A 4-channel, vector network analyzer microwave imaging prototype based on software defined radio technology

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ABSTRACT

We have implemented a prototype 4-channel transmission-based, microwave measurement system built on innovative software defined radio (SDR) technology. The system utilizes the B210 USRP SDR developed by Ettus Research that operates over a 70 MHz–6 GHz bandwidth. While B210 units are capable of being synchronized with each other via coherent reference signals, they are somewhat unreliable in this configuration and the manufacturer recommends using N200 or N210 models instead. For our system, N-series SDRs were less suitable because they are not amenable to RF shielding required for the cross-channel isolation necessary for an integrated microwave imaging system. Consequently, we have configured an external reference that overcame these limitations in a compact and robust package. Our design exploits the rapidly evolving technology being developed for the telecommunications environment for test and measurement tasks with the higher performance specifications required in medical microwave imaging applications. In a larger channel configuration, the approach is expected to provide performance comparable to commercial vector network analyzers at a fraction of the cost and in a more compact footprint.

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I. INTRODUCTION

Microwave imaging has been investigated for several decades as a potential medical tool for detecting/diagnosing a variety of indications including breast cancer (Poplack *et al.*, 2007; Meaney *et al.*, 2013; and Klemm *et al.*, 2008a), cardiac conditions (Semenov *et al.*, 2000), stroke (Persson *et al.*, 2014), bone health (Meaney *et al.*, 2012a), and temperature during thermal therapy (Meaney *et al.*, 2003b; 2003a; and Haynes *et al.*, 2014). In these applications, the technique exploits the considerable dielectric property contrast between normal and diseased tissues (Joines *et al.*, 1994; Chaudhary *et al.*, 1984; Surowiec *et al.*, 1988; Lazebnik *et al.*, 2007a; 2007b; Sugitani *et al.*, 2014; Meaney *et al.*, 2012a; Semenov *et al.*, 2002; Persson *et al.*, 2014; and Gabriel *et al.*, 1996a) and property

variation as a function of temperature (Duck, 1990; Lazebnik *et al.*, 2006; and Ohlsson and Bengtsson, 1975). Microwave imaging offers associated benefits, namely, nonionizing exposures and relatively low cost. While some microwave imaging systems have advanced to phantom and animal experiments (Meaney *et al.*, 2003b; 2003a; and Ostadrahimi *et al.*, 2013) and even patient examinations (Poplack *et al.*, 2007; Meaney *et al.*, 2013; Klemm *et al.*, 2008b; Fear *et al.*, 2013; and Persson *et al.*, 2014), efforts have often been limited to numerical simulations (Shea *et al.*, 2010; Catapano *et al.*, 2009b; and 2009a) because of the high cost of vector network analyzers (VNA's) required to acquire the multichannel coherent microwave signals needed for medical imaging. More extensive reviews of the current state-of-the-art can be found in Nikolova (2014) and O'Loughlin *et al.* (2018).

Most medical microwave imaging systems are designed as either radar, holographic, or tomographic configurations and commonly acquire backscatter or transmission data (Meaney *et al.*, 1995; Semenov *et al.*, 2005; Hagness *et al.*, 1998; Sill and Fear, 2005; Gibbins *et al.*, 2009; Wang *et al.*, 2014; Tajik *et al.*, 2018; and Nikolova, 2014). Both designs require systems with high dynamic ranges and many channels to accommodate multiple antennas. Radar implementations leverage either backscatter- or transmission-measurements and require time domain data. Two prominent approaches to acquiring the information are either (a) to generate an actual pulse and transmit and receive time-domain data or (b) to acquire frequency domain data over a wide bandwidth and synthesize the recordings into the equivalent time domain pulse. For the former, prototype systems designed by Zeng *et al.* (2014), Porter *et al.* (2013), Kubota *et al.* (2014), Fasoula *et al.* (2018), Marimuthu *et al.* (2016), and Bialkowski *et al.* (2016) have been produced and tested. The Zeng *et al.* system generates a 100 ps pulse (translating to a 3.5 GHz bandwidth centered at 3.0 GHz) through connectorized components and records transmission data. Multichannel designs based on monolithic chip technology are planned with the goal of reconstructing 2D and 3D wideband dielectric property maps. Porter *et al.* have designed a 16 channel system based on a commercial pulse transmitter and a switching network to implement a multichannel imaging system with a bandwidth of 2–4 GHz. Kubota *et al.* have developed monolithic transmit and receive channels using CMOS technology with integrated bent dipole antennas having an effective bandwidth of 5.9 GHz centered at 10 GHz. A prototype system utilizing reflection measurements has been reported (Song *et al.*, 2017). The Fasoula *et al.* system uses 18 Vivaldi antennas circumscribing the target of interest and incorporates network analyzer-based technology and a switching network to configure the array for transmission measurements. Marimuthu *et al.* have devised a software defined radio (SDR) based system which utilizes a single antenna that is mechanically moved about the target. Measurements are performed over a wide bandwidth in the frequency domain and subsequently transformed to a pulse in the time domain. Bourqui *et al.* (2012) have developed a reflection based measurement system consisting of a single antenna which is also rotated about the target. Conversely, Byrne *et al.* (2017) reported a multichannel system that consists of conventional VNA's coupled to a micromechanical switching array for excellent cross-channel isolation. This system has undergone several iterations and is currently in use clinically (Preece *et al.*, 2016). Conventional multichannel VNA's can be prohibitively expensive for most research groups and the task of building a custom system from commercially available components can also be expensive and time-consuming (Li *et al.*, 2004 and Epstein *et al.*, 2014).

With respect to holographic and tomographic system implementations, three of the more prominent have been described by Tajik *et al.* (2018), Poltschak *et al.* (2018), and Ostadrahimi *et al.* (2013). The former utilizes transmit and receive horn antennas placed above and below the compressed breast (cranial-caudal orientation), respectively, and raster scans the region of interest while collecting data between 3 and 8 GHz. The current implementation, tested on phantoms, is slow (6 h) but plans exist to upgrade the system to an electronically switchable array. The Poltschak *et al.* system consists of transmit and receive ports placed on the same circuit board using commercially available surface mount

components to realize high dynamic range near 1 GHz. The system acquires transmission data as input to a 3D inverse scattering algorithm. Ultimately, it will involve a 160 element array intended for imaging stroke victims. Recent implementations by Ostadrahimi *et al.* reconstruct dielectric property images from data acquired with a standard VNA and multichannel switching to produce multiview data of the target. This system has been tested with phantoms and some anatomical targets.

Design challenges include accommodation of signal attenuation across most biological tissues which increases as a function of frequency (Gabriel *et al.*, 1996b; 1996c). Thus, high dynamic ranges are required to achieve image resolution. Commercial VNA's are attractive for microwave signal measurements because they collect the coherent data needed for image formation—generally both amplitude and phase information (Keysight, 2016). Unfortunately, their dynamic range is often limited (typically reaching –100 dBm). More recently, several manufacturers, including Rohde and Schwarz (Munich, Germany) and Keysight Technologies (Santa Rosa, CA) have developed multichannel systems that measure signals below –140 dBm, but their prices are proportional to the number of channels and unit costs are typically in excess of \$100K.

An additional design challenge is signal corruption due to multipath propagation—an effect that is particularly detrimental to near field imaging (Meaney *et al.*, 2012b). Common problems include surface wave generation along the outside of feedlines and interfaces [e.g., between illumination chambers and coupling medium, and between the target and coupling medium (Trizna, 1997 and Gao *et al.*, 2007)]. Channel-to-channel leakage is another form of multipath signal corruption. Gating strategies have been proposed to filter unwanted signals in the time-domain, but these approaches have been difficult to manage in practice. Using a single transmit/receive antenna in a backscatter configuration with mechanical motion to collect data from a full complement of directions is a design that minimizes the number of antennas and their mutual coupling, as well as surface waves along support and feed structures (Fear *et al.*, 2013). Other approaches have exploited lossy coupling materials (Meaney *et al.*, 2003c and Micrima, 2016) to eliminate undesired signals, but at the cost of increasing dynamic range requirements for realizing a fully functional data acquisition system.

Modern software defined radios (SDR's) have potential to address the measurement challenges in microwave imaging with a compact and low cost form. The technology has developed rapidly—RF and microwave tasks can now be performed by sophisticated monolithic chips such as the AD9361 agile transceiver developed by Analog Devices (Norwood, MA), which covers multiple communications bands between 70 MHz and 6 GHz. In fact, Ettus Research (Santa Clara, CA) has combined this capability with a powerful field programmable gate array (FPGA) processor (Xilinx Spartan 6 XC6SLX150, San Jose, CA) resident on a single circuit board (USRP B210) with the necessary power supplies and filtering features that is easily powered and controlled by a remote laptop computer via universal serial bus (USB) connection. The RF module includes two channels which are functional in both transmit and receive modes. The package is essentially a heterodyne design where signals are fully synthesized for precise frequency operation in transmit mode; and the cascade of components includes a low noise amplifier, variable gain amplifier, mixer for downconversion with reference against a

synthesized LO signal, and a 12 bit A/D converter in receive mode. While the technology is geared primarily for telecommunication applications, it incorporates the RF componentry required to realize a sophisticated test and measurement device at low cost (~\$1100 per two-channel unit). By designing an integrated, multichannel system with the B210 as the critical building block, much of the design task is converted from hardware fabrication to software programming (Mitola, 1992).

SDRs have two shortcomings which make them unacceptable for microwave imaging alone: (1) limited dynamic range, and (2) unreliable coherence between multiple boards (Ettus Research, 2018). The former results from an overall system noise figure of 8 dB, which is suitable for radio communications but too high for medical microwave imaging applications, but is overcome through the addition of a low noise amplifier at the receiver. The latter is remedied by incorporating B210 boards along with a modest amount of external microwave circuitry to ensure multiboard coherence. The following discussion describes the techniques we have implemented to overcome these B210 board deficiencies and presents a design for a 4-channel prototype operating in microwave signal transmission/reception modes. The concept is fully scalable to much higher channel count without performance degradation.

II. METHODS

A. System design

The Ettus B210 SDR incorporates an impressive amount of technological capability in a relatively compact, low-cost, and easily integrated package. The centerpiece of each board is the agile Analog Devices transceiver (AD9361) which operates over a bandwidth of 70 MHz–6 GHz and offers two channels that each act in both transmit and receive modes. In transmit mode, a range of power level settings is allowed to optimize transmissions for a given application. In receive mode, 12 bit A/D conversion is available for rapid sampling in combination with a variable gain amplifier (VGA) in the receiver chain (i.e., just behind the low noise amplifier, LNA) which provides linear performance over a wide dynamic range. Signal detection is configured in a heterodyne design where the IF frequency can be specified up to 500 kHz with a corresponding maximum sampling frequency of 61.44 MS/s. While individual boards contain the basic features required for a test and measurement device, they have important limitations that prohibit them from acting alone as a complete subassembly for coherent, multichannel, high dynamic range data acquisition with high channel-to-channel isolation. For example, while the VGA in combination with the A/D board provides a high dynamic range, the maximum gain is insufficient because the minimum detectable signal is not restricted by the receiver noise floor but by the discretization limitation of the A/D converter. In addition, when synchronized in a multiunit configuration utilizing the Octoclock-G feature, the board-to-board coherence is unreliable. In microwave tomography, signal coherence is critical. Thus, exploiting the B210 SDRs demands a level of external microwave circuitry to compensate for dynamic range and signal coherence limitations.

Figures 1(a) and 1(b) show a schematic and photograph of the 4-channel breadboard system that can be scaled to the number of channels required in a microwave tomography system. It

FIG. 1. (a) Schematic and (b) photograph of the 4-channel microwave data acquisition prototype configured from three Ettus B210 SDRs.

consists of a dedicated B210 board as the transmitter and two B210s as receivers providing two channels per board, additional LNAs for improving the channel dynamic range, switching modules for channel selection between transmit and receive mode, a dedicated RF reference signal for coordinating the transmitter board with associated B210 receivers, and the OctoClock-G (Ettus Research CDA-2990, Santa Clara, CA—not shown) which provides a 10 MHz reference signal and a 1 Hz PPS clock signal. Each B210 has two channels for which Rx (receive) and Tx/Rx (transmit and receive) ports exist. A single-pole/4-throw (SP4T) switch (UMCC SR-J010-4S, Universal Microwave Components Corporation, Alexandria, VA) selects the channel designated as transmitter. Single-pole/double-throw (SPDT) switches (MiniCircuits

ZASWA-2-50DR+) allow each channel to serve as either transmitter or receiver, and accompanying single-pole/single-throw (SPST) switches (MiniCircuits ZFSWHA-1-20+ not shown) add isolation to minimize leakage directly from transmitter to receivers. LNAs (MiniCircuits ZX60-P105LN+) improve channel noise figures (NF) while also boosting RF signal levels to maximize A/D converter ranges in conjunction with the B210 internal VGAs. Because measured signals during breast imaging rarely (if ever) exceed -50 dBm, adding an extra 20 dB gain stage at receiver inputs does not increase risk of damaging receiver-board RF componentry.

While many of the B210 settings (e.g., gain levels and operating frequency) can be reconfigured rapidly during operation, switching between transmit and receive modes is particularly slow. Accordingly, the second transmitting port is used to send a reference signal to each B210 to maintain coherence between transmit signal and receiver LO signals. The transmission signal is split with a power divider (MiniCircuits ZAPD-30-S+) and feeds one of the transceiver ports on each of the receive B210s. The OctoClock provides both a 10 MHz signal (for signal accuracy) and a 1 Hz PPS clock for simultaneous triggering of all B210 boards. The design allows each channel to behave coherently in both transmit and receive modes and achieves a dramatically improved dynamic range that competes favorably with high-end commercial VNA performance in a compact and modestly priced package.

B. System coherence

A single, multiplexed transmit signal is directed to each antenna; hence, transmission is guaranteed to be coherent from all antennas. An identical signal is transmitted from the second channel of the transmit board as a reference and fed into a single transceiving port of each B210 board via a power divider (MiniCircuits ZAPD-30-S+) after a high isolation switch (three PE4246 SPST switches in series, Peregrine Semiconductor, San Diego, CA). Here, the transmit signal only needs to be phase-locked with one of the receiver channels on a single B210 board since the LO signals of the associated two channels are already synchronized to each other. Each channel uses its receiver port for dedicated signal detection. With both ports of the transmitter sending out signals continuously, the boards and switches are first configured to acquire data from the reference signal at the transceiving port of Channel B [see Fig. 1(a), high isolation switch set to ON] and determine amplitude and phase. Once this task is completed, the isolation switch is set to OFF and internal switches of Channel B are set to receive signals only through the associated receiving port, LNA, and antenna. Subtracting the reference amplitude and phase from these signals produces values that embed coherence between transmitter and receiver channels. This process is repeated for each receive B210 and for each frequency. While acquiring the reference signal in a separate step may appear to increase data acquisition time, effort is dominated by sampling of the signal emanating from the antenna since its power level can be low and often requires significantly longer sampling periods to suppress the noise floor. Conversely, the reference signal can be set to an arbitrarily high level since it does not propagate through tissue. Thus, its detection time can be as short as a few IF signal wavelengths and still achieve a high signal-to-noise ratio (SNR) which is in stark contrast to the signal propagating through the tissue, which can require thousands of sampled wavelengths to maintain SNR.

FIG. 2. Diagram illustrating the associated noise figures and gains for the B210 receiver component cascade.

C. Dynamic range

The noise floor of the B210s by themselves is ultimately limited by discretization of the A/D converter. For our medical microwave imaging system, the goal is to transmit a signal at 0 dBm (1 mW) and detect it when it is as low as -140 dBm. B210s measure signals down to a noise floor of roughly -115 dBm, even when the signal sampling time is expanded. This restriction is largely due to limitations imposed by the 12 bit A/D converter. By incorporating an additional low noise amplifier with 20 dB gain in front of the receive B210s, and with a sampling time of 0.1 ms (bandwidth = 10 kHz), the theoretical noise floor is reduced to near -140 dBm.

In addition, the B210 noise figure is specified as 8.0 dB which further degrades its ability to detect signals relative to the noise floor. Integration of additional amplification (with a concomitant low noise figure) at the front end of the receiver improves the noise floor. Here, transmission losses of the SPDT switch, which is needed to alternate between transmit and receive modes, is included. The noise figure for the cascade of circuit elements in Fig. 2 is defined by

$$NF_{\text{cascade}} = NF_1 + \frac{NF_2 - 1}{G_1} + \frac{NF_3 - 1}{G_1 G_2}, \quad (1)$$

where NF_i is the noise figure of element i in the cascade and G_i is the gain of the i -th stage. Assuming noise figures of 1.0, 1.5, and 8.0 dB, and gains of -1.0 and 20.0 dB for the stages, respectively, the cascade produces an overall noise figure of 2.7 dB (relative to the 8.0 dB of the B210 by itself). The extra 5.3 dB is a noticeable improvement and allows the system to detect signals much lower than previously.

D. SDR programming

System control is accomplished through Matlab (MathWorks, Natick, MA) largely because of the overall versatility and power of the software and its ease of integration with LabView (National Instruments, Austin, TX) which controls the remainder of the hardware in our breast imaging system. Because Matlab is a sequential programming language, two instances are invoked—one for the dedicated transmitter (Tx) and a second for multiple receivers (Rx)—and are executed through the multiprocessor toolbox and involve interprocessor communications to coordinate Tx and Rx command sequences. Tx is set to the desired operating frequency along with a quadrature modulation of 100 kHz which is generated from a numerical sine wave to produce a complex sine wave. The carrier

signal is suppressed to at least 30 dB below the modulated tone. The transmit frequency and VGAs are set dynamically during program execution. The dedicated Tx process functions in a repeating loop to produce a continuous signal. The transmitter gains a range from 0 to 89.75 dB in increments of 0.25 dB. At 1500 MHz, a gain setting of 70 nominally generates a 1 mW signal.

On the detection side, operating frequency, VGAs, and channel selection are accessible dynamically. Signals are received and mixed with a reference LO having the same operating frequency and the resultant 100 kHz complex IF signal is sampled at 10 MHz to generate 10 000 data points spanning 100 complete cycles of the sine wave. Any DC component is rejected by the AD9361 during downconversion. Complex, fast Fourier transforms (FFT) of the IF signal are computed for analysis and extraction of signal amplitude and phase used in image reconstruction. Amplitude and phase are generated from the 100 kHz signal strengths as

$$\text{magnitude} = \sqrt{X_I^2 + X_Q^2} \quad (2)$$

and

$$\text{phase} = A \tan 2(X_I, X_Q), \quad (3)$$

where X_I and X_Q are the in phase and quadrature FFT values at 100 kHz.

In addition, tests are performed on each data set to ensure quality. Preliminary checks include confirmation that the sample length is correct (10 000 samples) and that no overrun or underrun flags are triggered during acquisition. Determination of signal saturation, signal relative to the noise floor, and signal transients is performed. For the noise floor test, the signal at 100 kHz is compared against the average of signals from 2.5 to 5.0 MHz within the FFT. The latter is a practical estimate of the noise floor. We determine that a signal is usable if its strength is at least four times greater than the average noise. Similarly, for the saturation test, the third harmonic of the 100 kHz signal is sampled, and if its strength exceeds four times the noise average, a flag for saturation is triggered. Finally, for instances of a transient in the sine wave, the FFT will exhibit spurious signals on both sides of the 100 kHz IF frequency and occasionally a substantial DC component will be evident. An average of the signal over the span from 50 to 99 kHz is computed and compared to the noise level. If the spurious signal average exceeds the average noise level by a factor of 10, the recording is rejected. Ultimately, 10 measurements of each signal are processed and a median filter is applied to select optimal values.

E. Calibration

The AD9361 utilizes 12 bit A/D which enables a 72 dB instantaneous dynamic range. Given the broad range of signal strengths encountered in our system during a breast exam including variations with respect to frequency and receive antenna location relative to the transmitter, signal levels readily span 100 dB. Accordingly, integration of VGAs is critical to achieving satisfactory performance. Amplifications span 89.75 dB in 0.25 dB increments in transmit mode and 70 dB in 0.5 dB increments in receive mode.

Unfortunately, the signal level is not necessarily linear as gain settings are changed. However, signal amplitude is linear for fixed transmit and receive gain levels (Sec. III C). Thus, we acquire full

sets of measurement data for the homogeneous bath which is already part of the calibration process (Meaney *et al.*, 2017). Based on this acquisition, gain levels are selected to achieve approximately 40 dB of instantaneous dynamic range above the measurement level and 30 dB below. These gain settings are achieved along with the corresponding signal amplitudes and phases. When the measurements are acquired with the breast in place, the same gain levels are used. During calibration—i.e., homogeneous amplitudes and phases are subtracted from breast signal amplitudes and phases—no ambiguity occurs relative to the amplitude and phase offsets associated with the different gain settings. This process results in a linear signal which is critical for image reconstruction.

III. RESULTS AND DISCUSSION

A. Signal quality

Figure 3(a) shows received 100 kHz I and Q signals downconverted from 1300 MHz signals that have been differentiated into two separate, real-valued sine waves over a small (100 μ s) epoch of the sampled sine wave (from 700 to 800 μ s of a full 1 ms acquisition). The transmit signal was originally quadrature modulated with a 100 MHz signal to produce a complex sine wave. In this case, the sine waves of both components are readily recognized, albeit with noise. Figure 3(b) shows Fourier transforms of the two signals with their main amplitude concentrated in the 100 kHz position. SNR is about 40 dB in this example. The phase and amplitude are extracted through equations described in Sec. II D. Conversely, Figs. 3(c) and 3(d) show corresponding time-domain I and Q signals and their associated Fourier transforms for the case of signal saturation. Here, the elevated third harmonic and subsequent higher order odd harmonics are evident. Careful selection of IF frequency and sampling rate readily discriminates each component in the Fast Fourier transforms (FFT). The FFT is computed efficiently for these modestly sized signals so that amplitudes and phases of the fundamental and third harmonic are extracted rapidly for on-line signal quality assessments.

Figures 3(e) and 3(f) show time domain signals and FFTs when a transient, or loss of signal, occurs during acquisition. In this case, the 100 μ s time window was shifted to 660–760 μ s to capture the dynamics of the transient fully. Here, noticeable increase in signal appears to the left and right of the fundamental in the frequency domain. When the average of these aberrant signals to the left of the fundamental exceeds a threshold, the software is triggered to repeat the measurement. In certain cases, a pronounced increase in the DC component occurs which also skews the overall measurement. Because of the speed and accuracy of the embedded FFT software, these diagnostic assessments can be analyzed quickly to minimize the number of repeat measurements.

B. Synchronization

Figure 4 shows a simplified schematic focused on the two ports of a single receive B210 board. Normally the main transmit signal is directed to an antenna channel via a multiplexing switch and is received by two antenna channels which are connected to LNAs and subsequently to Rx ports of the receive B210. For synchronization evaluation, we have replaced the antennas with a power divider and 30 dB coaxial attenuators (not shown) for testing purposes.

FIG. 3. Time domain (left) and frequency domain (right) representations of in-phase and quadrature components of the IF signal, respectively: [(a) and (b)] normal signal with transmit gain level of 40 dB and receiver gain set at 0 dB, [(c) and (d)] saturated signal with transmit and receiver gain levels of 50 dB, and [(e) and (f)] transient signal with transmit gain level of 50 dB, receiver gain set at 20 dB for antennas right next to each other.

FIG. 4. Schematic diagram of test setup for evaluating system synchronization.

FIG. 5. (a) Measurements of the Channel A (signal) and Channel B (reference) phases as a function of 10K sample repetitions for different receiver gain levels (0, 10, 20, and 30 dB, respectively). (b) Differences between values in (a) as a function of repetition for the same gain settings. Values have been unwrapped to fall within the range of -180° to $+180^\circ$ in (a) and (b).

FIG. 6. Repetitions of measured phase differences between Channel A and Channel B as a function of groups of 10K samples.

The second transmit signal is fed into the isolating SPST switch and then into one of the Tx/Rx ports of the receive B210. In this configuration, both transmit signals are initiated simultaneously with both receive data acquisition modes of the receive board RxA and RxB enabled, and the isolating SPST switch is set to OFF. Multiple sets of 10 000 samples of each are acquired for evaluation. Afterward, while both transmitter signals are still running, the reference signal is acquired on the Tx/Rx port of Channel B. In full system operation, the multiplexer would be set so that no signal is transmitted and the switches for each associated channel are set to OFF. To replicate that for this measurement, the amplitude of the transmitted signal is simply set to zero while still continuously running. It is important to note that when utilizing the MATLAB Communications Toolbox, it is not possible to directly acquire data through the Tx/Rx ports. However, because there is acceptable leakage of the signal presented on the Tx/Rx port to its associated Rx port, it is possible to accurately measure the reference signal via this leakage. The phase of the reference signal is then subtracted from that for the two signals previously acquired on the RxA and RxB ports to achieve the desired coherent measurements.

Figure 5(a) shows phases of the Channel A Rx signals along with the reference port signals for eight repetitions of different receiver VGA settings. For each repetition, the Matlab code was turned OFF and restarted to assess overall robustness of the process. Figure 5(b) reports two sets of difference signals corresponding to the cases in Fig. 5(a). As expected, raw phases for signals acquired

during the repetitions vary considerably over the range of -180° to $+180^\circ$; however, the desired measurement signals track the reference signals. In fact, the calibrated phases (calibrated by subtracting out the reference signals) are consistent within 1.5° over all repetitions. Because of the calibration process used in the imaging algorithm (Meaney *et al.*, 2017), differences are systematically canceled out and are not a concern as long as the gain settings are constant. Note that for receiver gain levels between 0 and 20 dB, the phase differences remain within a relatively narrow range between 71° and 79° . However, for the 30 dB gain setting, the phase differences are nearly 180° off. Because of the considerable phase offsets due to the internal VGA settings, the final implementation of our system will utilize exactly the same gain settings for the situations of the homogeneous bath and the bath with the target so that VGA phase variability cancels out. The 1.5° variation falls within the B210 board specifications of 1.5° consistency. While this level of phase accuracy might be problematic in some measurement applications, sensitivity analyses have demonstrated that the Dartmouth imaging algorithm can tolerate phase inaccuracies up to 8° – 10° and still produce high quality images.

In addition to the signals at RxA being coherent with the reference, it is important that the signals at RxB also be coherent. Figure 6 presents calibrated phase differences between the signals measured at Channel A and Channel B for the first ten sets of 10 000 samples repeated eight times. The roughly 219° difference reflects different cable lengths between the power splitter and the associated input ports. Similar to above, because of the calibration process used for the imaging algorithm, the absolute phase differences are not critical, but the fact that they are consistent in terms of their differences is vital. These differences are within 1.5° which is consistent with B210 specifications.

C. Dynamic range

Figure 7 shows a schematic for tests used to assess the system dynamic range. In these evaluations, operating frequency was 1300 MHz, transmitted power level was 0 dBm, and IF signals were averaged over 10 000 samples. Figure 8 shows a representative FFT of a 10 000 sample IF signal that is plotted over the 0–400 kHz bandwidth. The attenuator simulated signal loss that would occur during transmission from one antenna to another within the tank. To estimate the noise floor, Fourier transforms were computed for each measurement and the signal was averaged from 10 to 90 kHz and from 110 to 200 kHz. This band was chosen to exclude spectra nearest DC and 100 kHz where the IF frequency signal is located. To

FIG. 7. Schematic of the test setup for assessing system dynamic range.

FIG. 8. Example FFT of a normal IF signal showing the primary signal component at 100 kHz and the two regions to either side of the signal used to assess the noise level.

determine the signal strength level, the FFT bin corresponding to 100 kHz was sampled.

Figure 9 reports a series of signal and noise measurements as a function of receiver gain for five attenuation levels. For the 50 dB attenuation, signal and noise track linearly for the lowest gain settings, after which the signal levels off and large ripples appear in the noise data. The linear relationships are expected since both the signal and noise are amplified similarly by the receiver circuitry. Deviation at the high end of the signal is due primarily to saturation of the receiver front end. Within the linear range, SNR is about 75 dB. At the other extreme when attenuation is set to 130 dB, both signal and noise are largely linear from the 25 dB to 60 dB receiver gain settings. For this range, SNR is roughly 20 dB. Below the 25 dB gain threshold, both signal and noise components taper off at a rate that is no longer linear relative to the rest of the range and with a decreasing SNR. These results indicate that signals can be measured down to -130 dBm with an associated SNR of 20 dB.

D. Linearity

Figure 10 shows measured signal amplitudes as a function of input power level for a single B210 receiver. The 1300 MHz input is generated by an Agilent E4432B RF Signal Generator (Santa Rosa, CA) and amplitudes are controlled by internal gain settings. In this case, the signal from the generator is fed directly into the low noise amplifier in front of the B210, such that the maximum synthesizer signal amplitude is limited to -20 dBm (to protect the B210 receiver). The IF frequency was changed to 50 kHz for this test as the upper limit for the Agilent Signal Generator. Because the A/D converter in the B210 cannot measure the full range of signals, we used multiple receiver gain settings to account for amplitude and phase differences at the gain crossover points. Here, the amplitude remains linear down to input power levels of -130 dBm, which is consistent with earlier results (Li *et al.*, 2004 and Epstein *et al.*, 2014) where the corresponding systems achieved resolution down to -125 dBm. For the range of input amplitude levels between -30 and -120 dBm, the slope of the output amplitude with respect to the input amplitude is 0.99 with a correlation coefficient of 0.9999.

E. Performance comparison with commercial vector network analyzer

Finally, we perform a set of tests over the nominal operating band of 1–2 GHz to compare results of our new SDR measurement system with those from a commercial vector network analyzer (VNA)—the Agilent (Santa Rosa, CA) E5071B. For both the VNA and our system, we utilized a set of fixed coaxial attenuators and a Weinschel Associates variable attenuator, model 940-114-11 (Mount Airy, MD), to set the input power. For phase measurements, once the amplitude reached the desired level, we introduced a Pasternack PE8245 (Irvine, CA) variable phase shifter into the receive line. For each attenuation level, phases were measured at eight dial settings. Least squares fits were performed for the highest amplitude level so that the slopes and intercepts could be computed. These values defined a straight line against which the measured phase data was compared and a rms error was computed. The tests were performed over the range of -50 to -120 dBm for the VNA and -50 to -140 dBm for the SDR system. The upper limit was set at -50 dBm because measurements greater than this level are not encountered during breast imaging. The lower limits were found empirically based on levels where the respective systems were no longer able to measure amplitudes and phases, accurately. For the VNA acquisitions, IF bandwidth was set to 100 Hz with 50 averages which required approximately 5.5 s for the six frequency sweep (1.0, 1.2, 1.4, 1.6, 1.8, and 2.0 GHz, respectively). For the custom SDR system, measurements were performed at a single frequency at a time, utilizing an IF of 100 kHz, and sample sizes of 10 000 at a rate of 1 MHz. Because the A/D converter can only accommodate an instantaneous dynamic range of about 72 dB, we adjusted the receiver gain for three levels to accommodate the full range. In this case, the settings were 0, 30, and 60 dB, respectively. Differences in the actual gain levels of the B210 internal variable gain amplifiers were compensated through our calibration procedure. For input power levels of -70 dBm and higher, -80 dBm to -90 dBm, -100 dBm to -110 dBm, and -120 dBm and lower, the number of measurement averages were 1, 50, 100, and 200, respectively. The associated acquisition times for the 50, 100, and 200 averages were 0.4, 0.8, and 1.6 s, respectively.

Figures 11(a) and 11(b) show the amplitude and phase error as functions of input power for both the VNA and SDR systems. For the VNA, both the amplitude linearity and phase error recordings degrade noticeably by -100 dBm and are not usable by -120 dBm. Correspondingly, for the SDR system, the amplitude linearity and phase error values remain consistent down to roughly -130 dBm and degrade significantly by -140 dBm. The dynamic range limitation of the commercial VNA is due primarily to its internal gain, sampling time, and A/D converter.

F. Comparison with other SDR medical microwave imaging system implementations

In the context of utilizing SDR's for microwave imaging, the most advanced system to date was reported by Marimuthu *et al.* (2016). There are some similarities between the previous system and ours; however, there are considerable differences. While the motivation for utilizing low cost, advanced measurements systems is similar, the general implementation is considerably different. First, the Marimuthu system is designed primarily for a reflection-based,

FIG. 9. Output amplitude as a function of receiver gain for 0 dBm transmit signals and attenuation levels of (a) 50, (b) 70, (c) 90, (d) 110, and (e) 130 dB, respectively.

or monostatic, configuration where a signal is radiated by a single antenna and received by the same antenna. The Marimuthu system integrates a circulator to achieve this. The antenna is mechanically moved to acquire data for all 20 physical positions around the target with future plans to expand the system to multiple channels and switching between dedicated antennas at each location. Our system only utilizes transmission data—i.e., a multistatic configuration. In this manner, we deploy a single SDR for generating the transmission and reference signals and then multiple

SDR's for receive only. In this case, for the Ettus B210 units, each board has two channels so that it can accommodate two antennas simultaneously.

In terms of the dynamic range, the Marimuthu system appears to be primarily limited by the range of the 12 bit A/D converter. In a reflection mode, this is more than sufficient and is not adversely affected by the relatively poor noise figure for most SDR's—on the order of 8–9 dB. Conversely, the Dartmouth system explicitly deals with the potentially confounding multipath signals by employing a

FIG. 10. Graph of the output signal measured by a single B210 as a function of the Agilent E4432B signal generator. To create the curve for the full range of input powers from -135 to -20 dBm, data was combined from acquisitions invoking B210 receiver gain settings of 0, 20, 40, and 60 dB, respectively.

lossy coupling bath to suppress the unwanted signals. A consequence of this is that we need to be able to measure signals down to -130 dBm (when transmitting at levels near 0 dBm). This dynamic range is accomplished by a combination of the SDR's inherent 72 dB

range from the A/D converter, variable gain amplification (VGA) on both the transmitter and receivers, external amplification at the receivers, and increased sampling time to lower the noise floor. In this case, the theoretical noise floor is roughly -154 dBm at room temperature (10 000 samples utilizing a sampling rate of 1 MHz). The extra amplification has the added benefit of improving the overall receiver noise figure down to roughly 2.7 dB.

The overall speed considerations are impacted by several factors—the number of measurements, number of frequencies, the overall sampling time for each measurement (similar to the increased averaging discussed in Marimuthu *et al.*), and whether the measurements are acquired sequentially or in parallel. For the Marimuthu system, 95 frequencies were used to adequately generate a time domain pulse (synthetically produced using an FFT technique to transform the multifrequency data). The stated measurement time for each measurement was $64 \mu\text{s}$, which is averaged 100 times. In this case, the measurements were acquired sequentially with mechanical motion of the antennas to each location. Future plans include deploying separate measurement systems to each antenna to avoid the motion time costs. While the details regarding what are the time limiting steps are limited in their report, the overall measurement time was 45 min which is quite slow. The Dartmouth system only acquires data at 7 frequencies over a range of 700–1900 MHz in 200 MHz increments. 10 000 samples were acquired for each signal at a rate of 1 MHz for a total sampling time for each measurement of 10 ms. This produced a theoretical

FIG. 11. (a) Output power, and (b) phase error as functions of input power level for the VNA (top) and the SDR measurement systems (bottom), respectively.

noise floor of -154 dBm. In this case, the transmitter and receivers run continuously so there is no switching between modes or turning ON and OFF, which are quite slow processes taking up to 2 s per operation. All switching between channels is accomplished via the external switches which have rise times on the order of 10 ns. Switching between frequencies and gain levels (VGA's) can be performed on the fly and take no more than 1 ms per operation. In this case, all receivers are run in parallel, so the acquisition time for this 4 channel system for all seven frequencies is 15 s. This amounts to transmitting at all 4 channels and receiving at the three complementary channels for a total of 12 measurements for each frequency. In expanding to 16 channels, sequential implementation will involve transmitting at 16 channels and receiving at the remaining 15 channels for a total of 240 measurements. This would be an increase to roughly 5 min. However, because of the potential to easily parallelize the operation, expanding to 16 channels will not add appreciably to the original 15 s.

One of the more challenging aspects of using the SDR's for these measurements is the overall phase coherence. For both systems, an external configuration is necessary for synchronizing the signals since the internal transmit oscillators for the two different SDR's are not phase locked with the receiver local oscillators. For the Marimuthu system, they add an external switch and a matched load so that they can compare the known reflected signal with the test one to determine the necessary phase offsets that need to be added. The authors do not report any measures of the accuracy of this technique. For the Dartmouth system, the approach exploits the fact that for each channel on the Ettus B210 boards, there is a Rx and a separate Tx/Rx port (i.e., it can transmit or receive at this port). For the dedicated transmit SDR, one of the channels transmits a signal to a multiplexer for eventual transmission to the antennas. Its second channel is used as a reference signal. This is convenient because the reference signal is coherent with the other transmitted signal. This reference signal is then sent through a high isolation switch and through a power splitter before being fed into the Tx/Rx ports of the receive SDR's (note that the Rx ports for each channel are used for the signals fed from the antennas). The test and reference signals are then sampled simultaneously exploiting the fact that their LO signals are also coherent. The desired phase is then the difference between the test and reference signals. Results from Sec. III B indicate that the phase accuracy using this technique is on the order of 1.5° which is more than sufficient for our image reconstruction algorithm. Comparable phase accuracy specifications for the Rohde and Schwarz ZNBT8 estimates the phase accuracy on the order of 0.2° for a transmission coefficient of -40 dB and degrading to roughly 10° at a transmission coefficient of -90 dB. In this context, our SDR-based approach is reasonably competitive and more than sufficient for our algorithms.

IV. CONCLUSIONS

We have implemented a 4-channel measurement system based on the Ettus Research B210 software define radios as the principle building block. We have configured the boards to transmit single frequency, modulated signals that can be measured coherently down to signal strengths of -130 dBm. We have implemented procedures to synchronize receiver signals through external circuitry which is critical for tomographic breast imaging. Performance evaluation

suggests that signal coherence is maintained to within 1.5° , and linearity is achieved with a correlation coefficient of 0.9999. These units provide an effective and low cost measurement module that is readily expanded to 16 channels or more and are an attractive alternative to commercial VNA systems.

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