Microstrip Butler Matrix Design and Realization for 7 T MRI

Pedram Yazdanbakhsh MSc.,* and Klaus Solbach

This article presents the design and realization of $8 \times 8$ and $16 \times 16$ Butler matrices for 7 T MRI systems. With the focus on low insertion loss and high amplitude/phase accuracy, the microstrip line integration technology (microwave-integrated circuit) was chosen for the realization. Laminate material of high permittivity ($\varepsilon_r = 11$) and large thickness ($h = 3.2 \text{ mm}$) is shown to allow the best trade-off of circuit board size versus insertion loss, saving circuit area by extensive folding of branch-line coupler topology and meandering phase shifter and connecting strip lines and reducing mutual coupling of neighboring strip lines by shield structures between strip lines. With this approach, $8 \times 8$ Butler matrices were produced in single boards of $310 \text{ mm} \times 530 \text{ mm}$, whereas the $16 \times 16$ Butler matrices combined two submatrices of $8 \times 8$ with two smaller boards. Insertion loss was found at 0.73 and 1.13 dB for an $8 \times 8$ matrix and $16 \times 16$ matrix, respectively. Measured amplitude and phase errors are shown to represent highly pure mode excitation with unwanted modes suppressed by 40 and 35 dB, respectively. Both types of matrices were implemented with a 7 T MRI system and 8- and 16-element coil arrays for RF mode shimming experiments and operated successfully with 8 kW of RF power. Magn Reson Med 000:000–000, 2011. © 2011 Wiley-Liss, Inc.

Key words: branch-line coupler; Butler matrix; insertion loss

INTRODUCTION

Numerous methods have been proposed to mitigate $B_1$ inhomogeneity using multiple transmitters (1–5), but utilizing these additional transmit channels is an extremely challenging task. Arrays formed from the orthogonal modes of a Birdcage Coil have been shown to have beneficial properties (6). To access these modes simultaneously, a Butler Matrix (7) is used to drive the individual rungs of the Birdcage Coil in linear combinations to form the uniform birdcage mode and higher modes (5,8,9). The use of a Butler Matrix has several advantages over directly connecting the amps to the coil. The first advantage is that an array of $2N$ coils only needs to be driven by $N$ amplifiers because only half of the modes contribute to the correct circular polarization. The second advantage is that one of the modes is the conventional CP mode, which will require most of the power; therefore, the RF amplifier of an existing single transmit channel system can be connected to this mode. The third advantage is that the matrix forms naturally decoupled orthogonal modes, i.e., under uniform mutual coupling and uniform mismatch of the rungs, the excited modes do not couple. The fourth advantage is that a significant part of the power reflected from the coils is not reflected back into the transmit amplifiers but into dummy loads connected to the unused ports. The Butler matrix has also been found to provide reflection coefficients that are insensitive to the load (9). Recently (10), a Butler matrix was used as a variable power combiner improving the power utilization in a multitransmit-channel MRI system.

A Butler matrix has multiple inputs and outputs, where each input port produces a linear phase distribution at the output; the phase distribution can be selected by selecting the appropriate input port. Each output port is connected to a coil element. In high field MRI, RF power amplifiers operate at, e.g., 298 MHz (wavelength of 1 m) for a 7 T (bias) static and homogeneous magnetic flux density $B_0$ and typically provide a sum peak power of 8 kW (in eight channels), which is much less than in lower field MRI systems. Therefore, insertion loss of the Butler matrices is one of the main challenges for practical realizations while the size of the matrices is limited to allow mounting the units close to the MRI bore. Important further requirements concern the power handling capabilities of all parts of the matrices and phase/amplitude accuracies of output signals.

Butler matrices require the combination of three functions, namely of hybrid couplers, phase shifters, and connecting lines. “Bread Board” designs have been proposed using a mixture of technologies for the three functions, using packaged hybrids and coaxial connectors with coaxial cables or solder junctions and stripline/microstrip connecting lines. We decided to avoid connectors or other transitions between the network components as far as possible to achieve a low-loss, low-cost, repeatable, and reliable network. Therefore, a choice was made to use planar integration technology, known as microwave-integrated circuit, which combines microstrip circuits on a planar dielectric substrate. A major challenge in the design of complex networks using microwave-integrated circuit technology is the need for a high precision design tool so that each component can be realized with high accuracy. This is necessary since later modification is difficult due to the limited accessibility of components embedded inside the network. The second challenge is the production of relatively large size substrate boards to create a full matrix network in one board. However, the required board dimensions were...
compatible with the state of the art in microwave and antenna circuit fabrication with specialist manufacturers, and therefore, low cost production can be achieved.

MATERIALS AND METHODS
Choice of Material
Starting from the microwave-integrated circuit technology decision, the size and insertion loss of a Butler matrix depends critically on the choice of the substrate material (Laminate), i.e., on the permittivity and the laminate thickness: Large relative permittivity ($e_r$) reduces the length of transmission lines for a given phase shift, which reduces the length and width of hybrid couplers, phase shifter lines, and connecting lines and, therefore, also reduces the required board size for a given network. As at 300 MHz, the conductor dissipation loss dominates the attenuation constant of the microstrip transmission lines (dielectric attenuation loss is less pronounced with the low-loss laminates on the market), and shorter lines can also mean less insertion loss. However, this can be put into effect only, if the conductor cross-section is kept approximately constant (constant series resistance per length). Thus, when the circuits are designed for a fixed characteristic impedance ($Z_0 = 50 \Omega$), the substrate thickness has to be increased if the permittivity is increased. These considerations have led to the choice of dielectric substrates of $e_r = 10 – 11$ with 3.2-mm thickness, as will be discussed in this article.

Design Concept
The design effort was directed at the realization of a Butler matrix in a single board. Even with thick, high-permittivity material all network components require folding to minimize the board area for each component and to allow the layout of a full $8 \times 8$ matrix with the size constraints set by board manufacturer (maximum usable circuit area: 540 mm $\times$ 430 mm). All interconnect line and phase shifter line for this matrix can be integrated as planar microstrip lines without the need for crossings of lines when an unconventional layout concept, Fig. 1a is used, which is not constrained by a predetermined placement of input and output ports at opposite sides of the network, Fig. 1b, as is usual in antenna applications.

Input and output ports of the matrix are fitted with vertical microstrip-to-coax transitions (SMA flange adapters), which allow coaxial cables from, e.g., the 8-channel transmitter to be connected to the input ports and the 8-channel coil array connected to the output ports. In $16 \times 16$ Butler matrix, Fig. 2, two modified (different phase shifts) $8 \times 8$ matrices, networks #1 and #2,
are combined with two additional subnetworks, networks #3 and #4, using 16 coaxial connecting cables.

Central to the single-board concept for 8 × 8 matrices is the size reduction of connecting lines, phase shifter lines, and of the hybrid coupler. Size reduction was realized for a given electrical length of the transmission lines first by using high-permittivity dielectric substrate and second by meandering (folding) of lines.

**Branch Line Coupler**

The realization of the hybrid coupler is limited to microstrip circuits, which allow phase quadrature, 3-dB coupling. The branch line coupler (11) is best suited due to its simplicity and uncritical design and fabrication, although the rat-race coupler (11) could be also be used (with compensation for its 180° phase difference where necessary). A further alternative could be the Lange coupler (11), which exploits coupling of several interdigitated parallel quarter-wave lines with relatively small area consumption. However, this hybrid coupler design requires cross-over bridges, which add an extra production step, not covered in standard printed-circuit board production. The hybrid coupler was carefully optimized because it dominates the board area consumption: The principal design of a 3-dB branch line coupler with four microstrip transmission lines of quarter wavelength combined in a square is shown in Fig. 3a, where \( W_1 \) and \( W_2 \) are the widths of 50-Ω (system impedance \( Z_0 \)) and 35.4-Ω \( (Z_0/\sqrt{2}) \) branches, respectively.

To reduce its occupied board area, a substrate with high-permittivity of \( \varepsilon_r = 10.2 \) and a thickness of \( h = 3.2 \) mm has been used. Then, each of the branches was folded. The highest reduction was achieved in Fig. 3b, which required that the preliminary design and layout from the circuit design tool was imported to a 3D Planar EM simulator (vendor ADS Momentum and SONNET®) to optimize its layout details. Even though we used many degrees of freedom in the design optimization, the high reduction of board area consumption of about 28%, Fig. 3b, came at the price of degradations in match and isolation to only about \( -25 \) dB, Fig. 3e, compared with about \( -48 \) dB, which was achieved in the unfolded hybrid coupler in Fig. 3d.

High match and isolation of the individual hybrid couplers are critical in a large matrix network because of potential build-up of reflection and cross-coupling by constructive superposition of contributions of each hybrid. Superposition of signals increases by 6 dB per two equal phase and amplitude signals, i.e., in an 8 × 8 Butler matrix superposition can lead to 9.5-dB increase in a worst case of three equal signals superposing (each of one of three rows of the matrix). This applies also to unwanted signals at the isolated ports.

The degradations were due to the combined effects of coupling between opposing transmission lines and coupling between parts of the same transmission line, and it proved impossible to cancel or compensate the effect by

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**FIG. 2. 16 × 16-modified Butler matrix.**

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layout optimization. Consequently, with a slightly worse reduction of occupied board area (only 15% area reduction), the hybrid coupler shown in Fig. 3c with larger distances between the conductors was realized, which also in fabricated prototypes performed closer to the perfect coupler, yielding –32 dB of match and isolation, Fig. 3f.

To demonstrate the problem of coupled lines, the coupling of two short parallel microstrip lines on a substrate with high permittivity of $\varepsilon_r = 10.2$ and a thickness of $h = 3.2$ mm was simulated and plotted as a function of distance between the two lines, see Fig. 4.

It is seen that for a coupled line length of 10 mm (equal to $\sim$2.5% of $\lambda$, the wavelength on the microstrip line), we need at least a distance about 6 mm (equal to about twice the width of the strip lines) to achieve a coupling below $–40$ dB, which was the critical level for a folded hybrid coupler that has to achieve match and isolation below $–30$ dB. Based on the coupler design, shown in Fig. 3b, we also checked the effect of the choice of the relative permittivity $\varepsilon_r$ on the occupied board area, and on the insertion loss, see Fig. 5.

In Fig. 5b, $A$ is the occupied area of the BL coupler, written as:

$$A = A_1 A_2$$

where $A_1$ and $A_2$ are the dimensions as shown in Fig. 5a. It can be seen that the absolute occupied board area can be reduced with increasing $\varepsilon_r$, but for each chosen $\varepsilon_r$, the area increases when the substrate thickness is increased: The transmission lines can be folded more closely in a circuit on lower substrate because the microstrip line width also reduces and the distance between parallel lines can be reduced accordingly; see the comparison of designs on different substrates and different substrate thicknesses in Fig. 5e.

We define a new factor as BL coupler “compactness factor” (CF) and explain this factor as:

$$CF = \frac{A}{L_1 L_2}$$
where $A$ is the area of the BL coupler explained in Eq. 1, and $L_{op1}$ and $L_{op2}$ are the optimized lengths of 50-$\Omega$ and 35.4-$\Omega$ branch lines, respectively. Figure 5c presents the relative area occupation as CF for those folded line designs: It is seen that for a high-permittivity ($\varepsilon_r = 10$) substrate of 3-mm thickness, the occupied board area is about 25% below that of the design using straight lines (Fig. 3a), while the relative advantage for circuits on a low-permittivity substrate may reduce to 15%. This, however, is not a realistic comparison, because the optimum substrate thickness for lower $\varepsilon_r$ would be found below 3 mm; e.g., for $\varepsilon_r = 3.4$, we designed networks using a thickness of 1.5 mm leading to a relative area reduction of more than 30%. As both the length and width of the transmission lines of a branch line coupler reduce as we increase $\varepsilon_r$ and reduce the substrate thickness, the coupler insertion loss increases with reducing the substrate thickness for a given $\varepsilon_r$ and increases with increasing $\varepsilon_r$ for a given substrate thickness.

As the most critical factor in matrix design, the insertion loss (IL) of the BL couplers is of highest interest: To create an $n \times n$ Butler matrix, we combine $(n/2)\times d$ couplers, where $n = 2^d$ (12), and any path from input to output ports of the matrix includes $d$ couplers in cascade.
Therefore, the critical Butler matrix insertion loss is dominated by \( d \) times the insertion loss of the single BL coupler (some additional insertion loss is contributed from connecting lines and CP-mode phase shifters). The insertion loss of the BL coupler, shown in Fig. 3d, can be calculated as the ratio of output power and accepted input power using the scattering parameters as:

\[
\text{IL} = -10 \log(|S_{31}|^2 + |S_{41}|^2) \, \text{dB}
\]  

Figure 5d shows the theoretical insertion loss figures calculated for those various designs. We can conclude that a design on \( \varepsilon_r = 10 \) and thickness 3 mm will be superior to a design on \( \varepsilon_r = 3.4 \) and thickness 1.5 mm.

Interconnecting Lines and Phase Shifters

The interconnecting lines and compensating phase shifter lines are important components in a Butler matrix. Phase errors accumulate with increasing line length due to unavoidable tolerances. Therefore, lines should be kept as short as possible. The layout of these lines was performed using the vendor ADS circuit simulator using the microstrip circuit models to describe the S-parameters of straight microstrip lines, mitered bends, and coupled microstrip lines. It was found that effective meandering of lines with 10 and more 90°-mitered bends requires very precise optimization of the bend miter to keep the aggregated mismatch of the cascaded bends low enough for a well-matched network; therefore, we used the 3D planar EM simulation to optimize bend dimensions and manufactured and tested long meander lines to be sure that we achieve better than 55-dB reflection coefficient per bend.

In addition, it was found that long folding sections (delay lines) could create considerable mismatch due to mutual coupling of parts of the line running in parallel; an example is shown in Fig. 6a, where we plot the reflection coefficient from a microstrip line folded section (length is 10% of \( \lambda \)) with strip widths \( W \) of both parallel lines equal to the width of an isolated 50-\( \Omega \) microstrip line; for part of the range of strip spacing, compensation can be realized by reducing the coupled line section strip widths as indicated in Fig. 6a. Depending on the length and the spacing of a parallel line section, a compensation of the mismatch may be possible by reducing the microstrip line width (higher \( Z_c \)) or by placing open-circuited stubs at the input and output ports of the folded section.

Shield Structure between the Coupled Lines

Common to all three components of the matrix network is the problem of coupling between the microstrip lines when the network is created in a very compact layout where neighboring strips of different parts of the network come very close over a considerable length. In this case, from Fig. 4, we find that in situations with relatively long parallel sections (e.g., 10% of a wavelength) we would have to keep distances of many times the strip widths of the coupled lines to keep coupling at a sound −35 to −40 dB level; we can observe that the often cited “design rule” for a minimum distance of twice the substrate thickness turns out to be incompatible with the low level of coupling required for the complex network described here. Keeping the very large spacing of neighboring lines in the network would deny the ability to create an \( 8 \times 8 \) Butler matrix in one board and, therefore, we introduced a partial shield structure between the coupled lines created by rows of via holes coming up from the ground plane to the strip conductor plane, which are connected to form narrow lines, which are at ground potential: The shielding effect of this grounded fence structure is demonstrated in Fig. 7.

In Fig. 7, the coupling of two parallel microstrip lines with and without grounded fence between the lines as a function of the strip distance is shown; considerable relaxation of the allowable distance of strip lines can be concluded from the results.

RESULTS

\( 8 \times 8 \) Butler Matrix

First, \( 8 \times 8 \) Butler matrix realizations were based on low-permittivity substrate RO4003 (\( \varepsilon_r = 3.55 \)) and occupied such large board areas that a single board design was not practical—these networks were fabricated by dividing the matrix into several partial boards and interconnecting the boards by coaxial cables (12). Our later
designs, however, were realized based on the above-men-
tioned design concept with board dimensions of 530 mm
× 310 mm in Roger’s TMM10 ($\epsilon_r = 9.2$) and RO3010 ($\epsilon_r = 10.2$) substrate materials of 3.2-mm thickness (avail-
able board sizes up to 457 mm × 610 mm). The matrix
layout details can be seen in Fig. 8a. Typical measure-
ment results for the scattering transmission coefficients
toward all output ports when the matrix is fed from one
input port are shown in Fig. 8b: For the lossless ideal
matrix, we expect insertion loss of 9 dB, whereas
measured loss is shown to vary around an average of 9.6 dB at the cross-over frequency of 298 MHz. We find the eight input and eight output SMA-flange adapters at the back, distributed across the layout area; for easier handling, the board was mounted to a closed housing made of thin plastic boards with aluminum front plates carrying type-N panel receptacles, which are connected to the ports of the matrix by RG221 coaxial cables, which also serve as CP-mode phase shifter at the output side (coil), Fig. 8c.

The 8-channel Butler matrix has an overall mean phase error of $E = 1.44^\circ$, calculated from:

$$E = \frac{1}{n} \sum_{i=1}^{n} |E_i|$$

where $E_i$ is the mean phase error of each mode and $i = 1, 2, \ldots, n = 8$. E.g., $E_1$ is the mean value of the phase errors at the output ports when the CP$^+$ mode input port is excited. The phase errors are calculated from the phase of the transmission S-parameters as the difference of measured phase increments between neighbor ports and the nominal phase increment of the excited mode. Measure phase error ranged from $-4^\circ$ to $+4^\circ$ in the eight modes. An overall insertion loss of $I = 0.73$ dB has been found for this Butler matrix, computed from:

$$I = \frac{1}{n} \sum_{i=1}^{n} IL_i$$

where $IL_i$ is the insertion loss of the Butler matrix feeding mode $m_i$ and $i=1, 2, \ldots, n = 8$. Because of the insertion loss of the connecting coaxial cables and the interconnecting microstrip transmission lines, the overall insertion loss has been increased over the rough estimate of three times the insertion loss of a single BL coupler, which would give about 0.49 dB. Reduction of insertion loss of 0.1 dB could be realized if we replaced the flexible cables by semirigid type. But still, for MRI applications this value of insertion loss is an acceptable level, meaning that only 14.2% of the transmitter power is lost inside the matrix. This Butler matrix has been used to feed an 8-channel Tx head coil array (13), shown in Fig. 8d, at the full RF power of 8-kW peak without power handling problems.

**16 × 16 Butler Matrix**

The realization of the 16 × 16 Butler matrix uses two identical 8 × 8 submatrices, each on a separate board, mounted one over the other in a closed housing. The full network, as shown in Fig. 2, uses RG221 coaxial cable to combine the two larger 8 × 8 networks with the two smaller networks incorporating the output stages with hybrid couplers, and phase shifter lines for the creation of CP-modes. In addition, we integrated directional couplers for the monitoring of output power and coil reflection coefficient, Fig. 9a, not shown in Fig. 2. We produced one matrix on TMM10 and three matrices on
RO3010 substrate; a typical 16 × 16 Butler matrix fabricated on RO3010 has an overall mean error of 2.5°, calculated from Eq. 4 with $i = 1, 2, \ldots, n = 16$. The overall insertion loss has also been calculated from Eq. 5 with $i = 1, 2, \ldots, n = 16$ as 1.1 dB, which is about 0.55 dB more than the accumulated insertion loss of the four BL couplers connected in cascade for any path through the matrix. The interboard cables plus the cables connecting the input ports of the 8 × 8 boards to the input connector board, the extra length of transmission line for the directional couplers and the large number of connectors are responsible for about 0.4 dB of insertion loss. A further source of insertion loss is due to the aggregation of many small reflections inside the multistage network, which produces spurious output signals at the “isolated” input ports (levels found between −25 and −35 dB). Further improvement of the insertion loss can be achieved as in the 8 × 8 Butler matrix by using semirigid cables instead of flexible cable. Finally, this Butler matrix has been used to drive a 16-channel Tx body coil array (14), shown in Fig. 9b with volunteer. Figure 9c shows T₁-weighted FLASH 2D images (FoV 300 × 300, TR 30 ms, TE 3.8 ms, 384 × 384, 2 av., breath hold) acquired in CP⁺ and CP²⁺ mode of the 16-channel array.

**Phase Error Corrections**

Despite the careful design work, the uncertainty of the relative permittivity and the variation of the permittivity across the board as well as from board to board, the limited accuracy of the microstrip-models in our design software and the distributed effect of coupling of neighboring lines across the large network, residual phase errors were found after assembly and test of the matrices. Conventionally, it would be impossible to pin-point the source of such errors without placing internal measurement ports and thereby destroying the matrix. To allow for improving the phase accuracy of our networks without destruction, a de-embedding method that estimates the phase and insertion loss of each connecting line, phase shifter line, and hybrid coupler embedded in the network was applied. The method relies on the measured transmission scattering coefficients (from each input port to each output port) of the network and the block diagram of the network. From the network block diagram, the signal paths from input to output can be traced and the transmission scattering coefficients for each path (nth input port to mth output port) by a combination of the insertion phase and loss from each component traversed by the signal can be modeled. As
each network component is part of several different paths, a system of equations that force the measured scattering transmission coefficients to be approximated by an optimum choice of the properties of the various network components can be set up. This method was applied during the prototype development of the 8 × 8 matrices, and results were used for redesign; last improvements of the final design have been done by selective modifications to individual microstrip lines inside the fabricated matrices, see Fig. 6a: Adding positive phase by a few degrees was realized by shorting bends (filling), and adding negative phase was realized by adding open-circuited stubs to the lines (shunt capacitance) or cutting out small areas of stripline (series inductance).

Performance Evaluation

Matrix accuracy (amplitude- and phase-error) and insertion loss were evaluated from the scattering matrix measurements and results for earlier Butler matrix realizations (12) have been presented as average/maximum phase errors and average/maximum insertion loss (in this article we have given average values only). Although the magnitude of such errors (or loss) gives a good indication of the accuracy achieved in the implemented design, the meaning of the pure numbers is unclear in context with the application of a Butler matrix within an MR-system when driving a transmit coil array. Therefore, a new evaluation method for the measured scattering transmission coefficients of the Butler matrix has been derived, which allows for interpretation of errors in terms of CP-modes, which would be excited in the MRT coil system. This approach starts from the fact that each mode can be represented as a combination of equal amplitude signals at the output ports of a CP-Butler matrix with phase increments between consecutive ports fixed but different for each CP-mode. Therefore, in general, any given distribution of signals at the output ports can be represented as the superposition of an infinite number of CP-modes in the same way as a periodic signal can be represented as a combination of harmonic signals. The corresponding “Fourier”-series of the CP-mode decomposition is calculated from the measured scattering transmission coefficients (when one particular input port is fed) by correlating the measured distribution with the distribution of each CP-mode. The resulting mode amplitudes indicate the purity of the field, which would be created in a perfect coil array excited by the Butler matrix when the matrix is fed at one particular input port. Figure 10 shows the measured transmission coefficients and gives the calculated spectrum of corresponding CP-modes of the 8 × 8 and 16 × 16 Butler matrices when we fed the input port for the CP1 mode. We can recognize that the target mode is by far the most dominant mode and only slightly attenuated (amplitude smaller than 1/√8 or 1/√16); other “unwanted” CP-modes are kept below −40 dB relative to the target mode in the 8 × 8 Butler matrix and below −35 dB in the 16 × 16 Butler matrix. Since, the mode analysis also gives the relative phase of modes this result can be used for a compensation of the “unwanted” modes by feeding equivalent antiphase signals into the affected mode input ports. However, this would only be feasible for relatively low levels of “unwanted” modes and has limitations if amplifiers are only connected to half of the input ports. On the other hand, without such compensation, we can conclude from the resulting magnitudes of “unwanted” modes that the desired superposition of CP-modes in, e.g., an RF shimming scheme would be degraded by spurious modes. Therefore, the requirement specification for Butler matrix phase and amplitude errors should be based on the application and, in the case of RF shimming, based on the required dynamic range of the CP-mode superposition. In our presently limited experience of RF mode shimming (14), optimization hardly requires more than 25 dB of amplitude difference between excited modes. Realization of mode shim should rely on an accuracy margin such that we assume a mode purity of the Butler matrix of 35 dB should be sufficient for most cases.

DISCUSSION

In this article, the trade-offs of folded microstrip branch line couplers have been discussed with a view to the design and realization of planar-integrated Butler matrix networks for 7 T MRI systems. As a result, the adoption of substrate material of high permittivity and large board thickness is found optimum for the purpose. Successful realization of low-loss 8 × 8- and 16 × 16-Butler matrices was demonstrated with 1.1 dB of insertion loss for the 16 × 16 network and with total mechanical dimensions suitable for MRI system implementation.

Application of our results to MRI systems of higher field (e.g., 9 T with about 400-MHz RF) and lower field (e.g., 3 T with 128-MHz RF) is possible: We find slightly higher insertion loss for the lower frequency and lower insertion loss for the higher frequency. However, as the wavelength at 128 MHz is much larger than at 300 MHz (in our present 7 T application), the mechanical dimensions increase drastically for the 3-T application—therefore, we propose to use our planar integrated Butler matrix approach only for up to 8 × 8 channels for 3-T MRI systems.

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