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System and SAR Characterization in Parallel RF Transmission

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Abstract

The markedly increased degrees of freedom introduced by parallel RF transmission present both opportunities and challenges for SAR management. On the one hand they enable E field tailoring and SAR reduction while facilitating excitation profile control. On the other hand they increase the complexity of SAR behavior and the risk of inadvertently exacerbating SAR by improper design or ployout of RF pulses. The substantial subject-dependency of SAR in high field MR can be a compounding factor. Building upon a linear system concept and a calibration scheme involving a finite number of *in situ* measurements, the present work establishes a clinically applicable method for characterizing global SAR behavior as well as channel-by-channel power transmission. The method offers a unique capability of predicting, for any excitation, the SAR and power consequences that are specific to the subject to be scanned and the MRI hardware. The method was validated in simulation and experimental studies, showing promise as the foundation to a prospective paradigm where power and SAR are not only monitored but, through prediction-guided optimization, proactively managed.

Keywords

RF power; RF safety; SAR; parallel RF transmission; RF shimming; B1 shimming; RF coil; high field MR; system calibration; linear system

INTRODUCTION

During radiofrequency (RF) transmission the RF magnetic field B1 interacts with spins and induces MR signal. The concomitant electric field E, necessarily accompanying the B1 field according to the laws of electrodynamics, deposits RF energy in the subject and dictates specific absorption rate (SAR). Conventional RF transmission strives for a uniform B1 in a large volume, which tends to cause, regardless of the location or size of the imaged region, broad-reaching E field and unnecessarily high RF energy deposition. The advent of parallel RF transmission^{1, 2} brought about a new paradigm wherein RF pulses simultaneously drive distributed elements of a multi-port transmit coil to effect both spatial and temporal variations of the B1 and E fields. The markedly increased degrees of freedom^{1, 3} of parallel RF transmission as compared with traditional RF transmission were shown to enable tailoring of E field and containing of SAR while improving flip-angle profile control.

For a given coil-subject setup, exploitation of the degrees of freedom in parallel transmission for improving flip-angle profile and SAR control is realized through design of RF/gradient pulses. Practical control of flip angle profile *in vivo* is enabled by subject-specific B1 calibration, which captures the complex effects of coil-subject geometry, composition and interaction on the B1 maps and provides required guidance to the pulse design (based essentially on the Bloch equation).

However, due to a lack of subject-specific SAR information, guidance to pulse design for practical SAR control has not hitherto been possible *in vivo*. In fact, the extra degrees of freedom of parallel transmission, compounded by a generally poor understanding of SAR in high field MR, often spur concerns that improper design or playout of RF pulses may exacerbate SAR (e.g., by causing excessive constructive interference of electric fields), as opposed to reducing it. These safety and performance considerations underscore the importance of a practicable SAR prediction method – one that, given any set of RF shimming coefficients or RF excitation pulses, predicts the true SAR consequences globally and locally. Such a method would enable a paradigm shift from merely SAR monitoring (the current practice) to proactive SAR management. Extensive recent efforts to evaluate and predict SAR have relied upon electromagnetic field calculations in numerical simulations or experimental findings in “average” subjects. Yet it remains unclear if it will be feasible to adapt the details of the simulated coil-human setup or to extrapolate “average” findings such that the separately obtained SAR characteristics could closely track what happens or will happen to a subject undergoing a scan.

The present work follows the lead of a SAR prediction model¹, which, based on system linearity, relates global or local SAR to parallel RF pulses or RF shimming coefficients with a quadratic function. The model has been used in a number of parallel-transmission-related SAR investigations^{1, 3, 4, 5, 6, 7, 8, 9}, which demonstrated that design of RF pulses, when guided by the model, enables the extra degrees of freedom inherent in a parallel transmission system to be advantageously exploited, and SAR to be managed. The model has also been a foundational element for analyzing and approaching ultimate RF transmission performance⁸.

The SAR model indicates particularly that there is a structure to global and local SAR behavior in parallel transmission and that the uniqueness of a specific coil-subject setup can be encapsulated by a comparatively small set of parameters defining that structure. The present work focuses on devising a clinically applicable method to unveil the structure parameters that are specific to a coil-subject setup, and thereby to provide accurate SAR guidance for the actual MR examinations of individual subjects with a SAR tracking and prediction capability. The method is based on a finite number of *in situ* measurements, and does not require the use of simplifying assumptions about the subject or the scanner setup. Compared to a method relying on simulations or “average” results, this subject-specific approach eliminates a major source of error associated with inter-subject and inter-RF-apparatus SAR variations. In this paper the scope of the devised method is limited to global SAR and individual channel characterizations.

THEORY AND METHOD

RF energy dissipation and tissue heating

Pennes’ bio-heat equation describes thermal energy balance for perfused tissue:

$$\rho C \frac{\partial T}{\partial t} = \nabla \cdot (k \nabla T) + h_b + h_e \quad (1)$$

where ρ , C and k refer to tissue density, specific heat capacity and thermal conductivity, respectively, and h_b is the blood-to-tissue heat transfer rate. RF energy deposition accompanying RF transmission, which is captured in Eqn. 1 by the local RF energy deposition rate h_e , is a driving force of temperature rise and is proportional to the square of local E field strength: $h_e = \frac{1}{2} \sigma |E|^2$, where σ is electrical conductivity. Temporal and volume averaging of h_e , when further scaled by appropriate density or mass measurements, give, for example, head, torso, extremity or whole body average SAR as defined in FDA and IEC guidelines¹⁰.

Excessive tissue heating may potentially result from RF energy deposition. This risk tends to increase for high RF duty-cycle sequences and for high main magnetic field strength B_0 , in which cases monitoring and optimization of RF transmission become crucial.

A linear system perspective

An MR scanner commonly modulates both RF and gradient fields when exciting spins. Modulation of the RF field is achieved by updating, as specified by a designed RF pulse waveform, the magnitude and phase of a Larmor-frequency sinusoidal pulse that drives a transmit channel and a transmit coil. The updating occurs every Δt time increment, which is typically several microseconds in practice. This modulation is multiplied in parallel RF transmission, where a plurality of designed waveforms, sinusoidal pulses and transmit channels, as well as a multi-port transmit coil, are employed to provide considerably enhanced support for the RF field modulation, giving rise to an RF field that varies both in space and in time. Parallel RF transmission includes RF shimming as a special case.

A useful perspective to parallel transmission is to treat the transmit channels, the transmit coil and the subject as a single system. For any Δt interval, the magnitude-phase pairs specified by multiple RF pulse waveforms, expressed with complex scalars $w_p^{(n)}$ (n = port index and p = interval index), define the inputs to the system. There is one input configuration per Δt interval. The B_1^+ field and the RF energy deposition are important outputs of the system. During RF transmission the B_1^+ field interacts with the spins, forming the basis of MR signal induction (or spin excitation). The concomitant E field, necessarily accompanying the B_1 field according to the laws of electrodynamics, induces RF energy deposition in the subject.

To address the goal of creating a target excitation profile while keeping SAR low, one needs to design RF pulse waveforms, or system inputs, in an optimal way so that the system will respond with both 1) a properly modulated B_1^+ field that works with the gradient field in exciting spins, and 2) a properly restrained E field that steers clear of excessive RF energy deposition. A key prerequisite to the design optimization is accurate prediction of B_1^+ distribution and E-induced RF energy deposition *given any input configuration*. This is possible in practice, as described below.

Parallel RF transmission system hardware is normally set up in such a way that the electromagnetic field induced in the subject responds linearly to the inputs. This can be appreciated by considering the subject and the RF transmit coil as a multi-port network that interacts with a plurality of sources through the ports (Fig.1). The RF power amplifiers, each presenting to the multi-port network equivalently a voltage source in series with an output impedance, amplify modulated RF pulses and drive the ports. Provided that the amplifiers satisfy common linearity specifications, the voltage sources are related linearly to the modulated RF pulses that are fed to the corresponding amplifiers' input terminals. In this case, linearity of Maxwell equations dictates that the electromagnetic field responds linearly to the voltage sources and, in turn, to the modulated RF pulses. It follows that any

electromagnetic field component at a location within the object, as a function of time, can be expressed as

$$y(t) = \mathcal{L} \left(v^{(1)}(t), v^{(2)}(t), \dots, v^{(N)}(t) \right), \quad (2)$$

where $\mathcal{L}$ represents a linear mapping, $y(t)$ denotes the electromagnetic field component, $v^{(n)}(t)$ is the modulated RF pulse for port n , and N is the number of transmit ports.

Phasor notation is suitable for describing a time harmonic quantity within each Δt interval. For example, at location $\mathbf{r}$ inside the object and within the p th Δt interval, the E field can be expressed as $\mathbf{E}(\mathbf{r}, t) = [\alpha_x(\mathbf{r}) \cos(\omega_0 t + \phi_x(\mathbf{r})) \ \alpha_y(\mathbf{r}) \cos(\omega_0 t + \phi_y(\mathbf{r})) \ \alpha_z(\mathbf{r}) \cos(\omega_0 t + \phi_z(\mathbf{r}))]$. The corresponding phasor notation for E is a complex vector: $\mathbf{E}_p(\mathbf{r}) = [\alpha_x(\mathbf{r}) \exp(j\phi_x(\mathbf{r})) \ \alpha_y(\mathbf{r}) \exp(j\phi_y(\mathbf{r})) \ \alpha_z(\mathbf{r}) \exp(j\phi_z(\mathbf{r}))]$. For the same interval, the phasor notation for the modulated RF pulse for the n th port is the complex scalar $w_p^{(n)}$. The linear system behavior captured by Eqn.2 can thus be expressed in matrix form for the Δt interval:

$$\mathbf{y} = \mathbf{A} \mathbf{w}, \quad (3)$$

In Eqn. 3 $\mathbf{w} = [w_p^{(1)} \ w_p^{(2)} \ \dots \ w_p^{(N)}]^T$ is a vector collecting the definitions for the modulated RF pulses – the n th entry is the complex scalar corresponding to the pair of magnitude and phase values for the time interval and the n th port. $\mathbf{y} = [y_p^{(1)} \ y_p^{(2)} \ \dots \ y_p^{(L)}]^T$ is a vector collecting phasor representations of the time variables of interest for the same interval (e.g., B_1^+ field at L spatial locations). $\mathbf{A}$ is an L -by- N complex-valued matrix representing the linear mapping.

Local and global SAR models

Linearity allows decomposition of the net E field as a weighted superposition of E fields corresponding to the N individual sources, with the weights being $w_p^{(1)}, \dots, w_p^{(N)}$. In phasor

notation $\mathbf{E}_p(\mathbf{r}) = \sum_{n=1}^N w_p^{(n)} \mathbf{e}^{(n)}(\mathbf{r})$ where $\mathbf{e}^{(n)}(\mathbf{r})$ is the E field (the response) due to a $\mathbf{w}$ (the input) that has its n th entry being one and the rest being zeros. This leads to the following expression for local RF energy deposition rate:

$$\begin{aligned} h_e(\mathbf{r}) &= \frac{\sigma(\mathbf{r})}{2} |\mathbf{E}_p(\mathbf{r})|^2 \\ &= \frac{\sigma(\mathbf{r})}{2} \mathbf{E}_p(\mathbf{r})^* \cdot \mathbf{E}_p(\mathbf{r}) \\ &= \frac{\sigma(\mathbf{r})}{2} \left(\sum_{m=1}^N w_p^{(m)} \mathbf{e}^{(m)}(\mathbf{r}) \right)^* \cdot \left(\sum_{n=1}^N w_p^{(n)} \mathbf{e}^{(n)}(\mathbf{r}) \right) \\ &= \left[\dots \left(w_p^{(m)} \right)^* \dots \right] \underbrace{\begin{bmatrix} \dots & \dots & \dots \\ \dots & \frac{\sigma(\mathbf{r})}{2} \left(\mathbf{e}^{(m)}(\mathbf{r}) \right)^* \cdot \left(\mathbf{e}^{(n)}(\mathbf{r}) \right) & \dots \\ \dots & \dots & \dots \end{bmatrix}}_{\stackrel{\text{def}}{=} \mathbf{\Lambda}(\mathbf{r})} \begin{bmatrix} \vdots \\ \left(w_p^{(n)} \right) \\ \vdots \end{bmatrix} = \mathbf{w}^H \mathbf{\Lambda}(\mathbf{r}) \mathbf{w} \end{aligned} \quad (4)$$

where $*$ denotes complex conjugation, H denotes conjugate transpose and underscored indices are row or column indices for vector and matrix entries. Eq. 4 shows that local RF power deposition can be expressed as quadratic functions in $w_p^{(1)}, \dots, w_p^{(N)}$. In matrix form, $h_e(\mathbf{r}) = \mathbf{w}^H \mathbf{\Lambda}(\mathbf{r}) \mathbf{w}$.

It follows that global RF energy deposition rate can also be described by a quadratic model:

$$\xi_p = \int_V \frac{\sigma(\mathbf{r})}{2} |\mathbf{E}_p(\mathbf{r})|^2 dV = \left[\cdots (w_p^{(m)})^* \cdots \right] \underbrace{\begin{bmatrix} \cdots & \int_V \frac{\sigma(\mathbf{r})}{2} (\mathbf{e}^{(m)}(\mathbf{r}))^* \cdot (\mathbf{e}^{(n)}(\mathbf{r})) dV & \cdots \\ \cdots & \cdots & \cdots \end{bmatrix}}_{\stackrel{\text{def}}{=} \mathbf{\Phi}} \begin{bmatrix} \vdots \\ (w_p^{(n)}) \\ \vdots \end{bmatrix} = \mathbf{w}^H \mathbf{\Phi} \mathbf{w} \quad (5)$$

Matrices $\mathbf{\Lambda}$ and $\mathbf{\Phi}$ are named, respectively, local and global *power correlation matrices*. Both $\mathbf{\Lambda}$ and $\mathbf{\Phi}$ are N-by-N, and can be shown to be Hermitian and positive semidefinite. They capture the effects of field interference and tissue conductivity on RF energy deposition at regional and global scales. Total deposited RF energy during RF transmission is a time integral of ξ_p , expressed as:

$$\xi = \sum_{p=1}^P \Delta t \xi_p = \sum_{p=1}^P \left[\cdots (w_p^{(m)})^* \cdots \right] \mathbf{\Phi} \Delta t \begin{bmatrix} \vdots \\ (w_p^{(n)}) \\ \vdots \end{bmatrix} = \mathbf{w}_{full}^H \mathbf{\Phi}_{full} \mathbf{w}_{full}, \quad \mathbf{\Phi}_{full} = \begin{pmatrix} \mathbf{\Phi} \Delta t & & 0 \\ & \ddots & \\ 0 & & \mathbf{\Phi} \Delta t \end{pmatrix}_{NP \times NP} \quad (6)$$

where $\mathbf{w}_{full}$ is a vector collects all samples of the RF pulse waveforms. Similarly,

$$\xi_{local}(\mathbf{r}) = \sum_{p=1}^P \Delta t h_{e,p}(\mathbf{r}) = \sum_{p=1}^P \left[\cdots (w_p^{(m)})^* \cdots \right] \mathbf{\Lambda}(\mathbf{r}) \Delta t \begin{bmatrix} \vdots \\ (w_p^{(n)}) \\ \vdots \end{bmatrix} = \mathbf{w}_{full}^H \begin{pmatrix} \mathbf{\Lambda}(\mathbf{r}) \Delta t & & 0 \\ & \ddots & \\ 0 & & \mathbf{\Lambda}(\mathbf{r}) \Delta t \end{pmatrix}_{NP \times NP} \mathbf{w}_{full} \quad (7)$$

quantifies total deposited RF energy at location $\mathbf{r}$ over the course of the RF excitation.

It is to be noted that in characterizing SAR, even in the absence of inter-coil coupling, RF energy dissipation locally or averaged over an imaging volume cannot be treated by considering the individual sources in isolation. This can be attributed to the overlapping of E fields. Quantitatively this is reflected in the typically non-zero off-diagonal entries in $\mathbf{\Lambda}$ and $\mathbf{\Phi}$ – these correlation entries quantify the mutually interfering nature of the sources and encapsulate information instrumental for SAR reduction. In practice however, many existing RF shimming or RF pulse calculations attempt to follow RF power or SAR using a sum of magnitude squares of the waveform samples^{11, 12, 13}, which is equivalent to using an identity matrix in place of $\mathbf{\Phi}$ or $\mathbf{\Lambda}$. Such an approach to managing SAR can be subject to significant error. For instance, given a set of parallel RF pulses, one can add a 180° phase offset to one of the pulses without affecting the evaluation of $\mathbf{w}^H \mathbf{I} \mathbf{w}$ – use of the identity matrix $\mathbf{I}$ is incapable of capturing the SAR effect caused by constructive or destructive interference of the E fields driven by the pulses. Use of the identity matrix in place of true power correlation matrices further assumes no variation in response to the multiple sources, which is not true in any case where different array elements contribute differently to SAR.

Results represented by Eqns. 4 and 5 indicate that there is a compact structure to global and local SAR. With a proper experimental calibration scheme it is possible to unveil this structure and establish practical subject-specific SAR tracking and prediction capabilities.

B1 calibration as an example

An RF transmit system with linearity adequately maintained facilitates electromagnetic (EM) field calibration. Measuring EM field responses in M experiments of input

configurations $\mathbf{w}_{(1)}$ to $\mathbf{w}_{(M)}$ results in equations that can be compactly expressed in matrix form:

$$[\mathbf{y}_{(1)} \mathbf{y}_{(2)} \dots \mathbf{y}_{(M)}] = \mathbf{A}[\mathbf{w}_{(1)} \mathbf{w}_{(2)} \dots \mathbf{w}_{(M)}] \quad (8)$$

Eqn.8 pools a total of M Eqn.3-type equations together, with the experiment index shown as subscripts in parentheses. If $\mathbf{w}_{(1)}$ to $\mathbf{w}_{(M)}$ are linearly independent input configurations and $M=N$, then existence of the inverse of matrix $[\mathbf{w}_{(1)} \mathbf{w}_{(2)} \dots \mathbf{w}_{(M)}]$ is guaranteed and the EM field response to any input configuration can be predicted. This is because matrix $\mathbf{A}$ can be fully determined from the measured EM field responses and the corresponding inputs:

$$\mathbf{A} = [\mathbf{y}_{(1)} \mathbf{y}_{(2)} \dots \mathbf{y}_{(N)}] [\mathbf{w}_{(1)} \mathbf{w}_{(2)} \dots \mathbf{w}_{(N)}]^{-1}$$

A simple yet concrete example is calibration of transmit B1 fields with N experiments, where the n th experiment ($n=1,2,\dots,N$) involves driving the n th port with a unit-amplitude rectangular RF pulse (a sinusoid of unit amplitude and zero initial phase) and the other ports with zero-amplitude RF pulses. In this case $[\mathbf{w}_{(1)} \mathbf{w}_{(2)} \dots \mathbf{w}_{(N)}]$ is an identity matrix, and $[\mathbf{y}_{(1)} \mathbf{y}_{(2)} \dots \mathbf{y}_{(N)}] = \mathbf{A}$ stores individual channel field maps.

For improved robustness in the presence of noise or perturbations, more or tailored calibration experiments may be used, which is the subject of active research in the area of transmit B1 field mapping^{14, 15}. To illustrate with simpler notation, consider one row of Eqn. 8. It deals with EM field at one spatial location and can be rewritten as

$$\mathbf{b} = \mathbf{W} \mathbf{x} \quad (9)$$

where $\mathbf{W}$ is defined as $[\mathbf{w}_{(1)} \mathbf{w}_{(2)} \dots \mathbf{w}_{(M)}]^T$, $\mathbf{x}$ denotes the transpose of the l th row of $\mathbf{A}$, and for B1 mapping, $\mathbf{b}$ is a vector collecting $B1^+$ at the l th location measured in calibration experiment 1 through M ($M \geq N$). If the noise covariance matrix of measured $\mathbf{b}$ is $\mathbf{V}$, the classic solution to $\mathbf{x}$ is given by the best linear unbiased estimate¹⁶:

$$\mathbf{x} = (\mathbf{W}^H \mathbf{V}^{-1} \mathbf{W})^{-1} \mathbf{W}^H \mathbf{V}^{-1} \mathbf{b} \quad (10)$$

Clearly, for multi-port parallel RF transmission, calibrating $\mathbf{A}$ satisfies one prerequisite of the RF pulse design process. The calibration allows the spatiotemporal variation of the $B1^+$ field, and further, the spin excitation profile, to be predicted for any set of RF pulses, which in turn enables excitation profile control via appropriately-designed parallel pulse waveforms. This illustrated framework for establishing a model (and determining $\mathbf{A}$) points to a few important considerations, including physical basis of the prediction model, measurement noise handling, tailoring of calibration (specifically, design of the calibration configurations that $\mathbf{W}$ captures), and sensing or measurement scheme. These are examined in depth below in the context of power and global SAR characterization, the focus of the present work.

Subject-specific global SAR calibration and prediction

An MR system with linearity well maintained has RF power dissipation characterized by Eqns. 4 and 5, which facilitates SAR calibration and minimization. By the law of conservation of energy the net RF power injected into the N -port network should be equal to combined RF loss in the subject, in the transmission hardware (including the coil) and

through radiation. The latter two are normally related to $\mathbf{w}$ in the form of quadratic functions as well, giving rise to the following expression for total RF loss:

$$\text{net injected RF power} = \text{total RF power dissipation} = \mathbf{w}^H \overline{\Phi} \mathbf{w} \quad (11)$$

In Eqn. 11 $\overline{\Phi} = \Phi + \Phi_{\text{other}}$, where Φ_{other} , a positive semidefinite matrix, characterizes RF losses in hardware and through radiation¹⁷. Reduction of these losses that are outside the subject is an important goal of coil / shielding / component design and implementation. In cases where RF loss in the subject dominates the total RF loss, $\overline{\Phi}$ becomes a good approximation of Φ . In any case total RF power dissipation / $\overline{\Phi}$ represents an upper bound, or conservative estimate, of RF power deposition / Φ in the subject.

With power sensors at the ports capable of measuring forward and reflected power (Fig. 1), the difference between sum of forward power and sum of reflected power,

$\sum_n p_{\text{fwd}}^{(n)} - \sum_n p_{\text{rfl}}^{(n)}$, gives net injected RF power, allowing $\overline{\Phi}$ to be estimated through experiments. Specifically, given $\mathbf{w}_q$, an input configuration for the q^{th} experiment, measured forward and reflected power values are related to $\mathbf{w}_q$ by

$$\sum_n p_{\text{fwd},q}^{(n)} - \sum_n p_{\text{rfl},q}^{(n)} = \mathbf{w}_q^H \overline{\Phi} \mathbf{w}_q = \sum_{i,j} (w_q^{(i)})^* w_q^{(j)} \overline{\Phi}_{ij} \quad (12)$$

Eq. 12 is a linear equation with $\overline{\Phi}_{ij}$, the entries of $\overline{\Phi}$, as the unknowns, and product terms, $(w_q^{(i)})^* w_q^{(j)}$, as the coefficients. Carrying out calibration experiments with N^2 or more properly selected input configurations played out one at a time can probe the RF loss characteristic of the multi-port network. In particular, the coefficient and power values from the experiments allow assembly of Eq. 12-type equations and subsequent determination of all the entries of the N-by-N Hermitian matrix $\overline{\Phi}$. This process does not involve MR imaging and may be completed in a fraction of a second with an automated measuring system. Once the calibration is complete, the global SAR model $\mathbf{w}^H \overline{\Phi} \mathbf{w}$ predicts, for the specific subject, the SAR consequence of any input configuration or parallel RF transmit pulses. The APPENDIX shows an algorithm that was developed for prescribing input configurations applicable to the calibration. The algorithm maintains the robustness of the solution against perturbation or noise by constructing a complete set of Eq. 12-type equations with a condition number of $\sim N$.

The above scheme can be extended to model and predict individual channel forward or reflected power. In this case Eq. 12 is modified to assume the form:

$$p_{\text{fwd},q}^{(n)} = \mathbf{w}_q^H \overline{\Phi}_{\text{fwd}}^{(n)} \mathbf{w}_q = \sum_{i,j} (w_q^{(i)})^* w_q^{(j)} \overline{\Phi}_{\text{fwd},ij}^{(n)} \quad (13)$$

$$p_{\text{rfl},q}^{(n)} = \mathbf{w}_q^H \overline{\Phi}_{\text{rfl}}^{(n)} \mathbf{w}_q = \sum_{i,j} (w_q^{(i)})^* w_q^{(j)} \overline{\Phi}_{\text{rfl},ij}^{(n)} \quad (14)$$

As is clear from examining Eqs. 12–14, the same coefficient and power values for determining $\overline{\Phi}$ are sufficient for further determining $\overline{\Phi}_{\text{fwd}}^{(n)}$'s and $\overline{\Phi}_{\text{rfl}}^{(n)}$'s, and thereby characterizing the individual RF transmit channel's forward and reflected power. The predicted values for the n^{th} channel's forward and reflected power given an arbitrary input configuration $\mathbf{w}$ are $\mathbf{w}^H \overline{\Phi}_{\text{fwd}}^{(n)} \mathbf{w}$ and $\mathbf{w}^H \overline{\Phi}_{\text{rfl}}^{(n)} \mathbf{w}$ respectively.

Subject-specific planning and monitoring of power transmission

When applying parallel RF transmission in practice, coupling and interaction taking place in the multi-port coil structure as well as in the subject can significantly affect individual channel RF power transmission towards and away from the subject. Tracking and predicting the effects and proactively managing power transmission is important for ensuring a smooth scan.

The present method's capacity for predicting the forward and reflected power for any RF excitation are most relevant: 1) Given the peak power rating of the power amplifiers assigned to drive the parallel transmission channels, knowing in advance the peak power requirements for the individual channels allows the user to proactively adapt the excitation pulse design (e.g., by applying VERSE¹⁸) and/or reconfigure the transmit hardware (e.g., by updating the power combination scheme applied to the component amplifier units). 2) Given the reflected power handling capacity of the amplifiers / circulators on the parallel transmission channels, knowing in advance large peak reflected power for the individual channels similarly allows the user to implement software- and/or hardware-based mitigation strategies. 3) As a supplement to global power prediction and monitoring, checking the individual channel power predictions against actual measurements, or comparing the Φ matrices determined at baseline and that updated periodically afterwards, provides diagnostics that can detect in real-time system changes caused by, for example, hardware failure, system instability or patient position change, and can be used as triggers to suspend scanning as needed. In other words, using the power prediction models both in planning and in monitoring may avert amplifier peak power or voltage standing wave ratio faults, protection hardware breakdown, and excessive SAR due to system failure.

With these functionalities as well as global SAR prediction for guiding pulse design, the present method is henceforth referred to as a method of Power Prediction and Monitoring (PPM).

Simulation studies

Parallel RF transmission cases were studied with finite difference time domain (FDTD) simulations (xFDTD, Remcom, State College, PA). For each case FDTD calculations quantified EM field and net RF power dissipation corresponding to various input configurations, allowing assessment of global SAR calibration and prediction. Exclusion of coil loss and measurement noise is relatively straightforward in simulations, which facilitates an examination of the exact effects of field interferences and array coil loading on global SAR as well as an illustration of global SAR prediction accuracy.

A first study examined feasibility of global SAR calibration and prediction in 7T body imaging. The setup involved a human model with heterogeneous electrical properties and tissue density, and an RF coil composed of 4 loop elements placed approximately 2 cm from the back of the human model (Fig. 2a). The FDTD calculations used a cell size of $5 \times 5 \times 5$ mm³.

A total of thirty nine simulations were conducted employing the same coil-subject setup but different input configurations. For each input configuration the net power dissipation reading, as generated by the FDTD calculations, was recorded. To emulate a use scenario of calibration followed by actual scanning, the first twenty of the total of thirty nine experiments were conducted with input configurations prescribed for SAR model calibration as described above. These twenty input configuration specifications and their resulting net power dissipation values were used to assemble a set of twenty Eqn.11-type equations, which was solved with a least squares fit to give an estimate of Φ . Given the estimate of Φ and the specification of any one input configuration, denoted as $\mathbf{w}$, $\mathbf{w}^H \Phi \mathbf{w}$ provides a

prediction of net RF power dissipation. In the remaining nineteen experiments the input configurations were randomly prescribed. Plugging each of these nineteen input configuration specifications into $\mathbf{w}^H \mathbf{\Phi} \mathbf{w}$ gave a predicted net RF power dissipation, which was compared to the corresponding one that was directly quantified by the FDTD calculation. The comparison allowed an assessment of the accuracy of the model-based global SAR predictions and the validity of the prediction model.

To investigate impact of coil-subject setup on the global SAR model, in a second study two different setups were modeled each involving four identical loop elements of size $7 \times 7 \text{ cm}^2$ placed about 1 cm above a uniform water phantom (Fig. 3). Overlapping of elements was accommodated by displacing the elements slightly relative to one another in the anterior-posterior direction. The water phantom had dimensions of $24 \times 20 \times 24 \text{ cm}^3$, with conductivity 0.6 S/m and relative permittivity 80, approximating electrical properties of human tissues at 298 MHz. With either setup the set of input configurations included 16 programmed ones for model calibration and 8 randomly generated ones for model validation. Global SAR characteristics exhibited by the two setups were compared.

Phantom and *in vivo* studies

On a whole body 7 T scanner (Siemens Medical Solutions, Erlangen, Germany) with eight channel parallel RF transmission capability, experiments were performed to evaluate the global SAR calibration and prediction method under actual MR imaging conditions. Forward and reflected power on the 8 transmit channels were measured using a power sensor (NRP-Z11, Rhode & Schwarz, Munich, Germany) connected to directional couplers located at the output ports of the RF power amplifiers. An RF switch (National Instrument Dual 16x1 MUX) and customized application software for instrument control and data logging were additionally used to automate sequential collection of all power measurements using the single power sensor.

One set of power measurements was performed in a setup for parallel transmit MRI of a cylindrical phantom ($\sigma = 0.7 \text{ S/m}$, $\epsilon_r = 80$). The 4-port Tx-Rx array coil used in these experiments was comprised of four striplines on an enclosing former (Fig. 4a). (Additional striplines and a separate sodium birdcage co-located on the same former as part of a multinuclear coil designed for other purposes were not used in these studies.) Four pre-specified 18 msec-long staircase-shaped parallel RF pulses (Fig. 4c) were used to introduce thirty six input configurations, including sixteen programmed ones for model calibration, followed by twenty randomly generated ones for model validation. The pulses were played out several times to allow the automation software adequate time to communicate with the RF switch and the power sensor for sequential collection of all power measurements. The study was repeated in a more general coil-object setup, where four loop elements that formed a Tx-Rx array coil were arbitrarily placed on a head shaped phantom (Fig. 4b).

In vivo evaluations were conducted in human volunteers. In one study an 8-channel stripline array was used for parallel transmit MRI of a volunteer's left knee on the 7T scanner. Eight 40 msec-long staircase-shaped parallel RF pulses were used to introduce 100 input configurations, including 64 (designed with the algorithm shown in APPENDIX) for model calibration and an additional 36 randomly generated configurations for model validation. The automation program for instrument control and data logging was integrated with a data analysis script written in Matlab (MathWorks, Natick, MA) and C++.

It is appropriate to integrate into a scanner's existing PRESCAN the staircase pulsing scheme in which the SAR model is calibrated and validated prior to the MR scan. The established SAR prediction model can then be used to guide the calculation of SAR-minimized excitation pulses prior to the MR scan. Such proactive SAR management was

implemented in an actual imaging study where large-tip-angle parallel excitation pulses were used for phantom imaging on the 7T scanner. The pulses were designed with the prediction model incorporated¹⁹. The corresponding net power, quantified as sum of forward power measurements in all channels minus sum of reflected power measurements in

all channels $\left(\sum_n P_{\text{fwd}}^{(n)} - \sum_n P_{\text{rfl}}^{(n)}\right)$ was recorded as a function of time during payout. The experimental result was compared to the prediction given by the model to check the accuracy of the guidance the model had provided to the calculation of the SAR-minimized pulses (i.e., the agreement between what the design algorithm thinks is the SAR consequence of a design and the actual SAR consequence).

Evaluations of support for planning and monitoring

Experiments were further conducted to examine potential use of the power prediction models in planning and monitoring of parallel RF transmission on an MR scanner. Some modifications were applied to the phantom experiment setup described above. For this study instead of using the stock directional couplers that are located at the output ports of the power amplifiers, three MR compatible directional couplers were placed near the input ports of three loop elements. The latter formed a three-port Tx-Rx array coil, and were again placed around the head-shaped phantom in an arbitrary fashion. The modified directional coupler (power sensing) locations allowed tighter tracking of power dissipation in the subject as significant RF losses in the long cables linking the amplifiers and the coil are excluded from the power measurements and prediction models. (For instance, RF loss in cables account for over 50% of total RF power delivered by the RF power amplifiers on our 7T system.) Calibration and validation were performed as before.

Continued investigations examined diagnostic and monitoring potential of the present method by artificially introducing fault conditions. After calibration and validation under a baseline condition, the system hardware was altered to emulate two system fault scenarios. In Case 1 one capacitor in the loop structure of Element 3 was de-soldered, causing an open condition. In Case 2 an extra segment of coaxial cable was added to the cable linking Element 2 and its corresponding power amplifier, adding a phase offset of about 45 degrees. The actual measurements taken after the alterations were compared to the baseline measurements and the baseline model predictions, and differences in prediction models were further noted.

RESULTS

Results from the FDTD simulation using the human model are summarized in Fig. 2. Based on the net power values quantified by the FDTD calculations for the twenty input-configuration cases prescribed for SAR model calibration, Φ was determined. In each of the additional nineteen input-configuration cases, Fig. 2b shows that net power predicted with $\mathbf{w}^H \Phi \mathbf{w}$, the calibrated global SAR model, was in excellent agreement with net power quantified directly by FDTD. The average percentage difference between predicted and directly quantified net power was 2 percent.

In Fig. 3c and f, results from the second FDTD simulation study again indicate high accuracy achieved by the calibrated global SAR models. In addition, as shown in Fig. 3b, the magnitude of estimated Φ for Setup 1 shows significant off-diagonal power correlations between Elements 1 and 3, and 2 and 4, consistent with the interference patterns expected of the geometrical arrangement of the elements and the object (depicted in Fig. 3a). Similar observations may be made for Setup 2 (geometrical arrangement shown in Fig. 3d, magnitude of estimated Φ shown in Fig. 3e). For each setup, variation in the diagonal entries of Φ correlates with loading difference due to the elements' varying proximity to the

lossy object. In Setup 2 for example, Element 4 was the furthest from the object, while Elements 1 and 2 were the closest. Furthermore, notice that for the two setups the ratios of the largest to the smallest eigenvalues were, respectively, 3.1 and 6.3. These indicate that with either setup, different RF pulses with the same sum of magnitude squared could have significantly different SAR consequences (up to a factor of 3.1 and 6.3, respectively). An identity matrix-based SAR model would not be able to distinguish between such pulses.

For each of the two parallel Tx phantom studies, results demonstrated feasibility of the global SAR calibration and prediction method under actual MR imaging conditions (Fig. 4). The measurement hardware was supportive of global SAR modeling and prediction. The method handled both a conventional encircling Tx array coil and a more generally configured Tx array coil well, producing robust Φ estimates and giving model-based SAR predictions that closely matched actual measurements. Results for the 4-loop array coil setup are shown in Fig. 4d.

The measurement automation in the *in vivo* study streamlined the calibration and validation process. The entire process of RF pulse playout on the scanner, Φ determination, and measurement-versus-prediction display that allows the user to validate the calibration, was completed in less than 18 seconds. Fig. 5a shows that the predictions given by the calibrated global SAR model were in excellent agreement with direct measurements. Also notice that model calibration was performed at a low power levels, yet the validity of the prediction model held over a considerably wide test range, suggesting robust modeling as well as good system linearity. Similar results were achieved in an additional *in vivo* study. The mean $\pm$ standard deviation of the difference between the predicted and measured power were $3.68 \pm 4.39\%$ and $2.51 \pm 3.83\%$ for the two studies respectively. In both studies and in all validation cases the maximum prediction error was less than 11%.

For the phantom experiment employing designed large-tip-angle parallel excitation, Fig. 5c shows excellent agreement between predicted and measured global RF power deposition. The measured net power was nevertheless somewhat lower than that predicted at sharp peak locations. This was attributed, in part, to the timing offset between power measurement and RF pulse update (every 10 usec). Measurement imperfection was estimated to account for prediction error of up to $\sim 10\%$ ²⁰.

Fig. 6 shows one set of results from the study that evaluated the planning and monitoring potential of the present method. These results demonstrate that the present method was able to accurately predict not only the net power transmission but also the individual channel power transmission in both forward and reflected directions.

Three additional sets of results were obtained in the study, with setups that included a baseline and two that had artificially introduced faults. For the baseline case, calibration and validation steps were performed. Fig. 7a shows $\overline{\Phi}_{fwd}^{(n)}$, $\overline{\Phi}_{rfl}^{(n)}$ and Φ that were estimated from the nine calibrating steps. For the subsequent nine validating steps (each with an input configuration that was randomly generated) Fig. 7b shows a comparison between the model based predictions and the actual measurements. Again, high prediction accuracy was achieved by the present method. Similar calibration and validation steps were also performed for fault cases 1 and 2. Fig. 8b shows that the opening of Element 3 in Case 1 caused significant changes in Channel 3 reflected power measurements from baseline, especially for experiment steps whose input configuration significantly engaged Channel 3 (e.g., Steps 14, 16 and 17). Fig. 8e shows that the extra phase shift along Channel 2 caused significant changes in Channel 2 power measurements from baseline, especially for experiment steps where the input configuration significantly engaged Channel 2 (e.g., Steps 11, 13 and 18). As expected, the ease of detecting a fault based on a comparison of

prediction and measurement was input configuration-dependent. However, a systematic detection was possible by evaluating the eigenvalues or eigenvectors of Φ . In Cases 1 and 2, the altered system remained linear, which allowed power re-characterization by repeating the calibration and validation steps leading to updated $\overline{\Phi}_{fwd}^{(n)}$, $\overline{\Phi}_{rfl}^{(n)}$ and Φ . For the baseline, Case 1 and Case 2, the significant eigenvalues of Φ were, respectively, {4, 43}, {5, 29}, and {19, 41}. In further screening of the spectra of $\overline{\Phi}_{fwd}^{(n)}$ and $\overline{\Phi}_{rfl}^{(n)}$, sharp contrast was present between Case 1 $\overline{\Phi}_{rfl}^{(3)}$ and baseline $\overline{\Phi}_{rfl}^{(3)}$, and between Case 2 $\overline{\Phi}_{fwd}^{(2)}$ and baseline $\overline{\Phi}_{fwd}^{(2)}$, especially in terms of the largest eigenvalue (Fig. 8c and f). Similarly, differences in patterns of eigenvectors were noticed between the baseline and fault cases. These along with the pattern changes of eigenvalues were certainly reflected in $\overline{\Phi}_{fwd}^{(n)}$, $\overline{\Phi}_{rfl}^{(n)}$ and Φ (magnitude shown in Fig. 8a and d).

DISCUSSION AND CONCLUSION

From the perspective of advancing high field MR, it appears inevitable that one needs to perform *in vivo* SAR model calibration for managing SAR, just as one needs to perform B1 map calibration for managing excitation profile. Given the task of scanning a specific subject, to maximize performance one uses B1 and SAR calibration results acquired in the presence of the subject as key inputs to guide optimization of pulse or shimming calculations. Performance aside, to ensure a smooth scan it is highly desirable to accurately characterize RF power transmission and deposition, and thereby comply with safety and hardware limits.

These considerations on performance and safety point to a sensible *in vivo* MR paradigm: a PRESCAN process calibrates SAR prediction models specific to the subject and performs model-guided excitation pulse calculations to manage SAR for the intended scans; actual scans then proceed with the operator knowing in advance the SAR consequence, knowing that RF transmit has been performance-optimized for this specific subject and won't exceed safety limits or hardware capability, and knowing in real-time should any of the transmit chains malfunction.

Given the linearity of a common parallel RF transmission setting, an approach is presented that employs the concepts of a multi-port network and an overall linear system to *directly* link any RF pulses (system inputs) to RF fields and SAR in the scanned subject (system outputs). The accordingly established PPM method enables, for example, accurate characterization and proactive management of global SAR *in vivo* without forcing one to deal with subject, coil or hardware details. This is in contrast with other possible sensing and processing approaches that rely on voltage, current or field responses measured on or around the transmission hardware^{21, 22} with nontrivial sensor setup and phase-sensitive detection. While allowing one to infer some characteristics of RF transmission and SAR, these intermediate measurements may not lead to definite guidance to RF pulse design without additional calibrations that complete the determination of the link between the RF pulses and global SAR.

In a nutshell the PPM method, using judiciously defined input configurations, performs a finite number of calibration experiments and records the outputs of the overall system. The results of these calibration experiments are then used to build quadratic prediction models capable of relating any predetermined RF pulses to RF transmission and global SAR, and assisting in system error detection. The effects of field interferences and array coil loading on RF transmission and global SAR are captured by the models. These effects are nonexistent or less complex in a conventional RF transmission setting, and cannot be

correctly captured with pseudo RF power models (e.g., the common sum of magnitude squares of pulse waveform samples) that were originally introduced for regularization¹¹.

In characterizing subject-specific global SAR with the PPM method, power measurements taken at a set of discrete sensor locations together with the law of conservation of energy are used to monitor and predict an integral of non-negative local RF power deposition. RF pulse or shimming calculations that minimize the integral under the guidance of PPM tend to curb the occurrence of excessive local SAR. Structurally this tendency may be appreciated by noticing that minimizing $\mathbf{w}^H \Phi \mathbf{w}$ is equivalent to minimizing sum of $\mathbf{w}^H \Lambda(\mathbf{r}) \mathbf{w}$ over voxels (Eqn. 5) where Φ and $\Lambda(\mathbf{r})$'s are positive semidefinite – a $\mathbf{w}$ with excessive $\mathbf{w}^H \Lambda(\mathbf{r}) \mathbf{w}$ at a location tends (but does not guarantee) to elevate $\mathbf{w}^H \Phi \mathbf{w}$, which would steer the calculations to favor other designs.

Due to the integral nature overestimation of global SAR in the subject could result if the power measurements include hardware RF loss in addition to RF loss in the subject. However this effect does not impact the validity of treating the quantified global SAR as the upper-bound of the true global SAR. Mitigation of the effect is possible through improvements in hardware / measurement scheme (discussed below).

Subject-specific local SAR characterization is a considerable challenge, a practical solution of which is a prerequisite for addressing with confidence the possibly more limiting²³ local SAR limits. The PPM method, working with spatially resolved power or temperature measurements, has potentials, as explored in a preliminary investigation²⁴. In comparison a numerical simulation approach promises to provide an abundance of information that are difficult to obtain noninvasively. These include spatially resolved field, SAR or even temperature information that may enable the use of a multi-pronged strategy for mitigating local SAR²⁵. Possible SAR quantification errors resulting from a mismatch between the simulated and the real settings are difficult to analyze or manage however, which impacts validity of simulation-based SAR estimates and spells the need for an *in vivo* applicable scheme for assessing the match or for bounding the errors.

Implementation of PPM does not require special hardware. MR scanners equipped with parallel RF transmission channels already check in real-time, with RF power monitors^{10, 26}, forward power transmission in the individual cables linking the RF power amplifiers and the multi-port transmit coil, and are programmed to stop the scan upon detection of an excessive level of total forward power. The new implementation entails an upgrade of the RF power monitors' capacity to include individual channel reflected power, and a program that processes the forward and reflected power readings according to the descriptions described in the THEORY AND METHODS sections.

Since the transmit and receive periods of an MR experiment typically do not overlap, the parallel receive channels on the scanner could be utilized to demodulate directional coupler outputs and convert the results to power readings during RF transmission. The resulting parallel power sensing / monitoring capacity could facilitate the implementation of the PPM method.

While the built-in RF power monitors / parallel receive channels were not readily accessible on our system to allow further demonstrations, the present investigation did provide indications of the capability of a full fledged implementation: (I) Calibration completed within a second. The total amount of time required for playing out the calibration configurations scales with N^2 . It can be fit into 1 sec with the staircase-type pulses and parallel RF power sensing, if N is not substantially greater than 32. This makes the time cost of performing a calibration *in vivo* negligible. (II) Tighter characterization of global SAR, which can be expected in a sensing setup that has directional couplers located close to the

coil and better controlled power sensors. This improves the guidance provided to pulse design optimization and helps refine the SAR limit setting on the scanner. (III) Real-time monitoring and automatic protection. The individual channel power characterization and prediction capability, when leveraged to check channel integrity during the scan, helps ensure real-time detection of hardware failure. Such events as breakdown of components in the transmit hardware, alteration of coil geometry / placement, and subject posture change could all lead to changes in system characteristics, manifesting as a discrepancy between predictions and measurements, or, between characterizations done before and after the change. A system state change is a cause for intervention, including suspension of scanning for subject and scanner protection.

There is a link between the global SAR model presented here and the scattering matrix $\mathbf{S}$ of an N-port network²⁷. Given an $\mathbf{S}$ matrix that is determined from incident and reflected wave measurements (amplitudes and phases) taken at the N directional couplers during a calibration, one can subsequently express total RF power dissipation in the network as $\mathbf{a}^H(\mathbf{I} - \mathbf{S}^H\mathbf{S})\mathbf{a}$ for any $\mathbf{a}$, where $\mathbf{a}$ is an N-by-1 vector collecting the instantaneous incident wave measurements taken at the N ports. This leads to $\Phi = \mathbf{L}^H(\mathbf{I} - \mathbf{S}^H\mathbf{S})\mathbf{L}$, where matrix $\mathbf{L}$ captures the linear mapping from $\mathbf{w}$ to $\mathbf{a}$ as well as other scaling in the system. $\mathbf{L}$ is fixed for a given hardware-subject setup. An additional link between the global power correlation matrix and parallel receive noise covariance matrix Ψ was suggested in the past, based on the principle of reciprocity²⁸. In addition to offering alternative ways to calibrating a global SAR model, these links also suggest that the global power correlation matrix provides important information about $\mathbf{S}$ and Ψ , and vice versa. For example, knowledge of Φ and $\mathbf{L}$ allows determination of all of the scattering matrix $\mathbf{S}$'s eigenvalues. This example offers insights on the sensitivity of a Φ -based scheme in detecting system state changes.

The system perspective described earlier reinforces the rationale for a parallel transmission operation mode involving calibration + pulse design + transmission. It also helps pinpoint the limitations of some conceptions. As a simple example, in the early days of parallel transmission development some thought that B1⁺ of any element in a transmit array coil should be mapped while all other elements are detuned or deactivated. One issue here is that detuning or deactivation alters the system, making it different than what is encountered during the actual parallel RF transmission. Similarly, for schemes that attempt parallel transmission power calibration through $\mathbf{S}$ or Ψ measurement, it is important to make the measurement condition the same as what is in place during actual parallel RF transmission. In an example where Ψ is measured using the receiver chains, due to changes in impedance seen by the ports (from the low input impedance of the pre-amplifiers to the ~50Ω output impedance of RF power amplifiers), likely alteration of RF current and field patterns within the network (including the coil structure and the subject body) can result in changes in RF loss characteristics, rendering the derived SAR model less accurate predicting the RF loss characteristics for the actual parallel transmit MRI. In both examples above, system alteration can invalidate the application of (a linear system's) superposition principle and can cause errors in using calibration data for prediction or for guiding actual MR. Along the same line, the benefits of *in situ* measurements can be appreciated – for accurate models and results it is most desirable to calibrate the same system as that in effect during actual MR.

The present work is part of an effort to create a solution that quantitatively tracks and proactively manages SAR on a subject-specific basis. With the analysis shown and the performance demonstrated, there is the confidence that some significant improvements of high field clinical MR safety and performance are within reach.

APPENDIX

An algorithm that prescribes a complete set of input configurations for model calibration is provided here. For an N-port transmit coil, the algorithm gives N^2 input configurations (stored in input_config).

```
n_ports=8
input_config=sqrt(2)*eye(n_ports);
for id=1:n_ports-1
    building_block=zeros(2,n_ports);
    building_block(:,1)=[1; 1];
    building_block(:,id+1)=[1; j];
    tmp=building_block;
    for shift=1:n_ports-1-id
        tmp=[tmp; circshift(building_block,[2 shift])];
    end
    input_config=[input_config; tmp];
end
```

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Figure 1.

A system perspective for parallel RF transmission: The transmit channels, the transmit coil and the subject together form a system, with $w^{(n)}$, the parallel RF pulse waveforms, as inputs, and B_1^+ field and RF energy deposition as outputs. Parallel RF transmission system hardware is normally set up in such a way that the electromagnetic field induced in the subject responds linearly to the inputs.

(a)

(b)

Figure 2.

Global SAR calibration and prediction in a simulation study involving a heterogeneous human model and an RF coil composed of 4 loop elements. (a) Geometry of the human model and the array coil. (b) Power as a function of input configuration. Based on the net power values quantified by the FDTD calculations for the 20 input-configuration cases prescribed for calibration, the power correlation matrix Φ was determined. In each of the additional 19 input-configuration cases the net power predicted with the calibrated global SAR prediction model $\mathbf{w}^H \Phi \mathbf{w}$ appeared to be in excellent agreement with the net power quantified directly by FDTD, demonstrating the validity of the model.

Figure 3.

Investigation of the impact of coil-subject setup on the global SAR model. a,d) Two different simulated setups, each involving four identical loop elements of size $7 \times 7 \text{ cm}^2$ placed approximately 1 cm above a uniform water phantom. b,e) Absolute magnitude of estimated global power correlation matrix Φ for each setup. Off-diagonal entries indicate power correlations between distinct transmit elements, whereas variations in diagonal entries correlate with loading differences caused by the elements' varying proximity to the lossy object. The structure of Φ for Setup 1 (b), for example, indicates significant power correlations between transmit elements 1 and 3, and 2 and 4, consistent with expected interference patterns given the geometrical arrangement of the elements and the object. c,f) Power as a function of input configuration. With either setup the set of input configurations included 16 programmed ones for model calibration and 8 randomly generated ones for model validation.

Figure 4. Experimental global SAR model calibration and validation. (a) Photograph of a coil with four stripline elements used for 4-channel parallel transmit MR of a cylindrical phantom. (b) Photograph of a general coil-object setup with four loop elements arbitrarily placed on a head-shaped phantom. (c) Four pre-designed 18 msec-long staircase-shaped parallel RF pulses were used to introduce 36 input configurations, including 16 for model calibration (definitions shown in the table) and 20 randomly generated inputs to validate the model-based global SAR predictions. The inset shows the definitions of the calibration inputs. (d) For the 4-loop array coil setup, a comparison of model-based SAR predictions with actual measurements.

Figure 5.

Automated SAR model construction for *in vivo* parallel Tx MR resulted in an 18 second calibration and validation process. Results from an *in vivo* study indicated that the predictions given by the calibrated global SAR model were in excellent agreement with direct measurements (a), with a worst case prediction error of less than 10% (b). Results from a further study on the use of a calibrated SAR prediction model in RF pulse optimization indicated that the model offered reliable guidance to the pulse design algorithm – the predicted net RF power as a function of time was in excellent agreement with that actually measured during the playout of the designed parallel excitation (c).

- direct measurement
- calibration data fitting
- prediction

Figure 6.

Three channel parallel Tx study with a coil-object setup that had three loop elements arbitrarily placed on a head shaped phantom and formed a Tx array coil. Global SAR model calibration and validation was performed, which led to model predictions that accurately matched net power measurements (bottom plot). The power measurement data were further processed to calibrate predictive models of individual channel forward and reflected power,

which gave $\mathbf{w}^H \overline{\Phi}_{fwd}^{(n)} \mathbf{w}$ and $\mathbf{w}^H \overline{\Phi}_{rfl}^{(n)} \mathbf{w}$, n th channel forward and reflected power predictions respectively, for any input configuration $\mathbf{w}$. For the randomly prescribed input configurations (Steps 10 through 18), predictions from these models were in excellent agreement with actual measurements (plots labeled Ch n fwd and Ch n rfl).

Figure 7.

Baseline model calibration and validation for system fault monitoring. (a) Magnitude

displays of $\overline{\Phi}_{fwd}^{(n)}$, $\overline{\Phi}_{rfl}^{(n)}$ and Φ , which were calibrated using power measurements obtained under Input configurations 1 through 9. (b) Comparison of model predictions with power measurements for randomly generated Input configurations 10 through 18 indicated accurate prediction. These models / data serve as a reference for detecting system changes from baseline at a later time.

Figure 8.

Cases 1 and 2 in a system monitoring study. The baseline was altered to emulate two system fault scenarios. In Case 1 the baseline setup underwent a system change in which Element 3 was forced to open. Re-characterization was performed. (a) shows updated $\overline{\Phi}_{fwd}^{(n)}$, $\overline{\Phi}_{rfl}^{(n)}$ and $\overline{\Phi}$. (b) shows predictions, measurements, as well as predictions from baseline models (circles).

(c) shows maximum eigenvalues of Case 1 $\overline{\Phi}_{fwd}^{(n)}$ and $\overline{\Phi}_{rfl}^{(n)}$ (solid bars) vs. that of baseline $\overline{\Phi}_{fwd}^{(n)}$ and $\overline{\Phi}_{rfl}^{(n)}$ (non-filled bars). In Case 2 the baseline setup underwent a system change in which Channel 2 had an extra phase offset of about 45 degrees. Results are similarly displayed in d-f. Note the pattern change of a power correlation matrix's eigenvalues: For the baseline, Case 1 and Case 2, the significant eigenvalues of $\overline{\Phi}$ were, respectively, {4, 43}, {5, 29}, and {19, 41}. In terms the largest eigenvalues, further note the sharp contrast between Case 1 $\overline{\Phi}_{rfl}^{(3)}$ and baseline $\overline{\Phi}_{rfl}^{(3)}$, and between Case 2 $\overline{\Phi}_{fwd}^{(2)}$ and baseline $\overline{\Phi}_{fwd}^{(2)}$ (c and f).