WIRELESS MRI DETECTOR ARRAYS:
TECHNOLOGY & CLINICAL APPLICATIONS

by

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Dedication:

I’d like to dedicate this work to my family:

past, present, and future.
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<th>Full Form</th>
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<td>2D</td>
<td>two dimensional</td>
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<td>3D</td>
<td>three dimensional</td>
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<td>ADC</td>
<td>analog to digital converter</td>
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<td>AM</td>
<td>amplitude modulation</td>
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<td>BW</td>
<td>bandwidth</td>
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<tr>
<td>DC</td>
<td>direct current / zero frequency</td>
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<tr>
<td>DDS</td>
<td>direct digital synthesis</td>
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<tr>
<td>DR&lt;sub&gt;carrier&lt;/sub&gt;</td>
<td>dynamic range of carrier</td>
</tr>
<tr>
<td>DR&lt;sub&gt;sig&lt;/sub&gt;</td>
<td>dynamic range of signal</td>
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<tr>
<td>DR&lt;sub&gt;P1dB&lt;/sub&gt;</td>
<td>linear dynamic range defined by 1 dB compression point</td>
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<td>DSB</td>
<td>double sideband amplitude modulation</td>
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<tr>
<td>FA</td>
<td>flip angle</td>
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<td>FID</td>
<td>free induction decay</td>
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<td>FDM</td>
<td>frequency domain multiplexing</td>
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<td>FLASH</td>
<td>fast low angle shot</td>
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<td>FOV</td>
<td>field of view</td>
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<td>GRE</td>
<td>gradient echo</td>
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<td>IF</td>
<td>intermediate frequency</td>
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<td>MFM</td>
<td>magnetic field monitoring</td>
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<td>MR</td>
<td>magnetic resonance</td>
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<td>MRI</td>
<td>magnetic resonance imaging</td>
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</table>
$\text{NF}_{\text{sys}}$  
--- system noise figure

$\text{NMR}$  
--- nuclear magnetic resonance

$\text{OOK}$  
--- on-off keying

$\text{P1dB}_{\text{input}}$  
--- input-referred 1 dB compression point

$\text{PLL}$  
--- phase locked loop

$\text{RC}$  
--- resistor-capacitor

$\text{RF}$  
--- radio frequency

$\text{RMSE}$  
--- root mean square error

$\text{RX}$  
--- receive

$S_{21}$  
--- $S$-parameter: forward transmission coefficient

$\text{SAW}$  
--- surface acoustic wave

$\text{SNR}$  
--- signal-to-noise ratio

$\text{SSB}$  
--- single sideband amplitude modulation

$T1$  
--- longitudinal magnetization relaxation time

$T2, T2^*$  
--- transverse magnetization relaxation time

$\text{TE}$  
--- echo time

$\text{TH}$  
--- slice thickness

$\text{TI}$  
--- inversion time

$\text{TR}$  
--- repetition time

$\text{T/R}$  
--- transmit/receive

$\text{TX}$  
--- transmit

$\text{VCO}$  
--- voltage-controlled oscillator

$\text{wMFM}$  
--- wireless magnetic field monitoring
Wireless MRI Detector Arrays: Technology & Clinical Applications

Abstract
by
MATTHEW J. RIFFE

Magnetic resonance imaging (MRI) is an important mainstay in modern clinical care. Detecting the MRI signal with multiple detectors has several benefits, including improving the image signal-to-noise ratio (SNR) and decreasing its acquisition time. These benefits further improve as the number of detectors increases. Therefore, MRI manufacturers have been increasing the number of detectors in phased arrays, with as many as 128 elements. Currently, each detector requires its own receiver chain and cabling, so highly parallel phased arrays require a great deal of cabling. Large amounts of cabling can create multiple problems in MRI, so it is becoming difficult to continue increasing the size of phased arrays. Eliminating these cables through wireless technology would allow manufacturers to continue developing highly parallel phased arrays and further improve image SNR and acquisition speed.

This dissertation describes the development of a wireless system for MRI phased arrays. The system was based on Single Sideband Amplitude Modulation (SSB). While operating in the MRI scanner, a single SSB channel preserved the original image SNR by more than 97% for signal dynamic ranges up to 80dB. A simulation method was
developed to study the effects of interference when transmitting multiple SSB-modulated MRI signals simultaneously. This simulation was used to determine the optimal channel spacing in an eight-element detector array equipped with SSB, and SNR was preserved by 97% on average. The role of wireless array technology was also explored in two additional applications outside their typical imaging role. The first application was measuring the first order gradient dynamics of a MRI acquisition with an array of transceiver probes equipped with SSB. When expressed as a normalized k-space, residuals between wireless and wired measurements were within 0.55 Δk, and the majority were below 0.1 Δk. The second application was a handheld device with wireless micro-coils that controls the image plane during a MRI-guided needle intervention. Device operation was demonstrated by guiding a needle to a target and delivering contrast. The research presented in this dissertation paves the way for integrating wireless technology into highly parallel phased arrays along with other promising MRI applications.
CHAPTER #1

Introduction

Magnetic resonance imaging (MRI) is an important mainstay in modern clinical care. MRI was initially developed by expanding the well-established principles of nuclear magnetic resonance (NMR) spectroscopy (1) into a multi-dimensional spatially-encoded imaging modality (2). Shortly after it was proposed, the medical research community realized that a fully-developed MRI system had many impressive benefits over other medical imaging modalities: MR images would have superior contrast between soft tissues, the imaging plane could be acquired in any direction, and the imaging process did not require the use of ionizing radiation. These promising benefits led researchers to rapidly develop MRI and incorporate it into nearly every facet of modern clinical care. Presently, the MRI system has evolved into an indispensable and versatile medical imaging platform that provides physicians with not only superb anatomical images, but also detailed tissue perfusion and diffusion maps, comprehensive angiographic anatomy, real-time interventional guidance, and dynamic maps of neuronal network activity.

This dissertation focuses on the development of a wireless system that transmits the signals from multiple MRI detectors simultaneously. As will be described later in this Chapter, the incorporation of a wireless system will help spur the development of a MRI system that will possibly acquire images faster and better than the current MRI generation. This dissertation contains promising results that will serve as an excellent
starting point for a team of engineers developing the next generation MRI system. In the following chapters, a prototype wireless system is presented that is shown to transmit and preserve the quality of multiple MRI signals. The operation of this system is demonstrated in three different clinical applications that benefit from wireless technology: transmitting high quality diagnostic images, measuring magnetic field dynamics to reconstruct images, and controlling the orientation of images acquired during an interventional procedure. Before presenting these results, this introductory chapter provides the foundation and clinical motivation for developing a multi-channel wireless MRI detector system.

1.1 Basic Overview of the MRI Signal

In the presence of a static magnetic field (B₀), the nuclear magnetic moments of atoms align or anti-align themselves in the direction of B₀. A small difference will exist between the number of moments that are aligned and anti-aligned, which is typically on the order of parts per million (ppm) for living tissues for strong applied fields on the order of Teslas (T). This alignment difference results in a net magnetization vector for that group of atoms, and due to angular momentum, the magnetization precesses around B₀ at the Larmor frequency (ω₀) when excited.

\[ \omega_0 = \gamma B_0 \]  

Equation 1.1 reveals that ω₀ is directly proportional to the strength of B₀ with a proportionality constant γ, which is the nuclei-specific gyromagnetic ratio. While multiple species of nuclei have a nuclear magnetic moments, MRI primarily observes the
nuclei of $^1$H atoms ($\gamma = 42.58$ MHz/T), which have an extremely high physiological abundance.

When aligned with $B_0$, the magnetization is unable to be observed since it is superimposed on a magnetic field that is multiple orders of magnitude larger. To make this magnetization observable, the magnetization’s axis of precession is rotated to the plane transverse to $B_0$. This is typically done by applying a dynamic magnetic field that is transverse to $B_0$ and oscillating at the same frequency ($\omega_0$) as the magnetization. Since $\omega_0$ is typically in the radio frequency (RF) range, this is commonly referred to as “RF excitation”. The rotated magnetization, now transverse to $B_0$, is observed by placing an electrically conducting loop near the magnetization and measuring the induced electromotive force. The loop is typically a LC circuit, or a “coil”, that is tuned and matched to a 50 ohm connection at the resonant frequency $\omega_0$. The matched condition maximizes power transfer of the induced coil signal to a receiver system. The induced transverse magnetization signal is dependent on many different tissue-dependent parameters, including relaxation (T1, T2) and diffusion. The contributions that these parameters have on the signal can be controlled, and it is this control that gives the MR signal its vast breadth of soft tissue contrast mechanisms.

Assuming the main magnetic field $B_0$ is homogeneous, the MR signal is the superposition of many different species of magnetizations all rotating at the same frequency $\omega_0$. In this state, it is not possible to extract spatial information from the magnetization. By applying three orthogonal magnetic field gradients that superimpose with $B_0$, the individual magnetizations become spatially encoded with different rotational frequencies that have a known spatial-frequency relationship. This relationship is used to
separate the individual magnetizations from the bulk measured signal and assign them to a specific spatial location, forming an image. A further description on reconstructing an image from the gradient-encoded MR signal is found in Section 4.2.1.

While section presented only a brief overview of the MRI signal formation, an in-depth description of signal encoding, acquisition, and image reconstruction can be found in the textbook by Haacke et al (3).

1.2 Evolution of the MRI Detector Array

As mentioned in the previous section, a resonant coil is the principal tool used to detect and measure the MR signal (4). A large variety of coil designs have been shown to be effective in MR imaging experiments, and these designs are typically optimized for a particular imaging application. The majority of coils used in MRI are categorized into two categories: volumetric coils (most commonly realized as "birdcage" coils (5)) and planar coils (commonly referred to as “surface” coils (6)). A birdcage coil is an advantageous design that creates a device with a spatial sensitive profile that is both large in volume and mostly homogenous. On the other hand, a surface coil has a sensitivity volume that is concentrated locally near the coil and highly inhomogeneous. Surface coils are beneficial because they can produce images with high signal-to-noise ratio (SNR) in the vicinity of the coil. When positioned near the region being imaged, a surface coil receives a larger signal amplitude due its small diameter and its close proximity to the sample. Surface coils also have a reduced sensitivity to noise that is outside the region being imaged. These two components are the primary reasons why SNR is better with a surface coil that has limited spatial coverage. This demonstrates an important coil
tradeoff between image quality (SNR) and spatial coverage, and it must be considered when deciding on a particular coil design.

Starting in the late 1980s into the early 1990s, a novel acquisition system was presented, which allowed the signals from multiple coils to be collected simultaneously (7,8). This multiple-coil system is commonly referred to as a “phased array”, and this was an important development because it allowed coil engineers to circumvent the fundamental trade-off between image spatial coverage and SNR. For example, a large birdcage coil could be replaced with an equivalently sized array of multiple surface coils. The images obtained from each of these smaller coils can be combined to produce a single image, and this combined image would have the same spatial coverage as the birdcage coil but with the improved SNR performance of the smaller surface coils (9). This benefit can be directly seen in the images presented in Figure 1.1, where signal-to-noise ratio (SNR) between a volume coil and phased array are compared.

![Figure 1.1 - Comparison of Coil SNR Performance](image)

*Figure 1.1 - Comparison of Coil SNR Performance: Images acquired with two similarly-sized head coils on a Siemens 1.5 T Espree system demonstrate the SNR improvement obtained by using a phased array. In order to compare the images, each has been scaled in signal-to-noise ratio (SNR) units. The acquisition was a T₁-weighted spin echo sequence (TR = 500 ms, TE = 9.5 ms, Slice Thickness = 5 mm, 250 x 250 mm² field of view (FOV), 256 x 256 acquisition matrix).*
The construction of a phased array does require some additional hardware not necessarily present with only a single coil. In order for multiple coils to remain tuned and matched at the Larmor frequency, the individual coils must be electrically isolated from one another. This is typically accomplished using two separate methods, as illustrated in Figure 1.2. The neighboring coils are overlapped geometrically so that induced current on one coil will induce as little current as possible on the other (effectively cancelling their mutual inductance) (10). In addition, the magnitude of the current induced on each separate coil is reduced by using preamplifiers with low input impedances and a specialized matching network (7,11). This substantially reduces any remaining mutual inductance interactions between non-neighbor coil elements in the array. The disadvantage to constructing a phased array is the simple cost in increasing the hardware required for multiple coils. Each coil in a phased array requires its own preamplifier, cabling, and narrow-band MR receiver. Despite this cost drawback, the benefits of improved image quality and increased spatial coverage quickly made phased arrays a mainstay in modern MRI clinical care.
The introduction of parallel imaging (12–14) spurred another stage of phased array development. Parallel imaging is a reconstruction technique that dramatically reduces the time required to acquire an image. One of the major drawbacks to MRI continues to be the long time needed to acquire individual images, and thus the speedup provided by parallel imaging was very important to improving MRI's clinical utility. Typically, the time required to acquire an image in MRI is determined by the number of steps that must occur during the gradient encoding process to meet the Nyquist sampling criterion. However, parallel imaging can exploit the spatial variations between multiple coils' sensitivity profiles to reduce the aliasing that occurs when the Nyquist criterion is violated. This allows for the number of gradient encoding steps to be reduced in an acquisition, which directly reduces the image acquisition time. In order to use multiple

Figure 1.2 - Phased Array Schematic: A simple diagram of a phased array with four coils reveal the two common methods to electrically isolate coils from each other. Proper geometric overlap between neighboring coils cancels the mutual inductance between coils. Low input impedance preamplifiers reduce the interactions between the remaining coils. By using a matching network, the amplifier's low input impedance is transformed into a high impedance when observed by the coil. This reduces the induced currents on the individual coils, which therefore weakens any remaining interactions between the non-neighbor inductively-coupled coils.
coils simultaneously, parallel imaging requires a phased array to be used when acquiring the MR signal, and the extent of time reduction is limited by the number of independent coil elements in the direction of the gradient encoding. For example, decreasing the acquisition time by a factor of two requires at least two coils with unique sensitivity profiles, decreasing by factor three requires three coils, etc. The drawback to parallel imaging is that reducing the acquisition time also reduces the SNR of the reconstructed image. This places additional emphasis on using an array with a high coil count to obtain sufficiently high SNR images. This was recently demonstrated by a study by Keil et al. (15), where a comparison was made between two similarly-designed phased arrays with 32 and 64 elements. In normal acquisitions, the difference in SNR performance between the two phased arrays was not substantial, but once the acquisition was accelerated with parallel imaging, large gains in SNR were observed with the array with more coil elements.

With the promise of both decreased acquisition time and improved image SNR, there is a strong clinical demand for phased arrays with more and more detector elements. At present, highly parallel phased arrays (larger than 16 coils) are commonplace in the modern MR suite, and arrays are now being manufactured with as many as 128 individual coils elements (16,17). While the clinical benefits these large arrays provide are clear, the large amount of cabling and hardware needed for each coil in these arrays is beginning to become too burdensome for coil manufacturers. The coaxial cabling to transmit the signals from each coil is expensive and highly parallel arrays require large amounts of cable to transmit the signals from all the coils. These dense bundles of cables also have the potential for signal coupling, which can have undesirable effects on the
image reconstruction. In addition, large amounts of cabling pose a safety risk in MRI, where a single cable failure could result in serious RF burning to the patient during RF transmission (18–20). At the current state of technology, it is unlikely that manufacturers will continue increasing the number of coils in phased arrays and achieve higher rates of image acceleration much beyond the present state of the art given the current technological barriers.

1.3 Non-Coaxial Transmission of Multiple MR Signals

As described in the previous section, the large cabling requirement in phased arrays restricts the development and production of highly parallel arrays. Therefore, multiple research efforts have been made to develop technology that reduces the number of coaxial cables required for phased arrays (21–29). The primary constraint with reducing the number of coaxial cables is preserving the original dynamic range of the MRI signal. Dynamic range is the ratio of the maximum signal power to the minimum signal power, typically expressed in decibels (dB).

\[
DR_{\text{sig}} = 10 \times \log_{10} \left( \frac{\text{Power}_{\text{max}}}{\text{Power}_{\text{min}}} \right)
\]  

[1.2]

The dynamic range of a MRI signal (DR_{\text{sig}}) is dependent on many parameters including the scan’s acquisition parameters, coil size, body part being imaged, scanner magnetic field, etc. Any system processing the MRI signal should have a dynamic range greater than or equal to DR_{\text{sig}}. A recent investigation found that the majority of commonly performed clinical MRI acquisitions had DR_{\text{sig}} between 60-85dB (30). Therefore, any technology developed for cable reduction would need to satisfy this dynamic range requirement.
Time-division multiplexing signals from multiple coils into a single receiver port has been shown to reduce the number of coaxial cables required for an array (21), but the maximum sampling rates found in the analog to digital converter (ADC) within the narrow-band MRI receivers limit the extent of cable reduction obtained without compromising the received dynamic range. Direct optical modulation of the MRI signal replaces coaxial cables with optical cabling (22,31). While expensive, optical cabling does address concerns over cable coupling and safety. But, an analog optical link has been shown to have insufficiently high dynamic range (30). Hardware-based compression can reduce the transmission dynamic range of the MRI signal while preserving the reconstructed image quality (24). And by combining all these technologies, high dynamic range multi-channel optical transmission systems have been created that are suitable for phased arrays (25). However, these systems are highly complex and have yet to see widespread implementation.

For higher degrees of cable reduction, frequency division multiplexing (FDM) has been proposed to transmit all the signals in a large coil array using only a single cable (26) or wirelessly (27,29). A digital FDM implementation has been presented based on wireless protocol 802.11b (27). However, placing an individual digital-to-analog converter on each coil with sufficient dynamic range has the potential to become too expensive in cost and in power consumption. Digital technology also adds additional design concerns over introducing broadband noise into the MR environment from the required digital switching signals. For these reasons, an analog FDM technique is potentially a favorable alternative. Analog FDM techniques traditionally have high dynamic range and low power consumption when compared to its digital counterparts.
The drawback to analog FDM techniques is that they require precise synchronization between the transmitter and receiver systems to avoid cascading phase errors in the recovered signals (29).

When considering the bandwidth requirements of the MR signal, it would also seem that analog FDM methods would be superior to digital transmission. The gradient-encoded MRI signal can have a bandwidth up to 500 kHz, and in order to transmit all the signals from a highly parallel phased array, a FDM transmission system would need to transmit the signals in spectrally-efficient manner. Digital transmission with the 802.11b protocol requires 22 MHz spacing between neighboring channels to avoid overlap between the modulated signals (32). More advanced techniques, like 802.11n using Multiple Input Multiple Output and Orthogonal FDM (28), have been proposed to encode and transmit multiple signals in each channel. While this increases the number of channels that can be transmitted digitally, it also substantially increases the complexity and cost of the digital transmission system. In analog FDM techniques, such as single sideband amplitude modulation (SSB) (33), the ideal spacing between multiple encoded signals is only limited by the bandwidth of the MRI signal, allowing for a very dense transmission spectrum. When considering this and the additional benefits of low cost and small power consumption, analog FDM becomes the compelling solution to eliminate the coax cables in highly parallel phased arrays.
1.4 Additional Clinical Applications for Wireless MRI Detector Arrays

Outside their typical role in acquiring images, wireless MRI detector arrays could have additional clinical applications where the nature of wireless signal transmission is greatly beneficial. This dissertation investigates the feasibility of using wireless technology in two of these applications.

1.4.1 Measurement of Gradient Field Dynamics

As described earlier in this chapter, an image is reconstructed from the acquired MRI signal by utilizing the spatial-frequency encoding provided by the gradient fields. In order to reconstruct an accurate image, the gradient fields must not deviate from their expected operation. Modern MRI systems are rigorously designed so that only the smallest of perturbations occur during gradient operation. These perturbations are small enough that they rarely cause any measurable negative impact to the reconstructed images. But now, ultrafast MRI acquisitions are being developed that are prone to these effects. The temporal resolutions achieved by these acquisitions have promising clinical applications in visualizing cardiac motion (34) and angiography (35). In order to have correct image reconstructions, an accurate measurement of the gradient field dynamics must be performed concurrently with normal signal detection. A technique called magnetic field monitoring (MFM) has shown that an array of small solenoid coils can suitably measure the dynamics of the gradient fields (36–39), but it requires an additional specialized transceiver system to collect the signals from the MFM array. This additional hardware has prevented widespread integration of this technology clinically, and therefore, has also prohibited the use of clinically-relevant ultrafast acquisitions. In this
case, the incorporation of wireless technology is important because it allows easy integration of the MFM system into the standard MRI system. Wireless transmission allows the MRI receiver system to selectively listen to the signals that are needed at the time of imaging. If there are a limited number of available receiver channels (as is typical), the wireless MFM signals can be received only when necessary. Therefore, the presence of these wireless wMFM signals would not disrupt normal imaging procedures. Additionally, wireless technology could also facilitate the use of MFM probes based on non-hydrogen nuclei. By utilizing the frequency shifting properties of wireless FDM, the recovered wireless non-hydrogen signals could be collected by conventional narrowband MRI receivers. This has multiple benefits as described later in Section 4.1.2.

1.4.2 Image-Guided Needle Interventions

Image-guided needle interventions are procedures where a radiologist uses in vivo images to guide a needle to a target tissue within a patient. They are useful for biopsies and other clinical applications that requires safe and accurate needle delivery. While MRI potentially offers superior images for needle guidance, other medical imaging modalities (e.g. fluoroscopy and ultrasound) are often preferred by interventional radiologists. These preferred modalities give radiologists a means to quickly and intuitively control the orientation of the image being acquired. In MRI, it is difficult for a radiologist to simultaneously wield a needle inside the narrow magnet bore and simultaneously control the acquired image plane on a computer interface outside the bore. Therefore, there is a clinical need to develop an intuitive MRI tracking system that can operate inside the magnet bore.
One possible solution for such a MRI tracking system described here is a handheld device equipped with a wireless detector array. An array of wireless coils can be automatically localized by the MRI system (40). The position and orientation of the device can then be used to define the acquired image plane. This would provide the radiologist a control system similar to an ultrasound probe, where the orientation of the device would correspond to the image plane being acquired by the MRI. Acquiring the position of device takes less than 50 ms, so it can be interleaved with high temporal resolution imaging techniques, providing the radiologist near real-time image guidance. The wireless nature of the device allows it to be freely wielded inside the crowded MRI magnet bore. The wireless device would be small and entirely self-contained, so that it could be sterilized and incorporated into needle delivery apparatuses.

1.5 Overview of Dissertation

This dissertation on wireless MRI detector arrays has been organized into six chapters. This introduction chapter has described the clinical and engineering needs for a multi-channel wireless MRI detector system briefly. In the following chapter, the design and development of a wireless transmission system suitable for the MRI signal is presented. Then, three different clinical applications are presented that utilize the developed wireless technology. Finally, conclusions from these investigations are presented along with insights to future research directions. The following subsections summarize the remaining chapters of this dissertation.
1.5.1 Chapter 2: A Single Sideband Amplitude Modulation (SSB) System Suitable for MRI

Chapter 2 presents a single sideband amplitude modulation (SSB) transmission system that was designed to preserve the MRI signal and operate locally in the scanner bore (41). The SSB system was designed with minimum complexity, and a single upconversion stage was used to encode the MRI signal at the GSM-900 band for wireless transmission. A series of bench measurements and phantom imaging experiments were performed to evaluate the system's ability to preserve SNR when transmitting the signal from a single coil. The SSB system was found to sufficiently preserve SNR for signal dynamic ranges up to around 80 dB. Qualitative analysis of phantom images collected wirelessly reveals no structural artifacts introduced by the SSB modulation or system operation.

1.5.2 Chapter 3: Identification and Mitigation of Interference Sources Present in SSB-based Wireless MRI Receiver Arrays

Chapter 3 presents a study that investigated the feasibility of utilizing SSB in highly parallel wireless phased arrays (42,43). The study primarily focused on identifying and mitigating the interference experienced between multiple SSB-encoded MRI signals. A simulation method was developed to analyze the interference performance for any SSB system and to determine the optimal carrier frequency combination when transmitting multiple signals. Two different SSB systems were examined with the simulation method. The first SSB system was the minimally complex system presented in Chapter 2, and the second SSB system was developed to have improved filtering of interference sources.
Both SSB systems were used to wirelessly transmit the signals from an eight element phased array. When transmitting only a single coil, both SSB systems were shown to minimally impact the original image SNR. On the other hand, when transmitting all eight signals simultaneously, an average SNR loss was observed to be 12% in the first system, and the second system with more complex filtering was able to achieve a 3% loss in SNR. This work demonstrated that successful wireless transmission of multiple SSB-encoded MRI signals is possible as long as channel interference is properly managed through design and simulation.

1.5.3 Chapter 4: Magnetic Field Monitoring with a Transceiver Array Equipped with SSB

Chapter 4 presents a study that investigated the prospects of performing magnetic field monitoring (MFM) with a wireless transceiver array (44). A four-channel wireless MFM (wMFM) system, based on the SSB technology developed in Chapter 2, was constructed that was capable of measuring up to first order gradient field dynamics. The wMFM system consisted of four SSB-based transceiver modules that could both excite and transmit the signal from a small MFM probe. Additional technology was developed to send two synchronization signals wirelessly: a reference clock and a RF excitation trigger. A simulation was performed to determine the wMFM system performance requirements to achieve different levels of image quality. The wMFM system was used to measure ultra-fast spiral acquisitions at both axial and double oblique image orientations. When compared to other measurements from a wired MFM system and a standard method, the wMFM system was in good agreement. More specifically, when the field
dynamics were represented as normalized k-space trajectories, the residuals between the measurements never exceeded 0.55 in $\Delta k$, and the majority of the residuals were below 0.1 $\Delta k$. Phantom images reconstructed with the wMFM measurement showed no visible sign of artifacts or geometric distortion. This work demonstrated that an array of transceivers utilizing SSB technology is capable of transmitting field monitoring signals and using those signals to accurately reconstruct ultrafast MR images.

1.5.4 Chapter 5: Real-time Device Localization and Dynamic Scan Plane Control with a Wireless Active Tracking Array

Chapter 5 presents a study that investigated the possibility of performing device localization with a wireless detector array (45,46). A prototype device was constructed that contained three fiducial markers each coupled to an independent receiver coil equipped with the wireless SSB technology developed in Chapter 2. Acquiring orthogonal projections of these markers determined the position and orientation of the device, which was used to define the scan plane of an image acquisition. Since the device is used for localization and doesn’t require full imaging capability, the design of the SSB wireless system was simplified by allowing an asynchronous clock between the transmitter and receiver. The error caused by the lack of a shared frequency reference was quantified to be less than one pixel (0.78 mm) in the projection acquisitions. Image-guidance with the prototype was demonstrated with a phantom where a needle was successfully guided to a target and contrast was delivered.
1.5.5 Chapter 6: Summary, Future Directions, and Conclusion

Chapter 6 presents a summary of the research studies found in this dissertation. This summary includes the results from the studies along with a historical context of their innovation. Future directions for wireless detector array technology research are then explored. A second generation of the hand-held interventional prototype from Chapter 5 is described. Instead of controlling the image plane with a hand-held device, the new prototype system proposes to perform image-guidance with the positions of the operator's fingers. This new control scheme is potentially more intuitive and desirable to an interventional radiologist. It is also proposed that the new prototype could be improved by incorporating the transceiver technologies developed in Chapter 4. Finally, the conclusion of this dissertation is presented.
CHAPTER #2

A Single Sideband Amplitude Modulation System Suitable for MRI

2.1 Background

As discussed in Chapter 1, wireless phased array technology would be a benefit to modern MR imaging systems. Despite the advantages of wireless arrays, manufacturers have only begun to implement forms of cable-reducing technology (15,47), and an entirely wireless solution has proven elusive. For any cable-reducing technology, it is imperative that reconstructed image quality is preserved, and it is generally assumed that the engineering cost to accomplish this with any wireless technology is too high. Besides the high dynamic range requirements of the MRI signal, there are additional design challenges that must be considered which are unique to MRI, including the operation of electronics within large dynamic magnetic fields and the prevention of ambient electronic noise that will interfere with the received magnetization.

The goal of this Chapter was to develop a minimally complex wireless system and demonstrate that a single channel from the system was capable of preserving the MR signal dynamic range while operating locally in the MRI environment. The wireless technology used was single sideband amplitude modulation (SSB) (33,48,49). The theory of SSB modulation is presented along with bench measurements that were used to
characterize the system. The system's ability to preserve the MRI signal is presented through both bench measurements and phantom imaging.

## 2.2 Theory

### 2.2.1 Single Sideband Amplitude Modulation

Single sideband amplitude modulation (SSB) was first proposed by Carson in 1915 (50). SSB modulation offers improved efficiency in both power and bandwidth usage when compared to other amplitude modulation (AM) techniques. A representation of a coil’s MRI signal ('\(S_n(t)\)') can be written as:

\[
S_n(t) = \int a_n(r, t) \exp(-j2\pi (f_0 + \Delta f(r, t)) t) dr
\]

where ‘\(n\)’ is representative of the \(n\)-th coil in a phased array, \(f_0\) is the signal’s Larmor frequency, \('a_n(r, t)\'\) is a representation of the induced transverse magnetization signal, which includes the spatial distribution and relaxation behavior of the signal. This signal rotates at the Larmor frequency \((f_0)\), but this frequency is modified \((\Delta f(r, t))\) by spatiotemporally varying fields like the spatial-encoding gradient fields and signal off-resonances. In order to clearly explain SSB, Equation 2.1 is simplified by combining all the spatiotemporally varying terms into one term ('\(A_n(t)\)'):

\[
A_n(t) = \int a_n(r, t) \exp(-j2\pi (\Delta f(r, t) t) dr
\]

which can be used to substitute into Equation 2.1:

\[
S_n(t) = A_n(t) \exp(-j2\pi f_0 t)
\]

When representing the transverse magnetization as a rotating signal in a complex plane, a coil is able to observe a projection of that signal on the plane with an initial phase \((\theta_n)\).
Equation 2.3 can be further simplified by assuming that the coil observes the real projection of the plane. Using these simplifications, the MRI signal can be further simplified to:

\[ S_n(t) = A_n(t)\cos(2\pi f_o t + \theta_n) \] \[ 2.4 \]

By using this simplified signal representation, it is easier to describe how SSB can modulate and recover the MRI signal. In SSB, the MR signal is first multiplied by a monochromatic carrier with frequency \( f_c \) to create a double sideband modulated signal \( S_n(t)^{DSB} \):

\[ S_n(t)^{DSB} = S_n(t)\cos(2\pi f_c t) \] \[ 2.5 \]

which expands to after substitution of Equation 2.4:

\[ S_n(t)^{DSB} = \frac{1}{2} A_n(t) \left( \cos(2\pi (f_c - f_0) t + \theta_n) + \cos(2\pi (f_c + f_0) t - \theta_n) \right) \] \[ 2.6 \]

A spectrum representative of \( S_n(t)^{DSB} \) is shown in Figure 2.1. Typically, the carrier frequency is much higher than signal frequency. The modulated MR signal is now at two frequencies that surround the carrier frequency, which are referred to as sidebands.

Figure 2.1 - Spectrum of DSB Signal: The double sideband spectrum (solid black outline) is created after multiplication of the MR signal and carrier (dotted gray outlines). The carrier is depicted as simple delta function. The encoded MRI signal has been depicted as a Gaussian, but this shape will vary based on the acquisition and object imaged.
Since the two modulated sidebands contain the same information, it would be inefficient to transmit both of these signals wirelessly. In SSB, this DSB signal is filtered such that only a one sideband (the lower sideband) is selected for wireless transmission.

\[
S_n(t)^{SSB} = \frac{1}{2} A_n(t) \left( \cos(2\pi(f_c - f_0)t + \theta_n) \right)
\]  

[2.7]

In Equation 2.7, the lower sideband is selected for wireless transmission. Once this SSB-modulated MR signal is transmitted and received wirelessly with antennas, the original signal content of the coil is recovered by multiplying the modulated SSB signal with another carrier that has the same frequency and phase as the original carrier.

\[
S_n(t)^{Recovered} = S_n(t)^{SSB} \left( \cos(2\pi f_c t) \right)
\]

[2.8]

which leads to after substitution of Equation 2.7:

\[
S_n(t)^{Recovered} = \frac{1}{2} A_n(t) \left( \cos(2\pi(f_c - f_0)t + \theta_n) \right) \left( \cos(2\pi f_c t) \right)
\]

[2.9]

which expands to:

\[
S_n(t)^{Recovered} = \frac{1}{4} A_n(t) \left[ (\cos(2\pi f_0 t + \theta_n)) + (\cos(2\pi(2f_c - f_0)t + \theta_n)) \right]
\]

[2.10]

A spectrum representative of \(S_n(t)^{Recovered}\) is shown in Figure 2.2. Two sidebands exist in this recovered signal, where one of the sidebands is centered at the original Larmor frequency and another unwanted sideband is at a higher frequency \((2f_c - f_0)\). Low pass filtering \(S_n(t)^{Recovered}\) eliminates the unwanted sideband, which leaves only the original coil signal \(S_n(t)\).
There are some notable drawbacks to SSB modulation. Inherently, as revealed in Equation 2.10, the two carrier multiplication stages cause at least a 75% reduction in the original signal’s amplitude. Since signal power is related to the square of the signal amplitude, this would lead to a 94% decrease in power (or a 12 dB decrease.) In addition, further signal attenuation will occur during modulation and transmission (the wireless RF link, filtering, etc.), so this signal loss must be managed through proper design and amplification. Another important drawback in SSB is the synchronization requirement between the two carriers. If any jitter-based frequency error (Δf) exists between the two, this frequency error would cascade into the recovered MR signal. This error would interfere with the signal’s spatial-frequency encoding and manifest itself as image artifacts in the reconstructed image. This error must also be managed through proper design and electronic selection. The remaining topics in this section describe important engineering measurements that are necessary when quantifying these errors. These measurements determine the performance limitations of an SSB system.

Figure 2.2 - Spectrum of Recovered Signal: The recovered signal spectrum (solid black outline) is created after multiplication of the SSB modulated MR signal and carrier (dotted gray outline). The recovered signal contains two components. One is the recovered MRI signal centered at the original Larmor frequency with reduced signal power, and the other is an unwanted high frequency signal image that can be simply filtered.
2.2.2 Path Loss

In wireless transmission, it would be advantageous if the signal power delivered to the transmitting antenna would be equivalent to the signal power detected by the receiving antenna, as this would result in no additional loss to the SSB modulated signal. In reality, lossless transmission is infeasible, and path loss between the two antennas will occur. The principle component to path loss is propagation loss. The radiated power from the transmitting antenna spreads out as it travels to the receiving antenna and a fraction of the transmitted power is incident on the receiver antenna. Path loss is also comprised of additional lossy effects, which include antenna resistive loss, antenna efficiency, and multipath distortion. It is important to have an estimate of path loss when designing an SSB system.

2.2.3 Noise Figure

The incorporation of additional electronics between the preamplifier and the MRI system's ADC will introduce noise to the signal. This effect is typically characterized using the system noise figure (NF$_{sys}$). When expressed in decibels, NF$_{sys}$ is defined as the difference between SNR before and after the transmission system (51):

$$ NF_{sys} = SNR_{in} - SNR_{out} \quad [2.11] $$

NF$_{sys}$ is an important SSB measurement as it describes an inherent loss in signal quality that an SSB imposes on a signal, and it should be minimized with proper design.

When a transmission system contains multiple stages that each have their own unique gains and noise figures, the overall effect on system noise figure can be calculated with the Friis system noise figure equation (52):

$\sim 39 \sim$
\[
\text{NF}_{\text{sys}} = 10 \log_{10} \left( \frac{F_1 - 1}{g_1} + \frac{F_2 - 1}{g_1 g_2} + \ldots + \frac{F_N - 1}{g_1 g_2 \ldots g_{N-1}} \right)
\]  

[2.12]

Each subscript represents the individual stages of the transmission system. ’g’ is the gain in normal units and ’F’ is the noise figure in normal units, also known as noise factor. This equation is important because it reveals that \( \text{NF}_{\text{sys}} \) is dominated by the first stage noise figure as long as the gain is sufficiently large to minimize the noise figure impact from the remaining stages.

### 2.2.4 System Linearity

In SSB, it is important that the recovered signal is related linearly to the original MR signal. Image reconstruction in MRI relies on a linearity assumption between the signal's components, so any nonlinear amplification during modulation will result in image artifacts. Figure 2.3 depicts a typical system response curve of an SSB system. At higher levels of signal input power, the linear gain relationship between signal input and output begins to fail, which typically occurs due to a lack of power in the modulation circuitry. Linearity is assessed with the input-referred 1dB compression point measurement (\( \text{P1dB}_{\text{input}} \)). As shown in the figure, the \( \text{P1dB}_{\text{input}} \) is the input power level where the actual system gain (dashed line, Figure 2.3) falls 1 dB below the expected gain (solid line, Figure 2.3). In addition to input-referred, the -1dB compression point can also be output-referred (\( \text{P1dB}_{\text{output}} \)). When considering both the \( \text{P1dB}_{\text{input}} \) and signal noise floor (\( N_{\text{Signal}} \)), the linear dynamic range (\( \text{DR}_{\text{P1dB}} \)) of the SSB system can be determined, which is an important measurement because it will determine with the maximum SNR that can be linearly modulated.

\~40\~
2.2.5 Carrier Noise

Another important assumption in analog mixing schemes is that the carrier spectrum itself is a simple delta function. However, realistic carriers contain reference spurs and phase noise at frequency offsets surrounding the tone, which is depicted in Figure 2.4. When the MR signal is multiplied with a carrier such as this, these noise sources are now included in modulated MR signal. The dynamic range of the carrier (\(DR_{\text{carrier}}\)) is another measurement that determines the maximum SNR that can be encoded by a SSB system. If the MRI signal was mixed with a low \(DR_{\text{carrier}}\), the carrier's phase noise would interfere with the lower power spectral components of the MRI signal.

Figure 2.4 - Illustration of Carrier Noise: A realistic carrier contains phase noise and reference spurs surrounding the frequency tone, which determines the carrier's dynamic range (\(DR_{\text{carrier}}\)).
2.3 Methods and Materials

2.3.1 SSB Design

A SSB system was constructed for 1.5T MRI. The system was designed with the intention of implementing a minimally complex SSB system that was capable of preserving the high dynamic range of the MR signal while operating locally within the MRI environment. Operation of a single channel is summarized schematically in Figure 2.5. The coil preamplifier was supplied by a local coil manufacturer (Quality Electrodynamics, Mayfield Village, Ohio). The signal from the preamplifier was fed into a transmitter module, where SSB modulation was performed. The modulated signal was then transmitted wirelessly with a surface mount 1/4-wave monopole planar antenna (SP series, Linx Technologies). It was received immediately outside the magnet bore with a patch antenna. The patch antenna was based on a 1/2 wavelength rectangular design with a center coaxial feed point and air dielectric thickness of 1.27 cm (53). The received signal was then amplified (RF2314, RFMD), split by an 8-to-1 Wilkinson power splitter (not shown), and fed into a receiver module. The receiver module recovered the coil’s original signal content and filtered away the remaining signals from the other transmitters. The recovered signal was then fed directly into a MRI receiver port for image reconstruction as if it was a normal MRI coil signal.

![Figure 2.5 - Schematic of Single SSB Channel](image)

*Figure 2.5 - Schematic of Single SSB Channel: The amplified MRI signal was encoded with SSB modulation in the transmitter module (TX), transmitted wirelessly, and then the original MRI coil signal was recovered in the receiver module (RX). A highly stable clock reference was distributed to each module to ensure that the carrier frequencies were the same.*
The transmitter and receiver modules are depicted in Figure 2.6. The transmitter module contained an upconversion mixing stage (MAX2661, Maxim Integrated), an amplification stage (ERA-5XSM, Mini-Circuits), and a lumped-element, 7th order Chebyshev lowpass filter. The upconversion mixer performs the multiplication described in Equation 2.5. The lowpass filter completes SSB modulation by filtering the upper sideband of the mixed signal and any carrier leakage from the mixer. The amplifier preemptively compensates the modulated signal for path loss. The receiver module contained a downconversion mixing stage (MAX2682, Maxim Integrated) and a lumped-element 5th order Chebyshev bandpass filter. The downconversion mixer performs the multiplication described in Equation 2.8. The bandpass filter had a passband centered at the Larmor frequency. This filter removed the high frequency signal image at $2f_c-f_0$ along with any additional unwanted signal components. The Chebyshev filters were hand-built using a filter design program which is available online (54). When the modules were powered by a 10 V source (which was necessary to power the coil preamplifier), the transmitter module consumed about 800 mW, and the receiver module consumed about 300 mW.

**Figure 2.6 - Schematics of the Transmitter and Receiver Modules**: The transmitter module contained an upconversion mixer, amplifier, and lowpass filter. The receiver module contained a downconversion mixer and bandpass filter. The box labeled "$f_c$" is the site of carrier generation, and its schematic is depicted in Figure 2.7.
Carrier generation for both modules is illustrated in Figure 2.7. Each module’s carrier was generated locally using an integer-N Phased Lock Loop (PLL) (ADF4111, Analog Devices) and Voltage-Controlled Oscillator (VCO) (MAX2623, Maxim Integrated) combination. The loop filter had a bandwidth of 100 kHz and a phase margin of 45 degrees. The carrier frequencies produced were located in the 900MHz/GSM band, and they were limited to integer multiples of the clock’s 5 MHz due to phase instabilities present in the integer-N PLL. In order to eliminate excess cabling to control the transmitter modules, the carrier frequency was controlled locally with a microprocessor (PIC12F635, Microchip). The microprocessor had 1.75 KB internal FLASH memory that was used to store a look-up table of predefined settings to define the multiple carrier frequencies for the module to generate. External resistors and capacitors connected to the microprocessor defined a specific RC time constant, which selected a specific entry in the look-up table. Upon module power up, the microprocessor immediately probed the time constant and programmed the module. Following that, the microprocessor entered “sleep” mode, which allowed image acquisition to occur without clocking noise interference.

**Figure 2.7 - Schematic of Carrier Generation:** The carriers were produced using a PLL/VCO combination. The carrier frequency was programmed by a local microprocessor, which used the time constant of the RC network and an internally-stored look-up table.
To ensure phase stability and synchronization between the carriers produced by the transmitter and receiver modules, a highly stable 5 MHz clock signal was fed to both the modules. Using a spectrum analyzer (CXA series, Agilent), the phase noise of the clock was measured to be less than -90 dBc/Hz for frequency offsets above 100 Hz with -75 dBc spurious signal at 120 Hz offset. A screenshot of this spectrum measurement is shown in Figure 2.8. In addition, the frequency drift was measured to be less than 0.5 ppm over a 60 second interval. The clock was generated using a divide-by-two counter (SN74LS92D, Texas Instruments) and a 10 MHz crystal oscillator (CXOH20, Crystek Corporation). In this implementation, the clock signal was fed via coaxial cable to the array to eliminate any possible contamination from unstable reference sources.

![Figure 2.8 - Screenshot of the Clock Spectrum: (a)](image)

A screenshot from a spectrum analyzer reveals the low phase noise of the 5 MHz clock. The power measurement had a 5 MHz center frequency, 500 Hz frequency span, and 2 Hz frequency resolution. The phase noise at 100 Hz offset was -94 dBc/Hz. There are spurious signals at 120 Hz offset on both side of the 5 MHz clock.
2.3.2 Bench Measurements

The SSB system was tested with a series of bench measurements. The standard for evaluating a non-coaxial transmission system in MRI is the loss in image SNR. For a single MRI detector, an equivalent way to represent image SNR is through the acquired signal's dynamic range (DR\textsubscript{sig}). While DR\textsubscript{sig} is a variation on SNR, this dissertation uses SNR to denote image-space measurements and DR\textsubscript{sig} to denote frequency-space measurements. The mathematical relationship between these two values has been described by Gabr et al (30):

\[ DR_{\text{sig}} = 20 \log_{10} \frac{2}{\sqrt{N}} \sum_{n=1}^{N} \text{SNR}_n \]  \[ \text{[2.13]} \]

SNR\textsubscript{n} is the individual pixel complex-valued SNR in an image with N pixels. The value of DR\textsubscript{sig} is dependent on the same parameters as SNR, such as image contrast, coil size, proximity to patient, etc. In order to evaluate the performance in preserving DR\textsubscript{sig} and SNR, the SSB system was characterized with the measurements described in the theory section and described in further detail here.

The path loss between the two antennas was predicted by measuring it on the bench using a $S_{21}$ measurement on a network analyzer. The two antennas were positioned on the bench at the same distance and orientation they would be positioned in the MRI suite. Transmission coefficients were measured for the frequency range of 500 MHz centered at 868 MHz at intervals of 2.5 MHz.

$\text{NF}_{\text{sys}}$ was calculated for the SSB system. Assuming no nonlinearity effects to the signal were imposed on the signal, the definition of $\text{NF}_{\text{sys}}$ in Equation 2.11 can be modified as the difference between the acquired signal's dynamic ranges:
\[
NF_{sys} = DR_{\text{sig.in}} - DR_{\text{sig.out}} \tag{2.14}
\]

Therefore, any increase \(NF_{sys}\) with the SSB system will have a direct impact \(DR_{\text{sig}}\). Using the Friis equation in Equation 2.12, it is possible to calculate and predict the overall \(NF_{sys}\) for the preamplifier and SSB system together:

\[
NF_{sys} = 10 \log_{10} \left( \frac{F_{PA} + \frac{F_{TX} - 1}{g_{PA}} + \frac{F_{air} - 1}{g_{PA}g_{TX}} + \frac{F_{RX} - 1}{g_{PA}g_{TX}g_{air}}} \right) \tag{2.15}
\]

'PA' represents the coil preamplifier, and 'TX' and 'RX' represent the SSB transmitter and receiver modules. 'Air' represents the path loss experienced in the wireless link. It was assumed that there were no noticeable noise contributions from the MR suite environment during air transmission. 'Air' was regarded as a simple attenuation stage, where \(F_{air}\) was assumed to be equivalent to the inverse of the path loss \(g_{air}\). If significant, path loss could have a substantial impact on \(NF_{sys}\) even though it occurred after the first few stages of the receiver chain. To avoid this, it was important that the combined gain of the preamplifier and the transmitter module was sufficiently large to overcome the path loss. The increase in \(NF_{sys}\) for the SSB system was calculated using Equation 2.15.

\(DR_{\text{carrier}}\) for both the transmitter and receiver module was measured with a spectrum analyzer. Normalized phase noise relative to the carrier (dBc/Hz) was measured at frequency offsets of 1 KHz, 10 KHz, and 100 KHz from the carrier frequency. Reference spur power relative to the carrier (dBc) was also measured at integer multiple offsets of the 5 MHz reference using a bandwidth 500 kHz.

\(DR_{\text{P1dB}}\) for the SSB system was determined using a P1dB\(_{\text{input}}\) measurement. The system gain was recorded for varying amounts of 63.6 MHz input power, and P1dB\(_{\text{input}}\) was recorded when the gain deviated more than 1 dB. In order to simulate path loss, the
transmitter and receiver modules were connected directly with an attenuator that was equivalent to the measured path loss. Using the measured $\text{P}_1\text{dB}_{\text{input}}$ in decibels, $\text{DR}_{\text{P}_1\text{dB}}$ was calculated by subtracting $\text{P}_1\text{dB}$ with the coil's noise floor and the added noise from the coil preamplifier:

$$\text{DR}_{\text{P}_1\text{dB}} = \text{P}_1\text{dB} - (\text{-174 dBm} + 10\log_{10}(B)) - \text{NF}_{\text{PA}}$$  \hspace{1cm} [2.16]$$

where -174 dBm was the thermal noise in a 1 Hz bandwidth at 290 K, $B$ was the bandwidth of the acquired MRI signal, and $\text{NF}_{\text{PA}}$ was the noise figure of the coil preamplifier.

### 2.3.3 Phantom Imaging

Imaging experiments were performed on a Siemens 1.5 T Espree. The SSB system was connected to a rectangular coil (27.3 x 13.3 cm) mounted on a FR4 cylinder with a 25.4 cm outer diameter. The coil was actually one element from an eight element detector array, which is described in Chapter 3. The coil element was tuned and matched to the 1.5 T Larmor frequency (63.6 MHz), and the $Q_{\text{unloaded}}/Q_{\text{loaded}}$ measurement was 3.8 (250/65) (6). Transmission decoupling was performed passively with crossed diodes (56). The carrier frequency used was 935 MHz, corresponding to an 871.4 MHz SSB-modulated MR signal. One cable was used to transmit the clock and DC power to the transmitter module. The receiving patch antenna was placed at the end of the scanner patient table 80 cm from the transmitting antenna.

First, an imaging experiment was performed to determine if operation of the SSB system increased the noise floor of the MRI signal. A water phantom was placed inside the array, and the system’s body coil was used to acquire noise scans (where the body
coil transmitter voltage set to zero) when the SSB system was powered and unpowered. The standard deviation of the received signal for each acquisition was calculated and compared.

In order to fully characterize the impact to the MR signal, a set of imaging experiments was performed to measure $DR_{\text{sig}}$ before and after wireless transmission. To accomplish this, four separate images were acquired: (1) a “wireless image” collected using the wireless array, (2) a “wireless noise image”, which was the same as (1) but with the body coil transmitter voltage set to zero, (3) a “wired image”, which had the same imaging parameters as (1), but cabling was connected directly to the coil preamplifier output, and (4) a matching “wired noise image”. $DR_{\text{sig}}$ was calculated and compared for both the wireless and wired image acquisitions by doubling the peak k-space amplitude and dividing by the standard deviation of the noise acquisition as described by Gabr et al (30).

These experiments were performed with a large cylindrical water phantom (200 cm$^2$ cross-sectional area, 36 cm length) positioned inside the coil cylinder. A series of two-dimensional gradient echo (GRE) experiments were performed: TE = 10 ms, TR = 150 ms, FA = 47° (Ernst angle), FOV = 225 x 225 mm$^2$, Matrix Size = 128 x 128, BW = 100 Hz/Pixel. Slice thickness was adjusted to vary $DR_{\text{sig}}$. In addition, the same GRE experiments were performed with a higher acquisition bandwidth (BW = 1350 Hz/Pixel).

SNR maps were calculated for both the wired and wireless acquisitions and examined for artifacts. SNR maps were calculated by taking the magnitude of the reconstructed coil images and dividing by the standard deviation of the corresponding complex-valued noise image.
In order to estimate the error in the measurements described above, fifteen additional phantom GRE acquisitions were performed with a slice thickness = 2.5 mm and BW = 100 Hz/Pixel, where the signal from an individual coil was collected directly from the preamplifier. The standard deviation of DR_{sig} and the SNR map mean was calculated across the acquisitions.

2.4 Results

2.4.1 Bench Measurements

Figure 2.9 summarizes the path loss measurements performed on the bench. Path loss was measured at distances of 40, 80, and 160 cm. As the distance increased, the path loss worsened in average value due to increased propagation losses, and it also worsened in smoothness due to distortion effects becoming more dominant. Since the antennas were tuned and matched to 868 MHz, path loss continued to worsen outside the center bandwidth. When operating inside the MRI bore with a distance of 80 cm, this measurement estimated that path loss would be between approximately -25 dB and -30 dB for a frequency range of 840 MHz and 915 MHz.
The low noise preamplifier had a gain of 27 dB and a noise figure of 0.5 dB. The transmitter module had an estimated 10.0 dB noise figure and 24.7 dB gain. The receiver module had an estimated 9.5 dB noise figure and 7.2 dB gain. The receive module calculation included the antenna amplifier and Wilkinson distribution stage. Using Equation 2.15 and the worst case path loss of -30 dB, the expected performance of \( \text{NF}_{\text{sys}} \) with SSB was predicted to be 0.8 dB, which was only a 0.3 dB increase in the original coil noise figure.

The carrier noise of each transmitter and receiver module was measured. The maximum phase noise measured was -80 dBc/Hz at 1 KHz offset, -82 dBc/Hz at 10 KHz offset, and -80 dBc/Hz at 100 KHz offset. The maximum spurious power measured for the first system was -49 dBc at 5 MHz spacing and -63 dBc at 10 MHz spacing. Since the MRI signal is typically limited to 500 kHz, the reference spurs did not impact the

![Figure 2.9 - Path Loss Measured on the Bench](image)

(a) The transmission distances measured on the bench revealed how path loss worsened as distance between antennas increased. (b) A zoomed-in plot reveals that the path loss for the imaging experiments (distance = 80 cm) should be better than 30 dB.
modulated signal. Therefore, $\text{DR}_{\text{carrier}}$ was found to be 80 dB, limited by the carrier phase noise.

The $\text{P}_{1\text{dB}}$ input of a single channel with -30 dB attenuation was measured to be -43 dBm. The measurement was limited primarily by the linearity of the upconversion mixer in the transmitter module. When using Equation 2.16 and a 100 kHz signal bandwidth, the calculated $\text{DR}_{\text{P}_{1\text{dB}}}$ was 80.5 dB for the first SSB system. This value for $\text{DR}_{\text{P}_{1\text{dB}}}$ was on similar order to the value obtained for $\text{DR}_{\text{carrier}}$, but $\text{DR}_{\text{P}_{1\text{dB}}}$ will vary based on the acquisition used (which determines the MR signal bandwidth).

### 2.4.2 Phantom Imaging

The noise acquisitions collected with the body coil revealed no increase in noise standard deviation when either SSB system was powered, demonstrating no impact on the signal due to operation of the electronics.

The standard deviation of $\text{DR}_{\text{sig}}$ for the repeated phantom measurements was 0.1 dB, and the standard deviation of the average SNR was 1%. This demonstrates that the results in this study were reproducible within a narrow range. These standard deviations were used as the error estimate for results in this section.

Figure 2.10 summarizes the phantom imaging experiments performed with the SSB system. When comparing the difference in $\text{DR}_{\text{sig}}$, there was nearly identical performance between the wired and wireless measurements. At the smallest slice thicknesses, measured SNR preservation was approximately 101%. This value should never have exceeded 100%, and this discrepancy was likely due to signal losses in the
wired connection to the coil preamplifier. At larger slice thicknesses, compression of DR\textsubscript{sig} began to be observed.

Figure 2.11 shows examples of SNR maps acquired both wired and wirelessly. These maps confirmed that signal quality was almost completely preserved by the SSB system. Small structural artifacts were visible in the noise background of the SNR map for TH=10mm. These artifacts were likely due to slight non-linear amplification occurring at high DR\textsubscript{sig} levels. The SNR comparison (Figure 2.11b) further confirmed that SNR was nearly identical, and it also confirmed that the coil's sensitivity profile was not altered by the SSB system.
Discussion

The results presented in this study demonstrate that the SSB system successfully transmitted the MRI signal wirelessly without significantly impacting the signal quality. The system was characterized on the bench to have good SNR preservation for signals up to approximately 80 dB DR\textsubscript{sig}. This was confirmed in phantom imaging, where DR\textsubscript{sig} and mean image SNR were nearly identical when comparing wireless and wired acquisitions. Preservation was demonstrated for DR\textsubscript{sig} measurements up to 83.6 dB, but at this large signal level, compression of DR\textsubscript{sig} was beginning to manifest. Image artifacts in the

Figure 2.11 – SNR Maps: (a) SNR maps are presented from both the wired and wireless acquisitions for TH=2.5mm (top) and TH=10mm (bottom) at low BW (100 Hz/Pixel). The smaller images are the same SNR maps, but scaled for the bottom 2% of the original in order to reveal any artifacts introduced into the background noise. (b) Single lines from each of the SNR maps (vertical line, 35/128) are plotted.

2.5 Discussion

The results presented in this study demonstrate that the SSB system successfully transmitted the MRI signal wirelessly without significantly impacting the signal quality. The system was characterized on the bench to have good SNR preservation for signals up to approximately 80 dB DR\textsubscript{sig}. This was confirmed in phantom imaging, where DR\textsubscript{sig} and mean image SNR were nearly identical when comparing wireless and wired acquisitions. Preservation was demonstrated for DR\textsubscript{sig} measurements up to 83.6 dB, but at this large signal level, compression of DR\textsubscript{sig} was beginning to manifest. Image artifacts in the

~ 54 ~
background of SNR maps suggest that this was likely due to the acquisition’s $\text{DR}_{\text{sig}}$ approaching the $\text{DR}_{\text{P1dB}}$ limit of the SSB system.

The SSB system had multiple design features that made it uniquely compatible with the MR environment. In order to program the carrier frequency without additional cabling, a microprocessor with "sleep-mode" was used (41). This feature allowed local system control while avoiding ambient clocking noise typically associated with digital electronics, which would have potentially interfered with the MRI signal. Also, the MRI environment proscribed the use of electronic components with magnetic material. Large amounts of magnetic material are a safety hazard in MRI, as they would cause the device to become an uncontrollable projectile near the system magnet. But, even small amounts of magnetic material would have the potential to cause artifacts in the reconstructed images that result from local magnetic field distortion, magnetic saturation of electronics, or even vibration during the image acquisition. The lack of artifacts in the SNR maps demonstrated that the SSB system was designed sufficiently to avoid such artifacts.

Previous work has estimated that the majority of clinical MRI acquisitions have $\text{DR}_{\text{sig}}$ between 50 dB and 85 dB (30). The SSB system in this study was measured to have approximately 80 dB for both $\text{DR}_{\text{carrier}}$ and $\text{DR}_{\text{P1dB}}$, not covering the entire clinical range. For the high signal strength imaging experiments, the observed imaging artifacts had structural artifacts more likely corresponding to nonlinear signal compression and $\text{DR}_{\text{P1dB}}$. While the limiting effects of $\text{DR}_{\text{carrier}}$ were not clearly visualized in this study, $\text{DR}_{\text{carrier}}$ is still an important measurement in SSB, as it will likely be the limit of SNR preservation in more sophisticated designs.
The DC power consumption for the SSB system should be considered. Given the $P_{1\text{dB}}$ measurements, the maximum output power delivered by the transmitter module to the antenna was 7.4 mW. Yet, the transmitter module dissipated 800 mW to achieve nearly 80 dB $\text{DR}_{P_{1\text{dB}}}$. If a larger $\text{DR}_{P_{1\text{dB}}}$ system were designed (e.g. for 3D acquisitions), the required power would likely increase. It should be noted that the SSB design presented in this study was not optimized for efficiency. For example, the transmitter module power could have been reduced by nearly a factor of 2 if the coil preamplifier did not require 10V (the majority of the components required 5V or less). Even with an improved design, it is likely that the large power requirements in highly linear SSB systems will make long-term battery operation difficult in a highly parallel array.

The antennas used in this study were shown to have between -25 and -30 dB path loss with a distance of 80 cm. The transmission profiles shown in Figure 2.8 required precise positioning of the transmitting antennas. The antenna designs had highly anisotropic radiated power, so the transmission profile would vary considerably if a different orientation were used. The antenna positions found with the bench measurements served as guidance for operation in the scanner. The high levels of SNR preservation measured in this study demonstrate that the antenna positioning in the scanner was sufficient to replicate the measured bench performance.

An important topic is the clock synchronization between the modules and the MR scanner. Any additional phase noise or drifting from the lack of synchronization would have cascaded into the carriers and introduced additional phase errors into the acquired signal. A stable crystal oscillator signal was distributed to the modules via coaxial cable to ensure that the phase error accumulation did not drift outside the pixel bandwidth.
However, some other solution would have to be developed to move to a fully wireless coil array setup. The synchronization clock could be transmitted wirelessly to the array, but this high signal quality would need to be preserved even after path loss. Synchronous detection with the carrier could be used (29), but once again, high signal quality must be maintained during wireless transmission. Further investigation into the wireless transmission of a reference signal is explored in Chapter 4.

In conclusion, this Chapter described a single channel SSB system that was capable of transmitting the MRI signal and preserving signal quality while operating within the MRI environment. While single channel transmission is a noteworthy accomplishment, the advantages of wireless technology are relevant primarily when transmitting multiple coils’ signals in an array. In the next chapter, a feasibility study of using this SSB system in a highly parallel phased array is presented.
CHAPTER #3

Identification and Mitigation of Interference Sources Present in SSB-based Wireless MRI Receiver Arrays

3.1 Background

In the previous chapter, a SSB modulation system for MRI was presented that preserved a coil’s signal quality during wireless transmission. Signal preservation was only demonstrated with a single channel, where only the signal from one coil element was transmitted. The majority of the benefits associated with wireless coil technology come only when the signals from multiple coils are transmitted simultaneously. In this dissertation, SSB was investigated because of its believed ability to transmit multiple channels efficiently in a given bandwidth. Ideally, the spacing between channels in SSB is only limited by the bandwidth of the MRI signal. This would allow for a dense transmission spectrum, which is needed for highly parallel phased arrays. However, this advantage assumes that there are no potential sources of interference outside the encoded MR signal bandwidth.

The goal of this Chapter is to describe the interference between multiple SSB-encoded MRI signals. In this Chapter, potential sources of interference are presented, which include the increased bandwidth of the coil preamplifier noise and noise resulting from the carrier signal itself. A simulation method was developed that minimizes the
interference from these sources by identifying the appropriate channel spacing to be used when transmitting multiple modulated signals. This simulation method was used with the SSB system presented in Chapter 2 and another SSB system that was designed to have improved filtering of the identified interference sources. The optimal performance for both SSB systems was predicted in hypothetical transmission settings. Both SSB systems were used to simultaneously transmit the signals from an eight element receive array, and the simulation was used to determine the channel spacing. The performance of both systems was examined with bench measurements, phantom imaging experiments, and human head imaging. The majority of results from this Chapter have been published in an article in the journal Magnetic Resonance in Medicine (42).

3.2 Theory

3.2.1 Interference Sources of SSB-Modulated MRI Signals

Figure 3.1a is an illustration of a typical MRI coil signal modulated with SSB. If only the bandwidth of the MR signal was considered, the spacing between individual SSB channels would only need to be a few hundred kHz. Unfortunately, this modulated signal contains off-band noise that could interfere in adjacent SSB channels when they are combined during air transmission, which impacts the modulated signals' noise floors. Figure 3.1b is an illustration of only the noise from a SSB-modulated MRI spectrum. There is noise from the coil preamplifier that has a bandwidth much larger than that of the encoded MRI signal, and there is bleed-through of both the carrier and its phase noise from the upconversion mixer.
Figure 3.2 illustrates how these noise sources could interfere when two of these channels are transmitted simultaneously. If two of these spectra were separated by small frequency spacing (Figure 3.2a), the preamplifier noise would overlap into a neighboring MRI signal bandwidth, increasing its noise floor. Alternatively, if a wider spacing was used that was on the order of the Larmor frequency (Figure 3.2b), phase noise from the carrier would interfere with the neighboring signal bandwidth, also increasing the noise floor. Thus, one must take great care in both understanding and compensating for these interference effects when designing an SSB system and deciding the spacing between multiple channels.
A simulation method was developed in MATLAB to determine the channel spacing that yields the least amount of noise interference when transmitting multiple SSB-encoded MRI signals. The simulation generates the spectra from all the possible channel spacing combinations and calculates the expected noise increase in each of the encoded signal bandwidths. Channel spacing combinations are not limited to just uniform spacing, but could also contain mixed amounts of spacing between the transmitted channels. To determine all the possible spacing combinations, the simulation requires multiple input parameters: the number of SSB channels, the frequency step-size between possible carriers, the antenna transmission bandwidth, and the SSB noise power spectrum. The antenna bandwidth is the frequency range between the transmitting and receiving antennas where air transmission meets a certain path loss requirement determined by the SSB design. The SSB noise power spectrum should be representative of the typical SSB-modulated MRI signal when connected to an element of the coil array.

Figure 3.2 - Interference between SSB signals: This figure illustrates the potential interference that can occur when two SSB-modulated MRI signals are transmitted. (a) If two SSB signals have too small of channel spacing, preamplifier noise will interfere with the encoded signal bandwidth. (b) Similarly, if two SSB signals have too large of channel spacing, the phase noise from the carrier will interfere.
The spectrum should have a frequency span sufficiently wide enough to capture all the possible sources of noise interference and a frequency resolution on the order of the MRI signal bandwidth. With these input parameters, the individual SSB spectra for each possible carrier frequency can be generated and combined additively to create the spectra for possible channel spacing permutations. Using the combined spectra, the amount of interference in each channel is calculated by totaling the noise power increase in the encoded MRI signal bandwidth along with any additional bleed-through contributions that would occur during the MR receiver's down-sampling. The simulation records the maximum channel noise increase for each combination, and the overall channel spacing is chosen as the combination that creates the smallest increase.

3.3 Materials & Methods

3.3.1 SSB Design

To utilize the simulation method and investigate the effects of overlapping noise profiles in SSB-modulated MRI signals, two different SSB systems were constructed for 1.5 T MRI. The first system was presented in Chapter 2. The second system was designed to have a similar operation as the first system, but with an improved noise profile. Figure 3.3 illustrates the similarities and differences between the two systems. Generally, both systems had the same array configuration, receiver module design, carrier generation circuitry, and reference source. The primary difference between them was in the transmitter module design.
As depicted in Figure 3.3c, the transmitter module of the second SSB system was designed to have improved filtering of the interference sources present in the first SSB system. To filter the off-band preamplifier noise, an intermediate frequency (IF) stage at 43.6 MHz was included with a commercially available surface acoustic wave (SAW) filter (X6964M, EPCOS). At the time of design, an affordable commercial option for a.
63.6 MHz SAW filter was not able to be sourced. The 43.6 MHz SAW filter IF stage was chosen because when mixed with a high frequency carrier, there was still a large spectral gap present for SSB filtering, and the 20 MHz source was simply created through multiplication (PI6C4511, Pericom) of the 5 MHz reference clock. The drawback to the 43.6 MHz filter was that it has a relatively large -30 dB attenuation bandwidth of 7.6 MHz. The IF mixer was a low-power broadband mixer (LT5560, Linear Technology), and an extra amplification stage (ADA-4543, Avago Technologies) was necessary to compensate for SAW filter losses. The RF mixer was replaced with MAX2671 (Maxim Integrated), which had similar performance to MAX2661, but with additional conversion gain and carrier suppression. The SSB filter was replaced with a SAW filter from TriQuint Semiconductor. The particular SAW filter used depended on the module’s carrier frequency (part numbers used in this work: 855821, 856932, 855728, 856824.) This new transmitter module consumed about 1150 mW using a 10 V supply.

Figure 3.3a summarizes the configuration of both systems when utilized in a multichannel SSB array. The modulated signals from the coils were combined, amplified, and transmitted wirelessly with a single antenna. In this implementation, a single antenna pair was used to avoid potential negative effects from antenna coupling. The antenna pair was the same pair utilized in Chapter 2. The amplifier immediately before the transmitting antenna was ADA-4743 (Avago Technologies), and the amplifier immediately after the receiving antenna was RF2314 (RFMD). Signal combination and distribution was performed with cascaded series of lumped-element two-port Wilkinson power splitters.
Further design information regarding the receiver modules, carrier generation, reference source, antenna design, and first system transmitter module can be found in Section 2.3.1.

3.3.2 SSB Array Construction

Both SSB systems were used to transmit the signals from an eight coil element phased array. Eight similar rectangular coils (27.3 x 13.3 cm) were evenly placed on 25.4 cm outer diameter FR4 cylinder. Each element was tuned and matched to 63.6 MHz with an average $Q_{\text{unloaded}}/Q_{\text{loaded}}$ of 4.2 (256/61) (6). Neighboring coils were overlapped for nearest neighbor geometric decoupling, and low impedance preamplifiers provided decoupling for the remaining interactions (7). Transmission decoupling from the transmit pulse was performed only passively with crossed diodes (56). One cable was used to transmit the clock and DC power to each of the transmitter modules. The eight receiver modules were connected to a standard receiver cable, which could be directly plugged into the patient bed. Figure 3.4 shows a picture of the completed array.

![Figure 3.4 - Picture of Completed Array: The pictured array was equipped with the transmitter modules from the first SSB system.](image-url)
3.3.3 SSB Array Simulation

The channel spacing simulation was performed for both SSB systems when transmitting the eight coil signals from the array. The step-size between possible carriers was 5 MHz for both SSB systems, due to the use of an integer-N PLL. Using the path loss measurements found in Section 2.4.1 as a guide, a path loss requirement of -30 dB was chosen, and the usable bandwidth between the antennas was found to be 75 MHz (840-915 MHz). This allowed for 16 possible carrier values for the 8 channels to utilize. The SSB noise power spectra for both systems were acquired through two measurements. First, the noise power of the modulated MRI signal before SSB filtering was measured with a spectrum analyzer. The noise power was measured with a 0.5 MHz bandwidth at intervals of 0.5 MHz for a 200 MHz frequency span centered on the carrier. This fully characterized the carrier and preamplifier noise. Next, the profiles of the SSB filters were measured with S_{21} using the same frequency span and resolution. The two measurements were necessary to account for the variability of filter performance across a wide range of frequencies (particularly for the SAW filters in the second SSB system). Noise power for each channel was calculated as the noise power at the encoded center frequency with an additional 1% contribution from the upper sideband in the final downconversion stage (from 63.6 MHz to 1.4 MHz) in the MR receivers. This final fold-over occurs due to bleed-through in the down-sampling mixer (corresponding to 65.0-67.5 MHz.) Using these parameters, the simulation was performed for both SSB systems to find the optimal channel spacing, and the results were used in the array.
3.3.4 Antenna Bandwidth Simulations

It was clear that the upper limit of the number of channels in both the SSB systems was proportional to the total system bandwidth, which in this case was limited by the antenna bandwidth. Thus, to predict the maximal theoretical performance for other hypothetical SSB systems, additional simulations were performed for both systems using a variety of antenna bandwidths (75, 150, 250, 500 MHz) and a narrower carrier resolution of 0.5 MHz. With such a large number of channel spacing combinations to examine, it was impractical to go through every possible combination, so the wireless array simulation was modified to use simulated annealing (57) to estimate an approximately optimal channel spacing combination. The cost function for the simulated annealing was the average noise power increase experienced in all channels. Noise power for each channel was calculated as the average noise power over a 2.5 MHz signal bandwidth (corresponding to a 62.5-65.0 MHz center frequency) with the additional 1% signal contribution from the down-sampling stage in the MR receivers (65.0-67.5 MHz). Only a single representative SSB profile was used for each system. Since simulated annealing only examines a fraction of the total number of combinations, it was possible that the global optimum cost function may not have been found with only a single optimization. Therefore, the optimization was repeated at least 20 times for each arrangement of SSB system, antenna bandwidth, and coil number, and the minimum cost function was recorded.
3.3.5 Single Channel Bench Measurements

The bench measurements described in Section 2.3.2 were performed for both SSB systems. Please refer to that section for further details on the measurements for the predicted system noise figure ($\text{NF}_{\text{sys}}$), carrier phase noise and reference spurs, and the input-referred 1dB compression point ($\text{P1dB}_{\text{input}}$). These bench measurements characterized each system's ability to preserve the MRI signal dynamic range ($\text{DR}_{\text{sig}}$) and image signal-to-noise (SNR).

3.3.6 Phantom Imaging

The phantom imaging experiments described in Section 2.3.3 were performed with both SSB systems. Please refer to that section for further details on the experiments. Both SSB systems were tested with the body coil to examine for ambient signal interference. Wireless acquisitions collected with the SSB system were compared to counterpart wired acquisitions collected directly from the coil preamplifier. The gradient echo (GRE) acquisitions were performed first with only a single channel from each SSB system, and then, the same acquisitions were repeated with all eight channels operating simultaneously. $\text{DR}_{\text{sig}}$ and SNR maps were calculated for each imaging experiment.

3.3.7 Head Imaging

Both SSB systems were used to acquire anatomical brain images with the array. A GRE acquisition with an inversion recovery preparation pulse was used: $\text{TE} = 5$ ms, $\text{TR} = 2200$ ms, $\text{TI} = 1100$ ms, $\text{FA} = 15^\circ$, 250x 250 mm$^2$, 256 x 128 acquisition matrix, $\text{BW}$ =
180 Hz/Pixel, and TH = 10 mm. Each array element was normalized with respect to noise level and combined using an adaptive combination reconstruction (58).

### 3.4 Results

#### 3.4.1 Channel Spacing Simulation

Figure 3.5 shows the measured noise spectra of the modulated MRI signal for both SSB systems. The preamplifier input was connected to an unloaded coil element from the array. In the first system measurement, the preamplifier noise is visualized well, along with the bleed-through of the carrier and the unwanted upper sideband. In the second system measurement, the improved filtering provided by the SAW filters is seen in the reduced bandwidth of the preamplifier noise and the further SSB suppression.

![SSB Output Noise: System #1](image1.png) ![SSB Output Noise: System #2](image2.png)

**Figure 3.5 - Measured SSB Noise Spectra:** Example noise spectra from (a) the first SSB system and (b) the second SSB system reveals the differences in noise profiles. These spectra also illustrate the difference in the spectral gaps between the modulated signal and the carrier, highlighting the importance of simulation to manage the interference tradeoffs that exist between multiple SSB channels.

Figure 3.6 summarizes the channel spacing simulation results when using the operating parameters of the SSB array. The graphs in the figure are histograms of the
maximum noise power increase experienced in the channel spacing combinations. For the first SSB system, the best channel spacing combination was found to be when the channels were spaced apart by an even 10 MHz, while the remaining combinations have larger amounts of expected interference. For the second SSB system, the histogram reveals that all the combinations have improved interference performance when compared to the first system due to the improved filtering present in the second SSB system. Again, the best channel spacing combination was found to be an even 10 MHz spacing across channels, but there were additional combinations that could have been chosen with similar interference performance.

In order to further demonstrate the utility of the simulation method, Figure 3.7 summarizes the channel spacing simulation results when the parameters were modified so that nine channels from the array were transmitted instead of eight. In this case, channel spacing combinations with low amounts of interference no longer existed in either SSB

Figure 3.6 - Simulation Results for the Array: Histograms of the channel spacing simulation demonstrated that only a few channel combinations exhibited small amounts of interference when transmitting the eight channels of the SSB array. Because of the improved filtering in the second SSB system, the overall interference performance in all combinations was improved, and there was more than one channel spacing option for the optimal channel configuration.
system, because at least two channels must be spaced apart by 5 MHz, which causes the lower performance. This demonstrates how the simulation method can help determine the limits of an SSB system and assist in the design stage.

In order to show the most direct comparison between these systems in our scanner testing, a constant 10 MHz spacing was used for both systems in the array, where the receivers had carrier frequencies 905, 915, ... 975 MHz.

Figure 3.8 illustrates the agreement of spectra between the simulation and the actual SSB array. Figures 3.8a and 3.8c show the cumulative spectra predicted by the simulation. Figures 3.8b and 3.8d show the actual measured spectra from the SSB arrays. For both SSB systems, there was good agreement between the simulated and acquired spectra. In particular, the interference behaviors between neighboring preamplifiers agreed well. It should be noted that the individual channel gains of the first system were manually adjusted to lessen the impact of interference from carrier bleed-through. These adjustments (an additional gain of 2.5 dB to the 955 MHz channel and 5 dB to the 965

Figure 3.7 - Simulation Results for 9 Channels: The simulation parameters were modified to predict interference behavior if nine channels from the SSB array were transmitted instead of eight.
MHz and 975 MHz channels) made the predicted noise interference more uniform across the individual channels.

![Simulated vs Measured Power Spectra](image)

**Figure 3.8 - Multi-Channel SSB Spectra:** Both SSB systems were programmed with an even 10 MHz channel spacing. The simulated (a & c) and measured (b & d) spectra of all eight channels revealed similar behaviors. The smoother profiles in the simulated spectra were due to each data point representing the average power in a 0.5 MHz interval.

### 3.4.2 Antenna Bandwidth Simulation

Figure 3.9 illustrates the simulation results for multiple antenna bandwidths and a 0.5 MHz carrier resolution. For each antenna bandwidth, there is an abrupt increase in interference, which can be used as an indicator for the upper limit of performance for that particular configuration. For example, the predicted amount of interference for the first SSB system and a 75 MHz antenna bandwidth began to increase at eight coils,
corresponding to our observed results. As expected, a larger antenna bandwidth and better filtering would allow for more channels to be transmitted without a substantial amount of noise power increase. Given an appropriate antenna bandwidth configuration, more than 32 channels could be transmitted using either system. This figure shows how simulation can also be used to guide the selection of an appropriate antenna bandwidth and frequency band when designing an SSB system for MRI applications.

![Figure 3.9 - Results of Antenna Bandwidth Simulation](image)

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**Figure 3.9 - Results of Antenna Bandwidth Simulation:** The simulation results illustrate the relationship between antenna bandwidth, number of channels, and noise interference for both the first SSB system (a) and the second (b). Darker data points represent the average increase in channel noise for the optimal channel combination, and lighter data points represent the corresponding maximum noise power experienced in the combination.
In order to further demonstrate the utility of the simulation method, a hypothetical “third” SSB system was created and tested. This third system was modeled to replace the second system's IF stage with a 63.6 MHz SAW filter (831-SL63.6M-03B, Oscilent). This filter was too expensive to incorporate into the second SSB system design, but its inclusion would have eliminated the need for the IF stage and its associated drawbacks. The results of the simulation are shown in Figure 3.10, which demonstrate that the system would have similar behavior to the second SSB system.

![SSB System #3](image)

**Figure 3.10 - Additional Results of Antenna Bandwidth Simulation**: Additional simulations were performed for a hypothetical third SSB system that utilized a 63.6MHz SAW filter instead of the IF stage found in the second SSB system. The simulated behavior of the third system performs similarly to the second system. Additional repetitions of the simulation would produce smoother curves.

### 3.4.3 Single Channel Bench Measurements

The low noise preamplifiers had a gain of 27 dB and a noise figure of 0.5 dB. As described in the Chapter 2, the transmitter modules from the first SSB system had a predicted 10.0 dB noise figure and 24.7 dB gain. The second SSB system had a predicted 11.5 dB noise figure and 28.5 dB gain. The difference between the gains of the two systems was due to the additional amplification in the IF stage and the difference in
attenuation between the Chebyshev and SAW filters in SSB filtering. The receiver modules had a predicted 9.5 dB noise figure and 7.2 dB gain. Using Equation 2.15 and the measured worst case path loss of -30 dB, the expected performance of $\text{NF}_{\text{sys}}$ was predicted to be 0.8 dB for the first system and 0.7 for the second system. Despite the gain difference between the two systems, the expected difference in SNR performance should still be small when transmitting a single channel. These results have been summarized in Table 3.1.

<table>
<thead>
<tr>
<th></th>
<th>Coil Preamp</th>
<th>TX Module</th>
<th>Estimated Path Loss</th>
<th>Rx Module</th>
<th>Overall</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>SSB System #1</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Gain (dB)</td>
<td>27.0</td>
<td>24.7</td>
<td>-30</td>
<td>7.2</td>
<td>28.9</td>
</tr>
<tr>
<td>NF (dB)</td>
<td>0.5</td>
<td>10.0</td>
<td>30</td>
<td>9.5</td>
<td>0.8</td>
</tr>
<tr>
<td><strong>SSB System #2</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Gain (dB)</td>
<td>27.0</td>
<td>28.5</td>
<td>-30</td>
<td>7.2</td>
<td>32.7</td>
</tr>
<tr>
<td>NF (dB)</td>
<td>0.5</td>
<td>11.5</td>
<td>30</td>
<td>9.5</td>
<td>0.7</td>
</tr>
</tbody>
</table>

**Table 3.1 - Summary of NF_{sys} Calculations:** Summary of the noise figures and gains calculated for the elements of each SSB system.

The carrier noise of each transmitter and receiver module was measured. In the first system, the maximum phase noise measured was -80 dBc/Hz at 1 KHz offset, -82 dBc/Hz at 10 KHz offset, and -80 dBc/Hz at 100 KHz offset. The maximum spurious power measured for the first system was -47 dBc at 5 MHz spacing and -61 dBc at 10 MHz spacing. In the second system, the maximum phase noise measured was -78 dBc/Hz at 1 KHz offset, -78 dBc/Hz at 10 KHz offset, and -80 dBc/Hz at 100 KHz offset. The
maximum spurious power measured for the second system was -54 dBc at 5 MHz spacing and -61 dBc at 10 MHz spacing.

The $P_{1\text{dB}}^\text{input}$ of a single channel with 30 dB attenuation was measured to be -43 dBm for the first system and -49 dBm for the second system, both input-referred to the coil preamplifier input. In both systems, the measurement was limited primarily by the linearity of the RF mixer in the transmitter module. The difference between the measurements was due in part by the additional gain provided by the second system's IF stage. When using Equation 2.16 and a 100 kHz signal bandwidth, the calculated $\text{DR}_{P_{1\text{dB}}}$ was 80.5 dB for the first SSB system and 74.5 dB for the second system. When transmitting multiple SSB channels, the linearity limit of the system shifts to the downconversion mixer in the receiver module, because no filtering was used at the input of the receiver modules and all the wireless signals were fed into the mixer. The measured compression point of a single receiver module was -12.5 dBm.

### 3.4.4 Phantom Imaging Experiments

The noise acquisitions collected with the body coil revealed no increase in noise standard deviation when either SSB system was powered. The error estimates used for the figures in this section where the same found in Section 2.4.2 (0.1dB for $\text{DR}_{\text{sig}}$ and 1% for SNR).

Figure 3.11 summarizes the imaging experiments performed with the first SSB system. The single channel data is the same data reported in Section 2.4.2. When all eight channels were transmitting simultaneously, the $\text{DR}_{\text{sig}}$ acquired wirelessly was noticeably lower than the wired, with an average of 12% SNR reduction present for the smallest
slice thicknesses. Visual analysis of the SNR background maps reveal no structured artifacts in the TH = 2.5 mm acquisition, eliminating the possibility that this loss was due to system non-linearity and implying it was entirely due to interference from the other channels. Compression of $\text{DR}_{\text{sig}}$ begins to be observed at TH = 10 mm for single channel transmission and TH = 7.5 mm for simultaneous transmission. This was likely due to slight non-linear amplification occurring at these high signal levels, which is confirmed by the structural artifacts present in the SNR maps.

![Graphs and images showing measured DR (db) and fraction of mean SNR vs slice thickness for different transmission modes and BWs]

**Figure 3.11 - Imaging Experiments with the First SSB System:** Summary of the $\text{DR}_{\text{sig}}$ measurements are presented (a) when transmitting a single channel and (b) when transmitting all eight channels simultaneously. (c) The reduction in SNR was predicted using the differences between wired and wireless $\text{DR}_{\text{sig}}$. (d) SNR maps are also presented comparing the wired and wireless acquisitions for TH = 2.5 mm and TH = 10 mm at low BW (100 Hz/Pixel). The smaller images are the same SNR maps, but scaled for the bottom 2% of the original in order to reveal any artifacts introduced into the background noise.

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Figure 3.12 summarizes the imaging experiments performed with the second SSB system. When transmitting a single channel, there was again good agreement between the wired and wireless measurements. When all eight channels were transmitting simultaneously, only a 3% SNR reduction was present at the smallest slice thicknesses. DR sig compression at larger slice thicknesses was more pronounced for both single channel and simultaneous transmission. Visual analysis of the SNR background maps suggest that this was again likely due to non-linear amplification. This agrees with the bench measurements where the second system had both increased gain and reduced linearity. To further demonstrate that the compression during simultaneous transmission was primarily due to receiver non-linearity, an additional set of simultaneous transmission experiments were performed with an additional 3 dB attenuation at the input of the transmitting antenna (Figure 3.11e). At the expense of slightly reduced SNR preservation, the non-linearity losses at the larger slice thicknesses were reduced.
Figure 3.12 - Imaging Experiments with the Second SSB System: Summary of the DR\textsubscript{sig} measurements are presented (a) when transmitting a single channel and (b) when transmitting all eight channels simultaneously. (c) The reduction in SNR was predicted using the differences between wired and wireless DR\textsubscript{sig}. (d) SNR maps are also presented comparing the wired and wireless acquisitions for TH = 2.5 mm and TH = 10 mm at low BW (100 Hz/Pixel). The smaller images are the same SNR maps, but scaled for the bottom 2% of the original in order to reveal any artifacts introduced into the background noise. (e) An additional set of multi-channel imaging experiments were performed with an extra 3 dB attenuation at the input to the transmitting antenna.

Figure 3.13 summarizes the SNR performance of the individual array channels during simultaneous transmission for the low BW and TH = 2.5 mm acquisitions. Mean SNR was calculated as the average across the entire magnitude-based SNR map, including background pixels. For each SSB system, the eight wireless channels showed a similar scale of decrease. In the first SSB system, mean SNR was preserved on average.
by 88% (minimum 84%), and in the second system, mean SNR was preserved on average by 97% (minimum 94%).

Figure 3.13 - Mean SNR Measurements: A channel-by-channel comparison of the mean SNR values for (a) the first SSB system and (b) the second SSB system when collecting the low BW (100 Hz/Pixel) and TH = 2.5 mm acquisition. The difference between the wired SNR measurements was due the coils being slightly modified (new tuning and matching components) when the array was switched from the first SSB system to the second.

3.4.5 Head Imaging

Figure 3.14 presents the image reconstructions from the human brain acquisitions performed with both SSB systems. The difference in SNR performance between the two systems can be qualitatively observed in the two reconstructions. The improved image quality provided by the second SSB system reiterates the importance of managing interference when transmitting multiple SSB-modulated MRI signals.
The results presented here demonstrate that multiple SSB-encoded MRI signals can be simultaneously transmitted wirelessly with proper design and interference considerations. Preamplifier noise and carrier bleed-through were identified as potential sources of interference. Through both the simulation and filtering of these sources, good SNR preservation was obtained when transmitting eight SSB channels. When imaging a phantom with the smallest slice thicknesses, the first system preserved mean SNR by 88%, but the second system, with improved noise filtering, preserved mean SNR by 97%.

Simulation was used to evaluate the limitations of SSB systems. For example, if either of the SSB designs presented here were used in a receive array with more than eight channels (as illustrated by the simulation results in Figure 3.7), both systems would be incapable of maintaining such high levels of SNR preservation. The simulations summarized in Figure 3.9 demonstrate how either SSB system would need to be modified to allow for more channels to be transmitted. Interference can be reduced by either
implementing more aggressive filtering of the preamplifier noise and carrier noise or by implementing a larger antenna bandwidth by using a higher transmission frequency and/or additional sets of antennas. This will also require a narrower carrier frequency resolution by using a smaller reference frequency or a fractional-N PLL.

The designs presented here should provide researchers with a good starting point on their own development of a fully functional wireless SSB array for MRI. For example, the effects of non-linear amplification were clearly observed during the simultaneous transmission of higher power signals. The inclusion of a channel selection filter and a variable gain stage into the receiver design would address this problem (59). These components would also provide a means to mitigate the negative effects of air transmission that were not experienced in this work.

Many of the important observations made in the Chapter 2's Discussion (Section 2.5) are also relevant in this Chapter. Neither SSB system had a measured DR\textsubscript{P1dB} or DR\textsubscript{carrier} that covered the entire clinical range up to 85 dB. When compared to the first SSB system, the second system had a reduced DR\textsubscript{P1dB} and an increased power consumption, which suggests that further improvements in the SSB design will decrease efficiency and prevent long-term battery operation. In order to satisfy path loss requirement in the imaging experiments, the antenna pair had to be reoriented multiple times before adequate path loss was observed in all channels. The reference signal was again delivered via a coaxial cable to each of the transmitter and receiver modules. If the reference were to be transmitted wirelessly, this Chapter highlights the difficulties in preserving that signal quality while many other modulated signals are also being transmitted.
Many of the engineering challenges confronted in this Chapter were directly related to the goal of preserving 100% of the original MRI signal quality. As described in Chapter 1, there are other clinical applications for wireless coil arrays that do not require such high levels of signal preservation to achieve useful results. In the next two chapters of this dissertation, these applications are examined.
CHAPTER #4

Magnetic Field Monitoring with a Transceiver Array Equipped with SSB

4.1 Background

4.1.1 Magnetic Field Perturbations

In recent years, novel ultrafast high-resolution MRI acquisitions have been developed that have excellent spatial and temporal resolution while maintaining a high signal-to-noise ratio (SNR). These acquisitions have great potential in improving clinical care in multiple applications. For example, cardiac MRI requires high spatial and temporal resolution to accurately visualize the dynamics of myocardium motion and blood flow. With these acquisitions, radiologists are now able to diagnose many cardiac diseases with MRI that had been previously undetectable, such as infarct cardiac tissue and atherosclerosis. But, the drawback to these ultrafast acquisitions is that they require precise knowledge of the magnetic field evolution during the scan. In commonly performed clinical acquisitions, perturbations in the expected magnetic fields typically have minimal visible impact on the image reconstruction, but in ultrafast sequences, perturbations can cause artifacts to appear in the reconstructed images that severely degrade their diagnostic utility. Figure 4.1 illustrates the profound impact these field perturbations can have on the quality of a real-time cardiac image reconstruction (60).
The artifacts observed in Figure 4.1 were based on actual gradient delay measurements taken on commercially-available MR scanners (61). Field perturbations can originate from many different effects including gradient system delays, eddy currents, gradient coil coupling, and even thermal drift. Many of these perturbations are corrected by the MRI manufacturer during system installation, but many smaller perturbations still exist during individual scanning sessions. These smaller perturbations arise from multiple unique factors including the MR system, the type of acquisition, and even the subject being imaged. Therefore, if ultrafast acquisitions are to be become a clinical mainstay, field perturbations must be measured in tandem with the normal image acquisition.

Figure 4.1 – Field Perturbation Artifacts: Images were reconstructed (a) with an ideal k-space trajectory and (b) with a trajectory contaminated with single 3.5 us gradient delay. Cardiac image data was acquired with a radial TrueFISP acquisition (128x128 acquisition matrix, 300x300x8 mm³, TE/TR/FA = 1.32ms/2.64ms/70°, BW = 1116 Hz/Pixel). Images were reconstructed using Through-Time Radial GRAPPA and 16 projections per frame (60).
4.1.2 Magnetic Field Monitoring (MFM)

Magnetic Field Monitoring (MFM) is a technique that measures magnetic field perturbations concurrently with the normal image acquisition (36–39,62). MFM uses an array of miniature MRI coils with highly localized sensitivities (which are commonly referred to as probes), and these probes simultaneously measure the scanner's magnetic field evolution at different spatial locations (39,62). If the probes' locations have an appropriate and well defined spatial distribution, the field dynamics throughout the entire imaging volume can be calculated accurately (36). The signals acquired by the MFM probes must be highly localized, such that the signal sources have a size on the order of the image voxel. It is important that the probe's signal does not experience interference from the bulk MRI signal emanating from the patient which can be multiple orders of magnitude larger. Consequently, robust MFM systems utilize probes that are either electromagnetically shielded (37) or sensitive to a nucleus different than hydrogen, such as fluorine or deuterium (38). These modifications require that the MFM system have a transmitter system to excite the probes. In addition, the Larmor frequencies of "heteronuclear nuclei" (non-hydrogen nuclei) are typically outside the bandwidth of the narrowband MRI receiver, so a specialized multi-frequency receiver is also required to sample and process the signals from the MFM probes (63). Therefore, a robust MFM system requires a specialized multi-frequency transceiver system in addition to the array of NMR probes. The added cost and complexity of that transceiver system hampers the implementation of MFM into modern clinical MRI scanners.
4.1.3 Wireless Magnetic Field Monitoring (wMFM)

A wireless MFM (wMFM) system could transmit the signals from multiple MFM probes and utilize the narrowband MRI receivers to collect the probes' signals. Since the MFM probes are essentially a specialized coil array, the MFM signals could be transmitted wirelessly using the single sideband amplitude modulation (SSB) technology developed in this dissertation. In addition, the frequency-shifting properties of SSB and other FDM modulation techniques can be exploited such that hetero-nuclear probes can also be collected by the narrowband MRI receivers. Small enclosed wireless MFM probes would be permanently mounted on the interior of MRI coil arrays and can be used selectively only when needed. While these probes would be powered with a cabled connection in the array, the nature of wireless signal transmission provides flexibility in utilizing the limited number of available MRI receiver channels. If an acquisition required higher order MFM corrections, additional wireless probes could easily be included. This modular design allows for a flexible operation that can adapt to the demands of any scanning session.

In this Chapter, a study into the feasibility of wMFM is presented. First, a brief description is provided on the theory behind MRI image reconstruction, k-space trajectories, and MFM. A simulation is then presented that determines the accuracy requirements for a wMFM system. The design of a prototype wMFM transceiver system is presented, where a RF excitation system was incorporated to excite the probes and the SSB system was designed to be compatible with multiple Larmor frequencies at 1.5 T. The wMFM system was used to measure first order magnetic field dynamics with 

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hydrogen-based probes. System performance was evaluated by comparing the wMFM measurement with a cabled MFM system and another commonly used measurement technique. Further assessment was performed by comparing phantom images reconstructed with the wMFM measurement and the other measurement techniques.

4.2 Theory

4.2.1 MR Image Reconstruction and K-space Trajectories

The use of orthogonal gradient magnetic fields is the primary method to obtain spatial encoding of the MRI signal. As a simple one-dimensional example, when a single gradient field is applied that varies linearly in the "x" direction, the received magnetization signal now has a precession frequency ($\omega$) that is a function of the x-gradient strength ($G_x$) and the signal's x-position. More generally, the precession frequency for a MRI signal at position vector ($r$) exposed to orthogonal time-varying gradients ($G(t)$) can be written as:

$$\omega(t) = \gamma(B_0 + G(t) \cdot r)$$  \[4.1\]

When the MRI signal is mixed down to baseband by the system's receiver, the phase time-course of that signal ($\theta(t)$) can be written as:

$$\theta(t) = \omega_{baseband}t = \gamma(G(t) \cdot r)t$$  \[4.2\]

Equation 4.2 reveals the important spatial-temporal relationship that the gradients impose on the phase of the MRI signal. This is the relationship that allows spatial information to be collected temporarily.

In MRI, the time-course of the temporal frequency encoding of $G(t)$ is typically
represented as a k-space trajectory \((\mathbf{k})\) (64):

\[
\mathbf{k} = \gamma \int_0^T G(\tau) \, d\tau
\]  

[4.3]

This allows the bulk MRI signal incident on a coil \((s(t))\) to be written as (3):

\[
s(t) = s(\mathbf{k}) = \int d^3r \rho(r) \exp(-j 2\pi \mathbf{k} \cdot \mathbf{r})
\]  

[4.4]

In this simplified example, \(s(t)\) does not include any relaxation effects. \(\rho(r)\) is the spatial distribution of the bulk magnetization at time \('t'\). \(\rho(r)\) is also assumed to be weighted by the field sensitivity profile of the coil. Equation 4.4 is notable because it reveals that \(s(t)\) is simply the Fourier transform of \(\rho(r)\). An image of the spatially varying magnetization \((\rho'(r))\) can be simply reconstructed by performing the Inverse Fourier Transform on the sampled signal \((s_m(t))\).

\[
\rho'(r) = \int d^3k \, s_m(k) \exp(+j 2\pi \mathbf{k} \cdot \mathbf{r})
\]  

[4.5]

Up to this point, the both position and k-space vectors have been treated generally. MRI is typically limited to first order reconstructions where only linear gradients are used. With this simplification, a single two-dimensional image can be reconstructed using the following equation:

\[
\rho'(x, y) = \int dk_x dk_y s_m(k_x, k_y) \exp(j 2\pi (k_x x + k_y y))
\]  

[4.6]

The individual k-space locations \((k_x, k_y)\) are typically represented as a two-dimensional frequency space that graphically visualizes the sampling of the signal's different spatial-frequency components (65). Theses sampled k-space coordinates must satisfy the requirements associated with traditional frequency-domain sampling. For example, if the spacing between the individual k-space sampling coordinates \((\Delta k)\) is too large, then
aliasing artifacts may begin to appear in the image. And, if magnitude of the sampled k-space coordinates are not large enough, then the image might not have sufficient spatial resolution. Therefore, sufficient sampling coverage of these k-space coordinates is a primary factor in acquiring a high quality diagnostic image.

The different patterns to sample k-space are called k-space trajectories. There are many different k-space trajectories that can be used to sample the gradient-encoded MRI signal. The most popular strategy is to sample k-space in a simple rectilinear grid that satisfies both the resolution and field-of-view requirements for an image. This 'Cartesian' sampling method is very robust, and it allows direct use of the Fast Fourier Transform (FFT), but its drawback is that it requires a long time to traverse k-space and acquire the image. Multiple alternative sampling patterns have been proposed, and this work has focused on spiral-based 2D k-space trajectories (66,67). The spiral trajectory samples k-space in a fast and efficient manner, and any under-sampling leads to image artifacts with incoherent-like noise behavior in the image. These benefits make a spiral trajectory a popular candidate when acquiring dynamic images that require good spatial and temporal resolution. The drawback is that the trajectory is especially sensitive to errors in the k-space trajectory, leading to significant image degradation. For this reason, spiral trajectories were used to evaluate the performance of the wMFM system.

4.2.2 Magnetic Field Monitoring (MFM)

This section briefly overviews how a MFM system measures and calculates a k-space trajectory (\(k(t)\)). A MFM system measures the phase evolution of multiple signal sources distributed about the volume being imaged. In lieu of a more general formulation,
MFM is first described with a simple one-dimensional example. Given two resonant point sources at positions \(x_1\) & \(x_2\) and an applied linear gradient \(G_x\) for a time period \(\Delta t\), the phase difference (\(\Delta \theta\)) between these two sources is:

\[
\Delta \theta = \gamma(G_x \cdot (x_2 - x_1)) \Delta t + \Delta \theta_{t=0}
\]  

\[4.7\]

\(\Delta \theta_{t=0}\) is the initial phase difference between the signal sources. If the initial phase difference and the positions of the signal sources are known, it is possible to calculate the gradient field strength \(G_x\) through the measured phase difference of the received signals:

\[
G_x = \frac{\Delta \theta - \Delta \theta_{t=0}}{\gamma(x_2 - x_1) \Delta t}
\]  

\[4.8\]

By monitoring the time course of the gradient field, Equation 4.3 can be used to calculate the \(k_x\) trajectory.

This simple 1D example of MFM can be extended into a more general formulation, where multi-dimensional gradient dynamics are measured using a large number of probes located at positions \((r_i)\). During the acquisition, the baseband phase evolution of each probe \((\theta_i(t))\) can be written as:

\[
\theta_i(t) = k(t) \cdot r_i + \omega_{ref,i} t
\]  

\[4.9\]

\(\omega_{ref,i}(t)\) is the additional phase accrual from probe off-resonance, and the initial phase of the probe is assumed to be zero. More generally, as described by Barret et al.(36), the measured phases from all the individual probes can be written as:

\[
\theta_p(t) = Pk(t) + \omega_{ref,p} t
\]  

\[4.10\]

\(P\) is the "probe location matrix" containing basis functions of the probes' locations. The \(k\)-space trajectory \((k(t))\) can then be calculated using the probes' phase measurements, positions, and off-resonances (36):
\[ k(t) = P^+[0_P(t) - \omega_{ref,P}t] \]  \[4.11\]

where \( P^+ \) is the pseudo-inverse of the probe location matrix. The calculated k-space trajectory can then be used to reconstruct image signal acquired simultaneously with the measurement.

### 4.2.3 Other Trajectory Measurement Techniques

The primary limitation to implementing MFM is the specialized hardware required for the measurement. There are other proposed trajectory measurement techniques that require very little additional hardware (68–74). One of the most often used alternative techniques for trajectory measurement is the method proposed by Duyn et al (70). The "Duyn method" is easy to implement and can be applied to a wide variety of trajectories. The calibration scan can be performed using either a phantom or the object being imaged. In this method, the signals from multiple very thin slices are acquired at equidistant orthogonal directions from isocenter. For each slice, only one gradient is operational, which is orthogonal to the slice plane. By using an orthogonal gradient, these thin slices essentially act like one-dimensional MFM signal sources. The acquisitions are repeated with and without gradient operation to determine any phase offsets from field inhomogeneity or off-resonance. For a two-dimensional trajectory measurement, the Duyn method requires at least eight excitations for each k-space line acquired (four for \( k_x \) and \( k_y \)). Averaging of additional excitations is also performed to compensate for a lack of signal in the thin slices. The individual \( k_x \) and \( k_y \) measurements are then combined to predict simultaneous gradient operation of the gradients and form a k-space trajectory.
However, there are two primary drawbacks to these other techniques that make MFM a preferable alternative. One drawback is that they only measure the field dynamics that are repeatable. The measurement is performed before the actual image acquisition with a separate calibration scan. Assumptions are made that the long-term effects from system drift will have a minimum impact on the trajectory. Another important drawback is that they are not performed during simultaneous gradient operation. Coupling between multiple gradient systems could introduce additional field perturbations. MFM-based techniques address both these drawbacks by performing simultaneous measurement of the trajectory during the actual image acquisition.

4.3 Methods & Materials

4.3.1 wMFM SNR Requirements

A series of simulations were performed in MATLAB to evaluate the SNR requirements of the wMFM trajectory measurement. A mismatch between the expected and actual k-space locations will cause artifacts in the reconstructed images. Similarly, any error in the wMFM trajectory measurement will also produce artifacts in the reconstructed images. These artifacts would have a similar structure to the Gaussian white noise already present in reconstructed MRI images. Therefore, the trajectory error from the wMFM system measurement places an upper limit on the achievable SNR in reconstructed images. Simulations were performed to characterize this relationship for the spiral k-space trajectory explored in this study.

The MATLAB simulations were performed using the ideal 2D spiral trajectory (34) and a segmented 128x128 T1-weighted brain image (MNI152_T1_2mm_brain) from
the FMRI structural library (75). Synthetic MRI signal from brain image was derived using the spiral trajectory and the NUFFT (76). The ideal spiral trajectory was modified to simulate error from a wMFM measurement by randomly perturbing the trajectory coordinates by a standard deviation error in $\Delta k$. The simulation was performed for varying percentages of error equally in the $k_x$ and $k_y$ direction. These simulated "measured" trajectories were then used to reconstruct images using the NUFFT with the synthetic signal. The root-mean-square error was calculated between these images and an image reconstructed from the ideal trajectory ($\text{RMSE}_{\text{wMFM}}$). Additional simulations were performed to calculate the RMSE in images imposed with varying amounts of Gaussian white noise, which corresponded to a particular image SNR ($\text{RMSE}_{\text{SNR}}$). Both RMSE values were calculated with masked images to eliminate error contributions from background pixels. The $\text{RMSE}_{\text{wMFM}}$ values were matched $\text{RMSE}_{\text{SNR}}$ values that were less than or equal. The set of simulations was repeated ten times and averaged.

4.3.2 wMFM System Design

Figure 4.2 depicts a simplified block diagram of the wMFM system. The four wMFM probes are depicted as small solenoid coils containing a small signal source. Four probes can calculate up to first order magnetic field dynamics, which is sufficient to perform most MR image reconstruction techniques. Additional probes would either improve the accuracy of the measurement or be used to calculate higher order field terms (36). The wMFM system was designed to be compatible with probes at multiple 1.5 T Larmor frequencies (e.g. hydrogen and fluorine), but this work focused exclusively on hydrogen-based probes. Each probe was connected to a wMFM transceiver module.
Unlike previous designs in this dissertation, the wMFM transceiver module required a RF excitation system in addition to the SSB electronics. The local excitation offered multiple advantages, including the ability to measure three-dimensional acquisitions, to improve the isolation between probes and other interfering signal sources, and the ability to excite probes with different Larmor frequencies. The module was switched between excitation mode and reception mode with a T/R switch, which was synchronized with the MRI scanner through an external trigger signal. A reference clock was also required to generate the RF excitation signal and synchronize the upconverting and downconverting carriers. Additional wireless systems were designed to transmit both of these signals (the RF trigger and the reference clock). In this subsection, each component of the wMFM system design will be discussed in detail.

Figure 4.2 – Overview of wMFM System: The wMFM system consists of four solenoid probes, which allowed the measurement of first order field dynamics. Each probe was connected to a wireless SSB-based transceiver module. Each module requires two external signals: a reference clock and a RF trigger. The system was designed to receive these signals wirelessly as well. The SSB-encoded probe signals were received with the same multi-channel SSB receiver system used in Chapter 3, which demodulated the signals such that they were able to be fed directly into the MRI scanner.
Figure 4.3 depicts a block diagram of the RF excitation system in each wMFM transceiver module ("RF" block in Figure 4.2). A direct digital synthesizer (AD9851, Analog Devices) generated a carrier signal that was matched to the Larmor frequency of the wMFM probe. This frequency was controlled by a local microprocessor (PIC12F635, Microchip). The microprocessor required a reference clock to perform accurate frequency synthesis. The output of the synthesizer was filtered with a 5th order low pass Chebyshev filter and amplified (ADA-4543, Avago Technologies). The trigger signal was used to toggle the RF excitation signal between excitation and reception. In receive mode, the RF amplifier was disabled and the frequency of the DDS was set to zero. In excitation mode, the RF amplifier was enabled and the frequency of the DDS was set back to the Larmor frequency. Controlling the DDS frequency required both a 100 ns rising edge trigger circuit (77) on the DDS's frequency update pin (FQ_UD) and a 100 ns falling edge trigger circuit (77) for the DDS's reset pin (RST). Benchtop measurements that studied the dynamics of the transition are described later in Section 4.3.3.

**Figure 4.3 – wMFM RF Excitation System:** The direct digital synthesizer (circular object) was programmed with the wMFM probe's Larmor frequency using a local microprocessor and a reference clock. The synthesized signal was then low pass filtered and amplified for RF excitation. A trigger was used to switch the system between excitation and reception mode.
Figure 4.4 depicts the T/R switch used in each wMFM transceiver module ("T/R" block in Figure 4.2). The T/R switch changed the coupling of the wMFM probe between RF excitation mode and reception mode. A DC-biasing trigger (3.4/0 V) forward or reverse biased a PIN diode network. When the diodes were forward biased, excitation mode was enabled, where signal from the RF port coupled freely to the wMFM probe. The shorted PIN diode to ground provided high isolation from the RF port to the SSB-N port, and the quarter-wave phase shifter further increased the isolation between the ports. When the diodes were reverse biased, reception mode was enabled, where signal from the wMFM probe coupled directly to the SSB-N port. The trigger signal was coupled with a 1 µF inductor, and each port was coupled with a 100 pF capacitor. Benchtop measurements characterizing these transmission parameters are described later in Section 4.3.3.

![wMFM T/R Switch](image)

**Figure 4.4 – wMFM T/R Switch:** The wMFM transceiver module was switched between RF transmission mode and reception mode through a DC-biasing trigger signal and PIN diode network.

Figure 4.5 depicts a detailed block diagram of the probe preamplifier and SSB electronics in each wMFM transceiver module ("SSB-N" block in Figure 4.2). Unlike the previous chapters, the coil preamplifier was a wideband low-noise amplifier (MAR-6SM, Minicircuits). Wideband operation allowed for probes with different nuclei to be used
with the same module. The TX trigger controlled the biasing of the low noise amplifier's power such that it was off during RF transmission. SSB modulation was performed with an upconverting mixer (MAX2671, Maxim Integrated), a SAW filter (856932, TriQuint), and a two-stage amplification (ADA-4543 & ADA-4643, Avago Technologies). The additional amplification was necessary to overcome the reduced overall gain from switching to a wide-band preamplifier. Carrier generation was performed with a fractional-N PLL (ADF4153, Analog Devices) and VCO (MAX2623, Maxim Integrated). Unlike an integer-N PLL, the fractional-N PLL allowed for carrier resolution much finer than the reference clock frequency. This finer frequency resolution could be utilized to frequency shift hetero-nuclear signals such that they could be received by the narrowband MR receivers. The PLL loop filter had a bandwidth of 25 kHz and a phase margin of 45 degrees. As with past studies in this dissertation, the carrier frequency was controlled with a microprocessor (PIC12F635, Microchip) and an RC selection network (41).

Figure 4.5 – wMFM Preamplifier and SSB System: (a) After an initial amplification stage, the signal from the wMFM probe is SSB-modulated. The SSB-modulated signal is amplified further and sent to an antenna for wireless transmission. (b) The carriers for SSB modulation ($f_{c-n}$) are produced with a fractional-N PLL and VCO combination. The carrier frequency was controlled by utilizing a RC network to select among predefined settings stored in the microprocessor's memory.
The SSB-modulated signals from each wMFM probe were combined with a cascade of two-port power combiners (PD09-73, Skyworks Solutions) and transmitted wirelessly with a surface-mount 1/4-wave monopole planar antenna (SP series, Linx Technologies). The combined wireless signals were then received outside the MRI bore with an in-house constructed patch antenna and general-use SSB receiver unit. This was the same antenna and receiver unit used in the previous chapters of this dissertation, but the receiver did have a few notable changes. The received wireless signals were first filtered with the same bandpass filter used to perform SSB modulation (856932, TriQuint). This additional filtering prevented saturation of the subsequent amplifier (ADA-4543, Avago Technologies) from other interfering sources received by the patch antenna. Carrier generation for downconversion also used a fractional-N PLL, as depicted in Figure 4.5b. The downconverting mixer (MAX2682) and the 5th-order Chebyshev bandpass filter remained unchanged. The recovered 63.6 MHz signals (hydrogen's Larmor frequency at 1.5 T) were fed into the narrowband MRI receivers as if they were normal hydrogen coil signals.

Figure 4.6 depicts the wMFM wireless reference clock system. A 316 MHz clock signal was generated using a PLL/VCO integrated chip (ADF4360-9, Analog Devices) and a 22.579 MHz reference source (CCHD-957, Crystek Corporation). The output frequency was controlled by the same microprocessor setup utilized to generate the carrier signals. SAW filters were used to pass through only the 316 MHz fundamental (B3711, EPCOS). The PLL/VCO chip contained two phase locked outputs, where one was delivered to the wMFM system and the other to the SSB RX unit. Wireless transmission of the 316 MHz reference to the wMFM modules was performed with a pair
of tuned surface-mount antennas (SP series and HE series, Linx Technologies). A logarithmic amplifier with a limiter output (AD8309, Analog Devices) was used to recover the attenuated 316 MHz signal. The 316 MHz signal was divided to 158 MHz with a divide-by-two chip (HMC432, Hittite Corporation), and the new 158 MHz clock was buffered (CDCLVC1102, Texas Instruments) and distributed to the individual modules. This combination of 316/158 MHz was chosen because of the commercial availability of 316 MHz SAW filters and 158 MHz was a suitable reference frequency for both the DDS and the PLL.

The RF excitation trigger for the wMFM system was synchronized with the MRI scanner's RF trigger. The rising edge of the MRI trigger was detected with an Arduino Duemilanove unit (www.arduino.cc). Upon detection, the Arduino unit created a new

![Figure 4.6](image-url)  

**Figure 4.6 – wMFM Wireless Reference Clock System:** (a) A 316 MHz clock signal is created with an integrated PLL and VCO integrated circuit that has two synchronous outputs. SAW filters pass through only the fundamental frequency. (b) The clock signal can be transmitted wirelessly to the wMFM system (the two antennas in this figure are part of one wireless channel), where it is recovered by a high gain logarithmic amplifier. The SAW filters select only the 316 MHz clock signal. The recovered frequency is then divided by two and distributed to the individual wMFM transceiver boards.
trigger signal specifically tailored for the wMFM system and the acquisition being measured. This trigger signal had two programmable parameters: the duration of the trigger pulse and the delay between the MRI trigger and wMFM trigger. The duration of trigger pulse was adjusted to generate a suitable flip angle for the wMFM probes. Please see Section 4.3.6 for the calibration of this pulse duration. The delay of wMFM trigger pulse was also adjusted such that the excited probe signals would have minimal time to relax before the acquisition's ADC began. The trigger pulse was +5 V during RF excitation mode and 0 V during reception mode.

Figure 4.7 depicts the wMFM wireless RF trigger system. The trigger signal was modulated into a wireless On-Off Keying (OOK) signal (48). A 915 MHz signal was created with the same electronics used to generate the wMFM carriers as depicted in Figure 4.5b. The reference clock source for this signal was a small surface mount 5 MHz clock (FOX924B, Fox Electronics). The power of this 915 MHz signal was modulated with the Arduino's wMFM trigger signal by controlling the biasing of a wideband switch (ADG901, Analog Devices) and a series of amplifiers (ADA-4X43, Avago Technologies). When the trigger signal was high, approximately 18 dBm of 915 MHz was delivered to the transmitting antenna. When the trigger signal was low, only -60 dBm power was delivered. The transmitting antenna for the wireless trigger was another in-house constructed quarter-wave patch antenna, and the receiving antenna was a surfacemount 1/4-wave monopole planar antenna (SP series, Linx Technologies). Both antennas were tuned and matched to work at a center frequency of 915 MHz. The OOK signal was filtered with narrowband 915 MHz SAW filters (B3718, EPCOS). Incident 915 MHz signal power was measured with a RF power detector (LT5534, Linear Technology), and
a comparator (MAX998, Maxim Integrated) recovered the original trigger pulse from the power detector signal. The comparator input had a low-pass RC filter with a time constant of 100 ns, and 300 mV of additional hysteresis was added for noise immunity. The comparator would switch high when about -20 dBm was detected. 915 MHz was chosen for the OOK signal frequency because of the large commercial availability of 915 MHz technology and this frequency did not overlap with the modulated wMFM signals or the wireless reference.

![Diagram of wMFM Wireless RF Trigger System]

**Figure 4.7 – wMFM Wireless RF Trigger System:** (a) The wMFM trigger signal from the Arduino was converted into an On-Off Keying (OOK) signal at 915 MHz, which was then transmitted to the wMFM transceiver modules (the two antennas in this figure are part of one wireless channel). (b) The trigger signal was recovered by measuring the incident RF power at 915 MHz and comparing it with a reference signal equivalent to approximately -20 dBm signal power. The recovered trigger signal was then distributed to the individual wMFM transceiver boards.

### 4.3.3 wMFM Prototype Construction

In this feasibility study, only hydrogen-based probes were utilized for trajectory measurement. Therefore, the carrier frequencies were the same for both upconversion and downconversion. The four carrier frequencies were 911.72, 916.71, 922.25, and 927.24...
MHz. The carrier frequencies were chosen because their modulated signals were in the bandwidth of the SSB SAW filter and were spaced sufficiently to avoid interference in the MRI system's down-sampling stage. Since the noise profile of coil preamplifiers were broadband, there was no need to further optimize the carrier frequency spacing for interference reduction.

Figure 4.8a is a photograph of the wMFM system. This initial prototype system was designed so that all the individual components could be stacked. The system consisted of four wMFM transceiver boards (100 x 90 mm$^2$ each), the wireless reference clock board (95 x 40 mm$^2$), and the wireless RF trigger board (60 x 40 mm$^2$). The boards were stacked with a 15 mm vertical spacing. The boards for the wireless reference clock and RF trigger could be removed so that the signals could be fed directly with a wired connection. Each transceiver board had a connection port for the wMFM probe. In this study, DC power was delivered with a cable. This was chosen over battery operation primarily for convenience, since the wMFM system drew a large amount of power (see the Section 4.5 for further discussion on this topic). When powered with a 5 V source, the wMFM transceiver boards required approximately 315 mA during RF transmission mode and 170 mA during reception mode. The wireless reference board drew 120 mA, and the wireless trigger board drew about 8 mA. During scanner experiments, the wMFM system was positioned immediately inside the MRI bore.

Figure 4.8a also contains an example picture of a wMFM probe. The design of the probe was based on the previous work from Barmet (37). The probe coil was a four-turn solenoid coil that had an inner diameter of 1.7 mm and a length of 2.3 mm. It enclosed a glass capillary tube that contained 20 mM copper sulfate. Copper sulfate was
advantageous because it shortens T1, which leads to larger amounts of signal in faster acquisitions. The amount of solution in each probe varied from 12 to 17 µl. The column of solution was centered with respect to the coil.

Figure 4.8b is a photograph of the wMFM receiver unit. The individual receiver modules were connected to a standard array cable, which could be directly plugged into the patient bed. When operating in the scanner, the receiver unit was positioned approximately 80 cm away from the wMFM system. The antennas transmitting the wireless reference clock and RF trigger were positioned approximately 50 cm and 80 cm respectively away from the wMFM system. Since the wireless reference clock and RF trigger did not have precise path loss requirements, the antenna positions were determined by simply confirming that the wMFM system was operating correctly before performing a trajectory measurement.

Figure 4.8 – Picture of wMFM System: (a) The wMFM system consisted of four stackable transceiver modules, and their modulated signals were combined to be transmitted with a single antenna. Additional stackable boards were used to receive wireless reference clock and RF trigger signals. Each transceiver module was connected to its own wMFM probe. (b) The four-channel wMFM receiver unit was positioned at the end of the patient bed and plugged into MRI system as if it were a normal coil array.
4.3.4 Cabled MFM (cMFM) System

An in-house constructed cabled MFM system (cMFM) was used as a comparison to the wMFM system. Figure 4.9a depicts a block diagram of the cMFM system, and Figure 4.9b is a photograph of the system. When compared to the wMFM system, the cMFM system had the same design for the RF excitation system and the T/R switch. Since the system was originally constructed to work exclusively with hydrogen-based probes, the probe preamplifiers were narrowband low-noise amplifiers used in traditional phased arrays (Quality Electrodynamics, Cleveland, Ohio, USA). The system also had cabled connections for a RF excitation trigger and a reference clock (30 MHz source, CXOH7, Crystek Corporation). Using this separate system as a comparison was convenient, because it could be swapped quickly with the wMFM without disturbing the experiment setup in the scanner suite.

![Diagram of cMFM System](image)

**Figure 4.9 – Cabled MFM (cMFM) System:** (a) An in-house constructed cabled MFM (cMFM) system was used as a comparison. Each cMFM board was equipped with the same RF excitation system and T/R switch as in the wMFM system. The probe preamplifiers were narrowband low-noise preamplifiers (Quality Electrodynamics, Cleveland, Ohio, USA). (b) A photograph of the constructed four-channel cMFM system.

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4.3.5 wMFM Bench Characterization

The SNR preservation of the wMFM SSB system was characterized with the same measurements performed in Section 2.3.2. Noise present in the probe's signal would directly impact the error in the field estimation, so it was important to ensure that each wMFM transceiver module introduced small amounts of noise to the probe's signal. Please refer to that section for further details on the measurements for the predicted system noise figure ($NF_{sys}$), carrier phase noise and reference spurs, and the input-referred 1 dB compression point ($P1dB_{input}$).

The wMFM transceiver's RF excitation system was also characterized. Using an oscilloscope (DSO Series, Agilent Technologies), multiple measurements were made for both modes of transceiver operation including the excitation signal power, the transition period, delay from the RF trigger, and power consumption. These measurements were repeated for each transceiver module.

4.3.6 wMFM Calibration

Before trajectory measurements could be performed with the wMFM system, multiple calibration steps were performed. First, the three-dimensional spatial coordinates of the MFM probes were measured with the MRI system. The probes were mounted on a fixed apparatus so that the probes remained in the same position throughout the imaging experiments. When inside the magnet bore, the mounted probes were positioned so that they were evenly distributed around isocenter. Multiple high resolution GRE images ($TE/TR/FA = 10/100ms/30^\circ$, 5 mm slice thickness, $BW = 100 \text{ Hz/ Pixel}$, $325x325 \text{ mm}^2$ FOV, $1024x1024$ matrix size) that contained the probes were acquired using the body
coil. Probe coordinates were recorded from reconstructed images with and without the scanner's gradient distortion correction. Since the probes were placed away from isocenter, the gradient field strength experienced by the probes may vary from the gradient field strength experienced near isocenter. The uncorrected and corrected probe coordinates allowed for the calculation of approximate correction factors that scale a measured trajectory to a trajectory experienced at isocenter. Scaling factors were calculated for each orthogonal direction, where each scaling factor was defined as the average ratio of probe distances between the corrected and uncorrected coordinates for each orthogonal direction. This calculation assumed that the gradient field profiles and the probe positions were symmetric about isocenter.

Next, the flip angle of the wMFM system was calibrated. The wMFM system was connected to the probe apparatus, and the MRI body coil transmission system was disabled. Multiple free induction decay (FID) acquisitions were then performed. A FID is a specialized acquisition that has no gradient encoding, which allows for simple observation of the signal magnitude decay. A series of FID acquisitions (TE/TR = 3.3/2000 ms, 64 repetitions) were performed with a single wMFM channel, where the width of the wMFM RF trigger pulse was incrementally increased from 10 to 140 µs. The wMFM system 90 degree flip angle was determined to be the RF trigger pulse length that yielded the largest average FID magnitude rounded to the nearest microsecond.

The final calibration step was determining the off-resonance frequencies of the individual probes through a FID acquisition. Each probe's FID signal should only contain signal components from the probe's Larmor frequency. By using the phase evolution from the FID signals, the off-resonance frequency for each probe was calculated by
fitting the slope of the phase-time curve with a least-squares curve fit.

4.3.7 Spiral Trajectory Measurements

The wMFM system was used to measure a series of spiral trajectories. The spiral trajectory was based on previous work (78) and had the following parameters: 322x322 mm² FOV, 48 spiral arms, 660 ADC data points with a 3 µs sampling period. The first thirty points of the trajectory were without the gradients operating, so these points were used to establish the center k-space point (k=0). During each trajectory measurement, the MRI scanner's RF excitation system was deactivated to prevent any interference during wMFM probe excitation. This prevented imaging to be carried out concurrently with the trajectory measurement. Trajectory measurements were done with only one acquisition, so no averaging was performed over multiple acquisitions. In addition to the wMFM system, the cMFM system was also used to measure the same spiral trajectories.

4.3.7.1 Single Gradient Acquisition (Duyn Method)

A customized Duyn method sequence was developed with the spiral trajectory described above. As described earlier in Section 4.2.3, the Duyn method requires a specialized trajectory where gradient systems operate individually rather than simultaneously. The sequence had the following relaxation parameters: TE/TR/FA = 6ms/30ms/30°. The wMFM and cMFM systems were used to measure the trajectory of the customized sequence. Ideally, the MFM trajectory measurements would match the trajectory calculated from the Duyn method. In order to do this comparison, the Duyn measurement was performed with an additional acquisition. In this measurement, the
scanner's excitation system was re-enabled and a water bottle was positioned on patient bed so that two elements from the spine array could collect the signal. The Duyn measurement was repeated 16 times and the signal was averaged before calculating the trajectory.

4.3.7.2 Simultaneous Gradient Acquisition

The wMFM and MFM systems were then used to measure normal spiral trajectories with simultaneous gradient operation. A trajectory with an axial image orientation was measured with the following imaging parameters: TE/TR/FA = 0.8ms/4ms/20°, TH = 10 mm. Then, a double oblique image orientation was measured with the following parameters: TE/TR/FA = 1ms/5ms/20°, TH = 5 mm, Sag/Cor/Tra cosines = 0.5/-0.5/0.7.

4.3.8 Image Reconstructions

All the measured trajectories were used to reconstruct image data of a standardized 190 mm diameter phantom from the American College of Radiology (79). In the axial images, a specific region of the phantom (slice #5) was imaged that had a 10x10 grid of 14.4 mm squares. The phantom position remained the same for the double-oblique images. Image data was acquired with elements of the patient bed spine array and a flexible surface coil placed on top of the phantom. Images were reconstructed onto a 128x128 matrix using the NUFFT (76) and the first 600 points of the each spiral arm.
4.3.9 Cartesian Image Comparisons

As an additional comparison, another set of phantom images were acquired with a standard Cartesian trajectory with the same size and orientation as the spiral trajectory images. The reconstructed images were analyzed for geometric distortions and other artifacts. Images were compared by measuring both the maximum residual magnitudes and artifact power between the images. Artifact power was calculated with the following equation:

\[
\text{Artifact Power} = \sqrt{\frac{\sum_{x,y} (\text{Image}_1(x,y) - \text{Image}_2(x,y))^2}{\sum_{x,y} (\text{Image}_2(x,y))}^2} \times 100\% \tag{4.12}
\]

Image1(x,y) and Image2(x,y) are corresponding pixel values in two comparison images. According to Equation 4.12, artifact power is found by summing the square of pixel residuals and normalizing that value. A lower artifact power denotes that the two images are more similar. Calculations were made with masked images to eliminate contributions from background noise.

4.4 Results

4.4.1 wMFM SNR Requirements

Figure 4.10 summarizes the simulation results for determining the relationship between wMFM error and image SNR. The RMSE from both wMFM \( \Delta k \) error and image SNR were matched to create the curve in the figure. For example, if an image was reconstructed from a spiral trajectory with 10% \( \Delta k \) error, then the reconstruction would have a RMSE similar to that of an image with a SNR of 100. This graph provides a ~ 110 ~
means to estimate the upper limit of SNR performance for an image reconstructed with the wMFM measured trajectory. It should be noted that these simulation results were calculated using a simple brain atlas image, and the results may vary slightly when another image object is used.

4.4.2 Bench Measurements

The performance of the wMFM SSB system was characterized with multiple bench measurements. The wMFM probe preamplifiers had a specified gain of 20 dB and a specified noise figure of 2.3 dB. This low noise figure was due to the wideband operation of the preamplifier. The wMFM SSB transmitter system had a predicted 9.4 dB noise figure and 34 dB gain. The receiver modules had a predicted 6.4 dB noise figure and 16 dB gain. Using Equation 2.15 and a path loss of -30 dB, the wMFM NF$_{sys}$ was calculated to be 2.5 dB. The carrier noise for each transceiver and receiver module was measured. The maximum phase noise was -79 dBc/Hz at 1 KHz offset, -79 dBc/Hz at 10

Figure 4.10 - Simulation Results for wMFM SNR Requirements: The RMSE for images reconstructed with a certain Δk error in the spiral trajectory were matched with the RMSE for images with a various average SNR levels. Each point of the curve represents the average over 10 sets of simulations and the error bars are the standard deviation of that error.

\[ \text{Image SNR with Equivalent Error} \]

\[ \text{Error in } \Delta k \text{ (%) } \]

\[ 0\% \quad 5\% \quad 10\% \quad 15\% \quad 20\% \quad 25\% \quad 30\% \]

\[ 0 \quad 50 \quad 100 \quad 150 \quad 200 \quad 250 \]

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kHz offset, and -87 dBc/Hz at 100 KHz offset. The maximum interfering spurious emission was -54 dBc. Other more powerful spurious signals were present, but not reported due to them not interfering with the encoded signals. The P1dB<sub>input</sub> was measured to be at least -46 dBm for the transceiver modules and -21 dBm for the receiver modules. Assuming 30 dB path loss and four channels transmitting, the receiver module was the primary limiter of linearity, which created an equivalent transceiver-input P1dB of -51 dBm. When using Equation 2.16 and a 100 kHz signal bandwidth, the calculated DR<sub>P1dB</sub> was 70.7 dB.

The RF excitation system of the wMFM system was also characterized with bench measurements. When operating in excitation-mode, the output power delivered to the probe was approximately 0 dBm. The transition period was 5.0 µs when switching from reception to excitation mode and 1.5 µs when reverting back to reception mode. The delay between the MRI system trigger pulse and the pulse received by wMFM system ranged between 1 and 5 µs. This delay variation was primarily due to limitations of the Arduino.

### 4.4.3 Calibration Measurements

The wMFM system was calibrated for operation with the MRI system. The mean length of the RF trigger pulse necessary to generate a 90 degree flip angle was found to be 92 µs. While this measurement varies slightly across individual probes, the average provided a sufficiently accurate estimate to generate high SNR signals in each of the wMFM probes. When used in a 2D k-space acquisition, the probes had a rectangular arrangement centered around isocenter (delta x/y/z = 239/98/0 mm). When used in 3D k-
space acquisition, the probes had a tetrahedral arrangement centered around isocenter (delta x/y/z = 239/126/77 mm). The resolution of the coordinate measurements was approximately 0.3 mm. Errors due to this resolution would manifest as scaling errors in the individual k-space coordinates. Scaling errors could have been as high as 0.3% in the rectangular arrangement and 0.6% in the tetrahedral arrangement. The probe off-resonance measurements varied slightly from experiment to experiment, so they were reacquired after each series of trajectory measurements. These variations were likely due to the sample being repositioned within the solenoid between experiments. Each of the measured probe off-resonances were within the range of 100 to 300 Hz. The phase accruals caused from these off-resonances were subtracted from probes' data prior to trajectory calculation.

4.4.4 Trajectory Measurement Analysis

The error for \( k=0 \) was calculated for each of the trajectory measurements. First, the measured trajectories were normalized onto a 128x128 grid with \( \Delta k=1 \). Then, the standard deviations for both \( k_x=0 \) and \( k_y=0 \) were calculated using the first 30 points of the trajectory where no gradients were operating. Finally, the error in \( k=0 \) was calculated as the vector magnitude of these two values. When the probes were in a rectangular orientation, the standard deviation of \( k=0 \) was approximately 0.017 \( \Delta k \) for the wMFM measurement and 0.007 \( \Delta k \) for the MFM measurement. When in the tetrahedral arrangement, the standard deviation of \( k=0 \) was approximately 0.025 \( \Delta k \) for the wMFM measurement and 0.011 \( \Delta k \) for the MFM measurement. The increase in error with the tetrahedral arrangement was due to the system measuring an additional gradient in the z-
direction while not increasing the number of probes. The differences between the errors between the wMFM and MFM measurements were due in part to the differences in overall system noise figure (about 2.0 dB). Even with a tetrahedral probe arrangement and a larger noise figure, the wMFM system was able to preserve the probes' SNR sufficiently to obtain $k=0$ errors below 5% $\Delta k$.

Figure 4.11 summarizes the differences between the three trajectory measurements (wMFM, cMFM, and Duyn) in the modified single-gradient Duyn method sequence. Figures 4.11a illustrates a spiral arm measured by each of three measurement techniques. Figures 4.11b contain plots of both the residuals from a single spiral arm and the maximum residuals across all spiral arms. When comparing the wMFM measurement with the Duyn measurement, the residuals become larger during each ADC sampling period. These differences were possibly due a small gradient perturbations or small errors in the wMFM/cMFM system calibration. When comparing the wMFM and cMFM measurement, there residuals were much smaller, but small structured error was still observed. These differences could have been due to gradient variation and noise introduced by the wMFM electronics. In both comparisons, the residuals never exceeded 0.15 in both $\Delta k_x$ and $\Delta k_y$ ($\Delta k_{x,y}$) in a normalized 128x128 grid. Figure 4.11c contains histogram plots of the residuals. The full width at half maximums of the histograms are about 0.07 $\Delta k_{x,y}$ for the Duyn method residuals and 0.05 $\Delta k_{x,y}$ for the cMFM residuals. These small trajectory differences demonstrate that both the wMFM and cMFM systems agree well with the established Duyn trajectory measurement technique.
Figures 4.12 and 4.13 summarizes the trajectory measurement for the simultaneous gradient acquisitions (the axial orientation trajectory and the double oblique trajectory.) In both orientations, the residuals between the wMFM system and the other measurements were never larger than 0.55 in $\Delta k_x$ and $\Delta k_y$. Once again, the differences...
between these measurements could have been due to actual gradient perturbations, errors in system calibration, and noise introduced by the wireless electronics. The larger differences at end of the ADC periods were larger likely due to the reduced probe SNR during the ADC period. These acquisitions had a much faster repetition time (TR) than the Duyn acquisition, so the steady-state SNR of the probes would have been smaller. Even with this reduced probe SNR, the full width at half maximum for all the histograms were below $0.2 \Delta k_{xy} \Delta k_{xy}$. The residuals for the double oblique orientation were larger than the axial orientation, and this was expected because more gradients were being measured with the same number of probes. These plots demonstrate again that the wMFM measurement agreed better with the cMFM measurement.

![An Example of Residuals for Single Spiral Arm](image1.png)  ![Maximum Magnitude of Residuals Across All Spiral Arms](image2.png)

**Figure 4.12 - Trajectory Measurement Comparison (Simultaneous Gradients, Axial Orientation):** (a) Residuals are plotted for a single example ADC period in gray, and the maximum residuals across all the spiral arms are plotted in black. In both sets of measurements, the residuals between the measurements are less than 0.55 in $\Delta k_x$ and $\Delta k_y$. (b) All the residuals are plotted in histograms with a 0.02 $\Delta k_{xy}$ bin size.
4.4.5 Image Reconstructions

Phantom images were reconstructed with the measured trajectories. Figure 4.14 summarizes the reconstruction results from the axial orientation images, and Figure 4.15 summarizes the reconstruction results from the double oblique orientation. In both cases, the sets of reconstructed images are in good agreement. Difference images between the different trajectory measurements reveal little dissimilarity in image structure and contrast. The maximum residuals between the Duyn and wMFM images were 0.10 (axial) and 0.08 (double oblique), and similarly, the maximum residuals between the cMFM and wMFM images were 0.04 (axial) and 0.08 (double oblique). When using the wMFM images as a comparison, the artifact powers for the Duyn images were 4.6% (axial) and 5.2% (double oblique), and the artifact powers for the cMFM images were 3.5% (axial) and 4.5% (double oblique). The combination of these measurements and the difference images show that the images reconstructed with wMFM and cMFM trajectories were...
more similar than the image reconstructed with the Duyn measurement. These results demonstrate that the wMFM system was able to preserve the original probes' signals well enough to reconstruct an image that would be similar to one obtained with a direct wired measurement.

Figure 4.14 - Image Reconstructions (Axial Orientation): (a-c) The images reconstructed from all three trajectory measurements were very similar and revealed no qualitative differences. (d-e) Difference images with the wMFM image reveal that the wMFM and cMFM images were more similar than the Duyn image.
4.4.6 Cartesian Image Comparisons

In addition to comparing spiral trajectory images with themselves, all the images were compared to counterpart images collected with a standard Cartesian k-space trajectory. Figure 4.16 summarizes the results from this comparison. In both cases, the spiral images agree similarly well with their Cartesian counterparts. The difference images demonstrate that the images all contain very similar spatial information. The maximum residuals in the axial difference images were 0.21 (Duyn), 0.22 (cMFM), and 0.19 (wMFM). The maximum residuals in the double oblique difference images were 0.22 (Duyn), 0.14 (cMFM), and 0.19 (wMFM). Artifact power was calculated with the Cartesian image as the comparison. The mask was a simple circle with a diameter of 100

Figure 4.15 - Image Reconstructions (Double Oblique): (a-c) Again, the images reconstructed from the three trajectory measurements showed no qualitative differences. (d-e) Difference images reveal that the wMFM and cMFM images were more similar than the Duyn image.
pixels at the center of the image. The artifact powers for the axial images were 8.8% (Duyn), 9.5% (cMFM), and 9.6% (wMFM), and the artifact powers for the double oblique images were 9.0% (Duyn), 9.2% (cMFM), and 9.4% (wMFM). These results demonstrate that all three trajectory measurements were successful in producing images similar to a Cartesian acquisition, which validates the previous image comparisons in Section 4.4.5.

Figure 4.16 - Cartesian Image Comparisons: The spiral-trajectory images are compared with a Cartesian trajectory image with similar acquisition parameters. Both the axial images (a) and double oblique images (b) agree well qualitatively with their Cartesian counterparts.
4.5 Discussion

The results of this Chapter have demonstrated the feasibility of transmitting the signals from a MFM system wirelessly. A four-channel wMFM transceiver system was constructed that utilized the SSB technology developed in this dissertation. Successful operation of the wMFM system was demonstrated by measuring multiple first order spiral k-space trajectories. In each of the trajectory measurements, the wMFM system preserved SNR enough to obtain a $k=0$ error less than 5% $\Delta k$. According to simulation data, this error would not become dominant for reconstructed images with an average SNR up to 250. When used in a sequence modified for the Duyn method, the wMFM trajectory measurement agreed well with the measurements from both the cMFM system and the Duyn method, where residual magnitudes did not exceed 0.15 for both $\Delta k_x$ and $\Delta k_y$ ($\Delta k_{x,y}$) in a normalized 128x128 grid. In sequences with simultaneous gradient operation and a much shorter TR, the wMFM measurements still agreed well with the cabled measurements. The residual magnitudes did not exceed 0.55 $\Delta k_{x,y}$, and the majority of them were still below 0.1 $\Delta k_{x,y}$. Images reconstructed with these trajectory measurements were in good agreement, and the images had no visible artifacts or geometric distortion.

The wMFM system would have required three wired connections: power, a reference clock, and a RF excitation trigger. While the cable for power could be eliminated with battery operation, additional technology was developed to transmit the reference clock and RF excitation trigger wirelessly. In order to transmit the reference clock, a synchronized 316 MHz tone was transmitted to the wMFM system and recovered with a logarithmic amplifier. A potential drawback to this strategy was if the phase of the
received clock signal varied during the ADC period. This could have been caused by simple patient motion within the bore, since motion slightly changes the wireless pathways for the modulated signals. But, since the wMFM modules were stacked and the same reference clock was distributed to each modules, any phase variations would have been global and subtracted during trajectory calculation. In future wMFM realizations, where the transceiver modules operate separately, a more robust wireless clock system would need to be developed. The remaining cabled connection, the RF excitation trigger, was transmitted wirelessly with OOK modulation at 915 MHz. While OOK modulation was simple to implement, it was prone to oscillation when switching between RF excitation and reception modes. In order to prevent oscillation, great care was taken in designing the comparator hysteresis loop and positioning the antennas. Future realizations of this platform would need to utilize an OOK modulation frequency farther away from the modulated wMFM signals to increase selectivity and robustness.

Even though only hydrogen-based probes were utilized in this study, the wMFM system was designed so that it could collect signal from multiple types of nuclei (eg. fluorine and hydrogen) while still using the narrowband MRI receivers. The overall bandwidth of the system was limited by the T/R switch, which had a well characterized bandwidth between 20-80 MHz. Wideband operation required the use of a broadband coil preamplifier and a fractional-N PLL. The drawback to a broadband preamplifier was the reduced noise figure of 2.3 dB (a typical low noise narrowband preamplifier in MRI has 0.5 dB noise figure). Even with this reduced performance, the wMFM system preserved probe SNR sufficiently to agree well with the other trajectory measurements and have a small $k=0$ error.
One important aspect of the wMFM system that should be discussed is its power consumption. As with the previously presented designs, the SSB transmitter system required a relatively large 450 mW power to preserve and transmit a high dynamic range signal. Given its $\text{DR}_{\text{P1dB}}$ limit, the maximum output for a single SSB channel was only 9 mW. This inefficiency was only amplified with the additional electronics needed to make the module into a transceiver. The DDS and RF excitation amplifier required approximately 625 mW to deliver 1 mW of excitation power to the probe. The PIN diodes in the T/R switch required 100 mA to switch the system between operational modes. And, there were other digital electronics used to buffer the reference clock and control the RF trigger. In total, each wMFM transceiver module required 1.6 W during RF excitation mode and 0.85 W during receive mode. This large power consumption was why power was delivered via a cable rather than a battery. Battery operation would have been possible with a large enough battery pack. Since wMFM probes are likely to be mounted permanently inside coil arrays, they are more likely to be powered with a cabled connection. In either case, substantial design modifications would still be necessary to this prototype system to reduce the power consumption to levels that are appropriate for long-term operation.

Another important aspect of the wMFM electronic design was the selection of the DDS (AD9851, Analog Devices), which was used to create the RF excitation source. This particular DDS chip was chosen because of its commercial availability and its ability to directly generate Larmor frequencies for 1.5 T. The large power consumption of the DDS core was able to create observable ambient interference during reception mode. In order to eliminate this interference, the DDS was reset in reception mode to produce a
DC signal (zero frequency). The drawback to this solution was a loss of phase coherence between consecutive RF excitation pulses. This created random variations in the amplitude of probe signal from TR to TR. These variations were small, so they only posed a problem in measurements with very short TR and low signal. Future revisions of this system should include a DDS that can maintain synchronization when switching between operational frequencies and/or one that has low enough power consumption to not create observable interference. And for better performance, the new DDS should also be able to control the excitation phase from TR to TR.

The wMFM system was used to measure an ultrafast spiral trajectory with a 322x322 mm$^2$ FOV and 192 ms temporal resolution (48 spiral arms with 4 ms TR). This trajectory measurement was near the performance limit of the wMFM system. If a faster acquisition or higher gradient strength were used, the probes' signals would have experienced a significant amount of dephasing that would make it impossible to extract correct phase information. As already noted in the previous paragraph, this problem would be addressed in part by incorporating a phase-synchronized RF excitation system, but a more impactful solution would be to improve the quality and size of the wMFM probes. The probes could have improved susceptibility matching, which would increase the sample's T2* relaxation time along with the magnitude of the sample's signal (39). The probe size could also be reduced, which would allow for higher gradient strength acquisitions with less signal dephasing. Improving either of these probe characteristics would increase the range of trajectories that could be measured by the prototype wMFM system.
The wMFM trajectory measurements were compared to trajectory measurements performed with a wired MFM system and the Duyn method. The comparison with the Duyn measurement was very useful, because it demonstrated that the trajectory measurements from both the wMFM and cMFM systems were accurate. Previous research has claimed that the MFM trajectory measurement method should be more accurate than the Duyn measurement (80), but the results of this study were insufficient to demonstrate this. The MRI system utilized in this study had surprisingly good field perturbation performance in the spiral acquisitions, so the differences between the trajectories measurements were small enough that it was not possible to attribute them to either actual field perturbations or small errors in the MFM system calibration. The differences between the reconstructed images revealed little difference in phantom structure and shape. Performing higher resolution trajectory measurements and image reconstructions would likely reveal observable improvements with the MFM measurements. In addition to this, another reason that prevented this validation was that the image data was acquired separately from trajectory measurement data in this work. The MFM probes in this work were unshielded, so the scanner's transmitter system had to be disabled during the trajectory measurements to avoid transmitter interference. Collecting image data simultaneously with the wMFM trajectory measurement would be necessary for observing improvements over the Duyn measurement.

This Chapter has presented the next step in SSB-based wireless MRI array technology from the wireless receive-only technology presented in Chapters 2 & 3. Transceiver technology equipped with SSB was presented along with technology for a wireless reference clock and wireless RF trigger. While there are potentially multiple
MRI applications that can benefit from wireless transceiver technology, trajectory measurement was an especially appealing application because a MFM probe only requires a small amount of excitation power. On the other hand, the wireless reference clock and wireless RF trigger could be extended to all forms of wireless array applications, such as the receive-only imaging arrays presented in Chapter 3. In the next chapter, a different clinical application for wireless arrays is investigated, and rather than requiring a more sophisticated wireless system, this application can operate successfully with a substantially simplified SSB system.
For the past 25 years, the role of magnetic resonance imaging (MRI) in image-guided needle biopsies has rapidly expanded (81). When first developed, MR-guidance was achieved by the simple manual alignment of the imaging plane with the susceptibility artifact produced by the biopsy needle (82,83). Now, sophisticated systems exist that give a physician the ability to dynamically control the imaging plane and monitor the needle trajectory in real-time. This is accomplished by tracking the position and orientation of the needle insertion apparatus and using that information to update the imaging plane between each acquisition.

There are three primary methods of tracking the position and orientation of a needle apparatus inside the MR bore. They all share the common goal of localizing multiple markers with fixed positions on the apparatus. Optical tracking (84–86) uses multiple cameras with different positions and orientations to measure the locations of optical markers. By using reflective optical markers the entire apparatus can become self-contained and wireless (86). The drawback to this technique is that the optical markers must remain visible to the cameras throughout the entire procedure, which can severely limit the ability of the physician to freely wield the device in the crowded magnet bore.
Passive fiducial marker tracking (87,88) uses the signal enhancement produced by individual tuned coils that couple directly to local fiducial marker. The coils that are already being used to collect the image (body coil, spine array, etc.) are also used to collect the enhanced signal. The fiducial markers are localized by utilizing specific sequence parameters to enhance the signal of the passive fiducial marker while only slightly perturbing other surrounding signal sources. This allows for an apparatus to be entirely self-contained and wielded within the bore without regard for camera visibility. The drawback of this technique is that the enhanced marker visibility and automated localization is highly dependent on a number of parameters including the marker position and orientation, which still limits the ability to freely wield the needle apparatus inside the bore. Active fiducial marker tracking (89–93) also uses tuned coils containing fiducial markers, but connects them directly to the MR scanner treating them as standard coils. This allows for easy integration with the MR system, and the apparatus can be robustly localized regardless of marker orientation and position as long as the marker remains inside the bore's imaging region. However, this requires coaxial cables to connect the markers to the MR system. Coaxial cables may hinder the wielding of the device, and more importantly, cables can present patient safety issues (94).

Active tracking could be greatly improved with the incorporation of wireless detector array technology. This would allow for a device that is entirely self-contained and that can be freely wielded inside the magnet bore without worry of visibility or cabling constraints. Localization of the device would be easily integrated with the MR system. In this Chapter, the feasibility of performing active tracking with a wireless detector array was investigated. An active tracking device equipped with a wireless SSB
system is presented. The SSB system was based on the design presented in Chapter 2, which was simplified to take advantage of the reduced requirements in active tracking. The device was evaluated through both bench and imaging experiments. A needle-guidance experiment was performed with the device, which demonstrated active tracking with a wireless detector array to be well suited for MR-guided biopsies. Finally, a human volunteer imaging experiment was performed with the device to further demonstrate its potential clinical utility. The results of this Chapter have been accepted for publication in Magnetic Resonance in Medicine.

5.2 Theory

5.2.1 Active Fiducial Marker Tracking

The paradigm for active tracking used in this study was introduced originally by Dumoulin (40). Active tracking requires the use of three orthogonal linear gradient fields and multiple receiver coils each coupled to their own small fiducial marker. When the fiducial marker is exposed to a linear gradient field, the precession frequency of the marker becomes a function of its spatial position. For a single gradient field in x-direction ($G_x$), the precession frequency ($\omega_x$) of the marker at position ‘x’ can be written as:

$$\omega_x = \gamma (B_0 + G_x \cdot x) \quad [5.1]$$

If each coil is sensitive to only the signal coming from its own fiducial marker, the magnitude of the acquired signal spectrum should contain a single peak at a position corresponding to $\omega_x$. Since the received MR signal is mixed down to baseband, the ‘x’ position the marker can be calculated with the demodulated location of $\omega_x$ ($\omega_{x,\text{baseband}}$) and the strength of the gradient field ($G_x$):

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Acquiring the MR signal with a single linear gradient field is known as a projection, and in active tracking, at least three projections are acquired with subsets of orthogonal gradient fields ($G_x$, $G_y$, $G_z$). These projections are then used to determine the marker’s spatial coordinates ($x$, $y$, $z$). Multiple marker positions can be acquired simultaneously by using an array of small coils and fiducial markers. Once all the spatial coordinates of the markers are calculated, a new imaging plane can be calculated based on those coordinates. In order to define a plane, the positions at least three markers must be acquired. The relationship between the marker positions and the scan plane is arbitrary, so it is typically tailored to the design of the device containing the markers. The calculated scan plane is fed back to the MR control console and used in the subsequent image acquisition. The acquired image is reconstructed and presented to the physician via an in-room monitor. The projections and calculations typically take less than 50 ms, so this scan plane update can be interleaved between high temporal resolution image acquisitions. This pattern creates a feedback mechanism for the physician that gives intuitive scan plane control during an intervention.

### 5.2.2 Backprojection Reconstruction

Localizing the position of the active fiducial markers with orthogonal projections is a simplified case of backprojection reconstruction (95). Recall that the Fourier Transform of a single projection dataset is the 2D integral of the excited object in the directions orthogonal to the applied gradient field. In order to obtain an estimate of the original
object, this projection data can be backprojected onto an 3D image space. This creates an estimate object that mathematically satisfies the acquired projection. Additional estimates can be found by acquiring projections at other unique angles, and these estimates can be combined to get an accurate depiction of the original object. There are multiple ways to combine these estimates. A simple method is simply superimposing the estimates together, and a more accurate method involves iteratively finding a solution that satisfies all the estimates. In any of these methods, the accuracy of the object's reconstruction is improved as additional estimates are included with unique projection angles.

Since the fiducial markers are spherical and received on independent channels, only three projections that aren't necessarily orthogonal are needed to accurately determine the locations of the fiducial markers. For each channel, the fiducial marker projection estimate will ideally be a plane in the 3D image space. Therefore, three projections will create three planes in each channel, and the intersection of these planes is the location of the fiducial marker for that channel. To illustrate this, Figure 5.1 depicts 2D localization with three non-orthogonal projections. Since this is a 2D representation rather than 3D, the image estimates are lines instead of planes. The accuracy of the marker coordinates is improved by the number and the uniqueness of the projection angles. Therefore, orthogonal projections give the most accurate coordinate estimation while also being computationally simple. Device tracking with non-orthogonal projections can be useful with radial k-space sampling as a means to combine image acquisition with the device tracking (96).
5.3 Methods & Materials

5.3.1 Modified SSB Design for Active Tracking

As presented Chapters 2 and 3, the wireless SSB systems used to transmit multiple MRI coil signals had numerous rigorous design requirements. However, the localization of active fiducial markers only requires detecting the maximum signal magnitude in a simple projection acquisition (40), without requiring large amounts of SNR or any coherent phase information. This greatly reduces the requirements for wireless transmission and enables the use of technologies that would otherwise be unsuitable for conventional wireless MRI. Therefore, the SSB modulation system presented in Chapter 2 has been simplified to highlight these reduced requirements.

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Figure 5.1 – Active Fiducial Marker Localization with Non-Orthogonal Projections: Non-orthogonal projections can be used to localize active fiducial markers. In this 2D depiction, three non-orthogonal projections are acquired of the original object at different angles. The estimate images from the projections are lines that coincide with the marker's position. The spatial location of the marker can be determined by finding the intersection of these lines in a combined image.

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Figure 5.2 depicts the SSB transmission system implemented in the three tracking coils used in this study. For each tuned fiducial marker, the signal was first amplified with a low-noise preamplifier (Quality Electrodynamics, Cleveland, Ohio, USA). The signal was then mixed (MAX2671, Maxim Integrated) with a unique high frequency carrier. The carrier was generated locally using an integer-N PLL (ADF4111, Analog Devices) and VCO (MAX2623, Maxim Integrated) combination with a local 10 MHz reference clock (FOX924B, Fox Electronics). The 10 MHz reference clock was chosen because of its small surface mount package and good frequency drift performance. The PLL loop filter had a bandwidth of 100 kHz and a phase margin of 45 degrees. The carrier frequency was controlled locally with a microprocessor (PIC12F635, Microchip), which was programmed to enter sleep mode immediately after programming the PLL (41). The modulated marker signal was amplified (RF2314, RFMD) and filtered to select the lower mixed sideband using a lumped-element 7th order Chebyshev lowpass filter. The Chebyshev filters were hand-built using a filter design program, which is available online (54). The SSB-encoded signal was then wirelessly transmitted with a surface-mount 1/4-wave monopole planar antenna (SP series, Linx Technologies). The three wireless signals were then received outside the MRI bore with an in-house constructed patch antenna (53). It was based on 1/2 wavelength rectangular design with a center coaxial feed point and air dielectric thickness of 1.27 cm. The received signal was amplified (RF2314, RFMD) and split evenly using a cascaded series of lumped-element two-port Wilkinson power splitters. The split signals were then mixed (MAX2682, Maxim Integrated) with the same carrier frequencies used for upconversion, each corresponding to an original marker signal. The mixers used for upconversion and
Downconversion were different because they were designed for these specific frequency translation applications. The down-converting carrier frequencies used a 5 MHz clock as the PLL reference frequency. The 5 MHz clock was generated away from the MRI scanner with a 10 MHz crystal oscillator (CXOH20, Crystek Corporation) and a divide-by-two counter (SN74LS92D, Texas Instruments), and it was delivered to the receiver unit with a coaxial cable. This was the same reference clock strategy used for the SSB receiver in Chapters 2 & 3. The demodulated signals were filtered with a 5th order bandpass Chebyshev filter centered at the Larmor frequency. The recovered tracking signals were then fed directly into the MRI receiver ports for signal processing and marker localization.

Figure 5.2 – Overview of SSB-Based Wireless Active Tracking Array: (a) Each coil’s signal was amplified, mixed with a unique carrier, filtered, and combined for transmission. The combined wireless signal was received and distributed to receiver boards that recovered the signal from one of the original coils. (b) The carriers were produced using a PLL/VCO combination programmed by a local microprocessor. There was a local clock controlling each individual PLL.
When SSB is used for conventional imaging, the MR image is encoded in both phase and frequency information, so it is important that both the up-convertting and down-convertting carriers are synchronous in both phase and frequency. This requires a master clock signal to be transmitted between the wireless transmitters and receivers. However, in the application of active tracking, the fiducial markers' spatial coordinates are contained in only the projection's frequency information, so instead of transmitting a shared clock signal, separate asynchronous reference clocks can generate carriers that have the same frequency but arbitrary phase. However, small differences between these two frequencies can affect the accuracy of the localization. These differences can be separated into two components: fixed frequency offset and frequency drift. Fixed frequency offset between the two carrier frequencies adds a fixed error in the encoded frequency information. This fixed error can be minimized though manual adjustment of the clocks during device construction. This will also appear as an offset in the position of a fiducial marker, which will typically be naturally taken into account by the user when positioning the device. Frequency drift between the carrier frequencies causes an additional random error in the projection's frequency information. The extent of this random drift error depends on the quality and stability of the clocks in both the wireless transmitter and receiver. Therefore, the desired overall accuracy of the scan plane updates must be taken into consideration when selecting the clock signal sources. On the bench, the extent of frequency drift was observed with a spectrum analyzer (CXA series, Agilent). A tone of 63.6 MHz was fed into each channel of the tracking device, and the maximum drift of the recovered tone was recorded over a two minute period using the marker tools on the spectrum analyzer. Following this measurement, a rare earth magnet
was placed close to the prototype to determine if a strong magnetic field introduced any additional drifting to the recovered signal.

5.3.2 Active Fiducial Marker Tracking

The paradigm for active fiducial marker tracking used in this work was introduced originally by Dumoulin (40). The algorithm was based on Siemens interactive real-time sequence with active tracking (97). A series of projections (FA = 5°, N = 512, FOV = 400 mm, BW = 400 Hz/Px) were acquired in three spatially orthogonal directions, and the encoded signals from three fiducial markers were collected wirelessly. The spatial locations of the markers were determined by the location of maximum signal in each projection and channel. This assumes that each tracking coil collects a clear maximum signal from only one of the fiducial markers. The positions of the fiducial markers were compared to the known geometry of the device, and used to calculate both position and orientation of the device. This information was fed back to the MR control console, and a new imaging plane was updated in relationship to the device location for the subsequent image acquisition. Selection of the new imaging plane is flexible, and for this work a plane was selected through all 3 markers (details below). Other possibilities for future work include varying the slice center offset relative to the device and varying the slice orientation. The time required to acquire all three projections, calculate marker positions, send feedback to the MR control console, and calculate the new imaging plane was approximately 30 ms. Other projection implementations, such as the Hadamard sequence for off-resonance correction (40), were not used in this work, but could be included in future realizations.
5.3.3 Prototype Construction

Figure 5.3a is a picture of the wireless active tracking prototype constructed for a Siemens Espree 1.5 T system. The device was designed for hand-held use with dimensions 12.7 x 3.5 x 6.0 cm$^3$, where the individual transmitter boards (12.7 x 3.5 cm$^2$) were placed 2 cm apart inside the device. The fiducial markers were small vitamin E spheres with a 0.75 cm diameter. Each fiducial marker was positioned at the top of an active tracking coil. The coils were made into a gradiometer with two adjacent counter-wound four-turn 0.65 cm diameter solenoids in series. This design was chosen because it ensured that the coils always received signal regardless of their orientation in the magnet. In addition, the coil's sensitivity range would be more restricted to only local signals when compared to conventional solenoids, which reduced the possibility of signal interference from the patient or operator's hand. The three coils were tuned and matched to 63.6 MHz. The carrier frequencies were chosen to be 925, 940, and 955 MHz, which corresponded to SSB-modulated tracking signal frequencies of 861.4, 876.4, and 891.4 MHz respectively. These frequencies were spaced 15 MHz apart to avoid interference from preamplifier noise (43) and centered at the transmission bandwidth between the two antennas (which was determined on the bench using a network analyzer $S_{21}$ measurement). Non-integer multiples of the 10 MHz reference clock were possible due to an integrated reference divider in the integer-N PLL. This reference divider had phase instabilities present that made it unsuitable for standard imaging applications in Chapters 2 & 3 (42), but no deleterious effects from these instabilities were observed when used with projection magnitudes. A connection for DC power was available so that either rechargeable batteries or an external power supply could power the prototype.
The wireless receiver base station is pictured in Figure 5.3b. This is the same eight channel SSB wireless receiver that was presented in Chapter 3. Three of the receiver channels were programmed to the prototype's carrier frequencies to receive the modulated signals. The receiver unit was placed at the end of the patient bed, resulting in a wireless transmission distance of approximately 80 cm. Bench measurements suggested that about 30 dB path loss was experienced by the wireless signals with this transmission distance.

Figure 5.3 – Picture of Wireless Active Tracking Prototype: (a) The picture of the wireless active tracking prototype shows the positions of three fiducial markers (labeled by the numbers.) (b) A close-up photo of the active fiducial marker shows the fiducial marker mounted on the coil. (c) The wireless signals were received by the general-purpose eight channel receiver constructed for SSB-based MRI.
5.3.4 Bench Characterization

The previous chapters using SSB-based wireless MRI arrays have required a thorough characterization of the wireless system in order to show that the MR signal was transmitted with minimum degradation to the original SNR. However, in the application of device localization, the projections of the active fiducial markers have a very high SNR, and the encoded spatial coordinates in those projections can be extracted at very low SNR levels. Therefore, the SNR preservation and dynamic range requirements for the SSB system can be greatly reduced when compared to a normal imaging application.

The SNR preservation of the tracking prototype's SSB system was characterized with the same measurements performed in Section 2.3.2. Please refer to that section for further details on the measurements for the predicted system noise figure (NF_{sys}), carrier phase noise and reference spurs, and the input-referred 1dB compression point (P1dB_{input}).

5.3.5 Wireless Localization Measurements

The performance of the tracking prototype was characterized in the scanner by measuring the precision of localization. By measuring precision, the errors introduced by the wireless system which would cascade into the inherent accuracy of projection-based localization were observed. The tracking prototype was placed near isocenter, and 100 sets of sequential projections were acquired with the device stationary in order to capture the full range of drifting error between any of the clocks. The maximum variation between measured coordinates was recorded.
5.3.6 Needle Guidance Phantom Test

The performance of the wireless tracking prototype was demonstrated in a needle-guidance phantom test. The needle delivery target was a kumquat placed in a 473 mL gelatin phantom. The phantom container was opaque, so sight of the target was completely obscured. Within the MRI bore, a researcher manipulated the wireless tracking prototype with one hand while operating a biopsy needle (MR2015, E-Z-EM Inc.) with the other hand. The prototype was used to guide the needle to the target and deliver 0.5 mL of gadolinium-based contrast. A GRE imaging sequence was used (TE = 5.5 ms, TR = 10 ms, FA = 15°, BW = 250 Hz/Px, TH=10 mm) with a 350 x 262.5 mm² FOV and 128 x 96 matrix size. Images were collected with two elements from the patient bed spine array and reconstructed with sum-of-squares. An in-room monitor was positioned at one end of the magnet bore so that the operator could view the acquired images during the procedure. The wireless tracking prototype was battery powered during the entire acquisition.

Using the prototype's position and orientation to define a scan plane is flexible, but to evaluate the device operation. The slice center was positioned approximately at signal source #3 with the phase encode direction created by sources #1 and #3, and the slice plane included all three fiducial markers. This allowed for a simple visual analysis of the acquired images to confirm accurate device localization and scan plane updates.
5.3.7 Demonstration with Human Volunteer

Operation of the device guidance was also demonstrated with a human volunteer. The imaging sequence used was a FLASH sequence: TE = 5.2 ms, TR = 9.4 ms, FA = 25°, FOV = 300x300 mm², and a 96 x 128 matrix size. Images were collected with a 32 element cardiac array. Due to limited channel availability on the MRI scanner, only 24 channels collected the imaging signal and the right posterior coil group was not utilized. Instead of being centered at the device, the center of the field of view was chosen to be 150 mm distal from the device in order to place the body near the image center.

5.4 Results

5.4.1 Bench Measurements

The SNR preservation of the prototype's SSB system was characterized by evaluating the increase in the system noise figure. The coil preamplifiers had 27 dB gain and 0.5 dB noise figure. The SSB transmitter electronics had an average 18 dB gain and 9.3 dB noise figure, and the receiver modules had an average 7 dB gain and 9.1 dB noise figure. Using these values with Equation 2.15, it was possible to calculate SNR preservation and NF_{sys} for different levels of wireless path loss. Under the normal operating conditions of 30 dB path loss, the overall system noise figure was only 1.45 dB. This corresponded to only a 20% drop in SNR when compared to a wired connection directly from the preamplifier. For a substantial amount of path loss (45 dB), the overall system noise figure became 9.72 dB, corresponding to 88% loss in SNR. These values would have varied slightly as the orientation of the device changed in the magnetic field.
(98). However, the active fiducial marker projections had such high SNR initially that localization still occurred reliably in the presence of these losses.

Multiple dynamic range measurements were also performed. The measured carrier phase noise for all carriers and frequency offsets never exceeded -90 dBc/Hz. The maximum spurious signal in the carriers was -55 dBc. The average P1dB was -54 dBm input-referred to the coil preamplifier. This value was limited primarily by the combination of both the up-converting mixer and the amplifier immediately following it. When using Equation 2.16 and a 20 kHz signal bandwidth, the SSB system had 76.5 dB linear dynamic range. These dynamic range measurements demonstrate that the SSB system was able to encode and transmit projection signals with very high SNR.

On the bench, the maximum frequency drift observed with the spectrum analyzer never exceeded 400 Hz. No additional drifting was observed when a strong rare earth magnet was placed in close proximity to the prototype. Since the tracking projections had a bandwidth of 400 Hz/Px, the errors caused by drift in marker coordinate calculation should never be more than 1 coordinate location or 0.78 mm (400 mm / 512 points). This error was confirmed when marker locations were measured with the stationary prototype. In the sets of 100 projections, the acquired spatial coordinates varied no more than 1 coordinate location. If the projections were acquired with a lower bandwidth, the SNR for the tracking projections would have increased even further, but at the expense of possibly increasing drifting errors to above 1 pixel.
5.4.2 Imaging Results

Figure 5.4 is an example tracking projection collected with the tracking device. In this example, the coil face was oriented perpendicular to $B_0$. Even at different coil orientations, no failure in peak detection was observed in this work. The combination of a fiducial marker with a high signal density and a coil with a highly localized sensitivity creates high SNR projections with easily identifiable maximum amplitudes.

![Graph showing tracking projection](image)

**Figure 5.4 – Example Tracking Projection:** An example tracking projection demonstrates that wireless transmission preserves the acquisition's high SNR, allowing for easy identification of the maximum amplitudes. In this particular projection, SNR was estimated to be more than 2000 by dividing the peak amplitude with the standard deviation of the leftmost 75 data points.

In total, 150 images were collected during the phantom needle guidance procedure. The spatial resolution of the frames has been increased four-fold through bicubic interpolation to more closely mimic the images displayed by the in-room monitor. Figure 5.5 presents selected frames from the procedure. First, the target was visualized with coronal slice positions (Figure 5.5a and 5.5b), where the operator's right arm is also visible. Then, the wireless tracking prototype was repositioned to view the target at a new angle and allow for better needle guidance and visibility (Figures 5.5c and
The needle was then delivered to the target (Figures 5.5e and 5.5f), contrast was injected (Figure 5.5g), and the needle was removed (Figure 5.5h). Finally, the target was visualized at different viewing angles to confirm contrast delivery. Figure 5.5i, which has the same image orientation as Figure 5.5b, shows the change in target contrast (along with views of the operator's arms and neck). Figure 5.5j has another slice orientation that captures the complete needle trajectory along with the contrast-filled target. The presence of the tracking markers in each image validates the operation and scan plane tracking accuracy of the wireless prototype.

Figure 5.5 – Images from Phantom Experiment: A selection of images collected during the needle placement procedure demonstrates accurate operation of the wireless prototype and successful guidance.
Figure 5.6 contains selected frames from the human volunteer procedure. As with the phantom images, the spatial resolution of the images have been increased four-fold through bicubic interpolation. In the axial images (Figures 5.6b and 5.6c), the absence of the right posterior coil array is noticeable through the loss in image quality. The cardiac receiver array was slightly sensitive to the signal from the fiducial markers. This caused the markers to alias into the image while the markers were actually outside the FOV. This aliased signal is seen at the bottom of the axial images. Despite these drawbacks, different anatomical features were visualized through the guidance of the device.

![Figure 5.6 – Images Collected from Human Volunteer Experiment: A selection of images collected during the human volunteer demonstrate device operation.](image)

### 5.5 Discussion

The results of this Chapter demonstrate that active tracking with a wireless coil array is possible. A wireless coil array with a simplified SSB design was used to localize marker coordinates with an accuracy of 0.78 mm. Successful operation of the wireless tracking prototype was demonstrated in a gelatin phantom, where a needle was guided to a target and an injection was delivered. Successful device operation was also
demonstrated with a human volunteer. It is clear that a plethora of other more sophisticated wireless designs could be used to achieve similar results, but the goal of this work was to demonstrate that even a basic wireless design could satisfy the relatively simple requirements for active tracking. In particular, transmission of a frequency reference signal was not necessary, so it was not included in the design. Wirelessly transmitting this reference would have added additional complexity to the SSB design, and it would have provided another opportunity for device failure.

In this realization, the wireless tracking prototype was separate to the needle insertion device. This required both the user's hands to operate the prototype and wield the needle. The advantage to this setup is that it gives an interventional radiologist a flexible way to control the imaging plane independent of the interventional device. Alternatively, the wireless active tracking fiducial markers could be integrated directly into the surgical device, which would then require only one hand to perform the intervention.

The wireless tracking prototype could benefit from further design improvements beyond solely improving the wireless electronics. An interface, placed locally on or near the device, could provide intuitive controls to toggle between scan plane orientations and positions. Incorporating a wireless gyroscope and accelerometer would allow for the device's orientation and position to be tracked outside the linear region of the gradient fields (99). Hybrid systems that include both optical and signal-based tracking have been shown to be effective in performing image-guided interventions immediately outside the MRI bore (100).
While not experienced in this preliminary study, wireless interference effects like multipath distortion could cause one of the wireless fiducial marker signals to experience very large amounts of path loss, and this would cause the received signal to fall below the receiver's electronic noise floor. This would disrupt localization and cause the tracking sequence to fail. A simple consistency check could be built into the scan plane adjustment to allow for immediate reacquiring of the projection set. Since destructive interference is typically narrowband and very sensitive to particular transmission distances, consecutive failures would be unlikely during actual operation, and this would cause minimal disruption to the real-time acquisition. If failure had occurred too frequently, the robustness of localization could have been improved by using additional receiving antennas, additional wireless fiducial markers, or an altogether different wireless platform that is less prone to destructive interference (for example, digital wireless transmission (27)).

Device localization with a wireless active tracking array does not only have applications in needle guidance. It could also prove beneficial in other active tracking applications, such as motion correction (101) and intravascular device tracking (102). But, simple wireless guidance devices, such as the one presented in this Chapter, are where wireless detector arrays will have immediate clinical relevance.
Chapter #6

Summary, Future Directions, & Conclusion

6.1 Dissertation Summary: Historical Reflections and Significance

The studies presented in this dissertation summarize nearly seven years of research. In this section, a brief historical summary of this research is presented. This perspective allows the individual accomplishments of this dissertation to be highlighted and reveal its significance to the field.

At the onset of this research, wireless detector array technology was still an emerging topic in MRI engineering. With the growing clinical demand for highly parallel arrays, major manufacturers were beginning to show interest in developing their own solution for wireless arrays. This trend was evident at the annual ISMRM scientific meeting in 2005, where the topic of the MR Engineering Study Group was titled "Wireless and Fiber Optic Cabling". At the time of this meeting, only a few preliminary studies with non-coaxial technology had been presented, and these only demonstrated operation of a single wireless channel on the bench (29,103).

Chapter 2 presents the development of a single channel wireless transmission system using SSB. While operating locally inside the MRI magnet bore, this system preserved nearly 100% of the original image SNR for signal dynamic ranges up to 80 dB. When originally presented in 2009 (41), this was the first non-coaxial transmission system used successfully inside the magnet bore positioned immediately adjacent to the
coil. At the time, the other investigated non-coaxial technologies utilized electronics that prohibited operation in a large magnetic field (22,27,31,103). The analog-based SSB technology in Chapter 2 did not introduce observable ambient noise to the MR signal environment, nor did it introduce susceptibility artifacts from excessive magnetic material. Equipping the SSB system with a local microprocessor also eliminated the need for additional cabling to control the system.

Chapter 3 presents a study into the interference sources present in SSB-based wireless MRI arrays (42,43). After finishing development of a single SSB channel, the next step was to demonstrate the SSB system transmitting multiple channels simultaneously. Non-coaxial multichannel transmission had only been demonstrated with four channels with optical technology (23,25). At that time, there was a general assumption that wireless systems would have scalability limited by only the bandwidth of the MRI signal, which would have allowed for a dense transmission spectrum containing many modulated signals. The content in Chapter 3 demonstrated that this is not necessarily the case. Interference sources were identified in SSB-modulated MRI signals, specifically the preamplifier noise and carrier bleed-through. This study demonstrated how to mitigate these interference sources through simulation and proper signal filtering. Two separate SSB systems were presented where the first system had a minimally-complex SSB design and the second had filtering designed specifically to reduce these interference sources. When transmitting eight channels simultaneously with a 10 MHz channel spacing, the first SSB systems preserved 88% image SNR on average, while the second system preserved 97% image SNR. This study was the first publication that demonstrated simultaneously wireless transmission of multiple MRI signals. In addition,
while only eight channels were demonstrated, the simulation method provides a means to design future SSB systems, where many additional channels could be transmitted while maintaining high levels of SNR preservation.

Chapters 4 & 5 presented two additional clinical applications for wireless detector arrays, both outside their traditional role of acquiring diagnostic images (44,45). Both clinical applications particularly benefited from the wireless nature of SSB-based transmission. And, because multichannel wireless MRI technology did not exist until recently, neither application had been previously proposed.

In Chapter 4, a four channel wireless magnetic field monitoring system (wMFM) was constructed using SSB. An array of small wMFM transceiver probes could be incorporated into standard manufactured coil arrays and provide a measurement of gradient field dynamics concurrently with the normal image acquisition. The wMFM system was used to measure first order gradient field dynamics in ultrafast two-dimensional spiral acquisitions (44). When compared to other accepted measurement techniques, the wMFM measurement had residuals no larger than 0.55 Δk in normalized k-space trajectories. The images reconstructed with the wMFM measurement showed no visible signs of artifacts or geometric distortions. This was the first publication to show magnetic field monitoring results collected with wirelessly transmitted probe signals. More generally, this was the first publication presenting a MRI transceiver array used in conjunction with wireless technology. Because of the power requirements for RF excitation, a completely wireless transceiver system will be realizable only in applications where small amounts of signal need to be excited. Additional wireless technology was also developed in this chapter to transmit both a reference clock and a RF
excitation trigger. While these technologies were useful to the wMFM system, they could be applied to many other wireless array applications. For example, the wireless reference clock technology could eliminate the clock cable needed by the SSB array in Chapter 3. This would have allowed that array to become truly wireless.

In Chapter 5, a handheld device with an integrated three channel wireless array was constructed, and the device was used to control the scan plane during an image-guided procedure (45). The wireless detector array permitted the device to be automatically localized with the MRI scanner through multiple orthogonal projection acquisitions (40). Device localization is a well-understood topic in MRI, and the majority of other research groups have focused on improving the accuracy of the localization (104). In this study, a different approach was taken, which emphasized the simplicity of device localization. Device localization only required the detection of peak amplitudes in received signal spectra. Therefore, the typically rigorous requirements for wireless transmission were significantly reduced, including SNR preservation and phase coherence. This allowed for the construction of an entirely wireless device that had multiple benefits over the existing technologies. These benefits include the wireless nature of the device, the ability to freely wield the device inside the MRI magnet bore, and easy system integration. Needle guidance was demonstrated in a phantom experiment where a needle was guided to a target and contrast was delivered. This study showed that wireless detector arrays have a promising application in interventional devices, and only a relatively small amount of engineering effort would be required to create a viable product.
Now, at the completion of this dissertation research, manufacturers have already begun making design decisions regarding their next generation of MRI receiver technology. For example, Philips Healthcare recently released their "dStream" system (47), where individual analog-to-digital converters are placed locally near the coil preamplifier and the digital signals are transmitted optically. Also, Siemens Healthcare has implemented a form of frequency domain multiplexing in their receiver systems (15), where signals from a pair of coils are multiplexed onto a single receiver cable. While not wireless technologies, both of these are examples of coaxial cable reducing technologies, and they demonstrate the trend towards wireless. As these manufacturers move forward in their next design decisions, the studies presented in this dissertation will serve as an excellent guide in the development of a wireless platform for MRI detector arrays.

6.2 Future Directions - Improved Scan Plane Control Device

There are multiple opportunities to further develop the wireless applications presented in this dissertation. It is clear that multiple improvements can be made by incorporating sophisticated technology from the wireless communication fields. For example, the newly released ADF4351 from Analog Devices is an integrated fractional-N PLL and VCO that can output frequencies between 35 to 4400 MHz with phase noise better than -90 dBc/Hz (105). Or, many of these devices could be substantially miniaturized, as was recently done with an integrated downconverting system was recently presented for intravascular catheter tracking, where an of 1 x 2 x 0.74 mm³ integrated chip contained a tracking coil, preamplifier, mixer, PLL, and VCO (106). Many of these improvements have already been acknowledged in each chapter’s
discussion section. In this section, rather than identifying many of these incremental technological improvements, an advanced interventional scan plane control device is proposed. Many of the improvements in this device are derived from features developed throughout this dissertation.

A prototype version of the new device is pictured in Figure 6.1. Instead of incorporating the fiducial markers and tracking coils inside a device, they are placed on the individual finger tips of the operator. The coil design has been modified such that the orthogonal solenoid coils connected in series are wrapped around the fiducial marker. This coil design reduces the active marker size sufficiently so that it can be mounted on the finger. Short cables connect the coils to a wireless transmitter mounted on the operator’s wrist. The device is powered by small lithium ion batteries that are now available in completely non-magnetic packaging (107). This device would give the physician unique scan plane control that will likely be more intuitive than the original handheld device.
Figure 6.2 shows a simplified schematic of the transmitter electronics for advanced device. The SSB design is a blend of the two SSB systems presented in Chapters 4 & 5. A wideband low-noise amplifier (ERA-3SM, Minicircuits) is used, which allowed the device to be powered with a 5 V source rather than its previous 10 V source. This allows the device to be powered by two lithium-ion batteries. A fractional-N PLL (ADF4153, Analog Devices) is used to allow for arbitrary channel spacing. Two low-powered amplifiers (RF2314, RFMD) amplify the modulated signal. A quarter-wave monopole antenna (Linx Technologies, CHP Series) with an omni-directional radiation pattern is used. When compared to the original interventional device presented in Chapter 5, this improved design has reduced power consumption by 33% (400 mW) and increased gain by 10 dB (59.5 dB).

Figure 6.1 – Picture of New Scan Plane Control Prototype: (a) When compared to the original device, the new SSB modulation board has reduced board size (65x42 mm²), increased gain (59.5 dB), and decreased power consumption (400 mW). (b) The three modulation boards are powered by two non-magnetic lithium ion batteries. (c) The device is shown mounted on a wrist, where short cables connect to fiducial markers mounted on the operator’s fingers.
By placing the fiducial markers on fingertips, the markers are no longer in fixed positions and these new degrees of freedom can be utilized for additional forms of scan plane control. Switch-based controls can be realized by monitoring the spatial proximities of the individual fiducial markers, and these switches can control different image acquisition parameters. For example, moving the index finger close to the thumb can switch through predefined FOV shifts for the acquired image. Fiducial markers can be placed on more than three fingers, leaving markers that can be devoted entirely to controlling the acquired image.

Figure 6.2 – Schematic of New Scan Plane Control Prototype: (a) The electronic design of the new tracking device is similar to its previous design, but it does have improvements. In particular, only two batteries are required to power the device. (b) The carrier generation for the new tracking device still utilizes a local clock as a frequency reference, eliminating the need for an external wired source.

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The wireless reference technology presented in Chapter 4 could also be beneficial to the new design. The original design eliminated the need for reference transmission by using a local clock and asynchronous carriers. This lack of synchronization slightly reduced the accuracy of device localization. By utilizing a wireless reference, the accuracy of device localization could be improved while still maintaining an entirely self-contained device. The improved accuracy would allow for finer scan plane placement, which may become necessary in certain interventional procedures.

The prototype could also take advantage of incorporating the transceiver technology developed in Chapter 4. Local excitation helps prevent interference from the operators hand or other fiducial markers, and it could be accomplished with either a shielded probe (37) or a heteronuclear probe source (38). In addition, the fiducial marker and coil could be substantially shrunk in size similar to the probes developed for wMFM. Such a small probe size would allow the entire apparatus to be comfortably worn underneath a sterilized surgical glove. The primary drawback to local excitation is the increased power consumption, but this can be lessened by reducing the excitation power through the smaller active marker design and by distributing many of the redundant electronics from just a single source (eg. only have one amplified excitation source, wireless trigger, and wireless reference).

Wireless interventional scan plane control devices, such as the one proposed here, have immediate potential for clinical impact. And as demonstrated in this dissertation, the device's reduced wireless transmission requirements make its design simple in both cost and engineering hours. A suitable wireless receiver could be easily incorporated in the next generation of modern MRI scanners. And once a wireless
receiver system is built into the MRI system, the path should be opened for many additional wireless applications to be developed.

6.3 Future Directions - Highly Parallel Wireless Phased Arrays

As described in Chapter 1, the original motivation for developing wireless receiver array technology was to eliminate the large number of cables needed in highly parallel phased arrays. While this dissertation only presented arrays with a relatively small channel count, the lessons learned from these studies will enable future researchers to accomplish this goal. The simulation results presented in Chapter 3 demonstrated that one of the primary limitations in preserving SNR is controlling the interference between multiple channels in a given antenna bandwidth. While interference can be reduced by improving the noise filtering of SSB-modulated signals, the MR signal bandwidth will eventually become the limiting factor in channel count. Instead, it would be wise to take advantage of advanced communication technology already developed for the super-high frequency band (3 to 30 GHz). Since antenna bandwidth is typically a percentage of the antenna center frequency, a higher channel count can be simply realized by using a high frequency band and therefore larger antenna bandwidth. Please refer back to Figures 3.9 & 3.10 for simulation results on antenna bandwidth, interference, and channel count. Large channel availability will become even more important in the future as additional wireless MRI devices are incorporated into modern scanners (like the tools developed in Chapters 4 & 5). Future research should look into developing new transmission systems at the higher frequency bands like the 5.8 GHz ISM frequency band and further explore
this topic. This would be a necessary step in attaining highly parallel wireless phased arrays and the wide-scale integration of wireless technology in future MRI scanners.

6.4 Future Directions – Alternative Modulation Strategies

Before presenting the conclusion to this dissertation, this section will briefly reflect on the lessons learned about SSB and its potential use in highly parallel phased arrays. SSB was investigated originally because it could achieve nearly 100% SNR preservation while using a wireless transmission bandwidth no wider than the MRI signal. This small transmission bandwidth requirement would have theoretically allowed for many modulated signals to be transmitted simultaneously while maintaining high SNR preservation. In practice though, multi-channel SNR preservation proved difficult to achieve, which is detailed in Chapters 2 & 3. If a future researcher were to begin developing a next-generation wireless platform for highly parallel phased arrays, the results from this dissertation suggest that exploring an alternative wireless modulation technology may be preferable.

Amplitude modulation techniques, like SSB, are very sensitive to the effects of path loss. As described in Chapter 2, excessive path loss can create large amounts of SNR degradation. In this dissertation research, the antenna transmission distance was maintained at 80 cm to ensure that the modulated signals experienced less than 30 dB path loss across a fixed antenna bandwidth. While not experienced in this dissertation, narrowband interference effects, like multi-path distortion, could have created very large amounts of path loss despite having the relatively short transmission distances. These interference effects make it difficult to robustly satisfy a path loss requirement in highly
parallel phased arrays. Interference effects can be addressed by increasing the gain in the wireless transmitters, but this solution only reveals the other major drawback to amplitude modulation techniques. Increasing the gain (and therefore output power) in a high DR $P_{1\text{dB}}$ transmission system is very power inefficient. For example, the wideband amplifier ADA-4743 from Avago Technologies has a maximum 50 mW output power ($P_{1\text{dB output}}$) but it consumes 230 mW. This power inefficiency is quite large, and it makes long-term battery operation less feasible. Therefore, simply increasing the gain of the transmitter is an unsuitable solution for addressing path loss and interference effects in amplitude modulation.

An alternative solution to addressing path loss would be utilizing an entirely different modulation technique that is less prone to the effects of path loss. This would allow for high SNR preservation while also having more efficient transmitter operation. To this end, frequency modulation (33) would be a good technique for future research to investigate. When compared to amplitude modulation, frequency modulation is much more robust to the effects of path loss, but it has the drawback of requiring a transmission bandwidth much wider than the MRI signal. In other common frequency modulation applications, the modulated signal bandwidth is truncated for multi-channel transmission. This truncation is done at the expense of some loss in SNR preservation, but this inherent SNRR loss may prove more advantageous than dealing with the pitfalls of amplitude modulation and path loss. Future research should investigate the feasibility of utilizing frequency modulation with multiple MRI signals. If the SNR preservation and bandwidth usage prove sufficient, it may very well be a more practical wireless solution for highly parallel phased arrays.
6.5 Conclusion

This dissertation has demonstrated the feasibility of realizing wireless MRI detector arrays with single sideband amplitude modulation (SSB). The utility of wireless MRI detector arrays was demonstrated in three clinical applications: diagnostic imaging, gradient field measurement, and interventional guidance. By adapting sophisticated electronics from the communications field, robust scalable wireless systems can be developed for many MRI-based applications including highly parallel wireless phased arrays. Simple wireless devices, such as one constructed for image-guided needle placement, can have immediate clinical impact without substantial engineering refinement. This dissertation is only the foundation for many new and exciting wireless applications that will be developed in future MRI systems.
Appendix A: Circuit Schematics

Appendix A contains a selected set of the electrical schematics that have been described in this dissertation. The schematics were created using EAGLE PCB design software by CadSoft (www.cadsoftusa.com). This program has a basic version that is available online for free to students. Therefore, if one needs further details regarding these schematics, please download the EAGLE software package to view the schematic files. All the files are located in the back-up files of Matthew Riffe. They are located in his "Eagle" folder. Please contact Matthew at matthew.riffe@gmail.com if you have any questions.
SSB Transmitter Module: SSB System #1
SSB Transmitter Module: SSB System #2 (Part A)
SSB Receiver Module

The SSB receiver module schematic was used in each of the applications described in this dissertation, including the two SSB systems, wMFM, & the handheld tracking device.
**wMFM Transceiver Module (Part A)**

The circuitry for the wMFM probe preamplifier & SSB modulation are not shown due to their similarity with the schematics presented in SSB Systems #1/#2.
wMFM Transceiver Module (Part B)

The circuitry for the wMFM probe preamplifier & SSB modulation are not shown due to their similarity with the schematics presented in SSB Systems #1/#2.
wMFM Wireless Reference Receiver

Power regulation components not shown for simplicity.
wMFM Wireless Trigger Receiver

Power regulation components not shown for simplicity.
Handheld Tracking Device Module

Single transmitter module is shown.
Bibliography


~ 171 ~


50. Carson J. US Patent 1,449,382: Method and means for signaling with high-frequency waves. 1915;


~ 175 ~


75. FMRIB. FSL FMRIB Software Library, http://fsl.fmrib.ox.ac.uk/fsl/fslwiki/. 2012;


98. De Zanche N, Roberts B, Fallone BG. Variation of Preamplifier Noise Figure With B0 Field Strength. Proceedings of the 18th Scientific Meeting of ISMRM, Stockholm, Sweden 2010;3916.


