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Parallel-Plate Waveguide for Volume Radio Frequency Transmission in MRI

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Abstract

Purpose—Radiofrequency coils in MRI are narrowband resonant structures. The high-quality factor makes their performance sensitive to loading and other operating conditions. We developed a new coil to provide volume transmit in a broad frequency band.

Methods—The proposed coil is a parallel-plate waveguide. When operating in the transverse electromagnetic mode, a uniform magnetic field is generated between the plates for imaging. The impedance of the transverse electromagnetic mode is matched by a broadband matching circuit. The two plates are actively driven with broadband Wilkinson power dividers and combiners. Re-entrant design is further applied to improve the overall efficiency.

Results—We developed a prototype of a coil to image the human forearm with a bandwidth of more than one octave. Its feasibility was demonstrated by proton imaging and phosphorous spectroscopy at 7T.

Conclusions—The proposed coil structure is compact and relatively easier to construct. It can be applied for volume radiofrequency transmit in MRI in a broad frequency band.

Keywords

traveling wave; broadband; multi-nuclear spectroscopy

INTRODUCTION

Conventional radiofrequency (RF) coils such as loop, birdcage, and transverse electromagnetic (TEM) coils are resonant structures tuned to specific Larmor frequencies (1–3). Outside a certain frequency range, RF power input is reflected back to the generator due to impedance mismatch (4). This narrowband feature is typically characterized by the quality factor (Q-factor) (1). On the one hand, high-Q coils are preferred for their low coil/sample losses. On the other hand, it also indicates the sensitivity of a coil to its operating condition. The resonant peaks of high-Q coils can drift with different loadings. This problem is especially pronounced in high-field applications, where the design of transmit coils can be a challenging task (5–8).

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From the electromagnetic perspective, high-Q resonators are sufficient but not necessary for effective RF transmission. The use of traveling-wave structures, which are nonresonant, has been studied extensively in antenna engineering (4). Due to the lack of sharp resonant peaks, broadband antennas require no frequency tuning within its operating bandwidth, and are less sensitive to environment variations and manufacture defects (4).

The use of traveling-wave structures in MRI has been studied by different groups in the past. Brunner et al. (9) demonstrated the feasibility of using the RF shield of an MRI scanner as a waveguide to support traveling waves. The main waveguide structure does not require reactive components for tuning, and a circularly polarized patch antenna was used for RF excitation and signal reception. This concept was further extended by using dielectric inserts to reduce impedance mismatch between air/tissue boundaries (10,11). Zhang et al. (12) performed a thorough comparison of waveguide-based transmitters and the TEM coil via numerical simulations. The investigators found that wave attenuation and reflection from body boundaries caused substantial B_1^+ inhomogeneity, although a high degree of uniformity was achieved inside an empty scanner bore.

One of the main features of waveguide-based transmitters is that the scanner bore size needs to be large enough so that the cutoff frequency of the desired mode is below the Larmor frequency (9). Otherwise, electromagnetic waves are evanescent, not traveling. Because the bore size of regular human scanners is barely large enough to support traveling waves at 7T (9), this type of structure is not suitable for clinical MRI. Tang et al. (13) reported that the TEM mode of a coaxial cable was applied in order to avoid the cutoff frequency issue. However, the requirement of a center conductor makes human imaging difficult. Zhang et al. (14) developed a surface transmitter using the TEM mode supported by two closely spaced conductors; Pan and Zhang later developed a volume transmitter for parallel transmission using an array of such coils (15).

In this study, we investigated the feasibility of using a different traveling wave structure (i.e., parallel-plate waveguide) for MRI. A parallel-plate waveguide is a transmission line that consists of two conductors with appropriate terminations that match its wave impedance. Just like coaxial cables, its wave impedance is frequency-independent. The use of transmission lines as broadband antennas can be dated back to the 1930s (16). This technique enjoyed a long history of success due to its simplicity and, more importantly, its effectiveness. As will be demonstrated, much of these advantages can be retained in MRI. A transmitter as simple and compact as two actively driven parallel conducting plates can provide broadband RF excitations. It should be noted that the proposed method is fundamentally different from the one proposed by Vazquez et al. (17), which applied a birdcage coil and two floating parallel conductors. By leaving both ends of the floating conductors open-circuited, a narrowband transmission line resonator rather than a broadband waveguide is formed (18).

METHODS

Transmitter Design

The electromagnetic field between two parallel plates can be expressed as the combination of different modes, which are not necessarily uniform except for the TEM mode (Fig. 1a) (18). To be useful in MRI, non-TEM (or higher-order) modes need to be suppressed. According to electromagnetic theories, higher-order modes can be classified as TE_n^z and TM_n^z modes, where superscript z represents wave propagation direction, which is also the direction of B_0 field (Fig. 1a), and subscript n is the mode index.

For TE_n^z modes, there exists a non-zero magnetic field component (18)

$$H_x = (j\omega\epsilon/k_c)A_n \cos\left(\frac{n\pi y}{d}\right) e^{-j\beta z}, \quad [1]$$

where A_n is the mode amplitude, d is the distance between the plates, $\beta = \omega \sqrt{\mu\epsilon}$ is the angular wavenumber, and $k_c = n\pi/d$ is the cutoff wavenumber. By Ampere's law, currents flowing on the two plates are in the same direction (common mode) when n is odd, and in opposite directions (differential mode) when n is even. For TM_n^z modes, there exists a non-zero electric field component

$$E_x = (j\omega\mu/k_c)B_n \sin\left(\frac{n\pi y}{d}\right) e^{-j\beta z}, \quad [2]$$

where B_n is the mode amplitude. A virtual ground exists in the middle of the two plates only if n is even, which corresponds to differential modes. Thus, both TE_n^z and TM_n^z are differential when n is even. Because the TEM mode is differential, in theory only higher-order differential modes can be simultaneously excited when the TEM mode is intended. In practice, balanced feeding structures are preferred in order to suppress common modes (4).

The suppression of higher-order modes can be further facilitated by impedance mismatch (19). For the TEM mode, its frequency-independent wave impedance is (18)

$$Z_{TEM} = \eta \frac{d}{w}, \quad [3]$$

where $\eta = \sqrt{\mu/\epsilon}$ is the characteristic impedance between the plates and w is the width of each plate. For TM_n^z mode, its wave impedance is

$$Z_{TM} = \eta \sqrt{1 - \left(\frac{k_c}{k}\right)^2}. \quad [4]$$

For TE_n^z modes, its wave impedance is

$$Z_{TE} = \frac{\eta}{\sqrt{1 - \left(\frac{k_c}{k}\right)^2}}. \quad [5]$$

In both cases, the wave impedances are frequency-dependent. If the distance-to-width ratio is close to unity, the wave impedances of both TE_n^z and TM_n^z modes will be substantially different from that of the TEM mode. For an impedance matching network designed to match cable impedance to that of the TEM mode, higher-order modes are rejected due to impedance mismatch.

Equation [3] further indicates that the TEM mode is broadband in nature since it does not have a size-related cutoff or resonant frequency. Its impedance is determined by the dielectric property of the medium between the plates and the distance-to-width ratio, not the actual size. Thus, the effect of loading to the wave impedance can be approximated by using an effective permittivity (18). Because the human body does not have a regular shape, numerical simulations can be used for more accurate prediction of the wave impedance.

Coil Construction and Re-entrant Termination

Tapered transmission line is typically applied to connect a pair of parallel plates to coaxial cables in microwave engineering (20). However, it does not enable patient access unless the longitudinal axis (z-direction, Fig. 1a) is transverse to B_1^+ field, which results in a B_1^+ distribution not much useful in MRI. To this end, a tapered transmission line is split into two halves, with each half going to the side of the waveguide (Fig. 1b). This effectively transforms the impedance of the TEM mode to the parallel combination of two impedances.

A coil suitable for imaging the human forearm was constructed in the above manner. Copper foils (3M, Saint Paul, Minnesota, USA) 7 cm wide and 20 cm long were laid on the outside of an acrylic tube (McMaster-Carr, Elmhurst, Illinois, USA) with a 10.8-cm outer diameter. The corresponding distance-to-width ratio of the parallel-plate waveguide was 1.54. Each copper foil was split at the ends and tapered down to form a 1-cm gap (Fig. 1b). The entire coil structure does not have any reactive components for tuning.

A broadband Chebyshev matching network was connected across each gap to transform 50- Ω cable impedance to the input impedance at a gap from 100 to 300 MHz (21). It was implemented symmetrically for common-mode suppression (Figs. 1b and 2a). At one end of the waveguide, a broadband Wilkinson power divider was applied to split the RF input equally and in-phase (Fig. 2b) (22). Its outputs were connected to the Chebyshev matching network. At the other end of the waveguide, a broadband Wilkinson power combiner was used to combine the RF energy flowing out of the waveguide and feed it back to the input by using a third Wilkinson power combiner to form an re-entrant scheme (Fig. 1c) (23).

As shown in the diagram in Fig. 1c, the rightmost positive-intrinsic-negative (PIN) diode was switched on during RF transmit. After going through the diplexers, the proton and phosphorous circuits were separate, and the phase shifters shown at the top of Fig. 1c ensured that at 120.3 MHz (for phosphorous) and 297.2 MHz (for proton), the RF input

from one end of the Wilkinson power combiner was in-phase with that from the system power supply.

It should be noted that the Wilkinson power combiner works in even mode when its inputs are fed by equal amounts of RF power (18). Because the two inputs are at equal potential, the interconnecting resistors are superfluous without actual currents flowing through and, therefore, without power loss (18). This is different from bench measurement conditions, where only one input is connected to the RF power supply and the other is terminated by passive load. In that scenario, the Wilkinson power combiner works in the combination of even and odd modes, with half of the RF power lost in the interconnecting resistors (18).

The design of all front-end circuit was facilitated by using ADS (Agilent Technologies, Palo Alto, California, USA). All capacitors were from the ATC-100B series (American Technical Ceramics, Huntington Station, New York, USA) with the exact component values denoted in Fig. 2. The inductance was implemented by 164- and 165-series tunable inductors (Coilcraft, Cary, Illinois, USA). Two types of PIN diode, i.e., UM9415B (Micro-semi, Aliso Viejo, California, USA) and MA4P7470-1072T (M/A-COM, Lowell, Massachusetts, USA), were applied for transmit/receive (T/R) switches. Bench measurements were performed by using an Agilent E-5061A network analyzer (Agilent Technologies). A 1.25-L phantom, which was a 20-cm-high and 8-cm-wide bottle filled with 55 mM NaCl and 100 mL phosphate, was prepared for bench measurement and phantom imaging.

For comparison purposes, an eight-rung unshielded linear high-pass birdcage coil was constructed by taping a 3/8-inch-wide copper foil on an acrylic cylinder with a 10.8-cm outer diameter, which was the same as the parallel-plate waveguide. The length of the birdcage was also the same as the parallel-plate waveguide. It was tuned and matched to 297.2 MHz with the saline phantom by using 2.3 pF capacitors.

Receiver Design

Signal reception follows the principle of open-circuited transmission-line resonator (18). It is preferred to sense the induced voltage rather than the current for better signal-to-noise ratio (SNR) (24). This was achieved by adjusting the total electrical length of an open-circuited transmission line to be equal to multiples of half wavelength, and connecting one end of the open-circuited waveguide to a low-noise preamplifier (LNA) with a high input impedance (24). In Figure 1c, the rightmost PIN diode is switched off during signal reception, which creates one open end of the transmission line. At each frequency, the phase measured between the PIN diode and the input of the LNA through the Wilkinson power dividers/combiners, the parallel-plate waveguide, the diplexers, and the T/R switches were adjusted to be equal to multiples of 180° . The high-impedance LNAs were implemented by using an ATF-54143 low-noise enhancement mode pseudomorphic high-electron mobility transistor (HEMT) (Avego Technologies., San Jose, California, USA) for the first stage and a BFP-720 low-noise silicon germanium bipolar transistor (Infineon Technologies, Neubiberg, Germany) for the second stage at 120.3 and 297.2 MHz, respectively. At both frequencies, the quiescent point of the ATF-54143 was 3V/ 40mA and that of the BFP-720 was 3V/15mA. The input matching network of the first-stage transistor was adjusted to provide large impedance when looking into it.

Computer Simulations

Full-wave simulations were performed by using FEKO (EMSS S.A. Pty Ltd, Stellenbosch, South Africa) and XFDTD (Remcom Inc., College Station, PA). FEKO was applied to determine the input impedance of the parallel-plate waveguide with the 20-cm long and 8-cm outer diameter cylindrical phantom used in experiments ($\epsilon_r = 78$, $\sigma = 0.657$ S/m). Resistors were connected to the four gaps of the waveguide and uniformly adjusted until the reflection coefficient at one gap was below -10 dB from 100 to 300 MHz.

XFDTD was used to examine the B_1^+ uniformity and the local specific absorption rate (SAR) inside the forearm of the Duke human model (23). The geometry of the parallel-plate waveguide was the same as in FEKO simulations. The mesh size was 2 mm isotropic. Ten layers of perfectly matched layer absorbing boundary condition were applied in each direction and separated from the coil model by a 40-layer buffer zone. Tissue properties were specified according to literatures at 120.3 and 297.2 MHz, respectively (25). Two in-phase 1-V voltage sources were applied across the gaps at one end of the parallel-plate waveguide with source impedances equal to that determined by FEKO. The two gaps at the other end of the waveguide were passively terminated by using the same amount of resistance. Time-domain simulations were run long enough to ensure numerical convergence. The frequency-domain electromagnetic fields were obtained by the Fourier transform and applied to compute B_1^+ maps and the local SAR. The B_1^+ was normalized by the square root of the total RF power, which represents the transmit efficiency. The 10-g averaged local SAR was normalized to 1-W RF power absorption by the forearm. The radiation efficiency (i.e., the ratio of radiated power to the sum of radiated and absorbed powers) was also calculated.

The eight-run unshielded linear birdcage coil was also simulated at 297.2 MHz by using the XFDTD. The coil conductor was modeled as wire filaments. The capacitors on the end rings were adjusted to 1.6 pF for the birdcage mode to occur at 297.2 MHz in simulations, which is lower than its hardware implementation. This is mainly because modeling the coil conductor as filaments increases the coil inductance and, therefore, decreases the required tuning capacitance. A 1-V source with 50- Ω source impedance was applied for excitation. The rest of the simulation procedure was the same as that for the parallel-plate waveguide. This procedure was repeated at 120.3 MHz with the forearm model. The tuning capacitance was 12.1 pF. In addition, both the parallel-plate waveguide and the birdcage coil were simulated with the saline phantom and a gel phantom ($\epsilon_r = 2$, $\sigma = 0$ S/m) at 297.2 MHz.

MRI Measurements

A 7T MRI scan was performed on a Siemens MAGNETOM scanner (Siemens Medical Systems, Erlangen, Germany) at Auburn University. A proton B_1^+ map in the saline phantom was acquired by using Siemens' product sequence based on the double-angle method. The scan parameters were as follows: pulse repetition time (TR) = 1000 ms; echo time (TE) = 14 ms; field of view (FOV) = 300 $\times$ 300 mm; and matrix size = 256 $\times$ 256. The turbo spin echo (TSE) image of the same phantom was also acquired using the following parameters: TR = 3600 ms; TE = 87 ms; FOV = 179 $\times$ 199 mm; matrix size = 362 $\times$ 448; and one average. The TSE image of the human forearm was acquired by using the same

parameters but a slightly shorter 83-ms TE. For comparison, the gradient echo (GRE) image of the forearm was acquired using the following parameters: TR = 8.6 ms; TE = 3 ms; FOV = 250×250 mm; matrix size = 512×512 ; and one average.

The proton transmit efficiency was measured with the 1.25-L phantom by using a free induction decay sequence with a 1-ms rectangular pulse, 1500-ms TR, and 1.1-ms TE. The RF power was increased incrementally until maximum signal intensity was detected. The finest voltage resolution was 2 V (peak RF voltage reading from scanner). The same procedure was repeated for the unshielded linear birdcage coil with a dedicated proton interface box that consists of a T/R switch and a low-impedance LNA. The power loss of each front-end circuit was considered when calculating the peak RF voltage required for maximum signal intensity. Phosphorous spectroscopy of the human forearm was performed using a Siemens' product chemical shift imaging (CSI) sequence. The scan parameters were as follows: TR = 1500 ms; TE = 0.1 ms; bandwidth = 5 kHz; cubic volume size = 25 mm; and 16 averages.

RESULTS

By using an effective relative permittivity of $\epsilon_r = 39.5$, which is the average of those of air and water (18), the impedance of the parallel-plate waveguide was estimated to be $Z_{\text{TEM}} = 92 \Omega$ using Equation [3]. Accordingly, the input impedance at each feeding port was estimated to be 184Ω . When terminating each port by a $200\text{-}\Omega$ resistor, the simulated reflection coefficient was less than -10 dB from 100 MHz to 300 MHz. In this condition, nearly uniform current distributions along the longitudinal direction were observed (Fig. 3a and 3b). In the transverse direction, currents mostly flow along the edges.

A broadband Chebyshev matching network was designed to match 50Ω (cable impedance) to 200Ω (Fig. 2a). The measured reflection coefficient was lower than -15 dB from 100 to 300 MHz (Fig. 4a). The bench measurement of the broadband Wilkinson power divider is shown in Fig. 4b. The power dividing ratios were -3.3 dB and -3.4 dB and the phase differences were 1° and 3° at 120.3 and 297.2 MHz, respectively. The measured insertion loss of the diplexer was -39 dB in the stop band and -0.3 dB in the pass band. The overall loss of the RF front-end circuit for the parallel-plate waveguide was -2.0 dB and -2.5 dB at 120.3 and 297.2 MHz, respectively. The loss of the RF front-end circuit for the linear birdcage coil was -0.3 dB at 297.2 MHz.

The measured reflection coefficient from one termination of the parallel-plate waveguide when loaded with the saline phantom was -15.6 dB and -15.9 dB at 120.3 and 297.2 MHz, respectively; when loaded with the forearm, they were -17.4 dB and -16.6 dB, respectively. The measurement was performed by connecting a network analyzer to the input of the broadband Wilkinson divider and a $50\text{-}\Omega$ termination behind the broadband Wilkinson combiner on the opposite side. The reflection coefficient of the birdcage coil was -13.1 dB when measured with the saline phantom at 297.2 MHz.

Figure 5a–5d shows the simulated B_1^+ distribution of the parallel-plate waveguide inside the forearm model at 120.3 and 297.2 MHz. The B_1^+ distribution of the birdcage coil at 297.2

MHz is shown in Figure 5e and 5f, which was nearly identical to that of the proposed coil. Figure 6b and 6d shows the normalized electric field distributions of the proposed coil on a transverse plane at 120.3 and 297.2 MHz, which is similar outside the arm. At 120.3 MHz, the maximum voxel-wise B_1^+ magnitude was 4.3 and 5.8 $\mu\text{T}/\text{W}$ for the parallel-plate waveguide and the birdcage coil. At 297.2 MHz, it was 1.5 and 1.8 $\mu\text{T}/\text{W}$, respectively.

Figure 6 also illustrates the 10-g averaged local SAR at 297.2 MHz. The peak value was 4.8 W/kg/W. In comparison, the peak local SAR of the birdcage coil was 5.6 W/kg/W. In terms of peak local SAR, the maximum voxel-wise B_1^+ magnitude was 0.69 and 0.76 $\mu\text{T}/\text{W/kg/W}$ at 297.2 MHz for the parallel-plate waveguide and birdcage coil, respectively.

Figure 7a and 7b shows the measured B_1^+ map for proton and the acquired TSE image on the transverse plane of the 8-cm outer diameter saline phantom. It should be noted that the double-angle method is not optimal in regions of low signal intensity; thus, the B_1^+ map is qualitative, not quantitative. The simulated maximum voxel-wise B_1^+ magnitude in the saline phantom was 1.4 and 1.5 $\mu\text{T}/\text{W}$ for the parallel-plate waveguide and birdcage coil, respectively. The radiation efficiency was 6% for the birdcage coil and 8% for the parallel-plate waveguide when loaded with the saline phantom at 297.2 MHz. When loaded with the forearm, the radiation efficiency was 44% and 53% at 297.2 MHz for the birdcage coil and parallel-plate waveguide, respectively. At 120.3 MHz, both coils had the same radiation efficiency (21%).

The measured proton transmit efficiency is shown in Figure 8a. Maximum signal intensity was observed with a peak RF voltage of 76 V and 65 V (scanner reading) for the parallel-plate waveguide and birdcage coil, respectively. Considering the loss of the front-end circuit and coil mismatch, the maximum signal appeared at 57 V for both coils (Fig. 8b). The TSE and GRE images of the forearm and ^{31}P CSI results are shown in Figs. 9 and 10, respectively. Coverage of the entire forearm was observed on both GRE and TSE images. The low tip-angle GRE image appears to be more uniform than the TSE image on the transverse plane. The acquired ^{31}P CSI results in Figure 10 demonstrate the feasibility of using the proposed coil for both signal transmit and reception in a broad frequency band.

DISCUSSION

Broadband Transmit

Figure 3 shows that currents on the parallel plates have similar distributions at 120.3 and 297.2 MHz. These currents are responsible for the distributed inductance of the transmission line. Figure 6b and 6d shows that the electric fields outside the forearm are mainly concentrated between the two plates and vary slightly at 120.3 and 297.2 MHz; they are responsible for the distributed capacitance of the transmission line. Because both the distributed inductance and capacitance are similar at different frequencies, the impedance of the TEM mode does not vary much with the frequency. It is a real number that depends on the distance-to-width ratio and the properties of the dielectric insert, not the absolute size of the coil.

The measured reflection coefficient of the proposed coil is the combination of those of the parallel-plate waveguide, the broadband Wilkinson power divider/ combiner, and the Chebyshev matching circuit. It can be further reduced with optimized design of each component. On the other hand, the bandwidth explored in this study was intended to demonstrate the broadband feature of the parallel-plate waveguide. It may rarely be needed in practice except for multinuclear studies at high fields. Optimizing the coil performance around discrete Larmor frequencies is more relevant in practice, and the circuit design can be simplified greatly with a reduced bandwidth. In light of this, several well-established narrowband RF devices (e.g., phase-shifters, high-/low-impedance LNAs, and T/R switches) support the common functionalities of RF coils. Because these narrowband devices belong to the RF front-end, which usually locate on separate circuit boards, the proposed broadband coil can be applied at various frequencies without modifying the coil itself. Currently, the phase adjustment for re-entrant termination was performed separately for proton and phosphorous by using a conventional Pi-network. If a truly broadband system is desired, narrowband phase shifters can be replaced by various broadband designs, such as those described by Schiffman (26) and Zheng et al. (27).

The role of the re-entrant design is not to reduce the loss of the proposed coil, which is determined by radiation, power deposition in the subject, and component losses of the front-end. These losses are evident from the simulated transmit efficiency and bench measurement results. The re-entrant design is to recycle the part of RF power that flows out of the coil, which will otherwise go into a terminating load. Without recycling, a lossless transmission-line coil will have most of its input power being dumped into a terminating load. [See Carter (28) for additional discussion on the role of re-entrant design for travelling wave antennas.]

Coil Construction

The proposed coil is relatively easier to construct than birdcage coils. It is formed by two pieces of copper foils without any reactive components. Once the impedance of the TEM mode is known, it only requires appropriate matching to cable impedance. Tuning the correct operating mode to the desired Larmor frequency by using reactive components, which is mandatory for resonant coils, is no longer needed. Such tuning requires multiple iterations, if possible, at higher frequencies, and the result can shift with different loadings. The proposed coil is more robust with different loading conditions due to its lack of resonant peaks and reactive components. This is evident from the similarity of the measured S_{11} when loaded with a phantom and the forearm.

The parallel-plate waveguide requires the suppression of higher-order modes. This was accomplished by using balanced matching circuit for the TEM mode and impedance mismatch of higher-order modes. According to Equations [1] and [2], higher-order modes are more likely to occur at higher frequencies or with a larger distance between the plates. For 3T proton imaging, one wavelength in air is about 2.3 m. Higher-order modes are possible for an unloaded parallel-plate waveguide only if the distance between the plates is farther than 1.15 m, which is unlikely for existing MR scanners. When the effective permittivity introduced by the human body is considered, higher-order modes should not be a major concern, because the space between the parallel plates is partially filled.

Nevertheless, a more rigorous mode analysis of partially filled parallel-plate waveguide is needed in order to better understand and improve its performance.

B_1^+ Coverage and Profile

From the simulation results at 297.2 MHz (Fig. 5), the B_1^+ profile of the parallel-plate waveguide in the forearm is nearly identical to that of the birdcage coil. Figure 5c and 5e shows the familiar center-bright pattern of both coils. Because higher-order modes were intentionally suppressed, B_1^+ nonuniformity is likely due to the dielectric resonance effect. This is evident from the simulated flip-angle map in a gel phantom in Figure 7d, which shows a high degree of uniformity. According to Zhang et al. (12), traveling wave structures are not expected to prevent standing waves from forming inside a dielectric body. The results reported by Zhang et al. also indicate that wave attenuation could be a problem for traveling wave coils. Figure 5 shows that the longitudinal coverage of the 20-cm parallel-plate waveguide at 297.2 MHz is as good as that of the birdcage coil. This was further confirmed by the sagittal forearm image in Figure 9c and 9d. At 297.2 MHz, neither the simulation result in Figure 5d nor the sagittal image indicates strong B_1^+ attenuation.

The finite coverage in the longitudinal direction of the parallel-plate waveguide should not be attributed to dielectric resonance or wave attenuation. Figures 3a and 3b demonstrate that currents, which generate the magnetic field for imaging, eventually flow into the terminations. Thus the coverage of parallel-plate waveguide cannot extend beyond its length. This is clear from Figure 5b, where the dielectric resonance effect in the forearm at 120.3 MHz is much less pronounced than 297.2 MHz.

Local SAR and Transmit Efficiency

The local SAR of the parallel-plate waveguide is about 14% less than that of the birdcage coil. They are on the same order but higher than that of a regular volume transmit coil (29,30). This should be mainly due to the close proximity between the coils and the forearm, not intrinsic to the proposed coil structure. The shortest distance between the coils and the forearm was less than 1 cm in simulations. For practical volume transmit applications where a separate surface coil array is typically positioned between the transmit coil and the subject, the local SAR is expected to be similar to that of a conventional birdcage coil.

The simulated radiation efficiency of the parallel-plate waveguide is the same as that of the birdcage coil at 120.3 MHz. At 297.2 MHz, the parallel-plate waveguide radiates more than the birdcage coil, but the difference is insignificant. An RF shield would be needed in order to make the proposed coil more efficient for MRI. In this scenario, the two parallel plates and the RF shield form a multiconductor transmission-line system that needs to be analyzed in the future. It was also observed that the radiation efficiency of both coils is reduced by sevenfold when loaded with the 8-cm outer diameter cylindrical phantom. This is because the human forearm is thinner (but oblong) than the cylindrical phantom and tapers from the elbow to the wrist. The coils were less loaded and more RF power is radiated. For the coils loaded with the cylindrical phantom, both their simulated and measured transmit efficiency at 297.2 MHz are about the same.

Signal Reception

The parallel-plate waveguide is a volume coil in nature. Although it can be constructed as a transceiver, using receive-only surface coil arrays should be more appropriate if higher SNR is desired (24). Meanwhile, the proposed signal reception using an open-circuited transmission line resonator is not the only choice. Because the voltage at one open end of the transmission line is to be detected, whether the other end is open or short should be irrelevant. Therefore, one possibility is constructing a quarter-wavelength transmission-line resonator by grounding the rightmost PIN diode in Figure 1c during signal reception (18).

CONCLUSIONS

A parallel-plate waveguide for broadband MRI was introduced. Its design principle and hardware implementation were discussed. Full-wave simulations were performed to study its B_1^+ profiles and local SAR. 7T proton images and ^{31}P CSI results acquired using the same coil demonstrate its feasibility. Compared with resonant coils, the proposed coil is broadband in nature and relatively easier to construct. No reactive components are required for tuning. As a result, its performance is more stable with respect to loading conditions.

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FIG. 1.

a: Electromagnetic field distribution of the TEM mode of a parallel-plate waveguide. **b:** Prototype parallel-plate waveguide. The balanced Chebyshev impedance matching circuits are mounted across the gaps. **c:** Transceiver schematic. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com.]

FIG. 2.

a: Schematic of the Chebyshev matching network. **b:** Schematic of the broadband Wilkinson power divider.

FIG. 3. Simulated current distributions at 120.3 MHz (**a**) and 297.2 MHz (**b**) using FEKO. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com.]

FIG. 4.

a: Measured reflection coefficient from the Chebyshev matching network. **b:** Measured power dividing ratios of the broadband Wilkinson power divider.

FIG. 5.

The B_1^+ field of the parallel-plate waveguide (PPW) at 120.3 MHz (**a, b**) and 297.2 MHz (**c, d**). The maximum voxel-wise magnitude is 4.3 and 1.5 $\mu T/\sqrt{W}$, respectively. **e, f:** The B_1^+ field of the birdcage (BK) coil at 297.2 MHz. The maximum voxel-wise magnitude is 1.8 $\mu T/\sqrt{W}$ at 297.2 MHz.

FIG. 6.

a: Forearm and parallel-plate waveguide model in XFDTD. The 10-g averaged local SAR distribution of the parallel-plate waveguide (c) and the birdcage coil (e) at 297.2 MHz is shown in dB levels. The peak values are 4.8 and 5.6 W/kg/W, respectively. **b, d:** Normalized electric field distribution of the parallel-plate waveguide at 120.3 and 297.2 MHz. **f:** Normalized electric field distribution of the birdcage coil at 297.2 MHz.

FIG. 7.
a–c: Measured B_1^+ map (a), TSE image (b), and simulated flip-angle map (degrees) (c) inside a water phantom. **d:** Gel phantom generated by the parallel-plate waveguide. The same linear gray scale was applied.

a **b****FIG. 8.**

Directly measured (a) and calibrated (b) transmit efficiency of the parallel-plate waveguide and birdcage coil loaded with the saline water phantom. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com.]

FIG. 9.
TSE (**a, c**) and GRE (**b, d**) images of the forearm at 7T.

FIG. 10.
a: Phosphorous CSI result overlapped on localizer image. **b:** Enlarged CSI result inside a single voxel. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com.]