in Fig. 2.4), a 16-channel shifted dual-row head coil (Coil 5 in Fig. 2.4), and a 16-channel elliptical body coil (Coil 6 in Fig. 2.4) were investigated for proof of the concept. Each structure was investigated in the presence of two different loading schemes, cylindrical phantom (first row in Fig. 2.4) and human head/body model (second row in Fig. 2.4), at 3 T field strength. HFSS (ANSYS, Canonsburg, PA, USA) and Microwave Office (AWR Corp., El Segundo, CA, USA) were used as EM and circuit simulators, respectively. All computations and simulations were performed on a workstation with two quad-core Intel(R) Xeon(R) X5472 processors with a 3 GHz clock rate and 32 GB RAM.

2.2.9 Coil Construction

The eight-channel DBC was constructed on a plexiglass cylinder (Fig. 2.5a) with a diameter of 293 mm. The length of rungs was 250 mm, and the width of copper strips was 15 mm. The RF-shield of the coil was built on a larger plexiglass cylinder (380 mm diameter and 500 mm length) (Fig. 2.5b). Gradient-induced eddy currents were reduced by slitting the shield in 16 equally spaced locations along the z-direction and connecting them with 1 nF capacitors [102, 103] at the positions facing the end-rings of the coil. We used eight coaxial cables with adjusted bazooka baluns on each cable to carry the RF power from amplifiers to the coil.

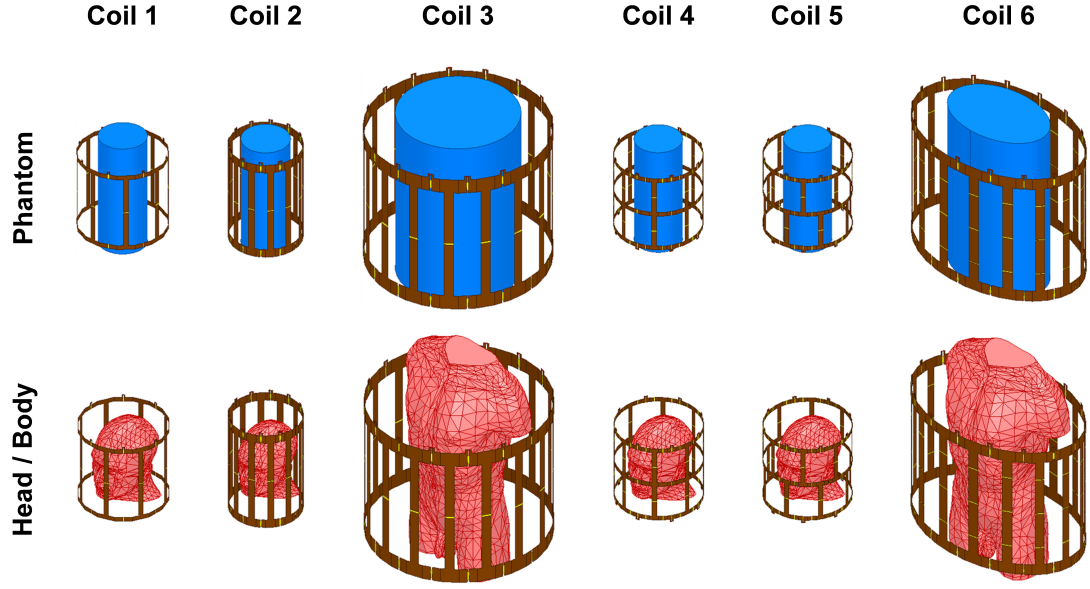


Figure 2.4: Comparison between the modified and original co-simulation method. Coil 1: eight-channel head DBC, Coil 2: 12-channel head DBC, Coil 3: 16-channel body DBC, Coil 4: 16-channel dual-row head coil, Coil 5: 16-channel shifted dual-row head coil, and Coil 6: 16-channel elliptical body coil. All structured were investigated in the presence of a cylindrical phantom (first row) as well as head/body model (second row).

The constructed DBC was also used as a receiver by placing a transmit/receive (T/R) switch between the coil, transmit amplifiers, and receiver system. The T/R switch structure and circuit schematic are pictured in Fig. 2.5c and 2.5d, respectively. There are two series-connected PIN diodes; two DC blocking capacitors; an RF choke (RFC), and a 90° low-pass π -type LC network [104]. The scanner provides DC control signals to change the states of the switch. The DC signals are connected to the board by the RFCs. In the transmit mode, the scanner turns on the PIN diodes with a 100 mA current. The receive port is isolated utilizing the high impedance created by the shorted 90° π -network, which acts like a shorted quarter-wave transmission line. In the receive mode, the PIN diodes are turned off with a -30 V signal so that the receiver port is connected to the coil over the low-pass LC network. The transmitting port is isolated using the high impedance created by the PIN diode, which is in the on-state. The isolation values are approximately 27 dB for both switching states, as determined by means of a bench-top measurement using a network analyzer (Agilent-E5061B, Baltimore, MD, United States).

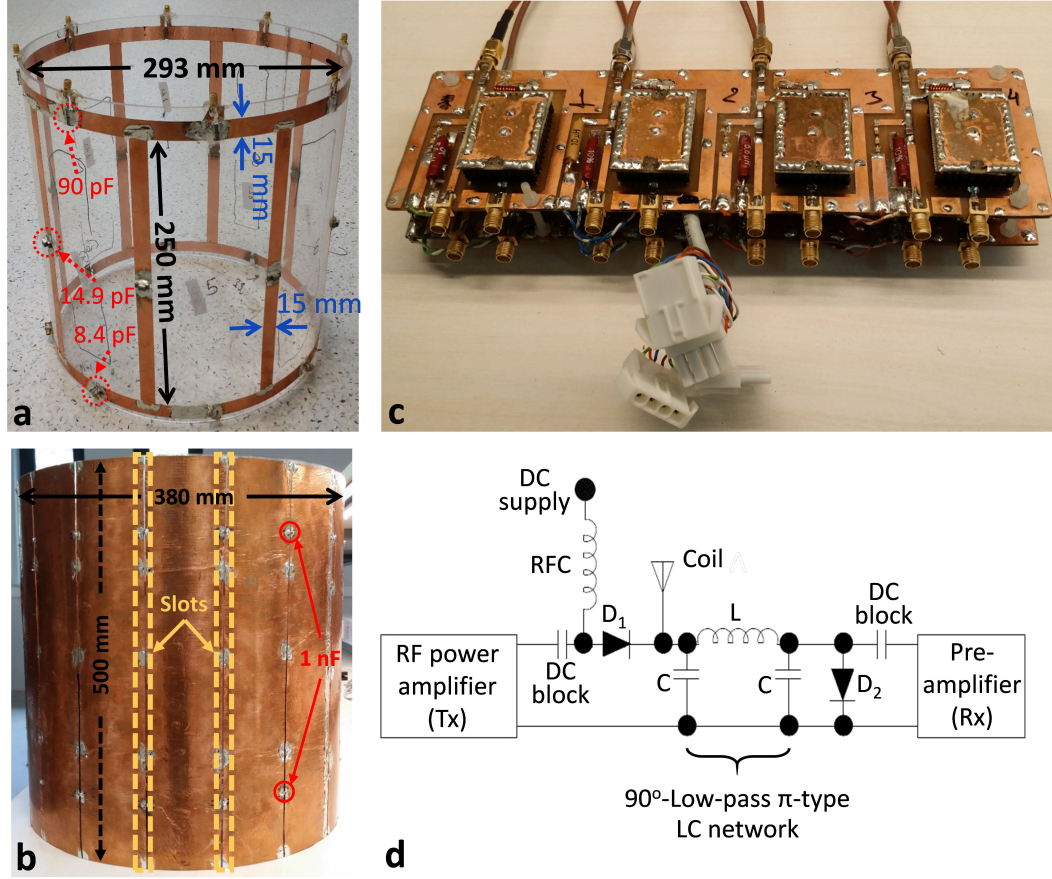


Figure 2.5: Experimental setups. (a) An eight-channel DBC, which was designed and constructed using the proposed method. (b) An RF-shield with several parallel slots in the z-direction on it (for reducing the eddy current effects); (c) The custom T/R switch used in the MRI experiment, (d) and the circuit model of the T/R switch.

2.2.10 System Configurations

The parallel excitation was performed on a 3 T Tim Trio system (Siemens Medical Solutions, Erlangen, Germany) equipped with an 8-channel transmit array system. Each channel used a separate 8 kW RF amplifier (Analogic Corp., Boston, MA, USA). All of the experiments were performed using a 150 mm diameter cylindrical SNR phantom (3.37 g/L $\text{NiCl}_2 \cdot 6\text{H}_2\text{O}$ and 2.4 g/L NaCl), which possessed electrical properties of $\epsilon_r=80$ and $\sigma=0.62$ S/m to mimic the human head. The conductivity of the phantom was measured using a magnetic resonance electrical properties tomography (MREPT) [105, 106] experiment, and its relative permittivity was assumed to be the

same as water.

2.2.11 B_1 Field Mapping and Sequence Parameters

B_1^+ -maps were acquired using the Bloch-Siebert (BS) approach described by Sacolick et al. [107] A gradient-echo (GRE) pulse sequence was modified to apply Bloch-Siebert (BS) shift to spins by an off-resonance Fermi pulse. The duration and the off-resonance frequencies were selected to be 8 ms and 2 kHz, respectively. The other relevant imaging parameters were TR=100 ms, TE=12 ms, NEX=1, 128x128, FOV=200 mm. Each map was acquired individually in transmit mode, i.e., both slice selection and BS pulses were applied to each channel in transmission mode, one by one. To prevent unreliable data (low-SNR) from affecting the B_1 -maps, a mask with a threshold of one-tenth of the maximum B_1 value was implemented on the maps. Afterward, a quadrature interpolation was applied to the maps to smoothen them. All of the channels were used for reception, and a sum-of-squares method was used for image reconstruction. Acquired data were analyzed with the method described by Turk et al. [108]. The B_1^+ relative phase maps were acquired by simply applying a GRE sequence with the same parameters, except for the Fermi pulse. The reference phase image was taken with the same GRE sequence when all channels were transmitting. The B_0 maps were acquired by a built-in field map sequence using the same imaging parameters, except with a TE₁ of 5 ms and TE₂ of 12.46 ms.

2.3 Results

The performance improvement in the proposed modification to the co-simulation technique is demonstrated by comparing the original method to the modified technique.

2.3.1 Comparison of the Proposed Method With the Original Co-Simulation Method

As a reminder, in the original co-simulation technique, the capacitor values are found using an optimization technique, and the cost function (the total reflected power when unit input is provided from one of the channels of the DBC) is a nonlinear function of the capacitor values and has many local minima. Because of this difficulty, the optimization problem is solved by repeatedly using the steepest-descent method with different initial guesses and reporting the minimum of each solution as the global minimum, as described in the literature [93, 94].

In our proposed method, to find the capacitor values of the DBC, the equivalent circuit model was used as described in the previous section. In the case of the eight-channel head DBC shown in Fig. 2.5a, analytic calculations led to coarse estimates of 12.9 pF and 24.5 pF for C_d and C_t , respectively. Then, an EM simulation of a single loop in the presence of the phantom was performed, and $R_{phantom}$ was determined to be 3.8Ω . Afterward, an interstage optimization was performed to obtain a fine estimate for the three capacitor values. At this stage, C_d , C_t , and C_m were computed and found to be 13.9 pF, 8.8 pF, and 89.5 pF, respectively. In the next step, all of the capacitors on the coil were replaced by 50Ω lumped ports in the simulation environment (Fig. 2.6a). The EM simulation was performed using the iterative meshing technique with 2.5×10^{-3} stop criterion (i.e., the difference between the S-parameters of two successive iterations) which led to 184000 tetrahedral meshes in the last meshing step. After performing the EM simulation, the resultant S matrix was exported to the circuit simulator. In the circuit simulator, each port was connected to the corresponding capacitor. Utilizing the optimization tool of the circuit simulator, which uses the gradient optimization algorithm, with the minimum reflected power constraints, the proper capacitor values were obtained. The fine estimates for the capacitor values from the previous stage were used as the initial guesses for the optimization tool at the current stage. The optimized capacitor values were 14.9 pF, 8.4 pF, and 90 pF for C_d , C_t , and C_m , respectively. Fig. 2.6b shows the S-parameters obtained from the circuit simulator.

To compare the efficiency of the proposed method to the original co-simulation

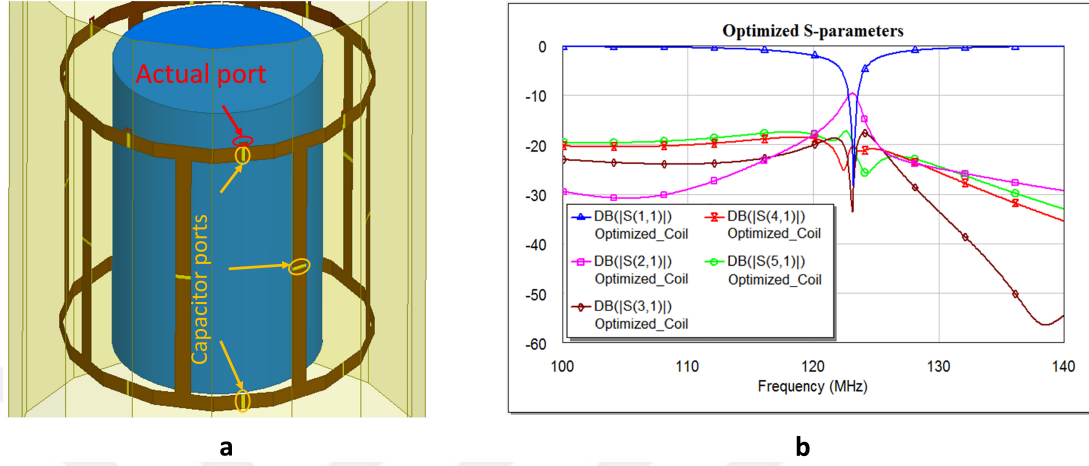


Figure 2.6: The procedure of the modified co-simulation method. (a) The EM simulation model of the DBC, including the phantom, used to extract the S-parameters when all capacitors were replaced with $50\ \Omega$ ports. (b) The optimized S-parameters using the circuit simulator with the constraint of minimum reflected power.

method, we repeated the original co-simulation design process as described in the literature [39,93,94].

All random values were chosen using the rand command in MATLAB (The Math-Works, Natick, MA, USA). By randomly choosing 1000 initial guesses (triple-sets of capacitor values in the range of 1 pF to 100 pF), the optimization tool converged to various final cost values. Fig. 2.7a shows the converged cost values for different initial guesses. All circle markers correspond to the randomly chosen initial guesses, and the asterisk marker represents the cost value that was determined by considering the fine estimate for the initial capacitor values. This shows that finding the global minimum using randomly chosen initial guesses is not guaranteed, and it was achieved only three times by seeding with 1000 initial points. On the other hand, the global minimum was found in a single shot when the calculated initial values for the capacitors were used. The total reflected power from all ports for a given global minimum cost value was found by both methods to be 0.25.

The duration of the optimization becomes an important factor for large problems. Fig. 2.7b demonstrates the number of iterations required for different initial guesses. The total number of iterations for 1000 randomly chosen initial guesses was over 794000. However, the use of the calculated initial value provides the same results

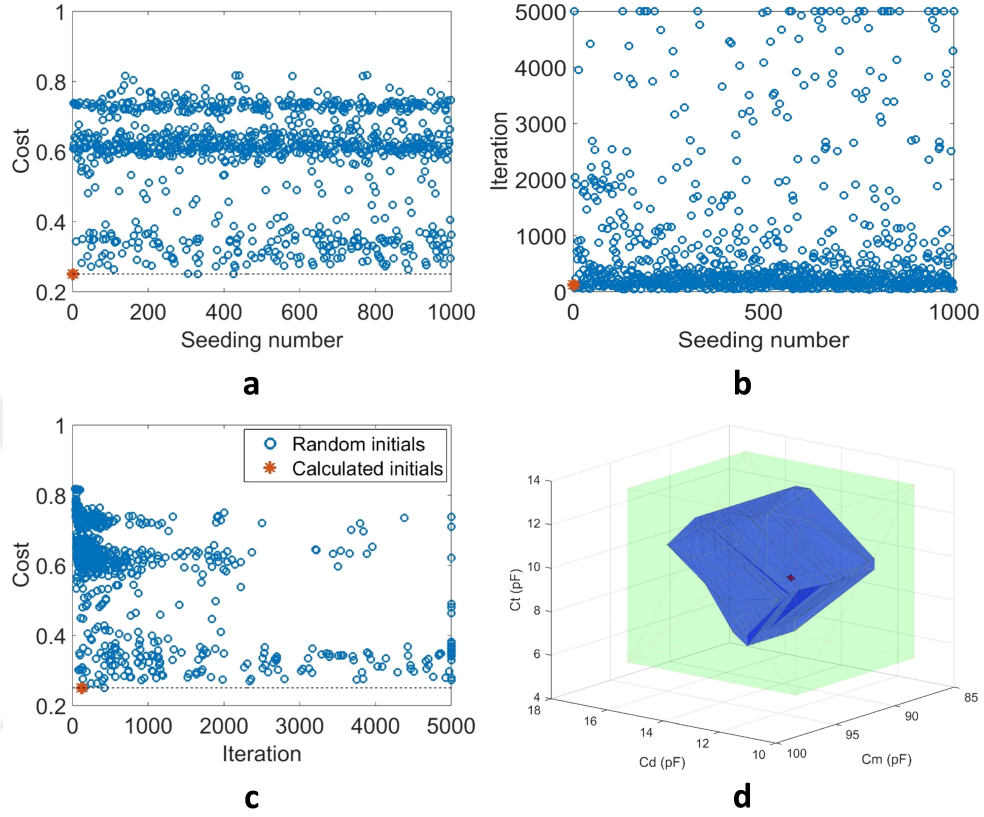


Figure 2.7: Analytically calculated initial value vs. randomly chosen initial values. (a) Final cost values achieved using various initial values in the optimization tool. (b) The number of iterations needed for the optimization tool to converge using various initial values. (c) The final cost values with respect to the numbers of iterations for different initial values. (d) The reliable initial guesses in the vicinity of the desired capacitor values that promised to converge to the global minimum.

in 123 iterations. Thus, in this case study, the proposed algorithm speeds up the optimization process by more than 6400 times.

In Fig. 2.7c, the final cost values for various initial guesses were plotted with respect to the number of iterations. The cross marker, corresponding to the calculated initial values, provides the minimum cost value achieved using the fewest iterations.

To prove the concept, six different structures, shown in Fig. 2.4, with two different loading schemes, were investigated. The duration for each stage of the design for both methods is given in Table 2.1. Modified and original co-simulation methods were abbreviated as “M” and “O” in Table 2.1, respectively. Coils 1 through 6 refer to the

structures shown in Fig. 2.4. The number of independent capacitors represents the number of independent variables that were optimized to achieve the minimum cost function.

In all cases, the number of initial capacitor values (seeds) was chosen to be 1000 to ensure that the global minimum would be achieved. This number is comparable to that used by Kozlov et al. [93,94] (3000 seeds). For each of the initial capacitor values (seeds), the gradient optimization tool, with termination constraints of a maximum of 5000 iterations and a convergence tolerance of 10^{-4} , was utilized.

As we argued, if the initial guess in a complex optimization problem is close to the global minimum, the problem can be solved using a simple optimization algorithm. In the case of the eight-channel DBC, to determine how close is close enough, the initial guesses for the three capacitor values are scanned between 4 pF to 14 pF for C_t , 85 pF to 100 pF for C_m , and 10 pF to 18 pF for C_d . For this example-case, the initial capacitor values that fall into the volume shown in Fig. 2.7d, the minimum value was obtained using the steepest descent algorithm is the global minimum. The red diamond shown in the figure represents the fine estimate

			Number of independent capacitors	Analytic calculations	EM simulation of the whole coil	Circuit simulation of a single loop	Interstage optimization	Final optimization	Total
Coil 1	Phantom	O	3	-	40 min	-	-	13.5 hrs.	≈ 14 hrs.
		M	3	< 1 sec	40 min	< 1 sec	15 sec	15 sec	≈ 40 min
	Head	O	17	-	35 min	-	-	46 hrs.	≈ 47 hrs.
		M	17	< 1 sec	35 min	4 sec	90 sec	200 sec	≈ 40 min
Coil 2	Phantom	O	3	-	80 min	-	-	18.5 hrs.	≈ 20 hrs.
		M	3	< 1 sec	80 min	< 1 sec	20 sec	25 sec	≈ 1.5 hrs.
	Head	O	25	-	115 min	-	-	53 hrs.	≈ 55 hrs.
		M	25	< 1 sec	115 min	5 sec	3 min	300 sec	≈ 2 hrs.
Coil 3	Phantom	O	3	-	140 min	-	-	22.5 hrs.	≈ 25 hrs.
		M	3	< 1 sec	140 min	< 1 sec	30 sec	30 sec	≈ 2.5 hrs.
	Body	O	33	-	220 min	-	-	71 hrs.	≈ 74.5 hrs.
		M	33	< 1 sec	220 min	5 sec	5 min	400 sec	≈ 4 hrs.
Coil 4	Phantom	O	4	-	140 min	-	-	26 hrs.	≈ 28.5 hrs.
		M	4	< 1 sec	140 min	< 1 sec	40 sec	40 sec	≈ 3 hrs.
	Head	O	34	-	195 min	-	-	93.5 hrs.	≈ 97 hrs.
		M	34	< 1 sec	195 min	< 1 sec	6 min	425 sec	≈ 3.5 hrs.
Coil 5	Phantom	O	4	-	140 min	-	-	25 hrs.	≈ 27.5 hrs.
		M	4	< 1 sec	140 min	< 1 sec	40 sec	40 sec	≈ 2.5 hrs.
	Head	O	34	-	160 min	-	-	97.5 hrs.	≈ 100 hrs.
		M	34	< 1 sec	160 min	< 1 sec	6 min	400 sec	≈ 3 hrs.
Coil 6	Phantom	O	12	-	240 min	-	-	55 hrs.	≈ 59 hrs.
		M	12	< 1 sec	240 min	< 1 sec	2 min	100 sec	≈ 4 hrs.
	Body	O	36	-	250 min	-	-	174 hrs.	≈ 178 hrs.
		M	36	< 1 sec	250 min	< 1 sec	6 min	450 sec	≈ 4.5 hrs.

of the initial capacitor values that is within this volume.

2.3.2 DBC Implementation

The nonideal construction of the coil and tolerances of the capacitor values result in slight changes in the resonance frequency (~ 1 MHz) and matching. The capacitors, placed on the upper ring, were used for matching purposes (C_m) and the value of each of these capacitors was 90 pF. The variable capacitors on the lower ring were utilized as tuning capacitors (C_t) and trimmed between 7.5 and 9.5 pF for fine-tuning. Finally, for decoupling purposes, 14.9 pF capacitors were used on the rungs of the coil (C_d).

Table 2.2 shows reflection and coupling coefficients in dB for the DBC at 123.2 MHz. All numbers in Table 2.2a were obtained by bench-top measurement of the S-parameters of the constructed coil using the network analyzer. The highest reflection coefficient (S_{nn}) was -28 dB. In the worst case, the mutual coupling coefficient (S_{mn}) for the adjacent channels was -11 dB. However, for nonadjacent channels, it was -18 dB. Table 2.2b shows the S-parameters obtained in the EM simulation environment for the corresponding DBC.

The ratio of total reflected power to input power was 18%

in the case of the constructed DBC but was 25% in the simulation environment. The

Table 2.2: Scattering parameters of the DBC. (a) Experimentally measured. (b) Obtained from the EM simulation.

	Channel	1	2	3	4	5	6	7	8
Experiment	1	-28							
	2	-11	-29						
	3	-29	-12	-29					
	4	-21	-29	-11	-30				
	5	-19	-20	-30	-11	-28			
	6	-20	-18	-20	-29	-12	-29		
	7	-30	-21	-19	-20	-30	-11	-29	
	8	-12	-29	-20	-18	-20	-29	-11	-28
Simulation	1	-30							
	2	-9.6	-30						
	3	-30	-9.6	-26					
	4	-21	-29	-9.6	-27				
	5	-19	-21	-29	-9.6	-28			
	6	-21	-19	-21	-29	-9.6	-25		
	7	-30	-21	-19	-21	-29	-9.6	-32	
	8	-10	-30	-21	-19	-21	-31	-9.6	-28

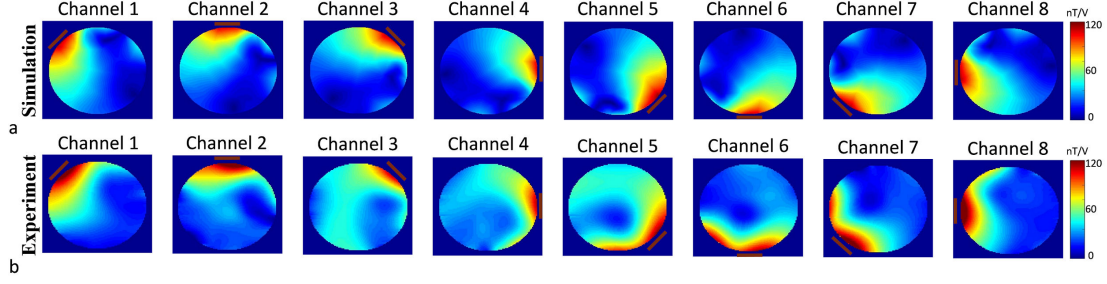


Figure 2.8: (a) Simulation (b) and MRI experiment results for B_1^+ -map of each channel.

difference between simulated and measured reflection values is possibly due to a mismatch between the geometry of the constructed coil and the simulation model. Besides, the capacitors were modeled as lossless in the EM simulations. The equivalent series resistance of the capacitors that we used may also play a role in this small error.

2.3.3 Field Maps

Once the B_1^+ maps are known, the TxArray system can be utilized for many applications, such as RF shimming [71–74]. The maps were acquired using the technique discussed in the Methods and Materials section. Fig. 2.8 shows both the simulated and measured maps.

To provide proof of proper operation of the DBC, we implemented RF shimming. To do so, we performed an EM simulation related to an ideal birdcage coil (BC) with the same size as the designed DBC. Quadrature excitation was used in the simulation environment to excite the circularly polarized (CP) mode of the BC (Fig. 2.9a). In the shimming process, we considered the B_1^+ profile corresponding to the CP mode as the desired profile. We optimized the phase and magnitude for each channel using the steepest descent method. In this proof-of-operation study, we did not include the power criterion in the optimization problem. Panels a and b of Fig. 2.9 demonstrate the simulated and measured B_1^+ -maps for the desired CP mode and the optimized profile, respectively. Fig. 2.9c is a gradient-echo MR image obtained using the optimized excitations.

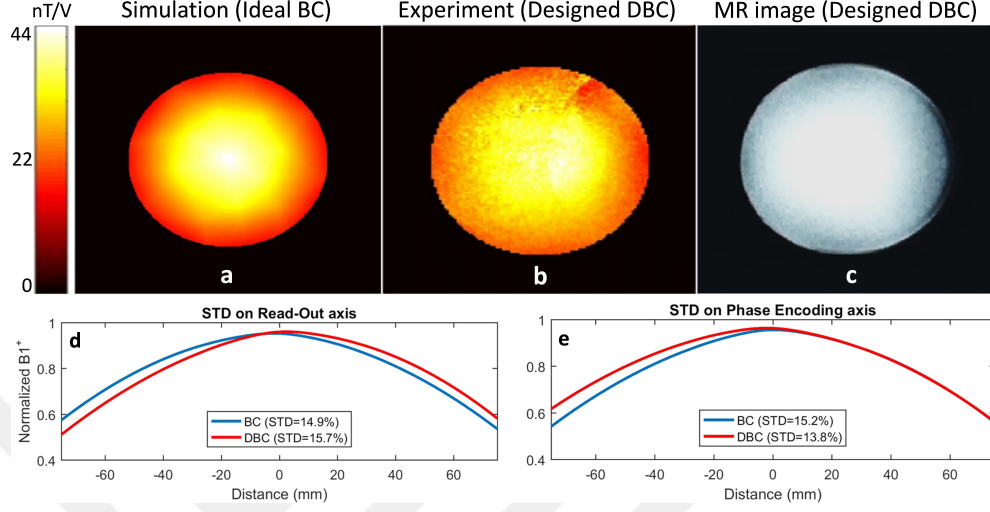


Figure 2.9: (a) B_1^+ -map of CP-mode of a standard BC that was achieved using the proper EM simulation. (b) CP mode that was generated using the constructed DBC. (c) The GRE image taken from a uniform phantom using the DBC. The sequence parameters are TR=100 ms, TE=12 ms, NEX=1, 128×128, FOV=20 cm, and slice thickness=5 mm. (d-e) B_1^+ field sampled over the readout and phase encoding axes corresponding to both BC and DBC. Calculated mean standard deviations (in percentage) verify a good performance of the DBC.

The comparison between the homogeneity performance of the BC and DBC can be investigated using the standard deviation of the B_1^+ inside the phantom. Fig. 2.9d demonstrates the relative standard deviation of B_1^+ on the readout axis for both the BC and DBC. Similarly, Fig. 2.9e shows the same parameter for both coils on the phase encoding axis. Accordingly, the relative standard deviation on the readout axis was 14.9% and 15.7% for the BC and DBC, respectively. On the phase encoding axis, this was 15.2% and 13.8%, respectively, for the BC and DBC. These results indicate that we managed to design DBC as desired, that is it has a mode of operation very similar to BC.

2.4 Discussion and Conclusions

In this chapter, accelerating the co-simulation method for the design of TxArray coils is studied using equivalent circuit models. In the original co-simulation method proposed

by Kozlov et al. [39], an EM simulation of the coil in the presence of an imaging object is performed while all lumped elements are replaced by excitation ports. Afterward, the resultant S-parameters are imported to a circuit simulator and a time-consuming optimization is performed on the lumped element values, due to the excessive number of local minima in the problem. In the proposed method, we show that starting from proper initial guesses for the capacitor values in the optimization process helps in quickly finding the optimum capacitor values. As an example, an eight-channel head-degenerate birdcage coil is constructed using the obtained values, during which the design process is accelerated by a factor of 20.

As shown in Table 2.1, the acceleration occurs in the final optimization stage of the design process. Correspondingly, the acceleration factor mostly depends on the number of unknown parameters (capacitor values) during the optimization. In the case of the coils in Fig.2.4 in the presence of the cylindrical phantom, the number of independent capacitors was decreased to three, three, three, four, four, and 12, respectively, due to the cylindrical symmetry. However, in the case of the loading with the human model, these numbers were 17, 25, 33, 34, 34, and 36, respectively. Note that, to assign the independent optimization parameter to the decoupling capacitors, we only considered the geometrical features of the coils regardless of the loading schemes.

Currently, degenerate birdcage coils are not used extensively in ultra-high field applications. However, these coils are one of the best choices for a 3T transmit array coil because it has a mode of operation similar to a conventional birdcage coil. Unfortunately, the complexity of their design is a disadvantage. This chapter partially addresses this problem, as well. Although we presented the circuit model for DBCs and used the corresponding circuit models for five other types of TxArray coils in this work, the proposed method can be easily extended and utilized in the design of other types of TxArray coils, as well, using a proper equivalent circuit model for the intended coil. The corresponding MATLAB scripts for the presented examples (i.e., six different coils with two different loading schemes) which include a relatively large variety of TxArrays to prove the concept are available online as the supplementary material¹.

¹https://github.com/UMRAM-Bilkent/Accelerated_CoSimulation.git

It should be noted that in the presented equivalent circuit model, the inductors represent the copper strips of the coil. However, this representation is only valid if the wavelength is much greater than the size of the strips. As the length of the wire increases, a multistage RLC network or even more complex models may be necessary to model a single conductive strip [104].

Furthermore, determining the optimum geometry of the TxArray coils in terms of either their dimensions or their shape is always demanding. Since tuning and decoupling of a candidate geometry is usually a time-consuming process, it is difficult to explore the many varieties of possible geometries. The proposed method can shorten the design process and determine the feasibility of a design, allowing a greater number of candidate designs to be examined.

The workstation used in this study is a mediocre system with enhanced memory. Clearly, more powerful systems can make the design process faster. Additionally, the gradient descent method we used as an optimization algorithm may not be the best choice for this problem. There are many sophisticated minimization algorithms for problems involving multiple local minima. As can be understood from Fig. 2.7, this problem has many local minima. Here, instead of optimizing the optimization algorithm, we chose to work on the minimization of the search interval. Our proposed solution should also help all other optimization algorithms since the initial guess is close to the optimum point.

In the construction of the coil, fine adjustment of the tuning capacitors was necessary. We believe this is mainly due to tolerances in the capacitors and imperfections in their construction. These can be handled by measuring the capacitance errors and 3D printing the coil-housing. Some additional errors are due to the errors of the EM simulations. To reduce these errors, convergence criteria can be reduced, though this significantly increases the simulation time and the memory requirement. However, the errors due to EM simulation are significantly fewer in number than those due to other sources of error.

Chapter 3

Improving Radiofrequency Power and Specific Absorption Rate Management with Bumped Transmit Elements in Ultra-High Field MRI

Preface

The content of this chapter was presented in part at the 26th and 27th Annual Scientific Meetings of International Society of Magnetic Resonance in Medicine (ISMRM) [79, 88] and it was published in Magnetic Resonance in Medicine [109]. The text and the figures of this chapter are based on the journal publication [109]. This is a collaborative work between UMRAM of Bilkent University and CMRR of University of Minnesota. Gregor Adriany, Gregory J. Metzger, Russell L. Lagore, Steve Jungst, Lance DelaBarre, Pierre-Francois Van de Moortele, Kamil Ugurbil, and Yigitcan Eryaman contributed to this study. Gregor Adriany, Gregory J. Metzger, Pierre-Francois Van de Moortele, Kamil Ugurbil, Ergin Atalar, and Yigitcan Eryaman were involved in study conception and design. Steve Jungst contributed to the construction

of the RF coils. Lance DelaBarre was responsible for B_1^+ -mapping and MR thermometry experiments.

3.1 Introduction

Demand for the ultra-high field (UHF, $B_0 \geq 7$ T) MRI is continuously increasing due to its numerous benefits such as a significant increase in signal-to-noise ratio [5–10] and improved susceptibility contrast [11, 12]. However, with an increasing static magnetic field (B_0), electromagnetic (EM) wavelengths decrease resulting in destructive and constructive interferences of radiofrequency (RF) magnetic fields causing degradation in overall image quality. In addition, constructive interference of the RF electric field can lead to elevated local specific absorption rate (SAR) hot spots which are an important patient safety concern at UHF. Nevertheless, it is shown that multi-channel transmit arrays (TxArrays) with improved RF performance can be used to alleviate both issues when optimal excitation strategies are utilized [15, 16].

In the literature, various types of the transmitting elements such as loop coils [46–51], transmission lines [10, 15, 52–54], and dipole-like structures [55–60, 87] are investigated for UHF-MRI. Thalhammer et al. [46] used TxArray of loop coils for cardiac imaging at 7 T. Adriany et al. [15] employed an array of the transmission lines for head imaging at 7 T. Vaughan et al. [53] built a transverse electromagnetic (TEM) coil for whole-body imaging at 7 T. Raaijmaker et al. [55, 56] introduced single-side adapted and fractionated dipole (FD) as new transmitter elements for body imaging at 7 T. Steensma et al. [57, 58] improved the SAR performance of the dipoles at 7 T and 10.5 T by introducing the snake antenna. Vaughan et al. [10] utilized a TEM coil for the head imaging at 9.4 T. Shajan et al. [48] performed head imaging at 9.4 T using the array of loop elements. Erturk et al. [59] proposed to use an array of fractionated dipoles for torso imaging at 10.5 T. RF coil development is essential for realizing the full potential of UHF-MRI. Improving the performance of RF coil arrays improves UHF imaging as demonstrated in numerous studies in the literature.

In this chapter, we propose to modify the geometry of individual coil elements by

placing a bump underneath the discontinuities (i.e., all lumped elements and excitation ports) on the coil [79]. This reduces the peak local SAR while maintaining comparable B_1^+ within the target region of interest (ROI). As a proof of concept, EM simulations were performed for three different types of coils: the loop coil and snake antenna at 7 T, and the fractionated dipole at 10.5 T. In addition, a bumped fractionated dipole is constructed and experimentally compared to a fractionated dipole design previously proposed for 10.5 T [59]. Moreover, two eight-channel body arrays are simulated using the bumped and regular fractionated dipoles and their Tx performance are compared to each other.

3.2 Theory and Methods

According to Eq. 1, the continuity equation in the electromagnetic theory, spatial variation of the current density produces electrical charges.

$$\nabla \cdot \vec{J} = -j\omega\rho \quad (3.1)$$

where ρ denotes the electrical charge density. Based on this basic principle we can predict that a high SAR region would occur underneath a conductor discontinuity on an RF transmitter. In other words, the discontinuity on the pathway of the current leads to charge accumulation, which results in an elevated electric field (so-called conservative electric field) and high local SAR in the tissue. As a solution for this RF safety problem, we propose to place a bump underneath the discontinuities and increase the distance between the accumulated charges and the body. This simple modification reduces the conservative electric field inside the tissue. On the other hand, the B_1^+ at a point which is sufficiently far from the coil is not affected by this modification.

3.2.1 Numerical Optimization of the Bump Geometry

As a proof of concept, simple bumps were placed underneath the discontinuities of a loop coil [46], the snake antenna [57], and the fractionated dipole [59], as shown in

Figure 3.1a-f. To determine the optimum height of the bump in each case, a deep-body target (e.g., prostate imaging) is defined, and accordingly, the following optimization problem is investigated by sweeping the height in a reasonable range.

$$\begin{aligned}
& \text{minimize} && [peak\ 10g\ averaged\ SAR] \\
& && [B_1^+ \text{remaining constant at the target range}] \\
& \text{subject to} && (i.e.,\ 50\ to\ 120\ mm\ depth),\ compared\ to\ the \\
& && original\ structure]
\end{aligned} \tag{3.2}$$

The height of the bump was swept from 5mm to 35mm for the loop coil (7 T), 60 mm for the snake antenna (7 T) and 60 mm for the fractionated dipole (10.5 T), moreover according to the original structures [46,57,59], the element-to-phantom separations of 15 mm, 20mm, and 20 mm were determined respectively. The EM simulations were performed using a numerical EM solver HFSS 18.1 (ANSYS, Canonsburg, PA, USA) based on the finite-element method (FEM). In all cases, conductors were placed on an FR4 printed circuit board (PCB) which was mounted on a block of polyethylene terephthalate glycol (PETG). A uniform rectangular phantom (200x200x400 mm³) with electrical properties similar to the average values of the human body ($\epsilon_r = 36$, $\sigma = 0.42$ S/m) were used in the simulations for optimization.

3.2.2 Experimental Element-wise Comparison

Both the bumped and regular FDs were constructed and matched to obtain a reflection coefficient -15 dB or less. Both dipoles consisted of conductors on an FR4 PCB mounted on a PETG block with the thickness of 10 mm. The regular [59] and the bumped fractionated dipoles, the latter with an optimized bump height using the methods detailed above, were positioned on a torso-sized elliptical body phantom (450x180x290 mm³) (Figure 3.1g) [110]. The phantom was filled with a NaCl doped hydroxyethyl cellulose (HEC) gel with electrical properties of $\epsilon_r=78.3$ and $\sigma=0.66$ S/m. The two FD elements being compared were used as transmitters in this setup while an additional three FD elements were placed on the phantom as receivers only (Figure 3.1g).

Experimental studies with the above setup were conducted on a whole-body 10.5 T Magnet (Agilent Technologies, Oxford, UK) and associated imaging system (Siemens Healthineers, Erlangen, Germany). The scanner is equipped with a 16-channel parallel transmit system with each channel driven by a 2-kW RF power amplifiers (Stolberg HF-Technik AG, Stolberg, Germany).

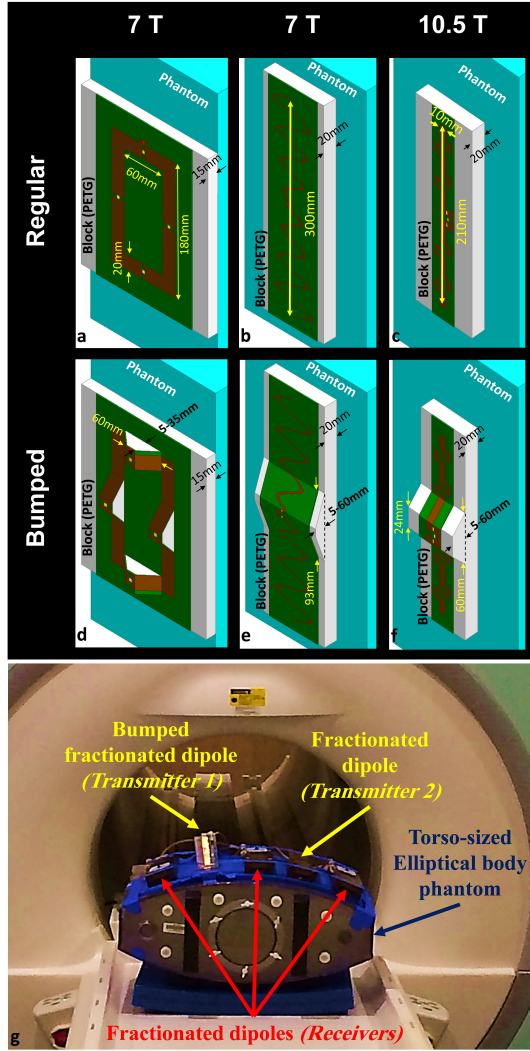


Figure 3.1: Regular vs. bumped elements. (a-c) Loop coil and snake antenna at 7 T, and fractionated dipole at 10.5 T. (d-f) bumped loop coil, snake antenna, and fractionated dipole. (g) The experimental setup used to compare the performance of the bumped and regular fractionated dipole.

TxArrays comprised of bumped or regular elements were simulated using a uniform

Transmit B_1^+ -maps were acquired using the actual flip-angle imaging (AFI) technique [111]. Temperature mapping was performed using magnetic resonance thermometry [112] (MRT) based on the proton resonance offset method [113] with a 3D multi-echo gradient-echo sequence. SAR values were calculated by finding the slope of the initial part of the heating curve and multiplying with the heat capacity of the HEC gel (4386 J/kg/°K). The experimental B_1^+ and SAR maps were compared against EM simulations of the same configuration performed using the HFSS.

3.2.3 Eight-Channel Body TxArray Coils: Numerical Simulations

To demonstrate the potential of the bumped geometry in a multi-channel excitation scenario, simulations were performed with arrays of the two element designs. Two separate eight-channel

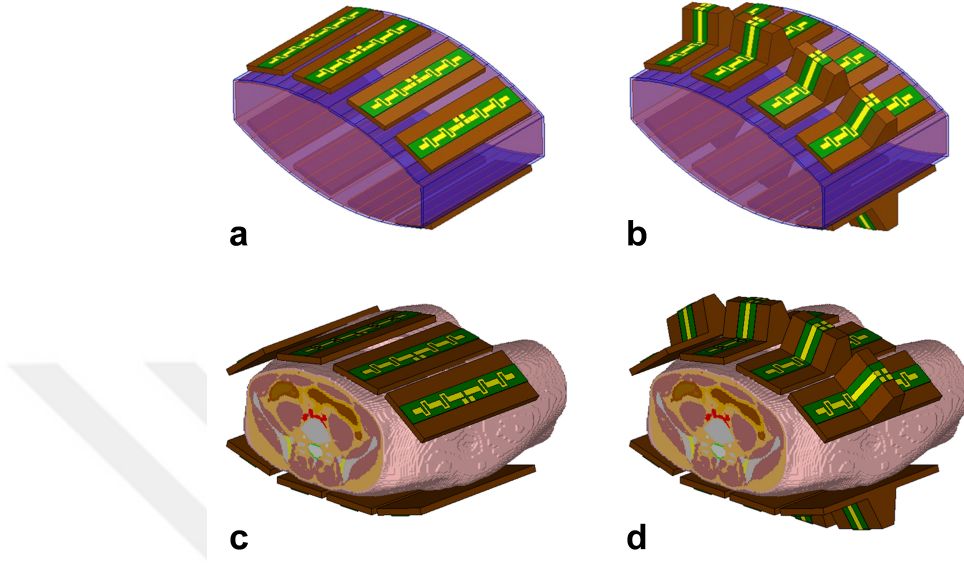


Figure 3.2: Comparison between the eight-channel TxArray of regular and bumped fractionated dipoles. (a) An eight-channel TxArray coil of regular and (b) bumped fractionated dipoles in the presence of a uniform elliptic phantom. (c) An eight-channel TxArray coil of regular and (d) bumped fractionated dipoles in the presence of a realistic human body model.

phantom and a realistic human body model. In both cases four anterior and four posterior dipoles were positioned circumferentially around the load. The blocks utilized in these simulations were 10mm-thick instead of 20 mm (i.e., utilized through the optimization procedure) in order to alleviate the lack of physical space due to the bore size limitations.

In the phantom simulations, all elements were matched to obtain a reflection coefficient of -20 dB or less. We maximized the B_1^+ -power efficiency at a prostate sized region at the center of the phantom using a phase-only shimming method without consideration of the local SAR [16]. B_1^+ -power efficiency at the position r is defined as

$$\zeta(r) = \frac{B_1^+(r)}{\sqrt{P_{acc}}} \quad (3.3)$$

where P_{acc} represents the accepted input power. The total input power was adjusted to enforce the same average B_1^+ value inside the ROI for both TxArrays.

For the body model simulations, the torso of the realistic human body model Duke

of the virtual population [114] was used (voxel size, $2 \times 2 \times 2 \text{ mm}^3$; height, 1.77 m; weight, 70.3 kg), as shown in Figure 3.2c and 3.2d. In this case, EM simulations were performed in CST studio 2018 (CST, Darmstadt, Germany) using a time-domain solver based on the finite-integration technique (FIT). All coil elements were matched to obtain a reflection coefficient of -15 dB or less. A phase-only shimming optimization was applied to maximize the B_1^+ -power efficiency on the prostate without consideration of the local SAR. Similar to the phantom simulations, the total input power was adjusted to enforce the same average B_1^+ value inside the ROI for both TxArrays.

3.3 Results

3.3.1 Numerical Optimization of the Bump Geometry

Figure 3.2a-c show the B_1^+ -power efficiency results sampled on a line perpendicular to the coil surface and passing through the coil's center for the loop coil, the snake antenna, and the fractionated dipole, respectively. The target range (i.e., 50 to 120 mm depth in the phantom) is shown with a shaded region in the figure. As can be seen, increasing the height of the bumps results in reduced B_1^+ -power efficiency at a region close to the coil. On the other hand, bumps did not have a significant effect on the B_1^+ -power efficiency at the target region. In Figure 3.2, the B_1^+ at a depth of 80 mm (i.e., the average distance from the transmit element to the prostate) corresponding to the regular structures was chosen as a reference. Then, the input power in the case of each height was adjusted such that the reference B_1^+ value has been achieved. Finally, the peak 10g-averaged SARs were determined and plotted for the adjusted input powers. As indicated on the graphs, the height of the bump that minimizes peak 10g-averaged SAR can be chosen as the optimum height. For the loop coil, there is a well-defined global optimum height. However, for the snake antenna and the fractionated dipole, the peak SAR plot reaches a plateau at a certain bump height. This allows us to pick a smaller bump size for those designs, improving the ease of fabrication and patient comfort without sacrificing B_1^+ -SAR efficiency.

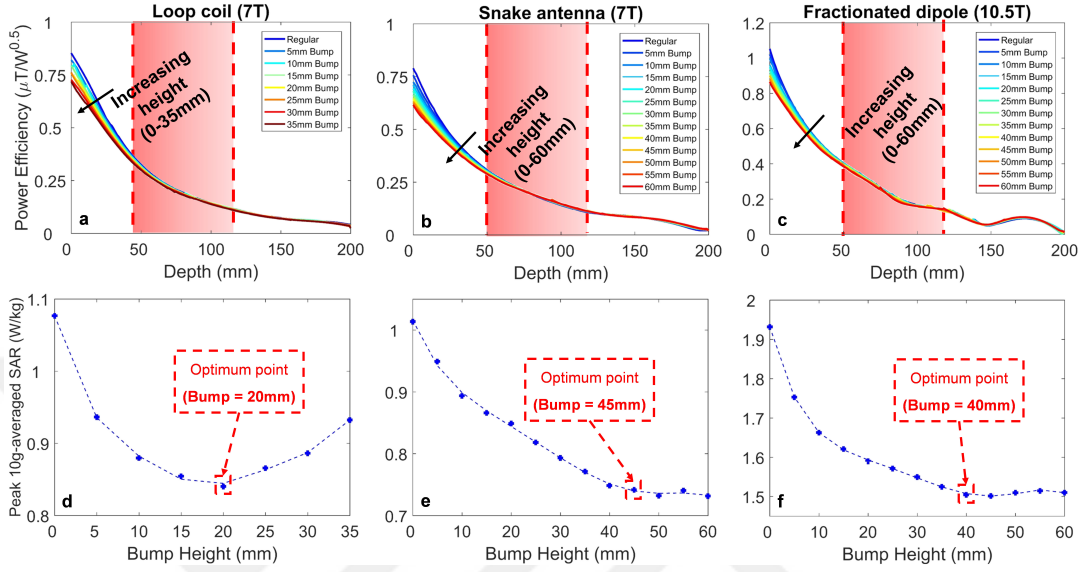


Figure 3.3: Optimizing bump height. (a-c) B_1^+ -power efficiencies of a loop coil, a snake antenna, and a fractionated dipole are calculated for various bump heights and sampled on a line perpendicular to the surface of the coil passing through the center of the coil. (d-f) Peak 10g-averaged SAR for each bump geometry is also calculated while the input power is adjusted to achieve the reference B_1^+ at the depth of 80 mm.

3.3.2 Experimental Element-wise Comparison

Using the phantom configuration of Figure 3.1g and the optimized bump height of 50 for the FD determined during optimization, simulation and experimental B_1^+ -power efficiencies were determined experimentally and from simulation as shown in Figure 3.4a and 3.4b, respectively. The B_1^+ -power efficiencies were sampled along the dashed lines perpendicular to the surface of the phantom and passing through the center of the dipoles in Figure 3.4a and 3.4b, then they were plotted with respect to the depth inside the phantom as shown in Figure 3.4c. Both simulation and experimental results demonstrate that the B_1^+ -power efficiency of the bumped FD approaches the regular FD and becomes almost identical at a depth 60 mm.

The simulated and measured 10g-averaged SAR maps determined from MRT are given in Figure 3.4d and 3.4e, respectively.

Figure 3.4f shows the simulation results for the B_1^+ -SAR efficiency of the bumped

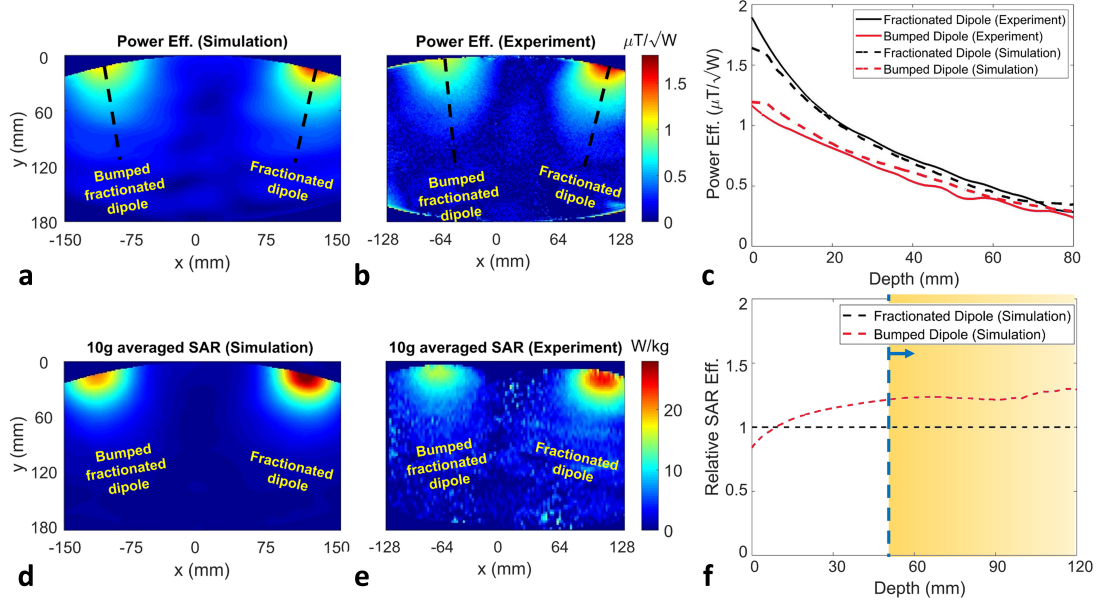


Figure 3.4: Simulation and experimental results for individual fractionated and bumped dipoles. (a-b) The B_1^+ -power efficiencies inside a uniform phantom. (c) The simulated and experimentally measured B_1^+ -power efficiencies on the dashed lines shown in Figure 3.4a and 3.4b. (d-e) The 10g-averaged SAR. (f) The B_1^+ -SAR efficiency of the bumped element (calculated relative to the regular fractionated dipole) on the dashed lines shown in Figure 3.4a.

FD relative to regular FD plotted with respect to depth. The bumped FD achieved a 25% improvement in B_1^+ -SAR efficiency at all depths greater than 50 mm.

3.3.3 Eight-Channel Body TxArray Coils: Numerical Simulations

The simulated B_1^+ -maps inside a uniform elliptic phantom obtained using the eight-channel TxArray of fractionated and bumped dipoles are shown in Figure 3.5a and 3.5b, respectively. The maps correspond to the phase-only shimming solution for the prostate-size ROI indicated by the black square. The input powers for both arrays were adjusted to achieve the same average B_1^+ (1 uT) inside the ROI.

The 10g-averaged SAR maps corresponding to the identical excitations utilized to obtain the B_1^+ -maps in Figure 3.5a and 3.5b are shown in Figure 3.5c and 3.5d. The presented SAR map is over the transverse plane which includes the location of peak

local SAR. The bumped dipole array had a 22% lower peak SAR than the regular dipole array, yet achieved the same B_1^+ on the prostate-size ROI.

Similarly, the simulated B_1^+ -maps inside the realistic human body model (Duke) are shown in Figure 3.5e and 3.5f. The phase-only shimming optimization was performed in the prostate on the transverse slice passing through the center of the dipoles.

The corresponding 10g-averaged SAR map on a transverse plane including the location of peak local SAR. Based on these results, the bumped dipole array can achieve a 27% reduction in peak 10g-averaged SAR while maintaining the same B_1^+ -power efficiency in the prostate.

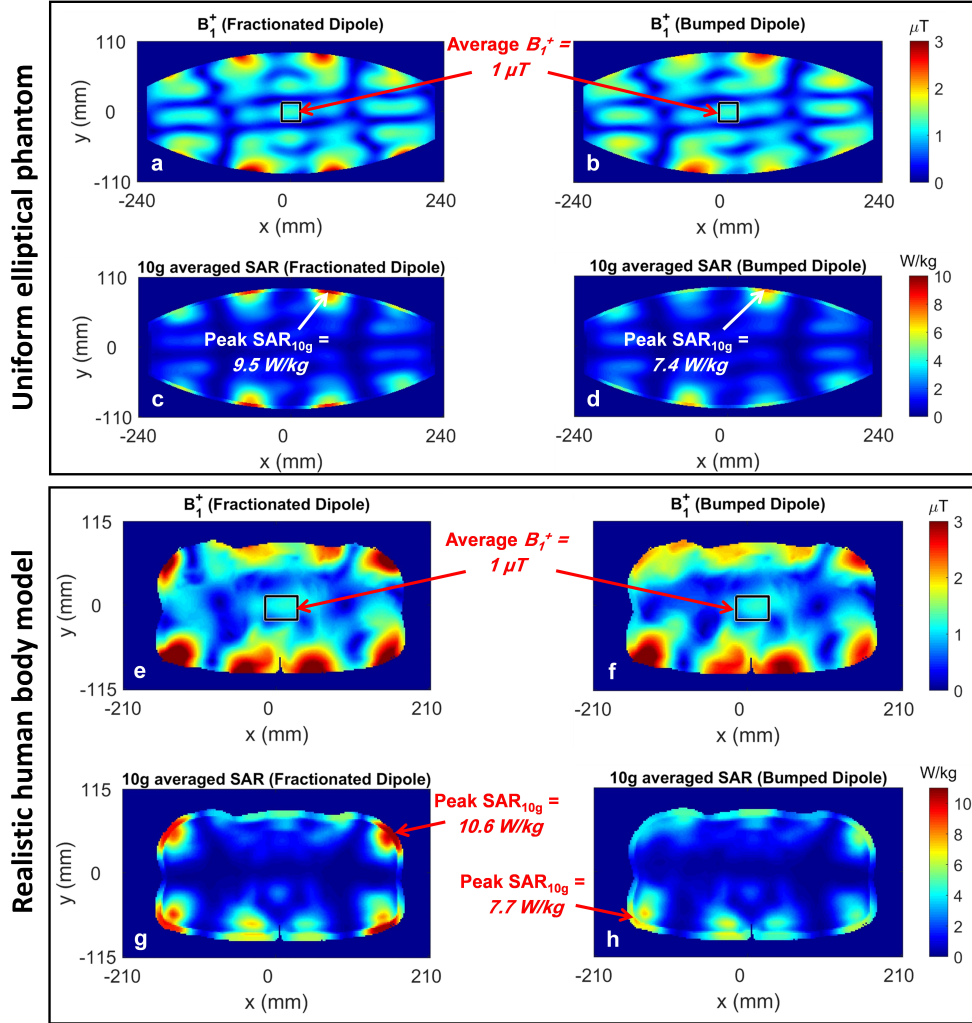


Figure 3.5: EM simulations corresponding to eight-channel fractionated and bumped dipoles in the presence of a uniform elliptical phantom and realistic human body model (Duke). (a-b) Phase-only shimming solutions for a prostate-size ROI at the center of the uniform elliptical phantom and (c-d) corresponding 10g-averaged SAR maps on the transverse plane which includes the peak local SAR. (e-f) Phase-only shimming solutions for the prostate and (g-h) corresponding 10g-averaged SAR maps on the transverse plane which includes the peak local SAR.

3.4 Discussion and Conclusions

In this chapter, we investigated a simple strategy based on adding bumps to a resonant transmit element underneath the discontinuities (i.e., lumped elements and excitation

port) in order to reduce the local SAR while keeping B_1^+ -power efficiency constant inside an ROI. This approach works especially well when the target of interest is relatively far away from the coil (i.e., prostate imaging). A similar advantage may also be present for head imaging studies at UHF [80, 88] especially for spectroscopy studies which aims to maximize B_1^+ -power efficiency at a single voxel near the middle of the brain [115].

We demonstrated the advantage of adding bumps for SAR reduction using three transmit elements; loop coils and snake antennas at 7 T, and fractionated dipoles at 10.5 T. The reason for such selections was to present the advantage of adding bump in a fair and unbiased manner since the selected structures are optimized and presented in previously published works [46, 57, 59]. This advantage is investigated for a fractionated dipole at 7 T [56] as well, and the results are presented in Figure 3.6. In principle, the approach is expected to work for all transmit elements with lumped matching and tuning elements. Optimizing arrays for the transmit performance in a collective manner is a challenging task, mainly because it requires modeling the complex interactions between individual transmit elements. As an alternative strategy, individual coil elements can be optimized for a given metric assuming that an array comprised of such elements will also yield optimal results [56]. Although this conclusion is not generally true for arbitrary coil arrays, it has been shown that optimizing the individual transmitters may result in improved array performance, especially when the EM interactions between individual elements are weak. In this chapter, we optimized the bump height for the individual elements. This optimization resulted in an overall improvement of the array performance as demonstrated with our simulation results.

Although the B_1^+ -SAR efficiency of transmit elements was improved in the ROI by adding the bump, the B_1^+ -power efficiency was slightly declined which indicate a degradation in SNR. The reduction in B_1^+ -power efficiency mostly can be tackled by increasing the transmit power, however, the SNR issue may remain unsolved if the bumped elements will be used as a receive array as well. A potential solution for such a problem during reception can be utilizing other types of receive elements along with the bumped elements (e.g., loop-dipole combination [60]).

As shown in Figure 3.3b, introducing the bump resulted in an increase in B_1^+ -power

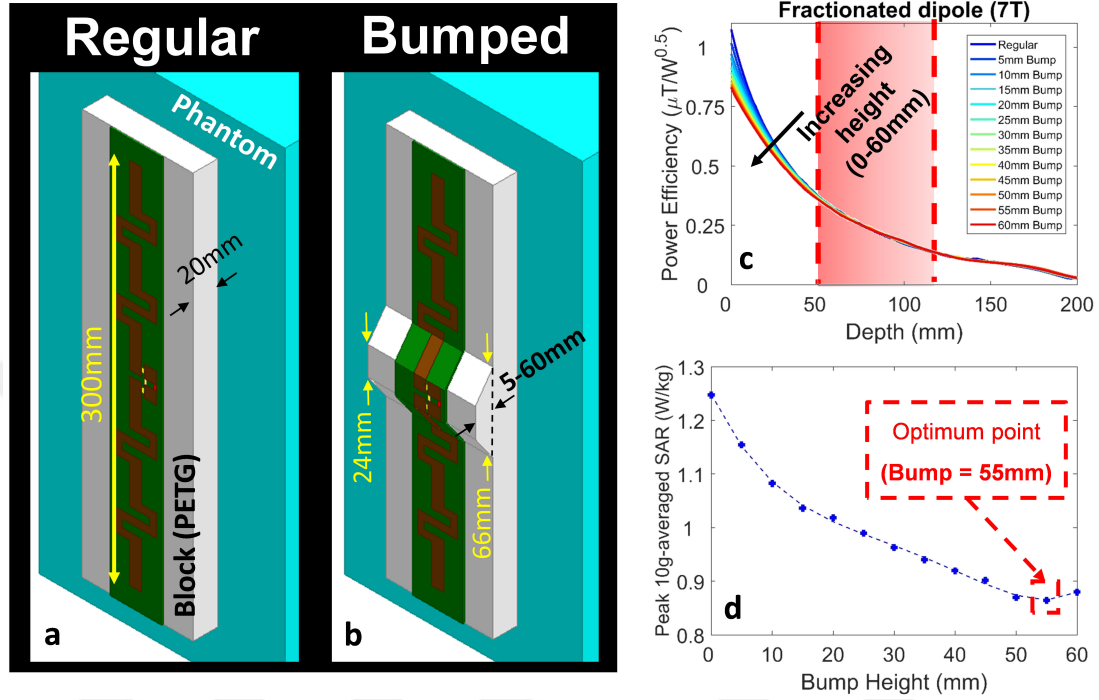


Figure 3.6: Regular vs. bumped fractionated dipole at 7 T. (a) Structure of the regular fractionated dipole used for the EM simulations. (b) Structure of the bumped fractionated dipole where the bump height was swept between 5 mm and 60 mm. (c) B_1^+ -power efficiencies of the fractionated dipoles are calculated for various bump heights and sampled on a line perpendicular to the surface of the dipole passing through the center of the dipole. (d) Peak 10g-averaged SAR for each bump height is also calculated while the input power is adjusted to achieve the reference B_1^+ at the depth of 80 mm. As a result, 31% reduction in the peak 10g-averaged SAR is achieved for a 55 mm-bump compared to the regular structure.

efficiency in the target depth-of-interest (shown with the shaded area). The snake antenna was originally designed to optimize the B_1^+ -SAR efficiency, however, B_1^+ -power efficiency was not optimized. Adding a bump to the antenna have altered the physical and effective length of the antenna resulting in an increased B_1^+ -power efficiency.

According to Figure 3.3, the optimal bump heights for the snake antenna and the fractionated dipole are 45 mm and 40 mm respectively, which makes the elements rather bulky. Using these elements to form a transmit array similar to Fig. 3.2 may have some drawbacks. For example, the bore size may be a limiting physical factor especially for subjects with a higher body-mass-index (BMI). In addition, the conductors at the bump will be closer to the gradient shield and this may affect matching of

the elements. However, both issues can be addressed by modestly sacrificing B_1^+ -SAR efficiency and choosing a lower bump height (e.g., 30 mm instead of 45 mm for the snake antenna and 40 mm for the FD which reduce B_1^+ -SAR efficiency by 3.7% and 1.5%, respectively).

Experimentally, we investigated the bumped fractionated dipole at 10.5 T comprehensively. We built regular and bumped fractionated dipoles and compared their transmit performances to validate our simulation results.

In this chapter, we investigated the effect of adding bumps to a resonant transmit element underneath the discontinuities (i.e., lumped elements and excitation port). The suggested modification reduced the 10g- averaged local SAR (25%) while preserving the B_1^+ -power efficiency at a distant point of interests. This preliminary work demonstrates that bumped elements are promising candidates as Tx coils for deep-targets (e.g., prostate) imaging at UHF. Future studies will focus on utilizing the proposed topology for in vivo imaging at UHF.

Chapter 4

In Vivo Human Head MRI at 10.5 T: A Radiofrequency Safety Study and Preliminary Imaging Results

Preface

The content of this chapter was presented in part at the 27th Annual Scientific Meeting of International Society of Magnetic Resonance in Medicine (ISMRM) [88] and it was published in Magnetic Resonance in Medicine [80]. The text and the figures of this chapter are based on the journal publication [80]. This is a collaborative work between UMRAM of Bilkent University and CMRR of University of Minnesota. Lance DelaBarre, Russell L. Lagore, Angel Torrado-Carvajal, Xiaoping Wu, Andrea Grant, Gregor Adriany, Gregory J. Metzger, Pierre-Francois Van de Moortele, Kamil Ugurbil, and Yigitcan Eryaman contributed to this study. Gregor Adriany, Gregory J. Metzger, Pierre-Francois Van de Moortele, Kamil Ugurbil, Ergin Atalar, and Yigitcan Eryaman were involved in study conception and design. Russell L. Lagore contributed to fabrication of the RF coil. Gregory J. Metzger, Lance DelaBarre, Xiaoping Wu, and Andrea Grant were responsible for B_1^+ -mapping, MR thermometry experiments, and human subject studies. Angel Torrado-Carvajal provided realistic human head models for the

EM simulations.

4.1 Introduction

Demand for ultra-high field (UHF, $B_0 \geq 7$ T) MRI is continuously increasing due to its numerous benefits such as significant increase in signal-to-noise ratio [5–10] and improved susceptibility contrast [11, 12]. However, increase in the static field strength (B_0) leads to shorter electromagnetic (EM) wavelengths. As a result, destructive interference of the magnetic fields causes signal inhomogeneity and degradation in image quality. On the other hand, both local and global specific absorption rate (SAR) increase with B_0 [6, 116–118]. Local SAR turns out to be the most critical radiofrequency (RF) safety concern as its limit is typically reached earlier than global SAR [118, 119] based on international guidelines [1]. Multichannel excitation solutions have been proposed in the literature to deal with these issues using hardware and software solutions [15, 16, 120].

Recently, a whole-body 10.5T scanner was installed in the Center for Magnetic Resonance Research (CMRR) at the University of Minnesota providing a unique opportunity to investigate human body and brain. Along with other coil designs [46, 55–57, 63, 95, 121], a bumped dipole coil array was proposed [79, 88] which demonstrated improved SAR performance compared to standard loop and dipole designs. However, confidently using the RF coils at such new field strengths (i.e., 10.5 T) essentially requires either a measurement or a proper prediction of the local SAR to ensure subject safety.

There exist different methods for assessing RF safety [114, 122–128], but the assessments based on generic models are the most common approaches [114, 126–128]. Usually a number of generic models are used to compute the EM fields of a particular RF coil and then the field solutions are used to calculate the power deposition in the tissue and to ensure safety under different RF excitation scenarios. Validation of the coil model is extremely important [127, 128], however, it can be a challenging procedure (especially for multichannel arrays) mainly due to difficulties in modeling cable

coupling and parasitic reactances, as well as inaccuracies in the geometrical model of the coil.

Improving the geometrical accuracy could improve the overall accuracy of the validation procedure. Previously, it was shown that computerized tomography (CT) images of an experimental set-up can be used to construct accurate simulation models for RF heating studies [129]. The difficulties related to modeling coupling and parasitic reactance may be overcome by using a previously described approach [130] which aims to improve the agreement between simulated and measured field patterns through the lumped elements' (capacitors) values in the coil model. This approach presented by Restivo et al. [130] used co-simulation [39] to determine the values in the simulation environment that minimized the root-mean-square (RMS) of the difference between the simulated and measured S-parameters rather than using the lumped element values as they are in the physical coil. To validate the RF coil model they demonstrated an agreement between simulated and experimentally acquired B_1^+ maps of a four-channel loop array at 7 T. Another work presented by Hoffmann et al. [127] described a validation procedure that included a temperature study, whereas the validation by Restivo et al. focused only on B_1^+ maps and did not involve the local SAR comparison. In both manuscripts, the excitation scenarios were limited to phase-only shimming scenarios and did not investigate the effect of the widely varying shims needed for practical applications [15, 16, 131] and which are potentially more revealing in highlighting inconsistencies between the simulation and experimental environments.

In this chapter, we implemented a coil modeling and validation workflow that expands upon the simulation approach proposed previously by Restivo et al. [130] and validation methods proposed by Hoffman et al.

The steps of the workflow are summarized below:

- 1) An EM simulation model is configured using high-resolution segmented computerized tomography (CT) images of the setup (i.e., the Tx coil and the phantom).
- 2) Matching and tuning lumped elements along with the length of the extension

cables are optimized using the co-simulation method to minimize the RMS of the difference between the simulated and measured S-parameters.

3) Validation is performed through experimental RF heating studies and B_1^+ measurements for different excitation scenarios in a phantom.

Using our workflow, we validated an eight-channel bumped dipole array [79] at 10.5 T. Using the validated coil model, we performed additional EM simulations with four realistic human head models to determine the safe RF power limits which were used to acquire the first human head images at 10.5 T.

4.2 Theory and Methods

4.2.1 Radiofrequency Coil and Numerical Validation

In this chapter, we evaluated an eight-channel bumped dipole array as an RF transmit/receive (Tx/Rx) head coil operating at 447 MHz (Fig. 4.1). The bumped dipole uses a previously demonstrated strategy of lifting the central feed point, where the highest E-fields exist, further away from the load with the goal of reducing peak local SAR [79]. Each element of the transceiver array consisted of a 210 mm-long fractionated dipole with an optimal bump height of 30 mm at the feed point [79]. The conductors were etched on an RO4003C (Rogers Corp.) laminate and mounted on a three-dimensional (3D) printed elliptical holder composed of polyethylene terephthalate glycol (PETG). All channels were matched to better than -7.5 dB and decoupled by at least 9.9 dB. To model and validate this RF coil at 10.5 T, we implemented a coil modeling and validation workflow to quantitatively evaluate agreement between simulated and experimentally measured results (See Fig. 4.2).

A jar-shaped cylindrical phantom filled with a gel was used for all experimental and simulation studies in the validation workflow. The gel consisted of 14 g/L Hydroxyethyl Cellulose (HEC) and 2.9 g/L NaCl with electrical properties of $\epsilon_r=78$ and $\sigma=0.5$ S/m was used. The electrical properties of the phantom were measured by a dielectric

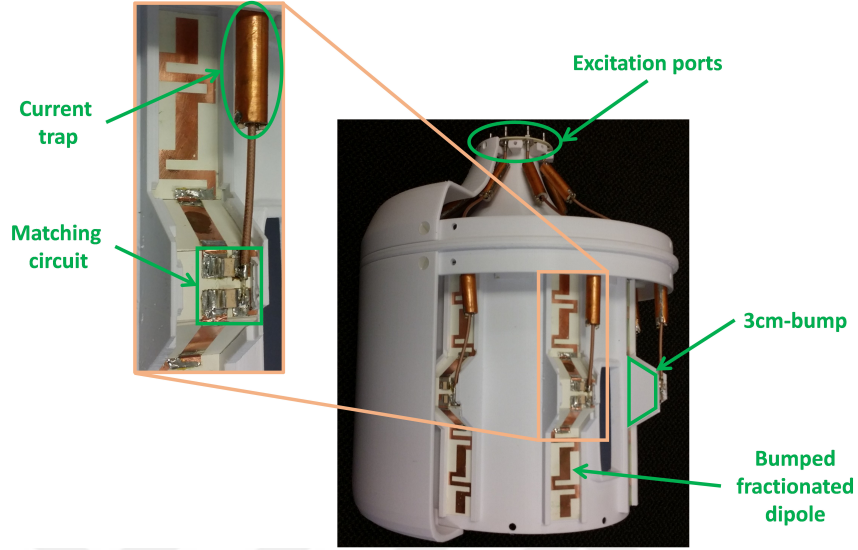


Figure 4.1: An eight-channel bumped fractionated dipole array used as an RF Tx/Rx coil for the human head imaging at 10.5 T.

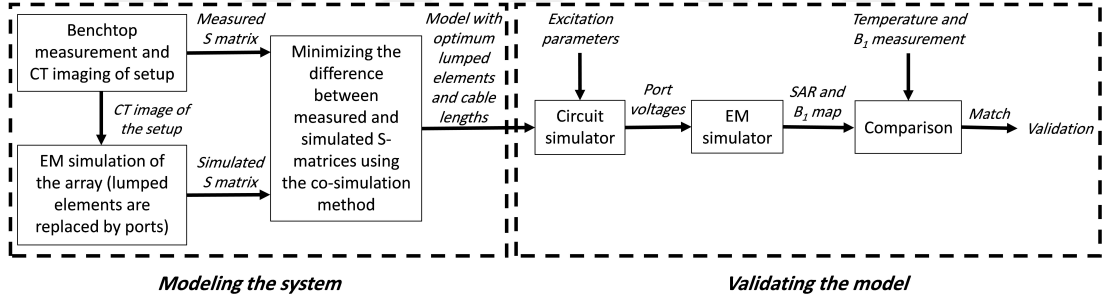


Figure 4.2: The expanded workflow to model and validate the EM model.

assessment kit (DAK-12, Speag, Zurich, Switzerland). With the phantom positioned in the RF coil, S-parameters were measured on the bench using a 16-channel network analyzer (Rohde & Schwarz ZNBT8, Munich, Germany).

To configure an EM model for simulation in which the relative position of the phantom in the coil could be accurately represented, the phantom and coil setup were transported to a CT scanner. 3D high-resolution (0.6x0.6x0.7 mm³) CT images were acquired using a biograph PET/CT scanner (Siemens, Erlangen, Germany) (Fig. 4.3a-c). The images were post processed to segment the RF coil and phantom and allow for a spatially accurate EM model (Fig. 4.3d) of the experimental setup. Experimental measurements of the phantom with the bumped dipole array were then performed at 10.5

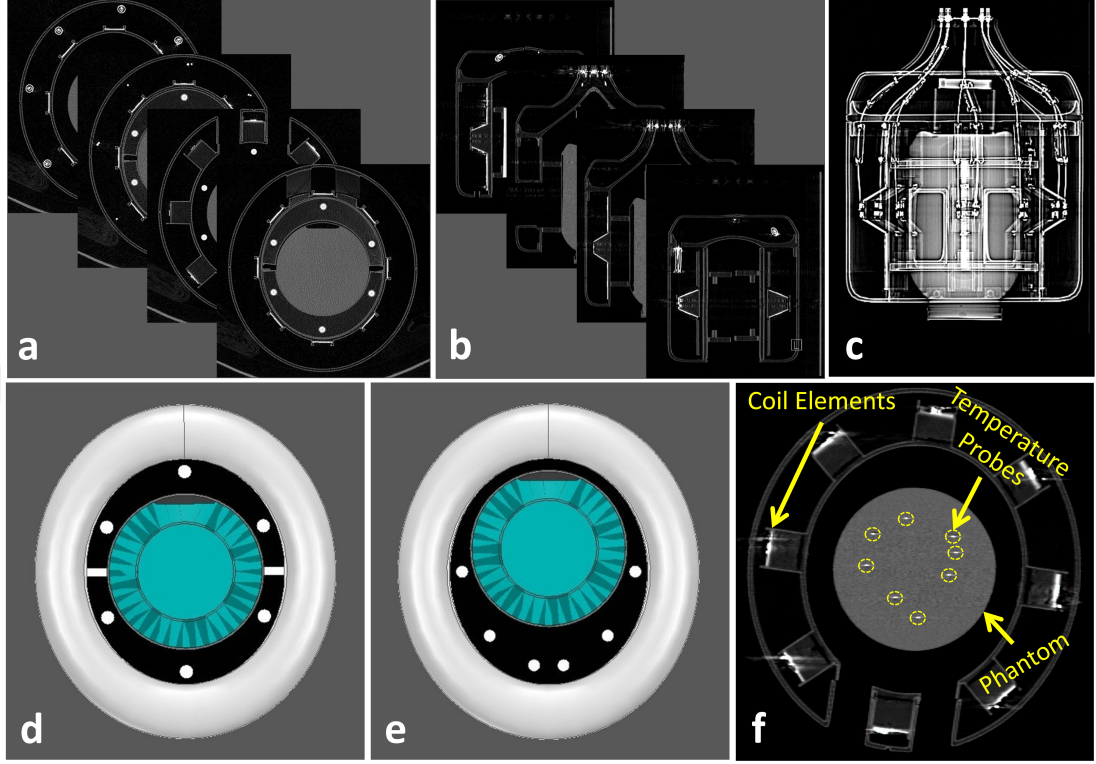


Figure 4.3: Modeling the experimental setup. (a-c) High-resolution 3D CT images of the experimental setup including the transmit coil and the phantom, (d) EM model of the coil and the centrally positioned phantom configured using the CT images, (e) EM model of the coil and the off-centered phantom configured using the CT images, and (f) CT image that shows the configuration of the optical thermal probes immersed into the phantom.

T (Agilent, Santa Clara, CA) equipped with a Siemens Magnetom console (Siemens Healthcare, Erlangen, Germany) and a 16-channel parallel transmit system. Transmit B_1^+ maps were experimentally acquired using various RF excitation patterns with two different phantom set-ups. For the first set-up, the phantom was centered in the coil (Fig. 4.3d) and for the second set-up it was placed off-center (Fig. 4.3e). In each configuration, four RF excitation patterns were used, namely: circularly polarized (CP) mode, linear mode, zero phase-difference mode, and a random excitation. To fix the phantom in the intended position and orientation during different stages of the validation, four half-elliptic holders were 3D-printed as indicated by black pieces in Fig. 4.3d and 4.3e. In order to keep these holders rigid and stable, six cylindrical rods (white dots in Fig. 4.3d and 4.3e) were 3D-printed as well and passed through the holes on the holders. The coil was driven by eight 2 kW RF amplifiers (Stolberg HF-Technik

AG, Stolberg, Germany) and the excitation parameters (i.e., magnitudes and phases) were measured using an eight-channel oscilloscope. To measure the RF excitation delivered by the system, the outputs at the MR patient table were connected to 50-dB directional couplers (Werlatone Inc, New York, NY) using LMR-400-type cables. We measured the voltage of the attenuated forward signal from the directional couplers with an oscilloscope for all excitation patterns that were used in the B_1^+ mapping study. We disconnected the directional couplers and oscilloscope after we performed voltage measurements.

In addition, to demonstrate the robustness of the EM model to the load variations, the phantom was shifted off-center in the coil and the same B_1^+ mapping experiments were repeated (Fig. 4.3e). All B_1^+ maps were acquired using the actual flip-angle imaging (AFI) technique [111].

For the experimental SAR results, we performed temperature measurements in an RF safety lab as described in a previous work [129]. Eight RF amplifiers (Communications Power Corp., Hauppauge, NY, US) with peak RF power of 500W, operating at 447 MHz. were utilized to deliver RF energy to the same coil-phantom setups, again with various RF excitation patterns matching those used for B_1^+ mapping. The output of the transmit system was confirmed using the same strategy described above using an oscilloscope. Eight fiber optic temperature probes (Lumasense Technologies, CA, USA) were used to measure the temperature inside the phantom at distinct locations (See Fig. 4.3f). The slope of the initial linear regime of the temperature progression curve was later used to estimate the local SAR at the probe tips.

An initial EM simulation was performed using HFSS (ANSYS, Canonsburg, PA, USA) with all lumped elements replaced by excitation ports, similar to the co-simulation method [39]. Subsequently, the root-mean-square error (RMSE) between the simulated and measured complex S-matrices with the centrally positioned phantom (i.e., setup 1) was minimized by optimizing the lumped elements' values using a circuit simulator (AWR Corp., El Segundo, CA, USA), as proposed by Restivo et al. [130]. For this purpose, the gradient descent method [132] was used as the optimization algorithm provided by the circuit simulator. The extension cables (i.e., cables connecting the dipole feed points to the T/R switches) were modeled using the length-adjustable

ideal transmission lines (TL) in the circuit simulator environment and the lengths were treated as variables for the optimization problem. The modeling of the RF coil was complete when the optimum values for the lumped elements and lengths of the TLs were determined.

In the next step, excitation voltages (measured during the B_1^+ and Temperature experiments) were imported to the circuit simulator in order to calculate the complex voltage at the feed point of the dipoles. The calculated signals were then used to drive the coil model in the EM simulation environment. Finally, the corresponding B_1^+ and SAR maps were generated for each of the EM simulation models and the simulated and experimentally measured results were compared. Note that no further EM simulation is required for calculating the B_1^+ and SAR maps since the initial simulation contains all necessary EM solutions. In other words, the calculated port voltages are used as the coefficients of initially solved electric and magnetic fields to calculate the B_1^+ and SAR maps.

4.2.2 RF Safety Analysis with Realistic Human Models

In order to obtain safe RF power limits for human subjects we performed additional EM simulations using the validated coil model and realistic human models. For this preliminary work, we limited the RF excitation to a CP mode. In order to investigate how local SAR levels would vary among different subjects, we ran EM simulations with four different realistic human head models using the CST Studio. The simulated human models included two females and two males: Ella of the virtual population (voxel size, 2x2x2 mm³; height, 1.63 m; weight, 57.3 kg) [114], Donna of CST voxel family (voxel size, 1.875x1.875x2 mm³; height, 1.76 m; weight, 79 kg), Duke of the virtual population (voxel size, 2x2x2 mm³; height, 1.77 m; weight, 70.3 kg) [114] and Gustav of CST voxel family (voxel size, 2.08x2.08x2 mm³; height, 1.76 m; weight, 69 kg). The EM simulations employed about 38x10⁶, 18x10⁶, 57x10⁶, and 17x10⁶ meshes for the models indicated above, respectively. We calculated the maximum 10 g local SAR among the four models and it was used to determine the safe total power limit for head imaging in CP mode.

4.2.3 In vivo Head Imaging and Pulse Sequence Parameters

FDA considers all MR scanners with $B_0 > 8.0$ T to be significant risk devices. Therefore, scanning human subjects at 10.5 T required an Investigational Device Exemption (IDE). Supported by the previous work and additional pre-clinical studies that were performed in our lab [129, 133, 134], the initial research study approved under this IDE had a primary objective of assessing the short term and long term effects of field exposure. This rather comprehensive study explored the cognitive, physiologic and vestibular performance of volunteers during an extensive set of tests totaling four visits over several weeks. During the fourth visit, we had the opportunity to perform a limited imaging study to accomplish the secondary objective, to explore the feasibility of imaging at 10.5 T. Of the 26 volunteers that completed the study, four of them were scanned using our head coil (age, average 31.5 (21-54) years old; weight, average 79 (63.5-90.3) kg; height, average 169.3 (165.1-172.7) cm; one male / three females). The local institutional review board (IRB) approved the protocol and written informed consent was obtained from all participants. The subjects who participated in the safety study were imaged with our safety-validated RF coil allowing us to acquire the first in vivo human brain images at 10.5 T.

We acquired T2-weighted images using the TSE pulse sequence with refocusing $FA = 200^\circ$, $TR/TE = 5000$ ms/72 ms, in-plane resolution = 0.4 mm, slice thickness = 3 mm, acquisition matrix = 512x408, TSE-factor 9, # of averages = 3, and pixel bandwidth = 488 Hz/pixel. In addition, T2*-weighted images were acquired using a GRE pulse sequence with $FA = 15^\circ$, $TR/TE = 200$ ms/20 ms, in-plane resolution = 0.4 mm, slice thickness = 3 mm, acquisition matrix = 512x512, # of averages = 8, and pixel bandwidth = 391 Hz/pixel. The CP mode was used for RF excitation in all imaging sessions. No RF shimming was performed.

4.3 Results

4.3.1 Numerical Validation

Fig. 4.4a-b show the magnitudes and phases of the S-parameters corresponding to the benchtop measurements. We ignored the interaction between the non-adjacent channels (i.e., the simulated maximum coupling between non-adjacent channels was -21 dB) and did not consider them in the validation process. Accordingly, the magnitudes and phases corresponding to S-parameters of non-adjacent channels are shown in gray in Fig. 4.4 and do not necessarily represent the actual values. Fig. 4.4c-d show the S-parameters of the numerical model obtained after optimizing the lumped elements. The RMSE between the measured and simulated S-parameters was 0.33 dB for magnitude and 9° for phase.

Fig. 4.5 shows the B_1^+ maps corresponding to four different excitations obtained from the experiments and simulations. The difference between the measured and simulated maps are given in the third row of Fig. 4.5. The results show a good agreement between simulated and measured B_1^+ maps. In this case, the phantom was positioned at the center of the coil.

To demonstrate the robustness of the validation method, the measured and simulated B_1^+ maps corresponding to the off-center phantom are shown in Fig. 4.6. The results demonstrate that the agreement in results are robust with respect to varying the load from the conditions that were used to match the S-matrix parameters.

Similar agreement was also observed between the B_1^+ maps corresponding to the individual channels obtained from the experiments and simulations while the phantom was positioned at the center of the coil (See Fig. 4.7).

In addition to the visual agreements between the presented B_1^+ maps, quantitative agreement also exists (See Table 4.1). As a parameter of similarity between the measured and simulated maps, the corresponding correlation was computed. Furthermore, normalized root-mean-square error (NRMSE) between the measured and simulated

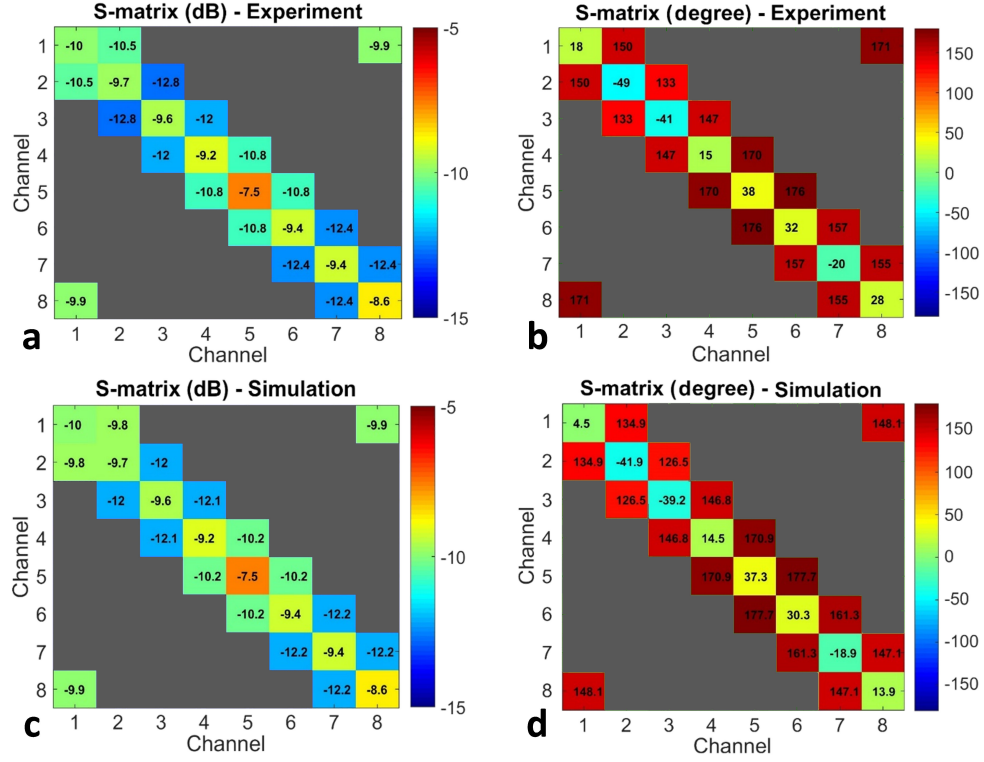


Figure 4.4: Measured and simulated S-parameters. The simulated S-parameters are obtained after the optimization of the lumped elements using the co-simulation model. (a) Magnitude of the measured S-parameters in dB, (b) Phase of the measured S-parameters in degree, (c) Magnitude of the simulated S-parameters in dB, (d) Phase of the simulated S-parameters in degree.

maps is given as a percentage. The NRMSE values show a typical 10% error corresponding to four different shimming scenarios while the typical errors for individual channels are slightly larger (i.e., 15%).

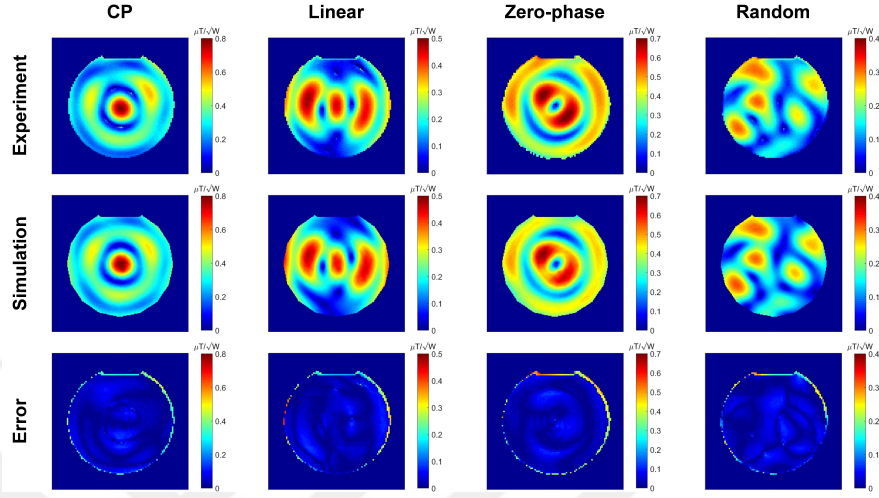


Figure 4.5: Experimental and simulated B_1^+ maps corresponding to four different excitations while the phantom is positioned at the center of the coil. The third row shows the difference maps between the experimental and simulated maps.

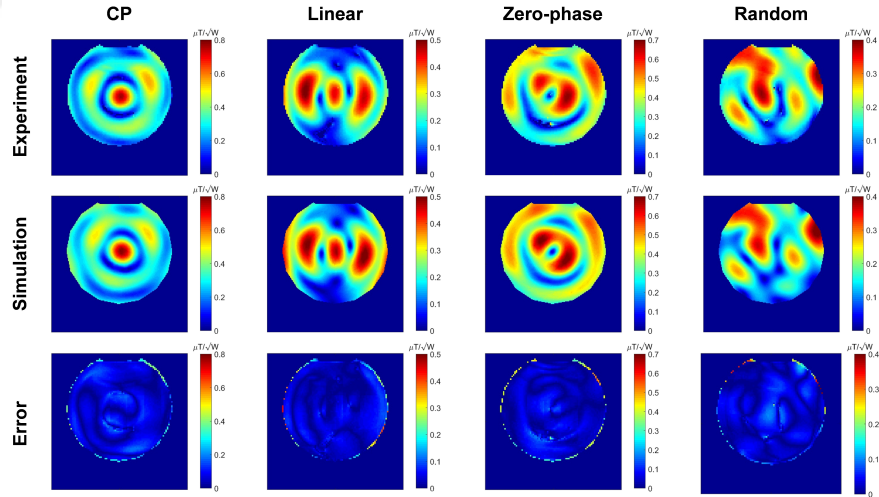


Figure 4.6: Experimental and simulated B_1^+ maps corresponding to four different excitations while the phantom is positioned off-center with respect to the coil. The third row shows the difference maps between the experimental and simulated maps.

Scatter plots of the simulated versus measured B_1^+ maps, where each blue dot corresponds to single voxel, demonstrate that a relatively good voxel-wise agreement between the measured and simulated B_1^+ maps was achieved (See Fig. 4.8 and 4.9).

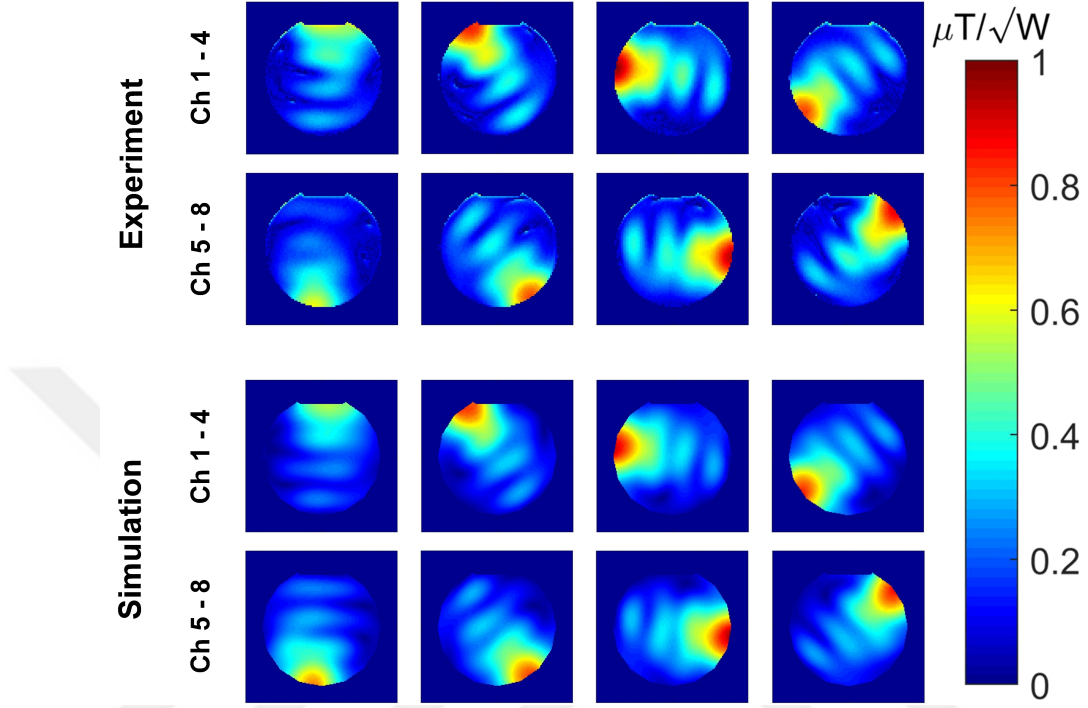


Figure 4.7: Experimental and simulated B_1^+ maps corresponding to eight individual channel excitations while the phantom is positioned at the center of the coil.

Table 4.1: Quantitative comparison between the simulated and measure B_1^+ maps corresponding to Fig. 4.7, Fig. 4.5, and Fig. 4.6.

Off-centered phantom	Random	0.96	8.0
	Zero	0.97	8.3
	Linear	0.97	7.9
	CP	0.94	8.5
Centered phantom	Random	0.94	10.4
	Zero	0.95	10.5
	Linear	0.95	8.8
	CP	0.94	7.5
Individual channels	Ch8	0.92	12.2
	Ch7	0.94	10.7
	Ch6	0.96	9.5
	Ch5	0.94	14.4
	Ch4	0.96	9.4
	Ch3	0.96	9.1
	Ch2	0.95	10.0
	Ch1	0.92	13.6
Correlation			NRMSE (%)

For the next step of the validation procedure, simulated and measured local SAR values were compared. Fig. 4.10a shows the simulated SAR-map while only channels 1 and 5 were excited using unequal signals. Similarly, Fig. 4.10c, and 4.10e show the simulated SAR-maps for two different random excitations. Fig. 4.10b, 4.10d, and 4.10f show a comparison between the numerically calculated and measured SAR values at the tips of the temperature probes. A good agreement between the simulated and measured SAR values was achieved (i.e., RMSE = 0.69 W/kg).

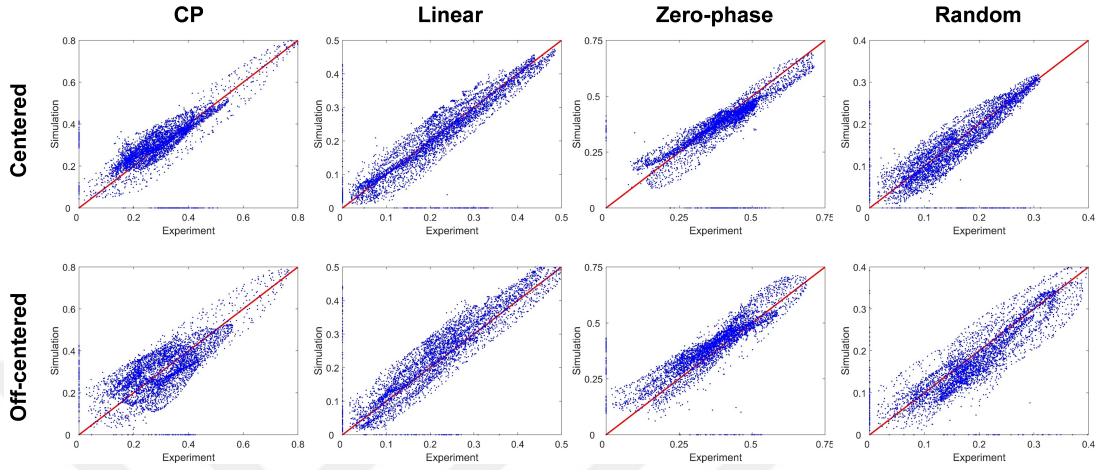


Figure 4.8: Simulated B_1^+ against measured B_1^+ values corresponding to four different excitation scenarios (i.e., CP mode, linear mode, zero phase difference, and random excitations) on the axial plane of the phantom centered (first row) and off-centered (second row) inside the coil.

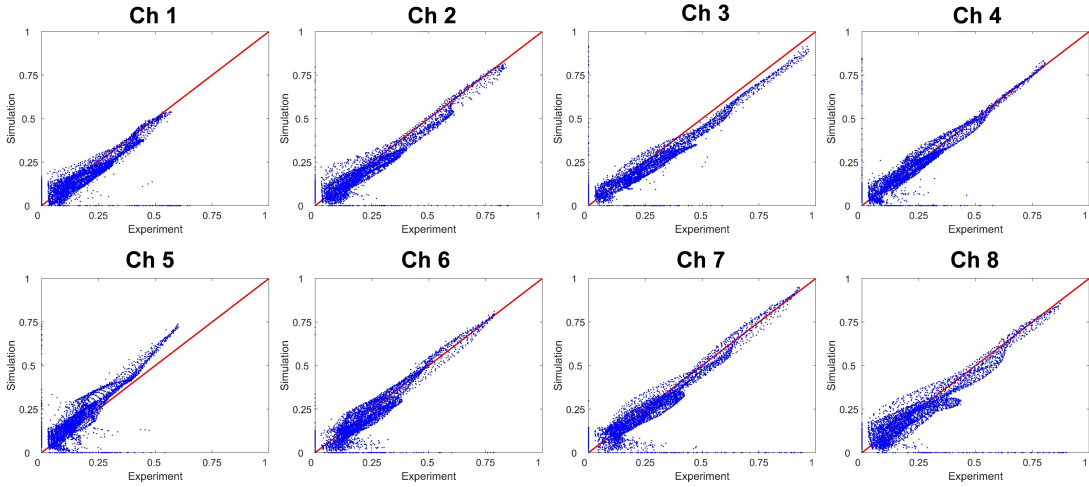


Figure 4.9: Simulated B_1^+ against measured B_1^+ values corresponding to eight individual channel excitations on the axial plane of the phantom centered inside the coil.

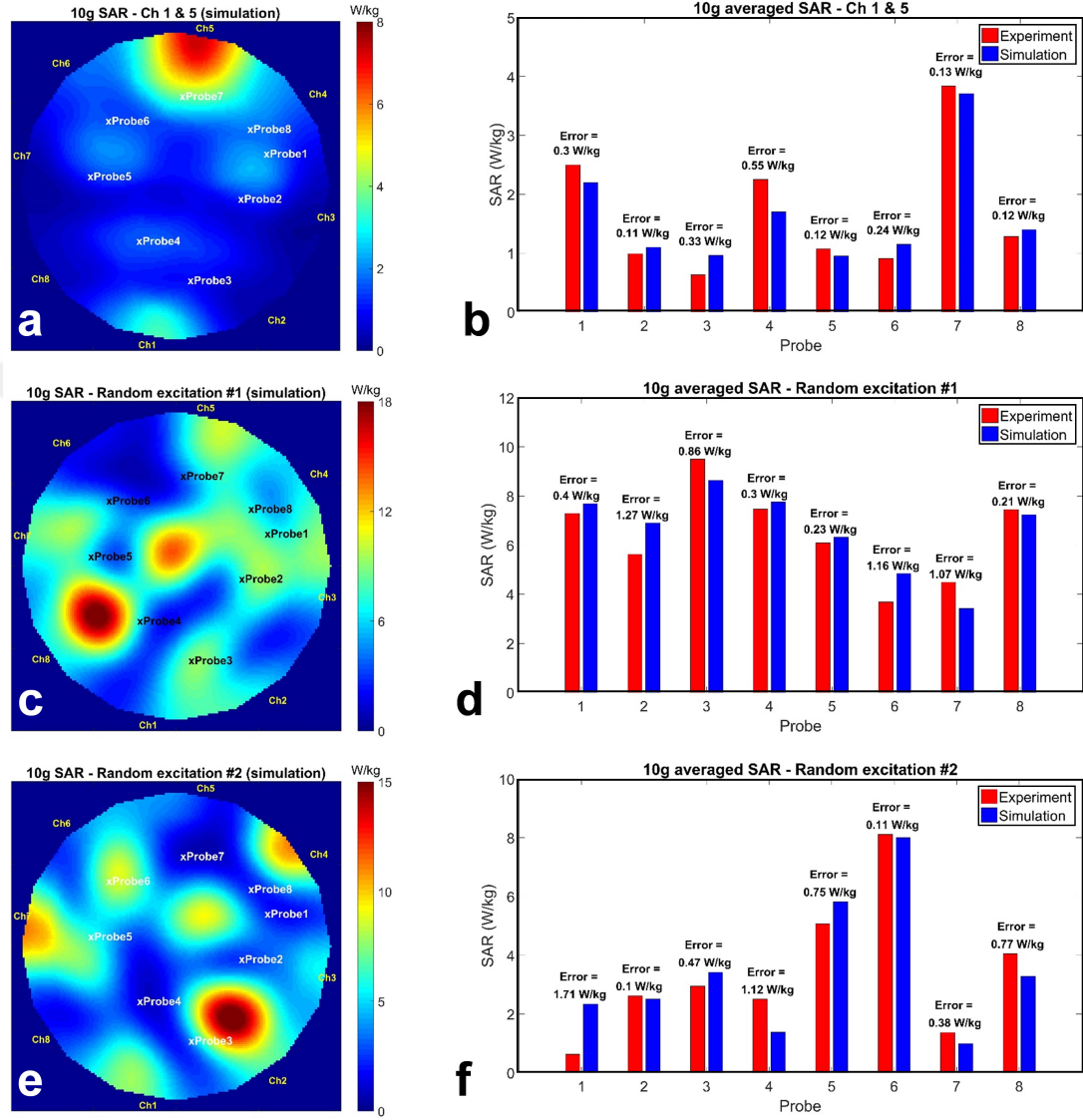


Figure 4.10: Validation using temperature measurement. (a) Simulated SAR-map while only channel 1 and 5 are excited, (b) Comparison between simulated and experimentally estimated SAR at the tip of the temperature probes corresponding to channel 1 and 5 excitation, (c,e) Simulated SAR-maps corresponding to two random excitations, and (d,f) Comparison between simulated and experimentally estimated SAR at the tip of the temperature probes corresponding to two random excitations. Cross signs on the SAR maps are indicating the positions of the temperature probes' tips.

4.3.2 Investigating RF Safety

In this chapter, CP-mode of excitation was chosen for imaging. In order to determine a safe power limit for human head imaging of various subjects with different genders using the CP mode of the transmitter, EM simulations were performed on four different realistic human body models. Fig. 4.11a-d show the 10g-averaged SAR maps for each model on the plane that contains the peak 10g-averaged SAR point for a total input power of 1 W. To be conservative, the CP mode power limit was determined based on the Duke model which exhibited the highest local SAR estimates. According to these results and considering 10 W/kg IEC guideline for local SAR in the head, the maximum allowed total input power to the transmitter was limited to be 25 W for subsequent in vivo human head imaging. In addition, Fig. 4.11e-h show the corresponding simulated B_1^+ maps normalized to a total input power of 1 W. The B_1^+ maps are presented on the axial plane which is fixed with respect to the feed point of the coil for all simulated human models.

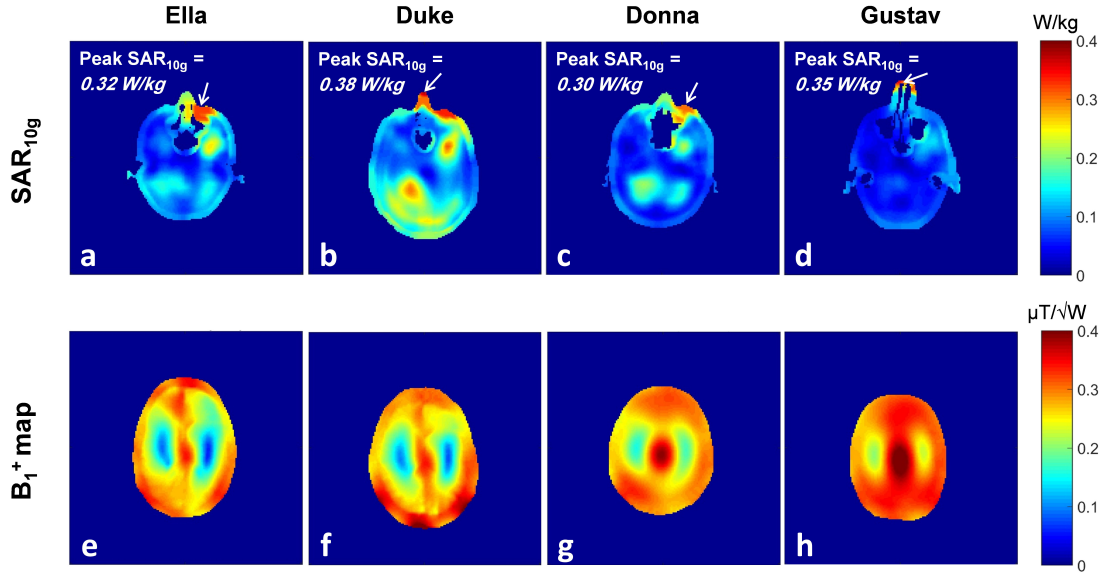


Figure 4.11: **First row:** Simulated 10g-averaged SAR maps for 1W total input power for different realistic human body models at the plane of peak local SAR. **Second row:** Simulated B_1^+ maps for 1W total input power at the same axial plane with respect to the coil elements (i.e., 20 mm above the feed points plane in inferior-superior direction). All maps were determined using the validated coil model with CP-mode excitation.

4.3.3 In Vivo Head Imaging

Using the CP mode of excitation and the total power limit of 25 W determined by the safety studies, we acquired MR images from volunteers at 10.5 T. We focused on two basic pulse sequences for these preliminary studies, namely, T2*-weighted GRE and T2-weighted TSE. Fig. 4.12 shows eight slices of the first T2-weighted human brain images at 10.5 T acquired using the TSE pulse sequence in 611 sec. Fig. 4.13 shows one slice among seven slices of the first T2*-weighted human images at 10.5 T acquired using the GRE pulse sequence in 430 sec.

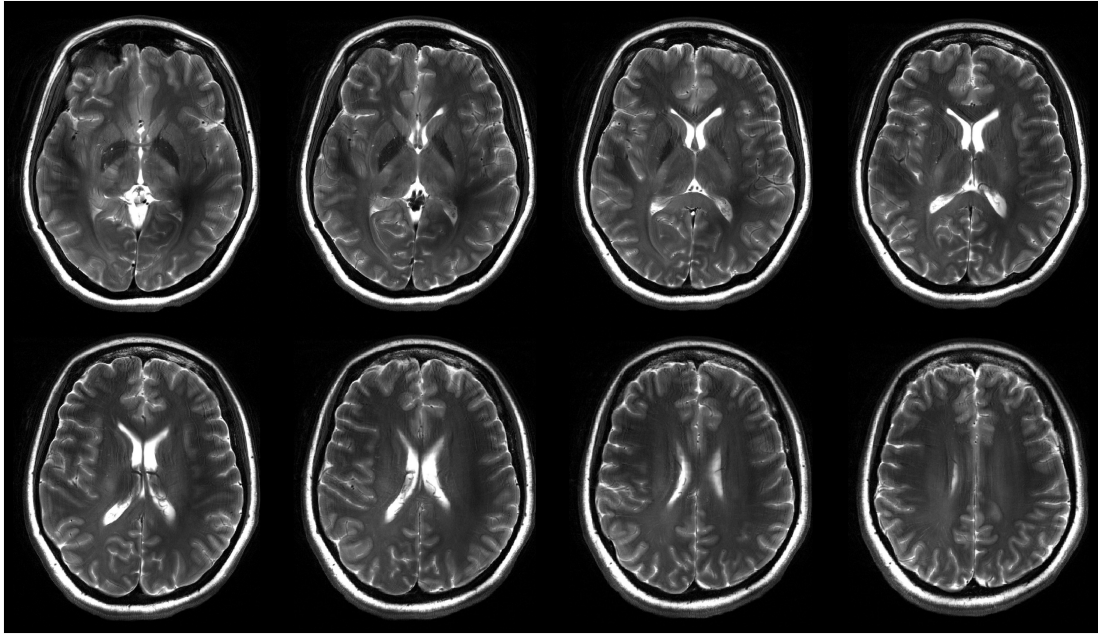


Figure 4.12: The first T2-weighted human brain image at 10.5 T acquired using the TSE pulse sequence with TR/TE = 5000 ms/72 ms. Matrix = 512x408, in-plane resolution = 0.4 mm, slice thickness = 3 mm, # of averages = 3, pixel bandwidth = 488, and total scan time of 611 sec.

The T2-TSE images presented with expected tissue contrast between white and gray matter. GRE images revealed some of the venous structures in the brain due to T2* contrast. For both studies, an acceptable signal homogeneity was achieved across the slices of interest.

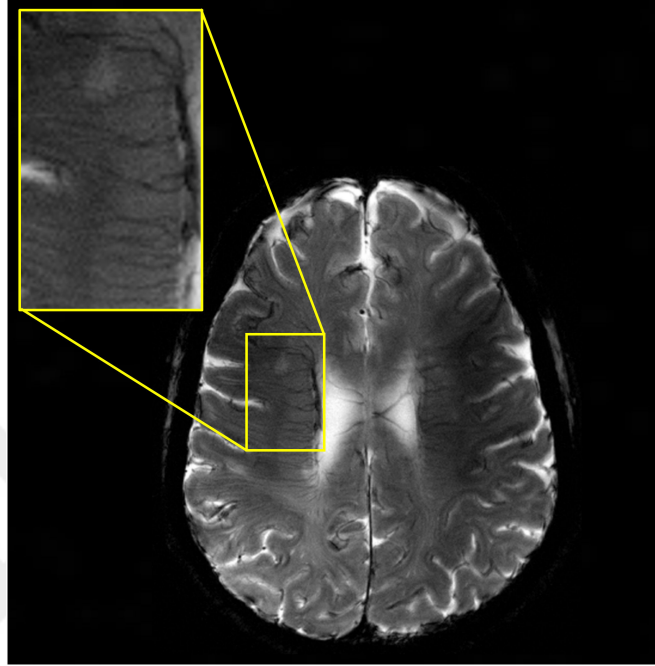


Figure 4.13: The first T2*-weighted human brain image at 10.5 T acquired using the GRE pulse sequence with $FA = 15^\circ$, $TR/TE = 200 \text{ ms}/20 \text{ ms}$, matrix = 512×512 , in-plane resolution = 0.4 mm, slice thickness = 3 mm, # of averages = 8, pixel bandwidth = 391, and total scan time of 430 sec.

4.4 Discussion and Conclusions

In this chapter, we acquired the first in vivo human head image at 10.5 T using an eight-channel Tx/Rx coil. As a part of the RF-safety assessment, we expanded a previously proposed workflow for validating EM simulations of multi-channel arrays. Using the validated simulation model of the coil, we calculated safe power limits for human head imaging and acquired the first in vivo T2- and T2*-weighted human brain images at 10.5 T.

The workflow requires constructing an accurate EM simulation model starting from high-resolution segmented CT images of the setup (i.e., the coil and the phantom). The lumped elements of the coil are optimized using the co-simulation method, as proposed by Restivo et al. [130], to minimize the RMS of the difference between simulated and measured S-parameters. This is a critical step that is necessary to accurately model

EM field quantities observed in the phantom. It is also necessary to capture the complex coupling behavior in multi-channel arrays. To avoid a large and time-consuming optimization problem, we imposed a limit of 50% deviation from the nominal lumped elements values used in the actual coil. In this work, the simulated capacitor values resulting from the optimization were between 10 to 15 pF, while the nominal (actual) values were 15pF.

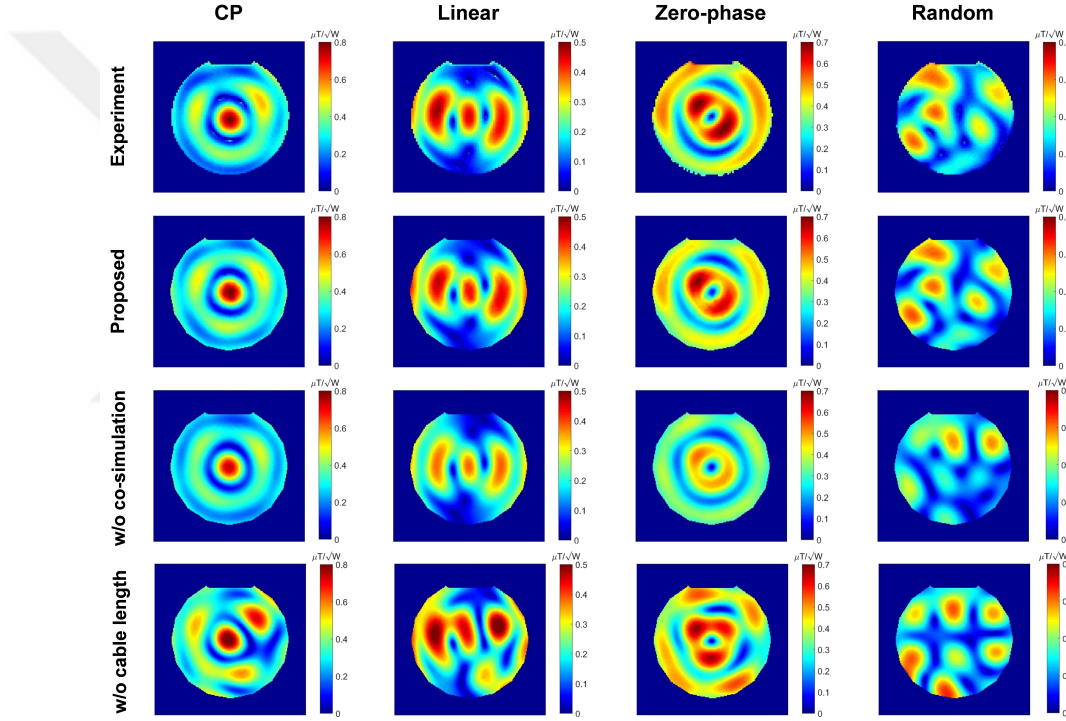


Figure 4.14: B_1^+ maps corresponding to four different excitations while the phantom is positioned at the center of the coil. First row: experimental results. Second row: simulated results obtained using the workflow proposed in this chapter. Third row: simulated results obtained by excluding the co-simulation optimization from the workflow proposed in this chapter. Fourth row: simulated results obtained using the workflow proposed in this chapter and ignoring the cable lengths.

To show the importance of optimization of the lumped elements, we omitted this optimization from the modeling part presented in the workflow. In other words, we plugged the nominal values of the lumped elements (i.e., the capacitors which are used for tuning and matching of the coil in the real-life) into the EM simulations without any further adjustment. The third row in Fig. 4.14 shows the B_1^+ maps corresponding

to four different excitations with the coil model resulting from using this method (i.e., excluding the co-simulation from the modeling part). The correlation and NRMSE between the B_1^+ maps in this scenario and the experimental B_1^+ maps are given in Table 4.2.

Table 4.2: Quantitative comparison between the simulated and measure B_1^+ maps corresponding to Fig. 4.14 (third and fourth rows).

	Without co-simulation				Without cable lengths			
	CP	Linear	Zero	Random	CP	Linear	Zero	Random
Correlation	0.91	0.74	0.53	0.47	0.83	0.7	0.51	0.46
NRMSE (%)	12.2	18.9	20.7	29.1	15.8	22.9	21.5	31.4

In addition to optimizing the lumped elements' (i.e., capacitors) values, we also considered the length of extension cables as an optimization parameter. Inclusion of the cable lengths in the optimization significantly af-

fects the phases of the complex S-parameters. The effect of minimizing the error between measured and simulated phases is directly reflected in the validation part as shown in Fig. 4.14. The fourth row in Fig. 4.14 shows the B_1^+ maps corresponding to four different excitations with the coil model resulting from excluding the cable length optimization from the presented workflow. The correlation and NRMSE between the B_1^+ maps in this scenario and the experimental B_1^+ maps are given in Table 4.2.

In the previous work presented by Restivo et al. [130], the major criterion for the validation of the EM model is the comparison between the simulated and measured B_1^+ maps corresponding to some phase-only shimming scenarios such as CP-mode excitation. However, relatively small errors in such B_1^+ maps may result in relatively large errors in the peak SAR predictions. Fig. 4.15 shows the statistical distributions of the error in the peak SAR predictions corresponding to the typical error values between the measured and simulated B_1^+ maps (i.e., NRMSE between simulated and experimental maps). These results were obtained using the Monte-Carlo method for four different types of excitations while the phantom was positioned in the center of the coil. For each excitation scenario, four different percentiles of errors in B_1^+ maps were propagated into the peak SAR prediction errors. For typical 10% B_1^+ error, the peak SAR underestimation ranges between 9% and 31% for the zero-phase excitation and the random excitation, respectively. The peak SAR underestimation becomes significantly larger if the error in B_1^+ maps increases. These effects were carefully considered prior to in-vivo studies while choosing a safety factor for our imaging studies.

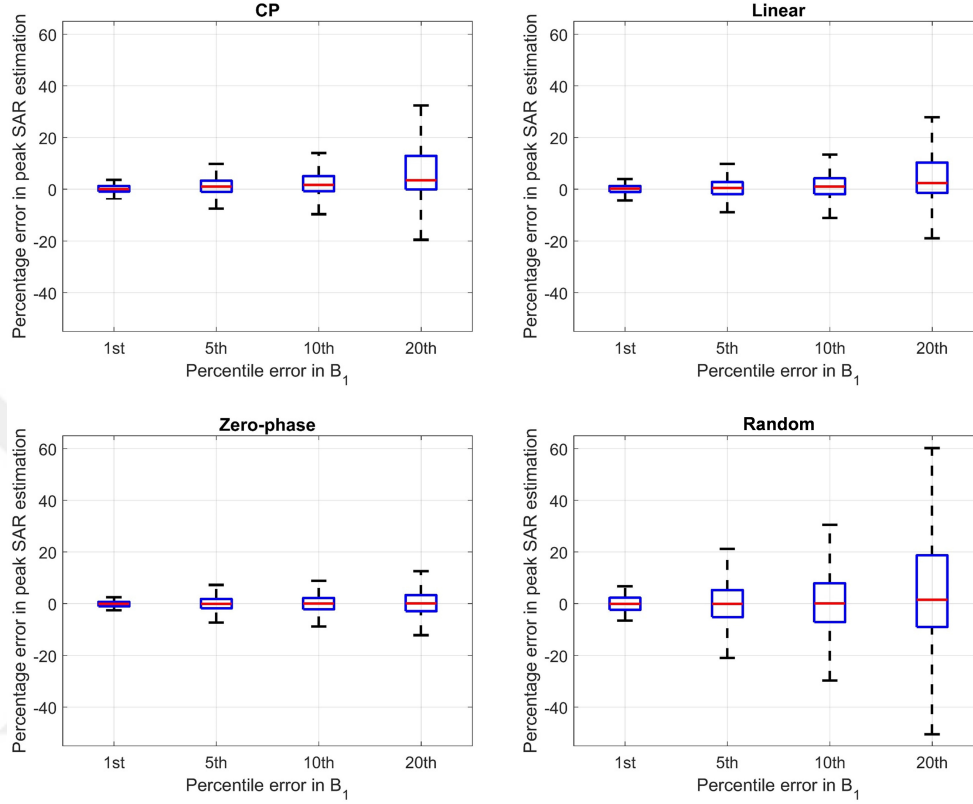


Figure 4.15: Statistics of the error in the peak SAR10g estimation corresponding to the possible error occurrence in B_1^+ map simulations for four different excitation scenarios (i.e., CP mode, linear mode, zero phase difference, and random excitations) inside the phantom positioned at the center of the coil. The Monte-Carlo simulation has been done using one million random perturbations in excitation signal of each individual channel (in each excitation scenario) then propagation these perturbations to the error in B_1 maps and peak SAR estimation.

The workflow is implemented to validate the EM simulation models of multi-channel coil arrays. Once the validation is complete, the coil model can be used to calculate field patterns in a different load. This concept was tested by using an off-centered phantom. The validated model accurately predicted B_1^+ field patterns in the shifted phantom without requiring additional optimization of the lumped elements a second time.

To demonstrate the accuracy of the workflow, we validated an eight-channel bumped dipole array [79] at 10.5 T. Using the validated coil model, we performed additional EM simulations with four realistic human head models to determine the

safe RF power limits. Although differences in the SAR distribution were observed from subject to subject, the inter-subject variation of peak 10 g averaged SAR was relatively small (i.e., 27 %) compared to the similar value, 76%, reported by de Greef et al. [135].

In previous studies, safety factors/margins were proposed based on inter-subject variation of peak local SAR [135–137] as well as variation with respect to EM properties of tissues and model translation [137–139]. In this work, we used a safety buffer based on the RF loss in the pathway from the RFPAs to the feed point of the dipoles, which was measured as 3.9 dB. Considering the inter-subject variability [135–139], this 3.9 dB safety buffer along with the calculated safe power limit based on our four-model worst-case scenario ensures safe use of the CP mode at 10.5 T in vivo imaging.

We achieved good signal homogeneity with the TSE sequence as shown in Fig. 4.12. The following factors contributed to this result:

- 1) With multiple refocusing pulses, the relationship between signal and flip angle is close to $\sin(\alpha/2)$. This allowed us to achieve a more homogeneous image intensity compared to a conventional spin-echo sequence [140].
- 2) The transmit profile is strongest in the center whereas the receiver profile is strongest in the periphery. The image intensity depends on both, allowing us to achieve a more homogeneous image intensity than either would produce individually.
- 3) Targeting slightly higher than 180° flip angle at the center of the brain resulted in an extended region in the brain with acceptable homogeneity.

In order to demonstrate this, we simulated (for the DUKE model) the expected image intensity variation in a T2-weighted TSE image with the CP mode. Fig. 4.16a shows the simulated Tx profile in CP mode in the presence of the realistic human head model (Duke). Fig. 4.16b shows the Rx profile obtained by sum-of-square of Rx sensitivity map of each individual channel. Fig. 4.16c shows the T2 map of the same slice. Fig. 4.16d demonstrates the simulated T2-weighted image obtained by multiplying the $\sin(\alpha/2)$, Rx profile, and $\exp(-TE/T2)$. Note we used the same TE as in Fig. 4.12. Fig. 4.16e shows the in vivo T2-weighted image on the similar slice.

As shown in Fig. 4.16, similar homogeneity and contrast is achieved in both simulation and experimental images.

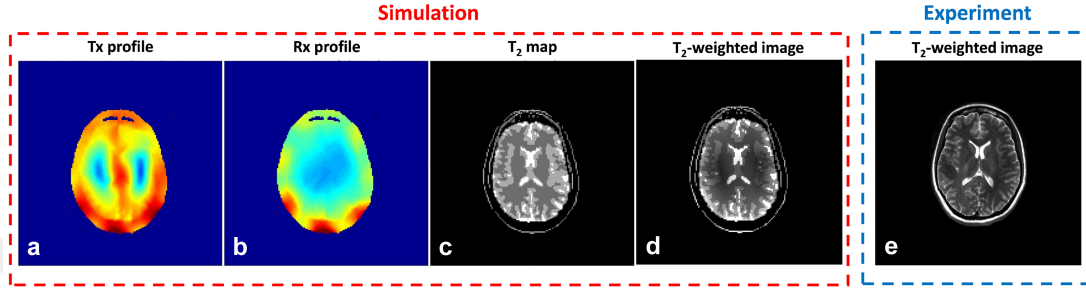


Figure 4.16: Simulated and experimentally acquired T2-weighted TSE image under the CP mode excitation. (a) Simulated Tx profile (b) Simulated Rx profile (c) T2 map of the realistic model (Duke) in the simulation environment (d) Simulated T2-weighted image (e) Experimentally acquired T2-weighted image.

In this chapter, we used bench-top measurements of S-parameters for modeling our setup and demonstrated good agreement between experimental and simulated results. An alternative approach is to measure the S-matrix inside the scanner as demonstrated by Restivo et al. [130]. However, this approach requires additional hardware modification on the scanner (i.e., the installation of directional couplers).

A previously presented workflow [130] is expanded for validating EM simulations of multi-channel arrays. The workflow is used to validate the safety of an eight-channel Tx/Rx coil and to calculate safe power limits for human head imaging at 10.5 T. Using the validated array, we acquired the first in vivo human brain images at 10.5 T. Future studies will focus on utilizing advanced RF parallel transmit [17, 141] methods as well as design and validation of novel coil arrays with higher channel count.

Chapter 5

A Highly-Dense Transmit Array Coil for 10.5 T: An Introduction to a Non-Uniform Dielectric Substrate (NODES) Antenna

Preface

The content of this chapter was presented in part at the 27th Annual Scientific Meeting of International Society of Magnetic Resonance in Medicine (ISMRM) [87] and Americas Committee for Treatment and Research in Multiple Sclerosis [142]. This is a collaborative work between UMRAM of Bilkent University and CMRR of University of Minnesota. Steve Jungst, Xiaoping Wu, Lance Delabarre, Mike Lanagan, Gregor Adriany, Gregory J. Metzger, Pierre-Francois Van de Moortele, Kamil Ugurbil, and Yigitcan Eryaman contributed to this study. Xiaoping Wu, Gregor Adriany, Gregory J. Metzger, Pierre-Francois Van de Moortele, Kamil Ugurbil, Ergin Atalar, and Yigitcan Eryaman were involved in study conception and design. Steve Jungst contributed to fabrication of the RF coil. Lance Delabarre and Gregory J. Metzger were responsible for B_1^+ -mapping and cadaver imaging. Mike Lanagan provided high-dielectric

materials.

5.1 Introduction

Ultra-high field (UHF, $B_0 \geq 7$ T) MRI is drawing considerable attention worldwide due to its numerous benefits, such as an increase in signal-to-noise ratio (SNR) [5–10] and significant improvement in susceptibility contrast [11, 12]. However, a higher magnetic field (B_0), which exploits the radiofrequency (RF) excitations with shorter electromagnetic (EM) wavelengths, results in highly variant excitation profiles, and consequently, degradation in overall image quality. On the other hand, constructive interference of the RF electric field can cause high local specific absorption rate (SAR) levels, which is a crucial safety concern at UHF MRI. Transmit array (TxArray) technology is essential to take full advantage of the UHF MRI as well as to manage SAR. Previous studies [78] demonstrated the impact of increasing the number of transmit elements in an array for improving SAR performance as well as flip angle homogeneity. However, the size of the elements and coil coupling pose practical limitations on the maximum number of elements that can be used in a TxArray coil.

In the literature, various types of transmitting elements such as loop coils [46–51], transmission lines [10, 15, 52–54], and dipole-like structures [55–60, 80] are investigated for UHF-MRI. Thalhammer et al [46][11] used TxArray of loop coils for cardiac imaging at 7 T. Adriany et al [15] employed an array of the transmission lines for head imaging at 7 T. Vaughan et al [53] built a transverse electromagnetic (TEM) coil for whole-body imaging at 7 T. Raaijmaker et al [55, 56] introduced single-side adapted and fractionated dipole as new transmitter elements for body imaging at 7 T. Steensma et al [57, 58] improved the SAR performance of the dipoles at 7 T and 10.5 T by introducing the snake antenna. Vaughan et al [10] utilized a TEM coil for the head imaging at 9.4 T. Shajan et al [48] performed head imaging at 9.4 T using the array of loop elements. Erturk et al [59] proposed to use an array of fractionated dipoles for torso imaging at 10.5 T. Sadeghi-Tarakameh et al [80] proposed to use an array of bumped

fractionated dipoles for head imaging at 10.5 T. RF coil development is still essential for realizing the full potential of UHF MRI. Improving the performance of RF coil arrays improves UHF imaging as demonstrated in numerous studies in the literature.

In this chapter, we propose a new dipole-like antenna [87] design utilizing a non-uniform dielectric substrate placed underneath the element to optimize B_1^+ -SAR efficiency. This modification increases the uniformity of the current distribution on the dipole and reduces local SAR. As a result, it enables us to design highly-dense transmit arrays using shorter elements. For proof of concept, we performed an optimization on a dipole at 10.5 T using the EM simulations. Also, we constructed the corresponding structure and conducted a phantom experiment on a whole-body 10.5 T scanner to validate the EM simulation results. In addition, we built a nine-channel transmit/receive (transceiver) coil and acquired human cadaver spine images at 10.5 T.

5.2 Theory and Methods

Decreasing the length of a dipole enforces the current to rapidly descend towards the ends (current vanishes at two ends of the dipole due to the high impedance at these points). On the other hand, based on Maxwell's equations, when the current rapidly changes along the conductor, the transverse electric field and SAR in the tissue is elevated as a result. Placing high-permittivity blocks at two ends of the dipole can potentially increase the effective capacitance between these points and the body. As a result, the corresponding impedance can be decreased. This effect in turn, should increase the uniformity of the current on the dipole and should reduce the electric field underneath the dipole. On the other hand, the B_1^+ generated by this current becomes considerably uniform in the z-direction.

5.2.1 Numerical Optimization of the NODES Antenna's parameters

To determine the optimum design parameters, a deep-body target (e.g., prostate imaging) is defined, and accordingly, the following optimization problem is investigated by sweeping the corresponding parameters (i.e., length of the conductor, width of the conductor at two end-points, the separation between the conductor and the sample, dielectric constants of the segmented substrate) in a reasonable range.

$$\text{maximize} \quad \text{Mean} \{ \xi(r) \text{ at the depth of } 50 \text{ to } 100 \text{ mm} \} \quad (5.1)$$

where $\xi(r)$ represents the B_1^+ -SAR efficiency at the position r , and can be defined as

$$\xi(r) = \frac{B_1^+(r)}{\sqrt{pSAR_{10g}}} \quad (5.2)$$

where $pSAR_{10g}$ is the peak local 10g-averaged SAR.

We performed 42875 fast EM simulations (i.e., coarse meshing) with a finite-element method (FEM)-based simulator (HFSS, ANSYS, Canonsburg, PA, USA). Such a high number of simulations came to the picture due to the wide range of sweep performed on the optimization parameters. As shown in Fig. 5.1a, six different parameters of the antenna, i.e., length of the conductor, width of the conductor at two end-points, the conductor-sample separation, and dielectric constants of the segmented substrate at two end-points, middle and the rest, were swept in the EM simulation environment with 5, 5, 5, 7, 7, and 7 values respectively.

As a result of the optimization, a non-uniform dielectric substrate (NODES) antenna was designed. Fig. 5.1a shows the structure we used for the optimization purpose as well as the search interval and the optimum values of the investigated parameters. A cubic phantom with relative permittivity of 78.3 and conductivity of 0.66 S/m was used to mimic the human body (these values imposed to the simulation to match with the experimentally measured phantom properties). Considering some space issues and also for the ease of fabrication, the element shown in Fig. 5.1b-d was built and validated.

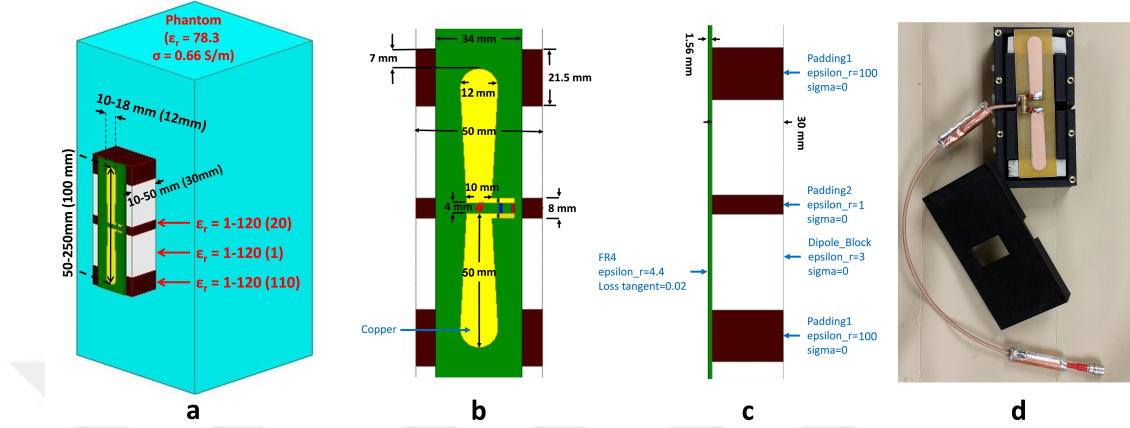


Figure 5.1: NODES antenna. (a) The structure used for optimizing the NODES antenna by sweeping the pre-determined parameters in the EM simulation environment. (b-c) The NODES antenna with the optimal parameters. (d) The constructed single element NODES antenna.

For the experimental setup, we employed a high-dielectric constant block made of TiO_2 ($\epsilon_r=100$, $\sigma=1.1 \text{ mS/m}$, measured at 447 MHz) in the construction of the NODES antenna (Fig. 5.1d). 30mm-thick housing of the NODES antenna was 3D-printed using polylactic acid (PLA) material with $\epsilon_r=2.8$, $\sigma=0.1 \text{ mS/m}$.

5.2.2 Element-wise Comparison

The comparison was performed between the NODES antenna, the fractionated dipole (FD) [59], and the loop coil [143], as shown in Fig. 5.2. For this purpose, a torso-sized elliptical body phantom (450x180x290 mm³) [110] fulfilled of hydroxyethyl cellulose (HEC) and 2.9 g/L NaCl with electrical properties of $\epsilon_r=78$ and $\sigma=0.66 \text{ S/m}$ was used. The fractionated dipole and loop coil consisted of conductors on an FR4 PCB mounted on a PETG block with a thickness of 10 mm. All three elements were matched to 50 Ω with a reflection coefficient of less than -15 dB.

To validate the numerical simulation results, B_1^+ -power efficiency maps of the three elements (see Fig. 5.2) were acquired numerically and experimentally on both axial and sagittal planes.

All computations and numerical simulations were performed using a commercial EM simulator (HFSS, ANSYS, Canonsburg, PA, USA) on a mediocre workstation with two quad-core Intel(R) processors with a 3.4 GHz clock rate and 128 GB RAM.

Experimental studies with the above setup were conducted on a whole-body 10.5 T Magnet (Agilent Technologies, Oxford, UK) and associated imaging system (Siemens Healthineers, Erlangen, Germany). The scanner is equipped with a 16-channel parallel transmit system with each channel driven by a 2-kW RF power amplifiers (Stolberg HF-Technik AG, Stolberg, Germany). Transmit B_1^+ maps were acquired using the actual flip-angle imaging (AFI) technique [111].

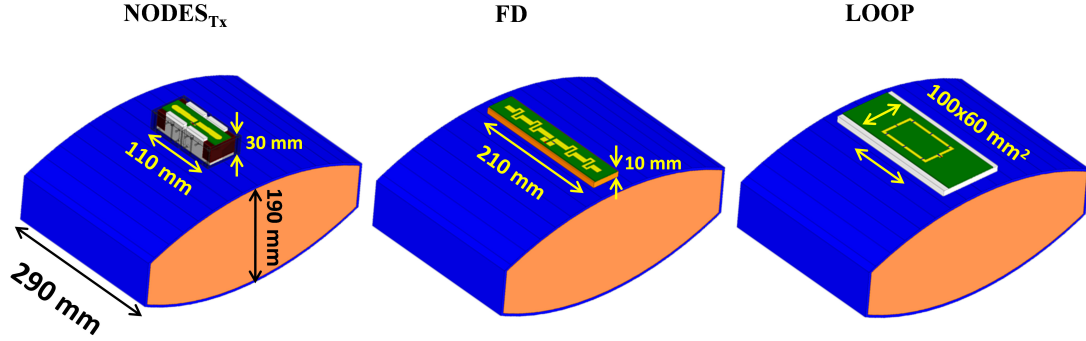


Figure 5.2: The NODES antenna, a fractionated dipole, and a loop coil were placed on an elliptical uniform phantom and compared to each other in terms of the transmission and reception performance.

To show the improved local SAR performance of the NODES antenna compared to the FD and the loop coil, B_1^+ -SAR efficiency of the three structures were obtained using EM simulations and compared numerically.

Furthermore, the NODES antenna with reduced patient-coil-separation (20 mm) was used as a receive (Rx) element in order to investigate the reception performance of it. Accordingly, the FD and loop coil, shown in Fig. 5.2, were selected to be compared to the receiver NODES antenna. For this purpose, the intrinsic SNR (ISNR) maps of the three structures were obtained numerically. ISNR, at the position r can be defined as [6]

$$\Psi(r) = \frac{2\omega M_0 B_1^-(r)}{\sqrt{4k_B T R}} \quad (5.3)$$

where ω is the angular frequency, M_0 is the magnitude of the longitudinal magnetization vector, B_1^- is the right-hand circularly polarized component of the RF magnetic field, k_B is the Boltzman constant, T is the sample temperature, and R is the real part of the input impedance seen from the ports of the coil. Consequently, the only coil-dependent parameters that affect the ISNR are B_1^- and R .

5.2.3 A Nine-channel Tx/Rx NODES Spine Array

Performance of a nine-channel Tx/Rx array composed of six NODES antennas with the 30mm-thick block (i.e., optimized for local B_1^+ -SAR efficiency) and three NODES antenna with the 20mm-thick block (i.e., improved SNR performance) in the spine imaging was investigated numerically using a FIT (finite integration technique)-based time-domain solver of an EM simulator, CST Studio Suite 2019 (CST, Darmstadt, Germany), as shown in Fig. 5.3a. We also built the nine-channel array, shown in Fig. 5.3b, for spine imaging. Channels 1 through 6 are NODES antennas with 30mm-thick block, and channel 7 through 9 are NODES antennas with 20mm-thick block. All elements were matched to better than -9.5 dB and the highest coupling between the elements was -12.5 dB. Two floating current traps were used on each coaxial cables connected to feed-points to prevent the unbalanced current flow on the outer conductors.

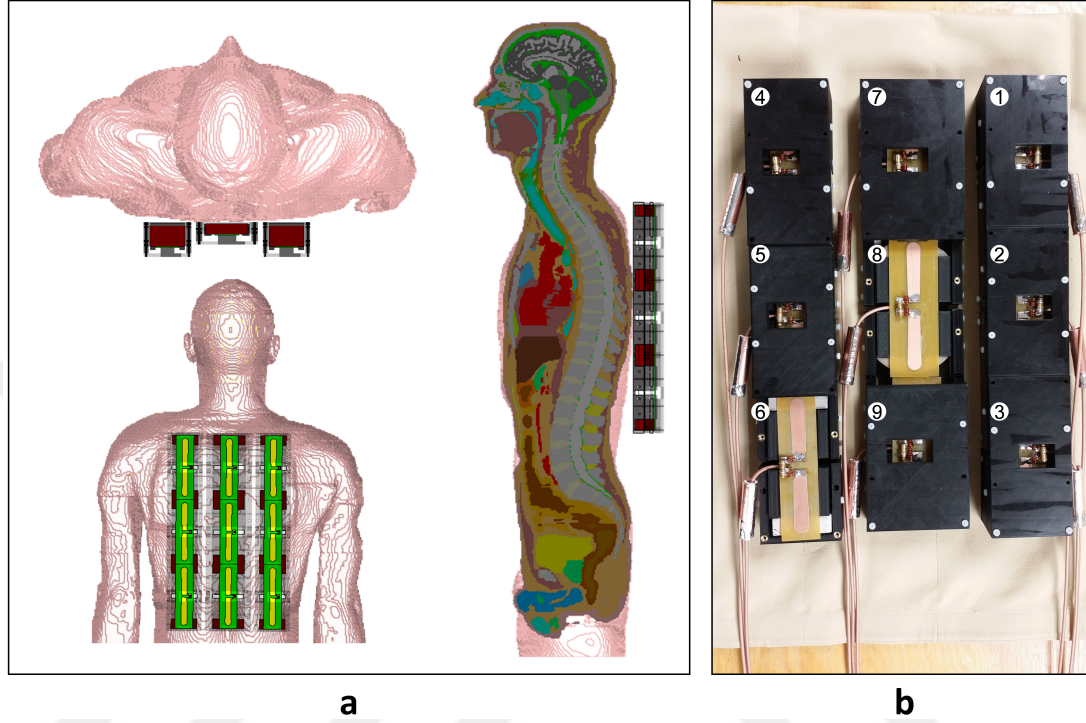


Figure 5.3: A nine-channel Tx/Rx NODES spine array. (a) EM simulation setup using the NODES spine array in presence of realistic human body model (Duke). (b) Home-built nine-channel NODES spine array.

5.2.4 B_1^+ Shimming and Human Cadaver Spine Imaging

We performed a phase-only shimming [16] with the goal of minimizing B_1^+ inhomogeneity over the spine (defined as the ROI) in both the simulation environment and the cadaver experiment.

We imported the realistic human model, Duke of the virtual population (voxel size, 2x2x2 mm³; height, 1.77 m; weight, 70.3 kg), to the CST Studio Suite and performed the corresponding EM simulations (see Fig. 5.3a).

We acquired whole lumbar and thoracic spine images in a single acquisition using the FLASH pulse sequence with refocusing FA = 20°, TR/TE = 168 ms/3.69 ms, in-plane resolution = 0.5 mm, slice thickness = 2 mm, acquisition matrix = 576x432, # of averages = 2, and pixel bandwidth = 212 Hz/pixel. In addition, T2*-weighted images

were acquired using a MEDIC pulse sequence with FA = 30°, TR/TE = 500 ms/19 ms, in-plane resolution = 0.24 mm, slice thickness = 2 mm, acquisition matrix = 640x640, # of averages = 4, and pixel bandwidth = 244 Hz/pixel.

5.3 Results

5.3.1 Element-wise Comparison

Fig. 5.4 and Fig. 5.5 show the B_1^+ -power efficiency of the three transmit elements (i.e., NODES antenna, FD, and loop coil) on axial and sagittal planes, respectively. In each figure, the first row represents the simulation results, whereas the second row corresponds to the experimental B_1^+ -power efficiency maps. A good agreement between the simulation and experimental results is achieved.

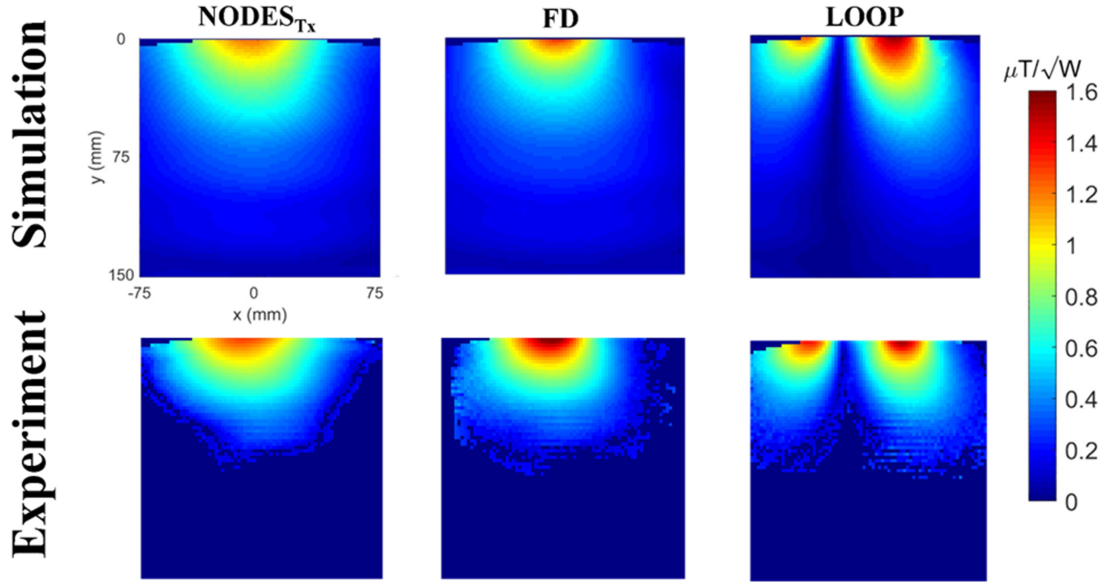


Figure 5.4: Validation of the numerical results. Axial B_1^+ -power efficiency maps of the three transmit elements obtained numerically and experimentally.

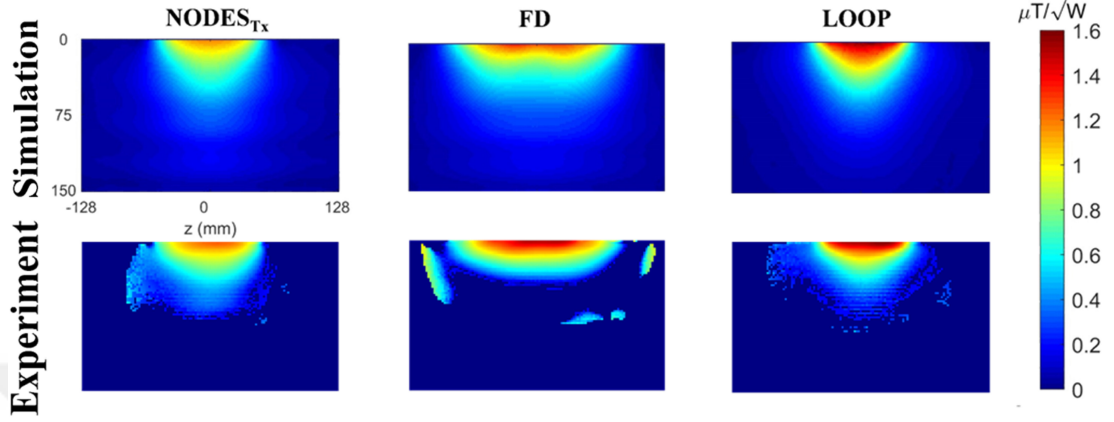


Figure 5.5: Validation of the numerical results. Sagittal B_1^+ -power efficiency maps of the three transmit elements obtained numerically and experimentally.

Fig. 5.6 a-c show the axial B_1^+ -SAR efficiency maps of the three transmit elements obtained using the EM simulations. To compare the SAR performance of these elements in-depth, the corresponding B_1^+ -SAR efficiency efficiencies are plotted over the dashed lines shown in Fig. 5.6a-c. Improvement in B_1^+ -SAR efficiency of the NODES antenna with respect to the fractionated dipole, and loop coil is shown in Fig. 5.6d. According to Fig. 5.6d, the NODES antenna outperforms both the fractionated dipole and loop coil at the depths greater than 50 mm (i.e., the depths-of-interest in the optimization problem).

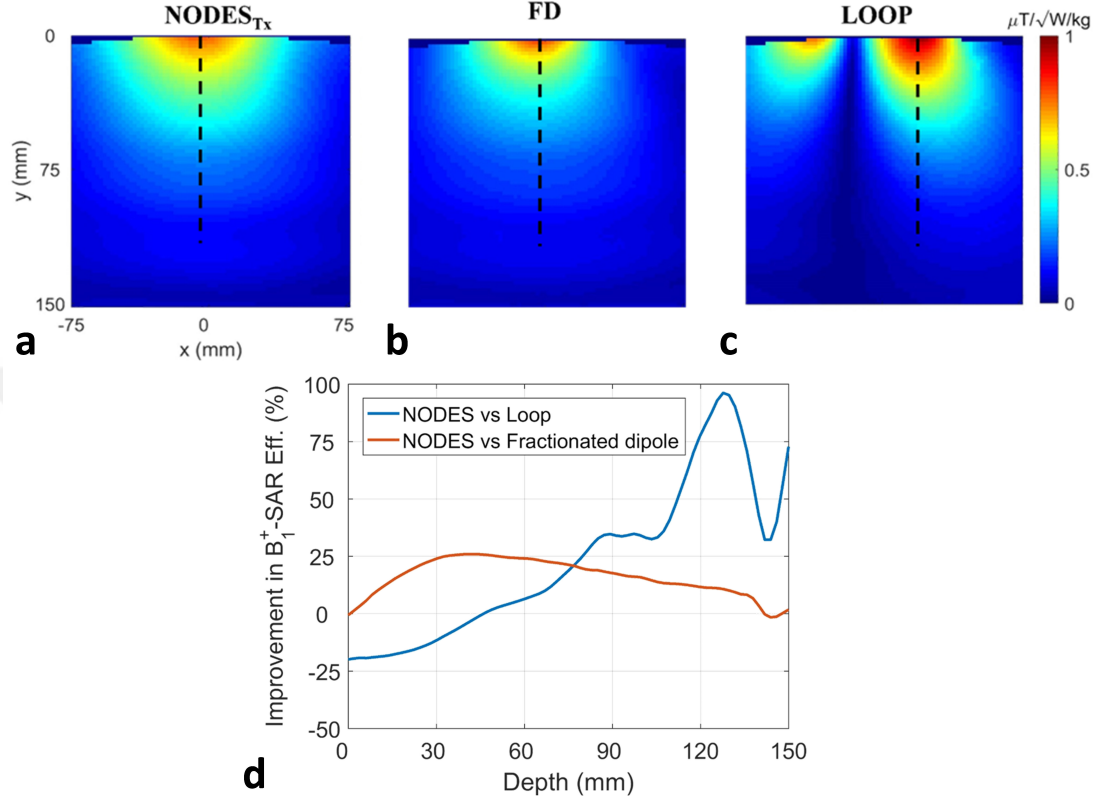


Figure 5.6: B_1^+ -SAR efficiency comparison. Axial B_1^+ -SAR efficiency map of (a) NODES antenna, (b) fractionated dipole, and (c) loop coil. (d) B_1^+ -SAR efficiency improvement of the NODES antenna in depth with respect to the fractionated dipole and loop coil.

Similarly, the reception performance of the three elements is compared through their intrinsic SNR (ISNR) plots. Fig. 5.7a-c show the axial ISNR maps corresponding to the NODES antenna (20mm-thick block), fractionated dipole, and loop coil, respectively. Improvement in ISNR of the NODES antenna with respect to the fractionated dipole and loop coil over the dashed lines (see Fig. 5.7a-c) is plotted in Fig. 5.7d. The NODES antenna outperforms both the fractionated dipole and loop coil in all depths.

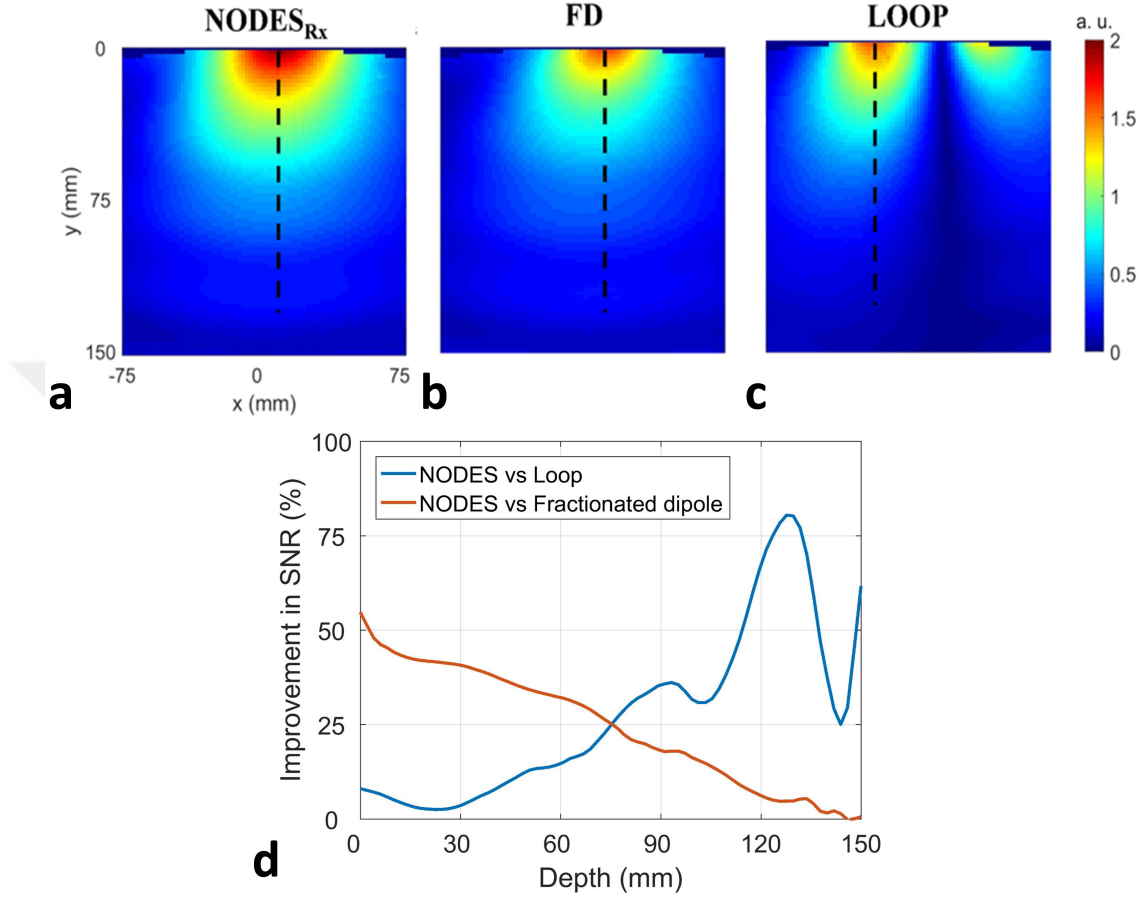


Figure 5.7: ISNR comparison. Axial ISNR map of (a) NODES antenna, (b) fractionated dipole, and (c) loop coil. (d) ISNR improvement of the NODES antenna in depth with respect to the fractionated dipole and loop coil.

5.3.2 Human Cadaver Spine Imaging

Using the phase-only shimming over the spine of a cadaver with the goal of minimum excitation inhomogeneity, we acquired MR images from a cadaver at 10.5 T. We focused on two basic pulse sequences for these preliminary studies, namely, FLASH and T2*-weighted MEDIC. Fig. 5.8 shows a sagittal view of the cadaver lumbar and thoracic spine image at 10.5 T acquired using the FLASH pulse sequence. Fig. 5.9 shows an axial view of the T2*-weighted cadaver spine image at 10.5 T acquired using the MEDIC pulse sequence.



Figure 5.8: The lumbar and thoracic spine image at 10.5 T acquired using the FLASH pulse sequence with $FA = 20^\circ$, $TR/TE = 168 \text{ ms}/3.69 \text{ ms}$. Matrix = 576×432 , in-plane resolution = 0.5 mm , slice thickness = 2 mm , # of averages = 2, pixel bandwidth = 212 Hz/pixel , and total scan time of 127 sec.

The FLASH image is presented with expected tissue contrast between cerebrospinal fluid (CSF) and spinal cord. This image presents the entire view of the lumbar and thoracic spine acquired in a single acquisition. MEDIC image interestingly revealed the motor nerve pathway in the spinal cord. For both studies, an acceptable signal homogeneity was achieved across the slices of interest.

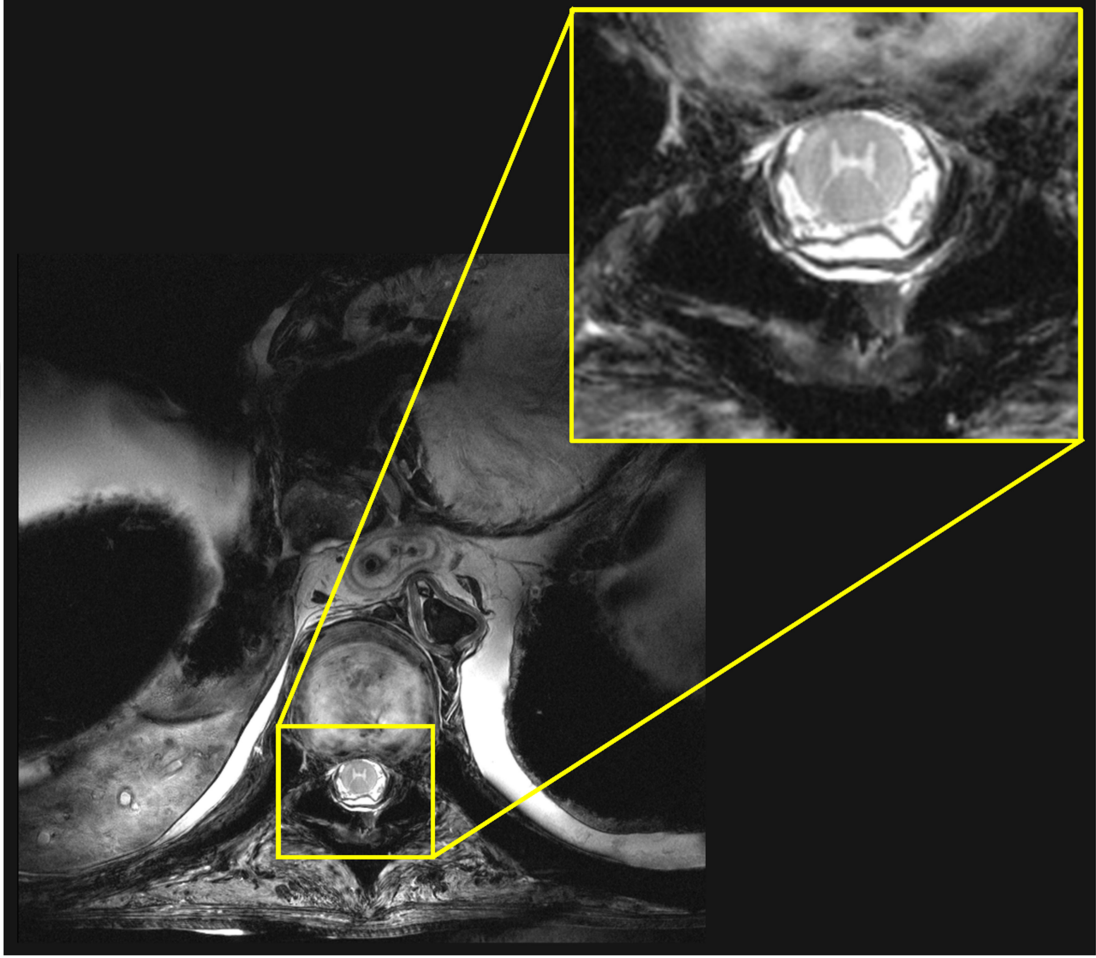


Figure 5.9: The T2*-weighted cadaver spine image at 10.5 T acquired using the MEDIC pulse sequence with FA = 30°, TR/TE = 500 ms/19 ms, matrix = 640x640, in-plane resolution = 0.24 mm, slice thickness = 2 mm, # of averages = 4, pixel bandwidth = 244 Hz/pixel, and total scan time of 963 sec.

5.4 Discussion and Conclusions

In this chapter, we introduced the NODES antenna, a short dipole with improved SAR performance mounted on a block with stepped dielectric constant, which enables us to manufacture an array with a significantly high number of elements. To prove the concept, we constructed a NODES antenna and compared its transmission and reception performance with the fractionated dipole [59] and loop coil [143] at 10.5 T,

numerically and experimentally. Results show that B_1^+ -SAR efficiency and ISNR performance can be significantly improved. We built a nine-channel Tx/Rx NODES array and acquired the cadaver spine MR images at 10.5 T.

We performed the optimization process over the pre-defined parameters with the goal of maximum B_1^+ -SAR efficiency in deep-body imaging. Consequently, we used the determined values for both Tx and Rx elements, however, they are not necessarily the optimal values for the reception elements. Investigating the optimum Rx elements is beyond the scope of this work and can be considered in future studies.

The NODES antenna built in this chapter is a prototype with the purpose of proof-of-concept. Therefore for ease of fabrication, the exact outcomes of the optimization problem are not utilized to construct the high-dielectric materials.

A short and novel Tx/Rx dipole-like element, a non-uniform dielectric substrate (NODES) antenna, with significantly improved SAR and SNR performance is introduced as a candidate for designing a highly dense Tx/Rx array for the MR imaging at 10.5 T. To prove the concept, a nine-channel Tx/Rx NODES array is constructed and utilized for the cadaver spine imaging at 10.5 T. The future studies will focus on designing and building denser array coils for in vivo human head and spine MR imaging.

Chapter 6

Ultimate Intrinsic Specific Absorption Rate Efficiency

Preface

The content of this chapter was presented in part at the 27th Annual Scientific Meeting of International Society of Magnetic Resonance in Medicine (ISMRM) [81]. Xiaoping Wu, Firdevs Ulus, and Yigitcan Eryaman contributed to this study. Yigitcan Eryaman and Ergin Atalar were involved in study conception. Xiaoping Wu contributed to the optimization process and advised in taking the full advantage of the RF shimming using VOP concept. Firdevs Ulus advised in implementation of the optimization technique.

6.1 Introduction

Demand for high-field (HF, $B_0 \geq 3$ T) MRI is continuously increasing due to its numerous benefits such as significant increase in the signal-to-noise ratio [5, 6] (SNR). However, peak local specific absorption rate (SAR) becomes a limiting factor in many

applications [144, 145]. Using transmit array (TxArray) coils with properly designed elements may provide a good solution to this issue [15, 16].

Various TxArray coil designs such as an 8-channel loop array at 3T [146], a 16-channel loop-dipole array at 7T [60], and a 10-channel fractionated dipole array at 10.5T [59] were proposed to overcome the SAR issue. However, researchers are still seeking for new designs to improve the SAR performance of the transmitters and introduced the B_1^+ -SAR efficiency [56] as a metric for this performance.

In this study, we utilize EM analytic relations to determine an upper bound for the B_1^+ -SAR efficiency inside a cylindrical homogeneous sample. We introduce this upper bound as the ultimate intrinsic SAR efficiency (UISARE). For proof of calculations, we compared the results at 1.5 T and 10.5 T with the results achieved using a commercial EM simulator.

6.2 Theory

The B_1^+ -SAR efficiency at the point of interest (POI) r_i can be defined as follows:

$$\xi(r_i) = \frac{\bar{s}^T(r_i)\bar{\alpha}}{\sqrt{\bar{\alpha}^H \bar{\bar{R}}(r_p)\bar{\alpha}}} \quad (6.1)$$

where $\bar{\alpha}$ is a complex vector contains coefficients of the cylindrical harmonics. $\bar{s}(r_i)$ contains H_1^+ 's corresponding to each cylindrical harmonic with unit coefficient at the POI inside the cylindrical sample. $\bar{\bar{R}}(r_p)$ is the local SAR matrix [147] at the position r_p where the peak local SAR occurs.

To maximize the B_1^+ -SAR efficiency, $\xi(r_i)$, at the POI, the following optimization problem can be defined [148]:

$$\begin{aligned} \min \quad & \bar{\alpha}^H \bar{\bar{R}}(r_p)\bar{\alpha} & \bar{\alpha} & \in \mathbb{C}^{2MN \times 1} \\ & & \bar{\bar{R}} & \in \mathbb{C}^{2MN \times 2MN} \\ \text{subject to} \quad & \bar{s}^T(r_i)\bar{\alpha} = 1 & \bar{s} & \in \mathbb{C}^{2MN \times 1} \end{aligned} \quad (6.2)$$

which minimizes the peak local SAR while keeping the H_1^+ constant at the POI. M and N are the numbers of the cylindrical harmonics in ϕ - and z -directions, respectively. The peak local SAR is defined as P_0 . Since $\bar{\bar{R}}$'s are positive semi-definite matrices, the minimization problem above is convex and can be solved for a global optimum. We applied the *Karush-Kuhn-Tucker* (KKT) conditions [149] to obtain the optimal coefficients and peak local SAR as follows:

$$\bar{\alpha} = \frac{\bar{\bar{R}}^{-1}(r_p) \bar{s}^*(r_i)}{\bar{s}^T(r_i) \bar{\bar{R}}^{-1}(r_p) \bar{s}^*(r_i)} \quad (6.3)$$

$$P_0 = \frac{\bar{s}^T(r_i) \bar{\bar{R}}^{-1}(r_p) \bar{s}^*(r_i)}{\left| \bar{s}^T(r_i) \bar{\bar{R}}^{-1}(r_p) \bar{s}^*(r_i) \right|^2} \quad (6.4)$$

Detailed calculations are provided in the Appendix.

In this work, a uniform cylindrical sample electrical properties of $\epsilon_r=80$ and $\sigma=0.6$ S/m was used for calculations. We investigated how many cylindrical modes would be needed to reach the UISARE for different B_0 field strengths ranging from 1.5 T to 10.5 T. Further, at each field strength, UISARE was calculated for different POIs inside the sample.

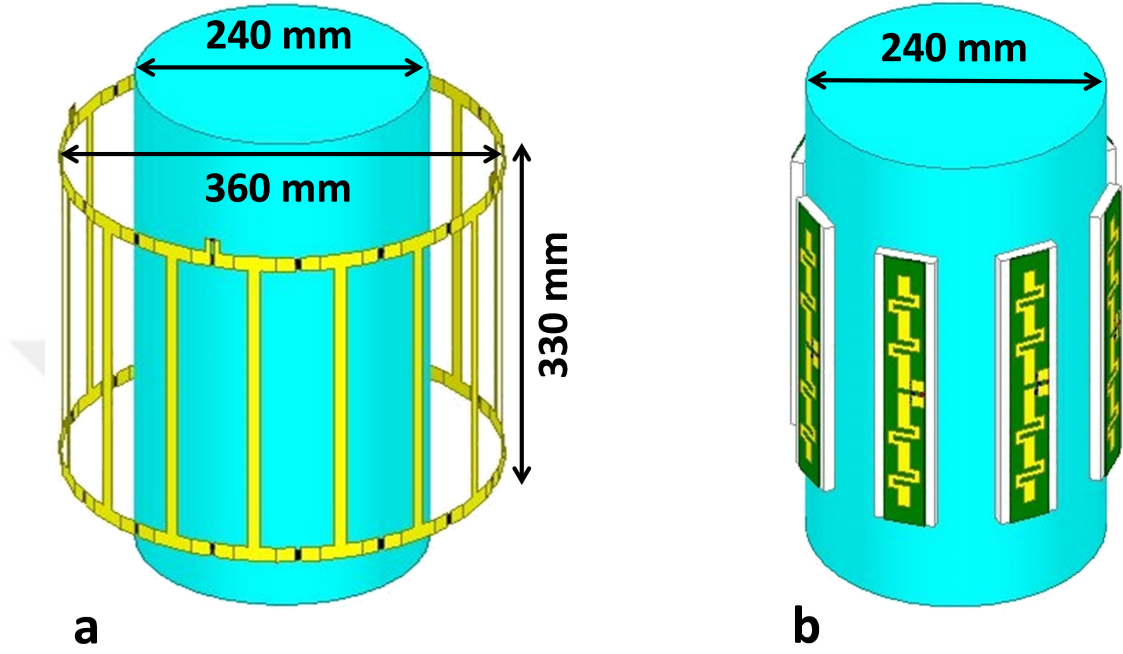


Figure 6.1: EM simulations for proof of calculations. (a) A birdcage coil at 1.5T with diameter of 360 mm and height of 330 mm and, (b) an eight-channel array of fractionated dipoles at 10.5T were numerically simulated in presence of a cylindrical phantom with $\epsilon_r=80$ and $\sigma=0.6$ S/m. The birdcage coil was excited in quadrature mode while the excitation of the array of dipoles was optimized to achieve maximum B_1^+ -SAR efficiency at the origin.

For proof of the calculations, the EM simulation of a quadrature-excited birdcage coil (Fig. 6.1a) was performed using an EM simulator (HFSS) at 1.5 T and the result was compared to the UISARE estimated for the center of the sample.

Eventually, the UISARE at the center of the sample at 10.5T was compared to the B_1^+ -SAR efficiency achieved using an 8-channel TxArray coil of fractionated dipoles (Fig. 6.1b). Note that, this array was driven by an RF shimming solution calculated with virtual observation points [150] to maximize B_1^+ -SAR efficiency at the origin.

For both 1.5T and 10.5T cases, to determine the limiting factor of the excitation, UISARE at the center was compared to the ultimate intrinsic transmit efficiency [151–153] (UITXE) as well. Furthermore, the 1g-averaging was chosen as the local-averaging to reduce the calculation time.

6.3 Results

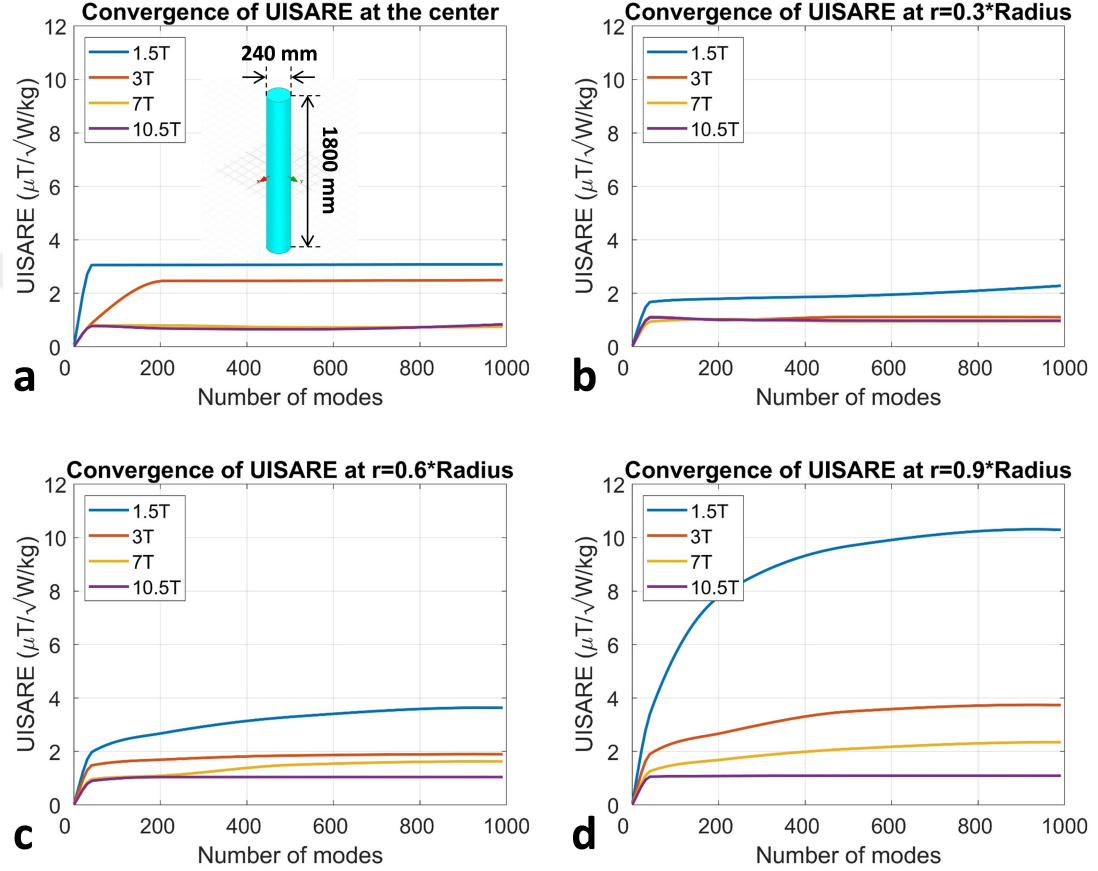


Figure 6.2: Convergence of the UISARE as a function of number of cylindrical modes for different B_0 field strength at (a) the center of the sample and the positions: (b) $r = 0.3R$, (c) $r = 0.6R$, and (d) $r = 0.9R$. The sample's radius is given by R .

Fig. 6.2 shows the number of cylindrical modes that is necessary for convergence of the UISARE in different field strengths for different POIs on the central transverse plane.

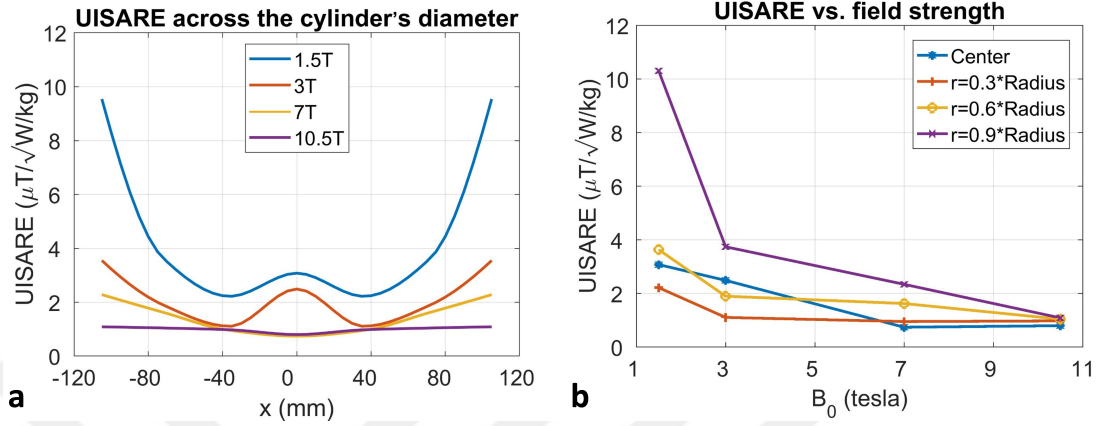


Figure 6.3: Calculated UISARE. (a) UISARE at some positions across the cylinder's diameter for different B_0 field strength. (b) UISARE at some B_0 field strength for different positions inside the sample.

Utilizing adequately large number of modes, Fig. 6.3a shows the UISARE at different POIs on the x-axis for each of B_0 values, separately. Correspondingly, Fig. 6.3b shows the UISARE versus B_0 values at different POIs.

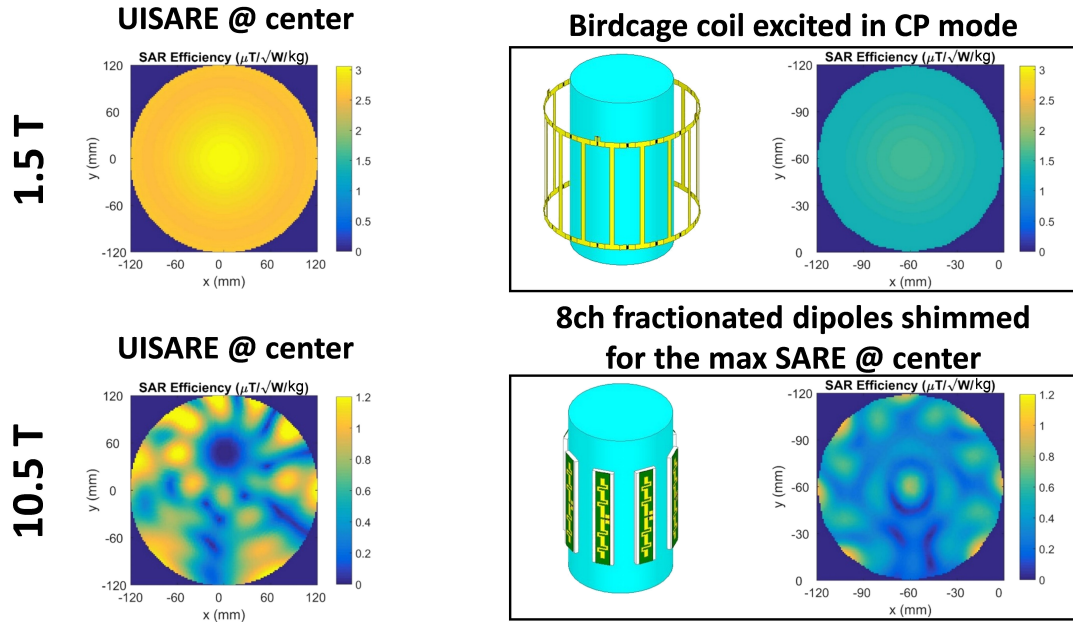


Figure 6.4: B_1^+ -SAR efficiency map obtained by considering the POI at the center and utilizing (a) UISARE at 1.5T, (b) simulated birdcage coil at 1.5T, (c) UISARE at 10.5T, and (d) simulated eight-channel TxArray of fractionated dipoles at 10.5T.

Fig. 6.4 shows the B_1^+ -SAR efficiency maps in 1.5T and 10.5T on a central transverse plane obtained by using the UISARE at the origin and performing EM simulations on the structures of Fig. 6.1.

Considering 2 W/kg and 10 W/kg limits [1] for the whole body and local SAR, respectively, Table 6.1 shows the upper bounds of B_1^+ that can be excited at the center of the 80kg

Table 6.1: Maximum B_1^+ can be achieved by putting a constraint on power corresponding to the upper limits of the whole-body and local SAR.

<i>max achievable B_1^+</i>	UITXE and UISARE theories		Birdcage @1.5T and Fractionated dipole @10.5T	
	Whole body SAR constraint	Local SAR constraint	Whole body SAR constraint	Local SAR constraint
1.5 T	8.4 μT	9.8 μT	7.6 μT	5.1 μT
10.5 T	11.4 μT	2.5 μT	8.9 μT	2.4 μT

cylindrical uniform sample with consideration of corresponding SAR limits in each case. These results show that confining to the UITXE and UISARE theories, the whole body SAR and local SAR upper limits restrict the maximum B_1^+ at the center at 1.5 T and 10.5 T, respectively. However, for the simulated birdcage coil at 1.5 T, local SAR is the limiting factor.

6.4 Discussion and Conclusions

In this chapter, we considered the B_1^+ -SAR efficiency as a metric for SAR performance of the transmitters, and calculated the UISARE as an upper bound for this metric. Results show that UISARE can almost be achieved in UHF ($B_0 \geq 7$ T) using the recently used TxArray coils with proper shimming strategy. However, at a lower field strength ($B_0 \leq 3$ T) the upper bound is significantly higher than the SAR efficiencies achievable with the commonly used birdcage coils. On the other hand, the results show that the limiting cases at 1.5T and 10.5T are whole body and local SAR constraint, respectively.

Our future work will be to characterize the ideal current patterns [153] for UISARE and study the UISARE in realistic body models [154].

Appendix

UISARE Calculations

A homogeneous cylindrical phantom with electrical properties identical to the average properties of a human body can properly mimic the body. Electromagnetic (EM) fields inside a such cylinder can be considered as a superposition of transverse-electric (TE) and transverse-magnetic (TM) modes which are solutions of the following scalar wave equation:

$$\left(\frac{1}{\rho} \frac{\partial}{\partial \rho} \rho \frac{\partial}{\partial \rho} + \frac{1}{\rho^2} \frac{\partial^2}{\partial \phi^2} + \frac{\partial^2}{\partial z^2} + k^2 \right) \psi(\bar{r}) = 0 \quad (6.5)$$

where

$$k^2 = k_\rho^2 + k_z^2 \quad (6.6)$$

The solution of Eq. 6.5 in the region that includes the origin is

$$Rg\psi_m(k_\rho, k_z, \bar{r}) = J_m(k_\rho \rho) e^{ik_z z} e^{im\phi} \quad m = 0, \pm 1, \pm 2, \dots \quad (6.7)$$

Rg stands for regular and it indicates the solution includes the origin. For the region does not consist of the origin, the cylindrical wave is

$$\psi_m(k_\rho, k_z, \bar{r}) = H_m(k_\rho \rho) e^{ik_z z} e^{im\phi} \quad (6.8)$$

Solving the Maxwell equations to find the transverse components of TE and TM modes leads to the following cylindrical waves [155]:

$$\begin{aligned} Rg\bar{M}_m(k_\rho, k_z, \bar{r}) &= \nabla \times (\hat{z} Rg\psi_m(k_\rho, k_z, \bar{r})) \\ &= \left[\hat{\rho} \frac{im}{\rho} J_m(k_\rho \rho) - \hat{\phi} k_\rho J'_m(k_\rho \rho) \right] e^{ik_z z} e^{im\phi} \end{aligned} \quad (6.9)$$

$$\begin{aligned} Rg\bar{N}_m(k_\rho, k_z, \bar{r}) &= \frac{1}{k} \nabla \times \nabla \times (\hat{z} Rg\psi_m(k_\rho, k_z, \bar{r})) \\ &= \left[\hat{\rho} \frac{k_\rho k_z}{k} J'_m(k_\rho \rho) + \hat{\phi} \frac{mk_z}{\rho k} J_m(k_\rho \rho) + \hat{z} \frac{k_\rho^2}{k} J_m(k_\rho \rho) \right] e^{ik_z z} e^{im\phi} \end{aligned} \quad (6.10)$$

For a cylindrical phantom of length L , a periodicity can be assumed in z -direction [156] such that

$$k_z = \frac{2n\pi}{L} \quad (6.11)$$

Therefore, the electric field inside the phantom can be written in the following form:

$$\bar{E}(\rho, \phi, z) = \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} A_{mn} Rg\bar{M}_m(k_\rho, k_z, \bar{r}) + B_{mn} Rg\bar{N}_m(k_\rho, k_z, \bar{r}) \quad (6.12)$$

Calculating the H -field from E -field leads to:

$$\bar{H} = -j \frac{1}{\omega\mu} \nabla \times \bar{E} \rightarrow \bar{H} = -j \frac{k}{\omega\mu} \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} (B_{mn} Rg\bar{M}_m + A_{mn} Rg\bar{N}_m) \quad (6.13)$$

In order to obtain the ultimate intrinsic SAR efficiency (UISARE), we need to solve the following optimization problem:

$$\begin{aligned} \min \quad & \|1g - averaged\ SAR(r)\|_{\infty} \quad r : \text{all voxels of the sample} \\ s.t. \quad & H^+(r_i) = 1 \quad r_i : \text{point of interest} \end{aligned} \quad (6.14)$$

Writing mn^{th} mode of the E -field in the matrix form

$$E_{mn} = \begin{bmatrix} \hat{\rho} \cdot Rg\bar{M}_m & \hat{\rho} \cdot Rg\bar{N}_m \\ \hat{\phi} \cdot Rg\bar{M}_m & \hat{\phi} \cdot Rg\bar{N}_m \\ \hat{z} \cdot Rg\bar{M}_m & \hat{z} \cdot Rg\bar{N}_m \end{bmatrix}_{3 \times 2} \quad \& \quad \alpha_{mn} = \begin{bmatrix} A_{mn} \\ B_{mn} \end{bmatrix}_{2 \times 1} \quad (6.15)$$

the E -field and SAR_{1g} can be expressed as:

$$\begin{aligned} [E]_{3 \times 1} &= \sum_{n=-\infty}^{\infty} \sum_{m=-\infty}^{\infty} E_{mn} \alpha_{mn} \\ \rightarrow SAR_{1g} &= \frac{\sigma/\rho}{1g} \sum_{m'} \sum_{n'} \sum_m \sum_n \int_{V_{10g}} \alpha_{m'n'}^H E_{m'n'}^H E_{mn} \alpha_{mn} dV \end{aligned} \quad (6.16)$$

where ρ represents the mass density of the sample. Defining $R(r)$ as follows

$$R(r) = \begin{bmatrix} E_{11}^H(r) \\ E_{12}^H(r) \\ \vdots \\ E_{MN}^H(r) \end{bmatrix}_{2MN \times 3} \quad \begin{bmatrix} E_{11}(r) & E_{12}(r) & \dots & E_{MN}(r) \end{bmatrix}_{3 \times 2MN} \quad (6.17)$$

SAR_{1g} can be simplified as follows:

$$SAR_{1g}(r) = \alpha^H R(r) \alpha \quad (6.18)$$

Similarly, $H^+(r)$ can be written in matrix form as follows:

$$H^+(r) = \frac{\hat{\rho} \cdot \bar{H}(r) - j \hat{\phi} \cdot \bar{H}(r)}{2} = \sum_n \sum_m -j \frac{k}{\omega \mu} \begin{bmatrix} 1 & -j \end{bmatrix} \begin{bmatrix} \hat{\rho} \cdot Rg \bar{N}_{mn} & \hat{\rho} \cdot Rg \bar{M}_{mn} \\ \hat{\phi} \cdot Rg \bar{N}_{mn} & \hat{\phi} \cdot Rg \bar{M}_{mn} \end{bmatrix} \begin{bmatrix} A_{mn} \\ B_{mn} \end{bmatrix}$$

$$\Rightarrow H^+(r) = \sum_n \sum_m s_{mn}(r) \alpha_{mn} = \begin{bmatrix} s_{11}(r) & s_{12}(r) & \dots & s_{MN}(r) \end{bmatrix}_{1 \times 2MN} \begin{bmatrix} \alpha_{11} \\ \alpha_{12} \\ \vdots \\ \alpha_{MN} \end{bmatrix}_{2MN \times 1} = s^T(r) \alpha \quad (6.19)$$

Eventually, the optimization problem turns into:

$$\begin{aligned} \min \quad & \left\| \alpha^H R(r) \alpha \right\|_{\infty} \quad r : \text{all voxels of the sample} \\ \text{s.t.} \quad & s^T(r_i) \alpha = 1 \end{aligned} \quad (6.20)$$

The problem above is convex if and only if $R(r)$'s are positive semi-definite. Assuming the convexity, the *Lagrange multipliers method* can be used for the minimization

problem. In order to get rid of the norm-infinity, the problem can be reshaped as follows:

$$\begin{aligned}
& \min \quad P_0 \\
& s.t. \quad s^T(r_i)\alpha = 1 \quad P_0 \in \mathbb{R}^+ \\
& \quad \quad \alpha^H R(r_1)\alpha \leq P_0 \quad \text{where} \quad \alpha \in \mathbb{C}^{2MN \times 1} \\
& \quad \quad \vdots \quad s \in \mathbb{C}^{2MN \times 1} \\
& \quad \quad \alpha^H R(r_v)\alpha \leq P_0 \quad R \in \mathbb{C}^{2MN \times 2MN}
\end{aligned} \tag{6.21}$$

where P_0 and α represent peak 1g – averaged SAR and coefficients of cylindrical wave modes, respectively.

Hence, the corresponding *Lagrangian* function is

$$L(P_0, \alpha, \nu, \lambda) = P_0 + \nu (s^T(r_i)\alpha - 1) + \sum_{j=1}^v \lambda_j (\alpha^H R(r_j)\alpha - P_0) \tag{6.22}$$

where v is the number of voxels inside the sample.

According to *Karush-Kuhn-Tucker* (KKT) theorem, if the cost function and inequality functions are convex and equality functions are affine, a solution is optimal if and only if it satisfies KKT conditions [149]. Applying the KKT conditions to the minimization problem 6.21 leads to the following statement: If there exists an $R(r_p)$ such that

$$\frac{s^T(r_i)}{|s^T(r_i)R^{-1}(r_p)s^*(r_i)|^2} \left((R^{-1}(r_p))^H R(r_j) - I \right) R^{-1}(r_p)s^*(r_i) \leq 0 \quad \forall j = 1, \dots, v \tag{6.23}$$

where r_p is the position of the peak local 10g-averaged SAR, then

$$\bar{\alpha} = \frac{R^{-1}(r_p)s^*(r_i)}{s^T(r_i)R^{-1}(r_p)s^*(r_i)} \tag{6.24}$$

$$P_0 = \frac{s^T(r_i)R^{-1}(r_p)s^*(r_i)}{|s^T(r_i)R^{-1}(r_p)s^*(r_i)|^2} \tag{6.25}$$

Chapter 7

Discussion and Conclusions

To serve one of the main goals of MRI, which is achieving the best image quality by improving SNR while confined to the patient's safety restrictions (i.e., SAR limits), it is beneficial to utilize TxArray technology. Correspondingly, the main focus of this dissertation was on designing the TxArray coils. More specifically, we introduced a novel and fast technique to design TxArray coils. Moreover, we presented some innovative transmit elements to develop a new generation of TxArray coils with improved SAR performance. We also proposed a new concept regarding the upper bound for the SAR performance of the transmitters and calculated this upper bound at different field strengths.

In the first study of this dissertation, accelerating the co-simulation method [39] for the design of TxArray coils was studied using equivalent circuit models. Although the co-simulation method [39] dramatically reduces the complexity of the design of TxArray coils, finding the optimum solution is not trivial since there exist many local minima in the optimization problem. We proposed to utilize an equivalent circuit model of the TxArray coil to obtain a proper initial guess for the optimization process of the co-simulation method. To prove the concept, six different TxArray coils (i.e., three degenerate birdcage coils (DBC), two dual-row head coils, and one elliptical body TxArray coil) with two different loading strategies (cylindrical phantom and human head/body model) at 3 T field strength were investigated theoretically; as an

example study, an eight-channel head-DBC was constructed using the obtained values. This approach accelerates the design process more than 20-fold for the coils that were investigated in this manuscript. The main finding of this study is a fast and accurate method for tuning and decoupling of a TxArray coil using its equivalent circuit model combined with the co-simulation method.

As another contribution of this dissertation, a strategy to reduce the local specific absorption rate (SAR) while keeping B_1^+ constant inside the region-of-interest (ROI) at the UHF-MRI was investigated. Locally raising the resonance structure under the discontinuity (i.e., creating a bump) increases the distance between the accumulated charges and the tissue. As a result, it reduces the electric field and local SAR generated by these charges inside the tissue. The B_1^+ at a point that is sufficiently far from the coil however is not affected by this modification. In this study, three different resonant elements (i.e., loop coil [46], snake antenna [57], and fractionated dipole (FD) [59]) were investigated. For experimental validation, a bumped FD was further investigated at 10.5 T. After the validation, the transmit performances of eight-channel arrays of each element were compared through EM simulations. Introducing a bump reduced the peak 10g-averaged SAR by 21%, 26%, 23% for the loop and snake antenna at 7 T, and FD at 10.5 T, respectively. In addition, eight-channel bumped FD array at 10.5 T had a 27% lower peak 10g-averaged SAR in a realistic human body simulation (i.e., prostate imaging) compared to an eight-channel FD array. The main finding of this study is a simple design strategy based on adding bumps to a resonant element to reduce the local SAR while maintaining B_1^+ inside an ROI. As an example, we modified an FD and performed EM simulations and phantom experiments with a 10.5T scanner. Results show that the peak 10g-averaged SAR can be reduced by more than 25%.

As one of the contributions of this dissertation, confining to the safety limits, the first human head images at 10.5 T is acquired. In order to ensure safety of subjects, we validated the EM simulation model of our coil. We obtained quantitative agreement between simulated and experimental B_1^+ and SAR. Using the validated coil model, we calculated RF power levels to safely image human subjects. We conducted all experiments and imaging sessions in a controlled RF safety lab and the whole-body

10.5T scanner in the Center for Magnetic Resonance Research (CMRR). Quantitative agreement between the simulated and experimental results was obtained including S-parameters, B_1^+ maps, and SAR. We calculated peak 10g-average SAR using four different realistic human body models for a quadrature excitation and demonstrated that the peak 10g-averaged SAR variation between subjects was less than 30%. We calculated safe power limits based on this set and used those limits to acquire T2- and T2*-weighted images of human subjects at 10.5 T. As the main finding of this study, we acquired the first in vivo human head images at 10.5 T using an eight-channel transmit/receive (Tx/Rx) coil. We implemented and expanded a previously proposed workflow to validate the EM simulation model of the eight-channel Tx/Rx coil. Using the validated coil model, we calculated RF power levels to safely image human subjects.

In another study of this dissertation, a novel short coil element, non-uniform dielectric substrate (NODES) antenna with improved SAR performance, is introduced which provides an opportunity to design highly-dense TxArray coils for 10.5 T. In the design of this new dipole-like antenna, we utilized a non-uniform dielectric substrate placed underneath the element to optimize the B_1^+ -SAR efficiency (i.e., B_1^+ normalized to the square root of peak local SAR). This modification increases the uniformity of the current distribution on the dipole and reduces local SAR. As a result, it enables us to design highly-dense transmit arrays using shorter elements. For proof of concept, we performed an optimization on a dipole at 10.5 T using the EM simulations. Also, we constructed the corresponding structure and conducted a phantom experiment on a whole-body 10.5 T scanner to validate the EM simulation results. In addition, we built a nine-channel transmit/receive (transceiver) coil and acquired the human cadaver spine image at 10.5 T. The main finding of this study is a novel transmit element, NODES antenna, with significantly improved SAR performance compared to the commonly used elements at 10.5 T. The short size of the NODES antenna enables us to build highly-dense TxArray coils.

In the final study of this dissertation, we introduced a new concept such that the B_1^+ -SAR efficiency was used as a metric to evaluate the SAR performance of transmitters, and we defined its upper bound as the ultimate intrinsic SAR efficiency (UISARE). We

used a uniform cylindrical sample for calculations and investigated how many cylindrical modes would be needed to reach the UISARE for different B_0 field strengths ranging from 1.5T to 10.5T. For proof of concept, we derived the UISARE values at 1.5 T and 10.5 T and compared to the SAR efficiencies of a 1.5 T birdcage coil and a 10.5 T eight-channel array of fractionated dipoles. The main finding of this study is determining an upper bound for the SAR performance of the transmitters at different field strengths. Our findings show that the limiting safety factors at 1.5 T and 10.5 T are whole body and local SAR constraints, respectively.

Overall, the findings of this dissertation including the novel techniques and innovative structures for designing the TxArray coils mainly contribute to improving image quality while confining to the patient's safety restrictions.

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