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An RF-gated wireless power transfer system for wireless MRI receive arrays

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Abstract

In MRI systems, cable-free receive arrays would simplify setup while reducing the bulk and weight of coil arrays and improve patient comfort and throughput. Since battery power would limit scan time, wireless power transfer (WPT) is a viable option to continuously supply several watts of power to on-coil electronics. To minimize added noise and decouple the wireless power system from MRI coils, restrictions are placed on the coil geometry of the wireless power system, which are shown to limit its efficiency. Continuous power harvesting can also cause a large increase in the background noise of the image due to diode rectifier up-conversion of noise around the frequency of the transmitted power. However, by RF gating the transmitted power off during the MRI receive time while continuing to supply power from a storage capacitor, WPT is demonstrated to have minimal impact on image quality at received power levels up to 11 W. The integration of WPT with a 1.5T scanner is demonstrated.

Keywords

Magnetic resonance imaging (MRI); Wireless Power Transfer; Inductive Coupling; Wireless MRI

I. INTRODUCTION

WIRELESS technology has impacted just about every aspect of modern life. As the number of portable electronic devices expand for phones, robotics, internet of things (IoT), and

wearable devices, the elimination of charging adapter cables, and battery-free operation has become a major technology driver. This is equally true for health care monitoring devices and the needs for mobility in the hospital environment. Wireless power transfer (WPT) has made major inroads for wireless charging of battery operated devices, and efforts are underway by companies like WiTricity for powering devices in the home [1].

Wireless power transfer has a long history, starting from the early 20th century with attempts by Nikola Tesla to implement large scale wireless power distribution using the earth's ionosphere [2]. Today, wireless powering is still accomplished by electromagnetic fields and can be categorized based on three frequency ranges used: 1) ferrous inductive coupling in the 100's kHz range for industrial power levels, 2) loose non-ferrous inductive coupling in the 6.7 and 13.2 MHz ISM bands for Watt-level transfer and 3) UHF and microwave frequency energy harvesting at the mW level. In recent years, the Wireless Power Consortium has been established to set standards for inductive coupling in the 100's kHz range and the Alliance for Wireless Power (A4WP) sets standards for inductive coupling in the ISM bands.

In MRI, there are many motives for cable-free receive arrays. There is a continuous drive toward a greater number of receiver channels in an effort to improve the signal-to-noise ratio (SNR) or imaging speed [3]-[8]. Increasing the channel count requires a commensurate increase in the number of RF cables and connectors for the array. The added bulk of these cables can cause increased patient discomfort, as well as requiring longer setup times which reduce patient throughput. Bulky cable traps are necessary to prevent the induction of RF current on the cables during B1 transmit, and for avoidance of RF burn hazards.

For wireless MRI arrays to be realized, a system must be developed for wireless data, wireless gating, wireless phase synchronization and wireless power transfer. Several groups have explored wireless patient coils as a way to avoid the cabling issues, with many focusing on the challenge of how to transmit data from the MRI receive coil [6], [9]–[13]. Preliminary methods of wireless phase synchronization and gating have been demonstrated [14]–[17]. An additional challenge with wireless coils is powering them inside the MRI [18], [19]. Non-magnetic batteries are one option, but it is difficult to ensure that they maintain a full charge, limiting the scan time. Using batteries also restricts the voltage and power available [20], [21].

We propose a wireless power transfer (WPT) system based upon resonant inductive coupling and explore the adaptations of WPT in terms of electromagnetic coupling, physical deployment and noise emissions for compatibility with MRI. This work expands upon our initial feasibility tests of [22]–[24]. The system uses RF gating to minimize the impact on the MR image by gating the WPT system off during the MRI receive time. A 10 mF storage capacitor is used to continuously deliver power while the system is gated on and off. Since wireless coils would require at least 100-300 mW per channel, a 32 channel array could require 10 Watts. Here we demonstrate wireless power transfer of up to 11 W suitable for powering such a system.

II. BACKGROUND

A. Wireless Power Transfer Theory

WPT over short distances, typically less than 1 or 2 meters, is based on the principle of resonant inductive coupling [25]– [29]. Resonant inductive coupling models the transmit and receive coils used for WPT as a loosely coupled transformer, shown by the equivalent circuit model of Figure 1 and the linear two-port network equation:

$$V = \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} R_1 + j\omega L_1 & j\omega M \\ j\omega M & R_2 + j\omega L_2 \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = ZI \quad (1)$$

Capacitors in series with each coil resonate out the leakage inductance at the frequency of the transmitted power, allowing for efficient power transfer even with lower coupling. Power is transmitted through a primary coil which generates a magnetic field that can couple to a nearby secondary coil and induce a current on that coil, thereby transferring power.

The coupling between coils can be used to model the secondary coil as an equivalent load impedance induced on the primary coil. The matrix equation of Equation 1 can be combined with the terminal voltage relation at port 2 and simplified to find the equivalent load seen by the primary coil:

$$V_2 = -\left(\frac{1}{j\omega C_2} + R_L\right)I_2 \quad (2)$$

$$Z_{eq} = \frac{\omega^2 M^2}{R_2 + R_L} \quad (3)$$

Equation 3 gives the induced load impedance when the secondary coil is at resonance. This impedance depends on the resonant frequency choice (ω), the mutual inductance between coils (M), and the parasitic resistance of the secondary coil (R_2) combined with the load resistance (R_L).

When the primary coil is also at resonance, the equivalent circuit model becomes a resistive divider between the parasitic resistance of the primary coil (R_1) and the induced impedance (Z_{eq}). Therefore, if both coils are perfectly resonant at the same frequency, we can derive the efficiency from the primary to secondary coil, η_{ps} , and from the secondary coil to the load, η_{sl} .

$$\eta_{ps} = \frac{Z_{eq}}{R_1 + Z_{eq}} \quad (4)$$

$$\eta_{sl} = \frac{R_L}{R_2 + R_L} \quad (5)$$

The overall efficiency, η_{sys} , of a WPT system depends on the inductive link efficiency, the power amplifier DC-RF conversion efficiency, and rectifier RF-DC efficiency giving:

$$\eta_{sys} = (\eta_{DC-RF})(\eta_{link})(\eta_{RF-DC}) \quad (6)$$

Equations 4 and 5 can be combined to provide the efficiency of the WPT inductive link when the coils are perfectly resonant, yielding:

$$\eta_{link} = (\eta_{ps})(\eta_{sl}) \quad (7)$$

$$= \frac{\omega^2 M^2 R_L}{\omega^2 M^2 (R_2 + R_L) + R_1 (R_2 + R_L)^2} \quad (8)$$

The auxiliary parameter

$$X = \frac{\omega^2 M^2}{R_1 R_2} \quad (9)$$

can be substituted into Equation 8 to simplify the η_{link} equation. Taking the derivative of η_{link} with respect to R_2 and setting it equal to zero allows us to solve for the R_2 value that maximizes efficiency. The optimal value of R_2 and maximum theoretical efficiency η_{max} of the coupled coils are then respectively:

$$R_{2, \eta_{max}} = \frac{R_L}{\sqrt{1+X}} \quad (10)$$

$$\eta_{max} = \frac{X}{(1 + \sqrt{1+X})^2} \quad (11)$$

This maximum efficiency can be achieved by matching the load resistance to the optimal load resistance found by rearranging Equation 10 and substituting X from Equation 9. From Equations 4 and 5, it follows that a higher load resistance will result in less power dissipated in the secondary coil but higher power dissipated in the primary coil, since the induced impedance on the primary coil is inversely proportional to the load resistance. As a result,

the optimal load resistance is dependent on both the parasitic resistances of the coils and the mutual inductance between the two coils, yielding:

$$R_{Lopt} = \sqrt{R_2^2 + \omega^2 M^2 \frac{R_2}{R_1}} \quad (12)$$

The parasitic resistances of the coils are limited by the conductor dimensions, and the load resistance is determined by the voltage rectification and regulation requirements of what is being powered. For high power levels the frequency of operation is limited to the low MHz range to meet RF exposure requirements [30]. As a result, the main variable in maximizing the efficiency is the mutual inductance. The mutual inductance between two coils can be calculated using Neumann's formula to integrate around the coil contours. The mutual inductance depends on the radius of each coil and the separation between the coils. This mutual inductance is maximized when the coils are the same size and the separation between the coils is minimized [31].

B. Primary Coil Geometry

Because the MRI imaging volume is so constrained, options for locating the primary coil are limited to solenoids or saddle coils concentric with the body coil, a large planar primary embedded in the patient table, and even arrays of small planar coils. Much of the early coil analysis for low field MRI (below 0.3T) can be directly applied to WPT because of the similar frequency options if patient SAR is to be minimized. We first review these options, and their suitability for efficient WPT inside the MRI bore.

1) Saddle Coil Built Around Birdcage Coil—To fully integrate the WPT system with the MRI environment, an ideal primary would seem to be a large saddle coil (or circularly polarized pair) that could be built concentric with the existing MRI birdcage coil to transmit power. A test prototype, shown in Fig. 2 was attempted. The saddle coils, 89 cm long and 62 cm diameter, were inserted in a 150 cm long RF shield of 82 cm diameter. A saddle coil creates an RF magnetic field that is approximately uniform and transverse to B_0 , allowing power harvesting with loops whose axes are orthogonal to B_0 much like MRI surface coils. By necessity, the primary coil will also be very close to the MRI shield, which can substantially reduce the transmit efficiency and coupling to the secondary coil.

The mutual coupling can be analyzed by methods first used for low field MRI SNR analysis [32]-[34]. For a saddle coil of radius a producing a field B_{xy} per unit current in free space, the addition of a concentric RF shield of radius c reduces the field by an approximate factor $(1 - a^2/c^2)$ over most of the interior volume. Conceptually, the shield induces an opposing electric image of the saddle coil of slightly larger diameter. For a circular n turn secondary of radius b , the mutual inductance then is simply

$$M = n\pi b^2 B_{xy} \left(1 - \frac{a^2}{c^2}\right) \quad (13)$$

In MRI, where shield diameters are about 10% larger than the body coil, the primary efficiency falls to 17% of its free space value, and the coupled resistance drops to 3%. Second, a whole body WPT primary would now see inductively coupled patient resistive losses which grow as $\omega^2 r^5$ in the case of a spherical load [33]. In our test platform, with practical conductor sizes and shield factor of 0.41, the coupled secondary impedance still failed to overcome the primary losses. It becomes quickly apparent that the concentric WPT primary is not a realistic geometry to transfer high power levels efficiently.

2) Body Coil Harvesting—Given the MRI transmit powers range from 10-30 kW, significant power harvesting is feasible at the Larmor frequency. To harvest power from the MRI transmit body coil, the secondary coil must be tuned to the same frequency as the MRI receiver coils, requiring physical separation of the secondary and receivers to avoid coupling. This also limits the size of the secondary coil, resulting in low coupling with the body coil and therefore low efficiency, though this may be acceptable given the high source power. Since the current being induced on the secondary coil is also at the MRI frequency, the secondary coil generates a magnetic field that would cause a local distortion in the B_1 field, leading to image artifacts. The amount of power that could be harvested from the body coil is also very dependent on the pulse sequence [35]. As a result, a small secondary coil that is compatible with MRI coils would only be able to harvest power levels of several 100 mW from the body coil, and therefore, is likely impractical for powering large receive arrays.

3) Planar Rectangular Coil—A third option for integrating a primary coil in the MRI environment is to have a planar rectangular coil that could be imbedded inside the patient table. To demonstrate this, we constructed a primary coil with approximately a width of 30 cm and length of 45 cm using copper tape wrapped completely over a 6.35 mm radius plastic pipe. The copper thickness of 36 μm was 1.7 skin depths at 10 MHz, giving similar RF resistance to thick copper pipe while being easy to solder. The secondary coil is designed to have a radius of about 9 cm and was made with 18 AWG (1.0 mm diameter) wire. Both coils are tuned to resonate at 10 MHz, the primary using a distributed series capacitance and the secondary with a series-parallel two capacitor matching network. Both coils also contain cable traps at the MRI frequency for coil decoupling. The parasitic resistances of the tuned coils were measured with a network analyzer to give a R_1 of 0.39 Ω and a R_2 of 0.46 Ω .

Figure 3A shows the coil simulation model in Ansys Maxwell to determine the mutual inductance between the coils at various positions. This mutual inductance and the measured coil resistances can be substituted into Equation 11 to find the maximum theoretical link efficiency of the coils at different positions. Figure 3B shows that while the secondary coil is centered over the primary coil, the efficiency drops as the separation between the coils increases. If instead, the secondary coil is held at a constant 3 cm above the primary coil and moved along the primary coil length and width, mutual inductance varies from 96 nH (center) to 146 nH (inside corner). The theoretical link efficiency (Fig. 3C) is about 87% centered and stays relatively constant while the secondary coil is inside the primary coil but decreases rapidly as it is moved outside the primary. This 87% efficiency with secondary is dependent on the secondary driving an optimal load of 6.55 Ω and translates to a primary

referred resistance of 5.19 Ω . Since the mutual inductance changes with coil placement, the optimal load will also change, making it difficult to achieve the maximum theoretical link efficiency. Even so, this coil combination gives us some freedom of placement of the secondary coil and allows the primary coil to be embedded a few centimeters inside the patient table while maintaining an overall high link efficiency.

4) Planar Multi-Turn Coils—A final option for consideration is an array of smaller planar coils. Multi-turn planar coils are commonly used in WPT to increase the mutual inductance between coils, allowing the use of smaller coils and increasing the impedance induced on the primary coil. However, in an MRI environment, it is very important to decouple the WPT coils from the MRI coils. Figure 4 shows the S_{11} of an MRI surface receive coil on a Smith chart to demonstrate the shift in resonance that can be caused by different coils. With the planar rectangular primary coil and single-turn secondary coil proposed above, simple 64 MHz cable traps can be inserted in the primary and secondary loops which minimizes their coupling effects when they are extremely close to the MRI coil. WPT multi-turn loops typically tend to be spaced to lessen peak field spatial variation. However, if such a three-turn coil is placed close to an MRI coil, there is a significant shift in the tuning of the MRI coil due to the coupling between the coils. The distributed windings create distributed parasitic coupling at 64 MHz that simple cable traps cannot eliminate to avoid image degradation. If multi-turn coils are to be incorporated, sufficient physical separation from MRI receivers would need to be assured.

As a result of the various WPT system interactions with MRI coils, the planar rectangular primary coil was determined to be the best option since its geometry allows for cable traps to decouple it from the MRI coils and it can be made to fit conveniently inside the patient table. It also allows for a high coil-to-coil efficiency over a range of positions, allowing for a flexible secondary coil that does not need to be in a fixed position.

C. Multiple Harvesting Coils

Given the difficulty of implementing large, tightly coupled coils for high efficiency in a practical WPT system, an alternative method is to use multiple secondary coils that harvest power simultaneously. The two-port network of Equation 1 can be generalized for a multi-port impedance matrix to examine the effects of additional coupled coils as:

$$\begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_n \end{bmatrix} = \begin{bmatrix} R_1 + j\omega L_1 & j\omega M_{12} & \cdots & j\omega M_{1n} \\ j\omega M_{12} & R_2 + j\omega L_2 & \cdots & j\omega M_{2n} \\ \vdots & \vdots & \cdots & \vdots \\ j\omega M_{1n} & j\omega M_{2n} & \cdots & R_n + j\omega L_n \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ \vdots \\ I_n \end{bmatrix} \quad (14)$$

The circuit model in Figure 5 shows a three-port network, with one primary coil and two loaded secondary coils, where there is now coupling between the secondary coils as well as coupling from each secondary coil to the primary coil. Figure 5 also shows the impedance induced on the primary coil by a single loaded secondary coil, two loaded secondary coils with no separation between them, and two secondary coils separated by a larger distance.

Instead of the expected frequency splitting, Figure 5 only shows an apparent shift in the resonant frequency when the secondary coils are close together. This is a result of measuring the coupled impedance through the planar rectangular coil that is much larger than the secondary coils. Figure 6 shows further work with unloaded coils to accentuate this phenomenon. In reality there is frequency splitting between two secondary coils, but it is only visible in 6C when one of the secondary coils is geometrically decoupled (through overlapping) from the primary coil. In double tuned circuits, currents in the lower frequency mode are in phase, while those of the higher frequency mode are anti-phase. With two secondary coils inside the primary in 6B, there is still frequency splitting but only the higher frequency mode is visible because the negative mutual inductance, combined with its anti-phase current, creates a spatial current mode that does couple to the primary field pattern. The lower frequency mode currents create a spatially anti-symmetric field causing a cancellation in coupling to the primary. In 6D the secondary coils are overlapped beyond their geometrically decoupled separation, resulting in a net positive mutual inductance, leaving only the lower frequency mode visible to the primary and the higher mode occult.

To mitigate this effect, the secondary coils must be placed sufficiently apart, perhaps a distance at least equal to the radius of the coils, to effectively decouple them. This implies the use of smaller secondary coils and it may be difficult to ensure the secondary coils maintain their distance in an actual system. In principle, to compensate for frequency splitting of under-lapped coils, each secondary coil could instead be tuned to resonate individually at a lower frequency such that the higher frequency mode resulting from splitting will match the WPT frequency [36], [37], but at a cost in reduction of maximum achievable efficiency. More practically, the WPT frequency could be tuned to the higher frequency mode, however harmonics are likely to be generated by the power rectifying circuit nonlinearities on each receiver. These harmonics, if very close to the MRI signal band, will be very difficult to filter in the MRI environment.

D. Resonant Frequency Choice

Industry standards for commercial WPT systems have been developed in recent years by the Alliance for Wireless Power (A4WP). A4WP standards apply to near-field WPT and set the ISM band for WPT to 6.78 MHz [38]. The ISM band at 6.78 MHz was chosen primarily for RF emission compliance, which is not a concern in the shielded MRI environment. Frequencies up to 15 MHz were considered in developing this standard, since at higher frequencies the specific absorption rate (SAR) in human tissue increases and the system would risk exceeding RF exposure limits [30].

A major concern for WPT in MRI is the potential for additional noise and RF interference in the image. Anticipating that the system will generate harmonics at multiples of the WPT frequency, we selected 10 MHz instead of 6.78 MHz which ensures the 6th and 7th harmonics will bracket the 1.5T MRI band. While 6.78 MHz could work for 1.5T scanners, the 19th harmonic is at 128.82 MHz, making it difficult to use at 3T. The use of WPT RF gating mitigates any issues with harmonics, but at a cost in duty cycle of power transfer. Fully continuous power transfer, in the future with additional filtering, will require a careful

drive frequency choice to ensure noise in the MRI bandwidth is dominated only by tissue noise.

E. Noise Added by WPT System

To employ wireless power transfer in MRI, it is important to minimize noise and harmonic interference that the WPT system adds to the image. Initial prototyping used a passive 5th order Chebyshev low-pass filter with 15 MHz cutoff to reduce the high power harmonics of the power amplifier before injection into the coupled coils. Figure 7B shows the filter response which has around 70 dB attenuation at the MRI frequency when terminated in an ideal 50 Ω load. In reality, the coupled coils present varying loads to the filter of much lower impedance. With a 10 MHz, 10 W signal injected by our power amplifier, Figure 7C shows the output noise/harmonic spectrum of the filter T-connected with attenuators to an analyzer while also driving the WPT coupling loops. While the higher frequency harmonics are significantly reduced, the noise around the 10 MHz signal is unfiltered. This has ramifications when power-harvesting nonlinearities are introduced.

One can then observe the noise spectrum at the output of an MRI preamplifier connected to a loaded MRI surface coil (Fig. 8). Here, the preamplifier (Advancedreceiver.com) is battery powered. The primary coil is underneath the MRI coil and transmitting increasing power levels but there is no secondary power harvesting coil present. As the power is increased, the power amplifier harmonics at multiples of 10 MHz start to spread out and at around 20 W drive, there is a noticeable increase in the distortion of the harmonics. However, the noise floor at 64 MHz remains constant for all of the power levels, due to the filtering and decoupling in our system.

Figure 9A–D demonstrates the impact of the addition of the secondary harvesting loop, with matching capacitors $C_a = 1.56$ nF and $C_b = 490$ pF, and rectifier (A) on noise and harmonics. The full-wave bridge rectifier uses Schottky diodes for fast switching and low voltage drop. In Fig. 9D, there is little difference in the noise floor at 64 MHz between the baseline noise (black) with WPT off and with the primary coil transmitting power without any secondary (red). However, the noise floor increases (blue) with the secondary coil present and rectifying RF. When the secondary coil provides power to the preamplifier through the preamp's internal 7805 linear regulator, the noise floor rises substantially near the harmonics around 64 MHz, even though we are using a very small secondary coil in this case. The elevated noise wings by harmonics with active wireless power looks very similar to the distortion around 10 MHz in the noise spectrum of Figure 7. It is likely that the secondary full-wave diode rectifier is nonlinearly mixing RFPA noise around 10 MHz and up-converting it to the higher harmonics, essentially acting like a mixer and causing a large back-emission noise in the MRI band when power is being recovered by the secondary coil. The filter resonators, when not seeing a 50 Ω load, may in fact cause resonant noise peaking of the noise floor around 10 MHz that gets up-converted.

The non-linear noise emissions prove difficult to filter in the MRI environment. The simplest solution is the use of RF gating to suspend power transfer during the MRI receive time, which is the most effective way to maintain image quality. A large storage capacitor can be

used at the output of the diode rectifier to provide continuous power to the load while the wireless power amplifier is cycled on and off.

III. FINAL SYSTEM IMPLEMENTATION

We chose a final configuration based upon a planar primary for its suitability for closer coupling and compatibility with patient table location. A diagram of the complete WPT system is shown in Figure 10. An RF switch, connected to the input of the power amplifier, gates the RF power transmission. The power amplifier is a KAA 2040 Class-AB wideband RF power amplifier capable of delivering up to 200 W into a 50 Ω load. An MFJ-929 automatic antenna tuner provides adjustable impedance matching between the power amplifier and the coupled coil impedance. Following the antenna tuner there is the 5th order Chebyshev filter described in the previous section, as well as an additional lumped element 10 MHz $\lambda/4$ filter inside the bore. This 10 MHz $\lambda/4$ block provides an immediate step-up of the coupled coil impedance before the long cable-run to the scan room, while its capacitors act as RF short circuits at 64 MHz.

The primary and secondary coils, as well as key components for our WPT system are shown in Figure 11. The primary coil was designed to be a 45 cm by 30 cm rectangular coil that could potentially be embedded inside the patient table. It was constructed using copper tape over 6.35 mm radius plastic pipe. The secondary coil was designed to be a 9 cm radius loop made of 18 AWG wire in order to be flexible and lightweight, allowing it to be part of the electronics that go on the patient, potentially in a wearable jacket. RF cable traps tuned to 64 MHz were added to the primary and secondary coils to prevent coupling to the MRI coils. To further prevent the large primary coil coupling to the body transmit, its cable connection point must appear as an open circuit at 64 MHz, but a 64 MHz RF short circuit must appear to external circuitry to attenuate residual body transmit coupling. To accomplish this, we use a 64 MHz quarter wave RG-58 cable (76 cm) to invert the 64 MHz short circuit impedance of the 10 MHz quarter wave block output capacitor. The quarter-wave cable also includes additional traps at 64 MHz and 10 MHz to further isolate the primary coil. The secondary coil is connected to a shielded full-wave diode rectifier, using Schottky PD3S140 diodes, with passive band-pass filters on its input and low-pass filter on its output.

Initially, imaging was done with a Medusa RF module [39] that could control an MRI gradient system and provided a completely independent transmit/receive chain to a MRI head coil to mitigate any risk to the clinical RF chain. The Medusa also sent a TTL signal to the RF switch to control the timing of the RF gating. Figure 12 shows the Medusa control setup, which first sent a long pulse to charge up the storage capacitor and then repeatedly gated the power amplifier off during the MRI receive time. Later imaging was done with the actual MRI gradients and scanner body coil and surface coil receive. For these images, the RF switch was controlled by a one-shot circuit (Texas Instruments CD74HC221E) that was triggered by the TX exciter unblank signal. Potentiometers on the one-shot adjusted the pulse width to match the MRI receive time window.

IV. RESULTS

To form a useful link, the coupled secondary load impedance must exceed the parasitic losses of the primary circuit. The 10 MHz quarter-wavelength block in Figure 11 provided an intermediate impedance step-up to the primary coil in addition to filtering. Figure 13 shows a test case of a 22 Ω dummy load on the secondary. The rectifier is absent to simplify network analyzer impedance measurements. With the quarter-wavelength connected inside the bore, the impedance can be stepped-up to around 23 Ω before the connection over long cables to the power amplifier and auto-tuner located outside the magnet room. Without the auxiliary step-up, the secondary load resistance still dominates the low parasitic resistance of the primary coil (0.39 Ω), but the total impedance is now only 2.5 Ω , which is small enough that the RG-58 cable losses to the RFPA become significant and degrade the efficiency of the system.

Efficiency was first measured with the input power measured at the output of the power amplifier and output power measured as the DC output power of the rectifier at a 22 Ω load. Table I shows that with the addition of the quarter-wavelength filter to immediately up-convert the coupled coil impedance, there is a large increase in the efficiency. This efficiency drops inside the MRI bore due to shield coupling with the drive loop, but is still sufficient to perform a high-power demonstration of harvesting several watts of power. This efficiency table encompasses primary, secondary and rectifier losses.

Figure 14 shows the test setup and results of imaging with a Medusa controlling an independent transmit/receive chain to an MRI head coil. The primary and secondary WPT loops were placed underneath the loaded head coil. We also added a red LED at the output of the rectifier to indicate when we were receiving power. A dummy load of 22 Ω was used. A GRE sequence was used for imaging, with a 10 cm FOV, 6 mm slice, a TR of 150 ms, and a TE of 10 ms. To compare the images and test the effect of the WPT system, we measured and annotated the background rms noise of each image, since all scans used the same image scaling and receive gain settings. This allowed direct detection of noise impairments. Image A was done with no power being transmitted or harvested, giving a reference for the rest of the images. In B and C there is power being transmitted with no secondary coil harvesting power. D shows the result of continuously harvesting power while E and F show the harvesting of increasing power levels while RF gating. Clearly, despite filtering attempts, the up-mixed noise and harmonics from continuous WPT overloaded the custom receive chain in D yielding image blackout.

Once confidence was gained that the WPT system did not cause adverse interactions with the main MRI hardware, additional testing was done using the whole body coil of the scanner and a 5 inch (127 mm) receive surface coil. The test setup is shown in Figure 15 where the WPT primary and secondary coils are underneath the MRI receive coil and phantom, and are powering a small 7 W light bulb. Figure 15 also shows the image results and RMS background noise numbers with no power (A), with 1 W of power harvested while RF gating the system (B), and with about 5 W of power harvested (C) while RF gating. SNR performance and image quality fall within normal variations that would exist even without

the WPT system. This test showed complete functionality with both body coil drive and WPT drive coexisting.

V. DISCUSSION

The MRI environment requires several compromises in WPT design that influence the achievable power transfer. The first of these was the use of single loop primary and secondary coils to avoid RF parasitic effects within the MRI frequency band and coupling with MRI coils. If high efficiency class E or D amplifiers are used, as well as low loss Schottky diode rectifiers, the overall efficiency of a WPT system, η_{sys} in Equation 6, will ultimately be limited by the inductive link since it is the stage with the lowest and most variable efficiency. Our efficiency results excluded the unknown class AB RF amplifier contribution, instead using total output RF power, but did include rectifier and all other effects along with the inductive link. Even so, the efficiency results of Table I fell well short of the theoretical 87% inductive link efficiency that accounts for primary and secondary losses.

This discrepancy between the theoretical and measured efficiency is potentially due to the efficiency of the inductive link being very dependent on the load resistance of the secondary coil. It can be difficult to achieve the optimal load and efficiency since the actual load resistance on the secondary coil is typically a voltage regulator, which can cause large changes in the load impedance [29], [40]-[45]. Even without a voltage regulator, the efficiency of the link and the impedance induced on the primary coil are dependent largely on the mutual inductance, which varies with relative coil location. Equation 11 relates the optimal load impedance to mutual inductance and coil losses that maximizes the link efficiency. Multi-turn loops allow transformer-like step-up of impedance, but the single loop design instead required extra lumped element matching. We had used a 22 Ω load and series/shunt capacitor matching network (Fig. 9A). Applying a shunt to series RC conversion with $C_b = 490$ pF yields 15.1 Ω effective series load, which should still have given 83% theoretical link efficiency, but induces only 2.34 Ω in the primary. The impedance step-up by a 10 MHz lumped quarter-wave network was essential to prevent significant losses over coax cable.

Inductive link efficiency requires both coils to be resonant at 10 MHz. Inside the MRI bore, the large rectangular primary coil still experiences some coupling to the shield, causing a frequency shift as well as RF field reduction, both of which can compromise the link efficiency if not anticipated. This off-resonance causes the efficiency of the system to drop when the coils are inside the bore. In our tests, some residual tuning sensitivity can still cause a low drive impedance which, over long RG-58 cable, created losses. Future work will include an MR compatible class-E power amplifier and auto-tuner that can be located by the MRI bore. This would eliminate the long BNC cables that are causing power loss between the coil and the power amplifier. An array of smaller primary coils, that can be switched to align with the secondary coil, could also improve efficiency through the increased coupling between similar sized coils, allowing for a greater range of placement of the secondary coil.

The large primary coil, being embedded in the patient table, is ideal for flexible wireless MRI coils that could be worn like a jacket, allowing the secondary coil to go on the patient's back with only a few centimeters separation to the primary coil. This would obviate the need to embed spine arrays in the patient table. For dedicated anatomical coils, or interventional and intraoperative applications, other topologies may be appropriate. The primary coil could be moved to the top of the bore, and segmented into an array to allow for better coupling with a small secondary coil on top of the patient. Such a primary array could compensate for the increased proximity to the RF shield to maintain efficient coupling. The WPT secondary will still be limited to one secondary per array, or per receive coil cluster – a WPT harvester per MRI coil will likely have inadequate coupling efficiency. Flexible primaries embedded in adjustable padding could provide yet another means of spatial coupling.

Harmonic and noise up-conversion by the diode rectifiers from the 10 MHz band to the MRI band is a key issue. The images of Figure 14 were consistent with those of the noise spectra of Figure 9. At relatively low drive power, up to about a 5 W threshold, with no secondary, there was almost no noise added to the image. When present, the secondary rectifiers back emitted noise that wiped out the image. RF gating the power amplifier signal off during the MRI receive time eliminates this issue, but reduces the duty cycle for WPT. However, in typical MRI sequences, the WPT duty cycle will almost always exceed 50%.

With RF gating, there is almost no change in image quality when the system is harvesting 1 W of power, and a 24% increase in background noise when the system is harvesting 11 W for the case of the Medusa custom receive chain (Fig. 14). Image F shows a minor slice shift from unintended table movement. Similar outcomes were achieved using the MR body coil for transmit and surface coil for receive (Fig. 15), however the observed noise increase of 2% is much lower than with the Medusa controlled head coil. This is likely the result of the Medusa custom transmit/receive chain (preamplifier and TR switch) proximity to both the head coil and WPT primary combined with cable routing from outside the shielded MRI enclosure. Figure 14 shows the complicated cabling and preamplifier supply routing near the head coil, making it susceptible to noise pickup. The normal scanner Tx/Rx chain avoided this pickup issue.

Using the MR body coil for imaging exposed the WPT system to the full transmit power. This demonstrates that our system can handle the large gradient and RF fields inherent in MRI and does not significantly couple to the MRI coils. In future, the RF gating could also play the role of control signal for Q-spoiling, while the WPT drive frequency could, at low RF gated levels, provide a phase synchronous reference, for use in wireless coils.

VI. CONCLUSION

Cable-free receive arrays are very desirable in MRI because increasing channel counts lead to a proportional increase in the number of RF cables and connectors needed for the array, adding to the bulk of the array. Wireless power transfer would be the ideal way to power these arrays, eliminating connections and limits on scan time and time lost charging batteries. In this paper, we have shown the limits as well as the potential of a practical WPT system in the MRI environment, where reducing coupling to the MRI coils and minimizing

added noise is of primary importance. This results in a tradeoff between the efficiency of the system and restrictions on the coil geometry to prevent coupling with the MRI coils. By employing RF gating of wireless power transfer at 10 MHz, we demonstrated up to 11 Watts of power transfer and that WPT can coexist in a 1.5T MRI scanner with minimal impact on image quality.

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Fig. 1.
Equivalent circuit model for two series-resonant coupled coils in a wireless power transfer system.

Fig. 2.

Concept of overlaying a quadrature WPT saddle coil primary over an 8-rung 64 MHz MRI birdcage resonator. A: Quadrature saddle coil elements shown by arrows. B: The saddle/birdcage assembly is inserted in a concentric RF shield. The saddle elements are 1/8 in copper tube segmented by series capacitors. C: Diagram showing an individual saddle coil around an MRI birdcage resonator.

A **B** **C****Fig. 3.**

Planar rectangular coil with a coupled secondary coil A: Simulated in Maxwell to find the mutual inductance, B: Efficiency variation as the secondary coil is centered over the primary coil and moved farther apart, and C: Efficiency variation as the secondary coil is kept a constant 3 cm above the primary coil and moved along the length and the width of the primary coil.

Fig. 4.
 S_{11} of an MRI surface receive coil on a network analyzer smith chart showing the impedance of the unloaded coil at 64 MHz when it is A: Next to a planar rectangular coil with cable traps, B: Overlapping a single-turn coil with cable traps, and C: Overlapping a three-turn planar coil.

Fig. 5. Multiple secondary coils harvesting power A: Circuit model showing coupling between all coils, B: When the two secondary coils close together, C: When the two secondary coils are far apart, and D: The measured impedance induced on the primary by the two secondary coils compared to a single secondary coil.

Fig. 6.

Real impedance induced on the planar rectangular primary coil by unloaded secondary coils A: With two geometrically decoupled secondary coils, B: Two secondary coils inside the primary, C: One secondary decoupled from the primary and the other secondary inside the primary, D: Two secondary coils largely overlapping each other, E: A single secondary, F: A single secondary decoupled from the primary, and G: The resultant frequency splitting for each case. Coupled coils create in-phase (lower frequency) and anti-phase (higher frequency) current modes. The net sum of these modes can create an integrated coupling to the primary, or result in cancellation of coupling.

Fig. 7. 5th order Chebyshev low-pass filter A: Schematic, B: Frequency response when connected to a 50 Ω load, and C: Noise spectrum at the output of the filter when the filter is connected to the coupled coils.

Fig. 8.

A: The noise spectrum measurement physical setup uses only a WPT primary coil with no secondary power harvesting coil present. B: Spectra are acquired from a loaded MRI coil through a preamplifier, boost amplifier and band-pass filter. C: The spectrum analyzer measurements are shown for increasing transmit power levels. In general, while 10 MHz harmonics increase, background noise is mostly invariant to drive level.

Fig. 9.

Spectral interference with power harvesting. A: The wireless power transfer secondary coil with a full-wave diode rectifier and filter. B: The spectrum is measured from a loaded MRI coil through a preamplifier, boost amplifier and band-pass filter. C: The noise was measured for three cases - 1) No WPT secondary coil, 2) the WPT secondary coil receiving power and the preamplifier battery powered, and 3) preamplifier wirelessly powered. D: The resulting noise spectrum for each case compared to the noise floor with the WPT system off.

Fig. 10.

Complete wireless power transfer system with an RF switch gating the signal to the power amplifier. The power amplifier is connected to an antenna tuner so that it sees a $50\ \Omega$ load while it is driving the coupled coils. There are multiple filters after the power amplifier and on the diode rectifier connected to the secondary coil. There is a 10 mF storage capacitor on the output to ensure continuous power delivery.

Fig. 11.

Physical components of wireless power transfer system, including the primary and secondary loops with 64 MHz cable traps distributed on the loops, the 64 MHz $\lambda/4$ cable for additional decoupling from MRI coils, and the 10 MHz $\lambda/4$ filter. Also shown connected to the secondary coil is the shielded diode rectifier and filters with a connector to the power terminals of an MRI preamplifier.

Fig. 12.

Control of RF switch gating of the power amplifier during the MRI receive time was done with a Medusa module when imaging with an MRI head coil and custom receive chain, or by and a one-shot circuit when imaging with the MRI clinical body coil and surface coil receive chain.

Fig. 13. Measured impedance of the primary coil and a coupled resonant secondary coil with and without a $\lambda/4$ filter connected to the primary coil.

Fig. 14.

Test setup and image results of WPT system near a Medusa controlled head coil with A: RF power off, B: Transmitting 1.9 W with no secondary coil, C: Transmitting 5 W with no secondary coil, D: Continuously harvesting 1 W, E: Harvesting 1 W with RF gating, and F: Harvesting 11 W with RF gating.

Fig. 15.

Test setup and image results with a body coil with RMS background noise numbers given for A: WPT off, B: Harvesting 1 W with RF gating on, C: Harvesting an estimated 5 W with RF gating on.

TABLE I**MEASURED SYSTEM EFFICIENCY**

Efficiency without $\lambda/4$ filter	11.5%
Efficiency with $\lambda/4$ filter	47.3%
Efficiency with $\lambda/4$ filter in MRI bore	33.5%

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