A Ferrite LTCC-Based Monolithic SIW Phased Antenna Array

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Abstract—In this work, we present a novel configuration for realizing monolithic SIW-based phased antenna arrays using Ferrite LTCC technology. Unlike the current common schemes for realizing SIW phased arrays that rely on surface-mount component (p-i-n diodes, etc) for controlling the phase of the individual antenna elements, here the phase is tuned by biasing of the ferrite filling of the SIW. This approach eliminates the need for mounting of any additional RF components and enables seamless monolithic integration of phase shifters and antennas in SIW technology. As a proof of concept, a two-element slotted SIW-based phased array is designed, fabricated and measured. The prototype exhibits a gain of 4.9 dBi at 13.2 GHz and a maximum E-plane beam-scanning of ± 28 degrees using external windings for biasing the phase shifters. Moreover, the array can achieve a maximum beam-scanning of ± 19 degrees when biased with small windings that are embedded in the package. This demonstration marks the first time a fully monolithic SIW-based phased array is realized in Ferrite LTCC technology and paves the way for future larger-size implementations.

Index Terms—Ferrite phase shifter, low-temperature co-fired ceramic (LTCC), phased array, substrate integrated waveguide (SIW).

I. INTRODUCTION

Antennas having beam-scanning functionality are required in radars and many modern communication and imaging systems. Over the years, many different implementations of beam-scanning antenna arrays have been carried out in different technologies (waveguides, microstrip, etc). In the past decade, Substrate Integrated Waveguide (SIW) concept has become a popular platform for realizing antennas and circuits [1]. SIWs combine the simplicity and low-loss nature of rectangular waveguides while being low-cost and easy to fabricate and integrate with planar circuitry.

Most of the developed SIW-based beam-scanning arrays in literature rely on a fixed beam-forming network such as a butler matrix or a rotman lens [2], [3]. The disadvantage of such approach is that the beam can only be switched over a discrete set of scan angles and that the size of the beam-forming network grows quickly with the number of required beam directions [4], [5]. The need of such complex networks can be eliminated by using tunable phase shifters and implementing a phased array with continuous beam-steering capability.

A problem in tunable phase shifters is that their integration with the SIW typically requires surface-mount components.

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Fig. 1: Conceptual sketch of a monolithic slotted SIW phased antenna array in Ferrite LTCC.

For example, the phase shifter implemented in [6] required using four p-i-n diodes as switches which are mounted on the SIW. Mounting tunable components such as varactor diodes or switches on the SIW increases the phase shifter complexity and adds parasitics which may become significant at high frequencies. In this work, we avoid the above-mentioned issues by using a controllable medium (ferrite) as a multilayer substrate for the SIW.

Generally, conventional ferrite materials have been extensively used for implementing tunable phase shifters in planar transmission-line (microstrip, CPW, etc) [7]–[10] as well as non-planar waveguide configurations [11]. But even though it has been over a decade since the invention of SIWs, only very few ferrite-based SIW phase shifters have been reported [12], [13] and these have not been demonstrated in any kind of phased array system. A possible reason for the limited demonstrations is the difficulty of integration of SIW and conventional ferrites. The brittle ceramic nature of ferrite substrates makes the realization of via rows without substrate fracture a real challenge. To circumvent the problem, researchers in [13], [14] used ferrite slabs to load a SIW that is implemented in a conventional duroid substrate. Although this approach might be acceptable if few devices are needed, it is not feasible for implementing medium and large phased antenna arrays due the need for extensive machining for each device.

A more elegant way to resolve ferrite-SIW integration problems is through the use of new technology called ‘Ferrite

Fig. 2: Conceptual sketch of (a) tunable SIW phase shifter biased using external circuitry and (b) tunable SIW phase shifter biased using embedded circuitry within the Ferrite LTCC package.

LTCC' [15]–[19]. In Ferrite LTCC, the traditional dielectric layers of conventional LTCC are replaced by ferrite ones. The conductor traces and vias are implemented in the unfired tape which is much easier to process. This makes it possible to have a multilayer ferrite package with printed conductors and vias and hence allows the realization of SIWs.

Using Ferrite LTCC, monolithically integrated slotted-SIW phased antenna arrays can be realized as demonstrated in Fig. 1. An SIW subjected to a variable static magnetic bias field serves as a tunable phase shifter for each slotted SIW antenna arm. This bias field could be supplied by circuitry external to the package in the form of a winding around a magnetic core (as in Fig. 2(a)). Alternatively, by adding more layers, bias circuitry can be embedded within the ferrite package as shown in Fig. 2(b). The embedded windings approach provides the highest level of integration, however the number of turns of the windings which is proportional to the generated bias magnetic field is generally limited by the number of available LTCC layers. On the other hand, external biasing coils, though they occupy more space, are generally not restricted to the LTCC stack size and therefore can be made with large number of turns.

In this work, as a demonstration, we implemented a 2x3 Ku-band slotted-SIW antenna array with each 1x3 subarray being fed by a tunable phase shifter. Each phase shifter consists of a SIW that is subjected to a controllable anti-symmetric biasing magnetic field whose strength determines the insertion phase of the SIW. Two biasing schemes have been realized to tune the phase shifters. The first employs a coil system external to the LTCC package while the second relies on windings embedded within the Ferrite LTCC package. The array operation has been experimentally verified using both biasing techniques showing a beam scanning range of $\pm 28^\circ$ in the case of external windings and about $\pm 19^\circ$ in the case of embedded ones. The obtained results are encouraging and suggest the feasibility of this configuration for implementing Ku-band phased arrays for applications such as satellite communication systems.

II. DESIGN AND SIMULATIONS

A 10-layer LTCC stack is used for realizing the antenna array. The stack is made of ESL-40012 ferrite tape system [20]. This tape has a $\epsilon_r \approx 14.6$, $\tan \delta \approx 5 \times 10^{-3}$ and a

Fig. 3: Simulated radiation efficiency of a single slot versus number of SIW layers (each is $110 \mu\text{m}$ thick). The ferrite material is simulated in demagnetized state ($M = 0, \mu_d \approx 0.68$ at 13 GHz) and silver is used for conductive traces and vias.

saturation magnetization ($4\pi M_s$) of 4000 Gauss that results in a magnetization frequency (f_m) of about 11.2 GHz [17]. The material is operated in the partially magnetized state ($0 \leq M \leq M_s$) to minimize the bias field required. A design frequency of 13 GHz is chosen to avoid the low-field losses that occur in such state at low frequencies and up to the material's magnetization frequency.

A. Layer Assignment

Since the 10-layer LTCC stack shall contain the antenna as well as the biasing circuitry, there will be a trade-off between the antenna performance (in RF) and dc biasing efficiency. Dedicating more layers for biasing circuitry would imply a denser winding with higher number of turns (N) that will decrease the current needed to produce a given required bias magnetic field ($H \propto NI$). However, dedicating more layers for the windings means less layers for the antenna which is expected to degrade the antenna performance (efficiency, gain, bandwidth). To examine how much the antenna efficiency is affected by the waveguide height, we simulated a single slot radiator (structure geometry similar to the model in Fig. 5 over different number of layers. The results, illustrated in

Fig. 3, reveals that the radiation efficiency of the slot antenna increases as the number of layers is increased and it reaches a maximum value of about 36%, relatively low due to the high ferrite permittivity, when all layers are dedicated for the antenna. It can also be noticed that the efficiency deteriorates quickly if less than 5 layers are used. Based on these results, it has been decided to dedicate the top five layers for the SIW and place the biasing circuitry in the bottom five layers.

B. SIW Phase Shifter

As mentioned earlier, the feeding SIW for each of the sub-arrays shall act as a phase shifter. The phase of the signal reaching each sub-array is controlled by biasing its feeding SIW-section. Biasing the SIW by a static magnetic field H_0 generates a magnetization level M inside the material that dictates the effective permeability sensed by the RF-signal propagating through. For a partially magnetized ferrite biased in the y-direction, the permeability is described by the following tensor [15], [21]:

$$[\mu]_{PMS}^{y-bias} = \mu_0 \begin{bmatrix} \mu & 0 & -j\kappa \\ 0 & \hat{\mu} & 0 \\ j\kappa & 0 & \mu \end{bmatrix} \quad (1)$$

Where the tensor elements are all functions of the magnetization level M established in the material and the signal frequency f as given by the equations below [22]:

$$\mu = \mu_d + (1 - \mu_d) \left(\frac{M}{M_s} \right)^{\frac{3}{2}} \quad (2)$$

$$\hat{\mu} = \mu_d \left(1 - \left(\frac{M}{M_s} \right)^{\frac{5}{2}} \right) \quad (3)$$

$$\kappa = n_k \frac{f_m}{f} \frac{M}{M_s} \quad (4)$$

In the equations above, μ_d is the material's permeability when completely demagnetized ($M = 0$) and its value is given by the expression in (5). For the off-diagonal component κ , n_k is a proportionality factor that is taken to be 0.5 at 13 GHz for the material under consideration based on the previous works of [15].

$$\mu_d = \frac{1}{3} + \frac{2}{3} \sqrt{1 - \left(\frac{f_m}{f} \right)^2} \quad (5)$$

Here, the phase shifting is achieved by applying an anti-symmetric bias (depicted in Fig. 4) to each sub-array feeder. This anti-symmetric configuration of biasing has been studied in [23] and theoretically shown to induce large phase shift per unit length and to achieve very high figure of merit (phase shift per unit loss). In our recent work [19], we demonstrated a phase shifter with a figure of merit of about 100 degree/dB at Ku-band using this anti-symmetric configuration and biasing the device through 8 layers of embedded windings. In this work, we implement the phased array using a phase shifter of the same concept as in [19] but with slight modification in the layer assignment and biasing circuitry.

For the present case, to estimate the attainable phase shift, simulations were carried out for the 5-layer SIW being biased anti-symmetrically to different magnetization levels. In our

Fig. 4: Simulated phase shift versus magnetization level of an anti-symmetrically biased SIW ($a_{SIW} = 4.5$ mm, $h = 550$ μ m, $L_{SIW} = 8$ mm) at 13 GHz.

Fig. 5: Simulated slot conductance at resonance and resonant length for different slot offsets at 13 GHz for five layers for a SIW width ($a = 4.8$ mm, $w_{slot} = 200$ μ m).

simulation model, the partially magnetized material permeability definition is implemented using the aforementioned equations in CST microwave studio. The simulation results for an 8 mm-long SIW feed section show that the forward insertion phase shift ($\angle S_{21}(M) - \angle S_{21}(M = 0)$) can reach about 180 degrees at 13 GHz when the material is magnetized to saturation as illustrated in Fig. 4.

C. Integrated Slotted SIW Phased Array

To design a slotted antenna sub-array, first the characteristics of a single slot need to be obtained. A longitudinal slot on the broad face of rectangular waveguide can be modeled by a shunt load [24]. The slot length L_{sl} will primarily determine the resonant frequency. While the slot offset from the SIW

Fig. 6: Optimized 2x3 slotted-SIW array. (a) 3D view. (b) side view. All slots are $230 \mu\text{m}$ wide, the rest of the dimensions are : $L_1=48 \text{ mm}$, $W_1=14 \text{ mm}$, $L_{sh} \approx 3\lambda_g/4 = 8.76 \text{ mm}$, $L_{sl}=4.87 \text{ mm}$, $p \approx \lambda_g/2 = 5.3 \text{ mm}$, $W_2=12.5 \text{ mm}$).

Fig. 7: Optimized 2x3 slotted-SIW array simulated reflection coefficient.

center (referred to using the symbol x) mainly dictates the slot conductance at resonance G_0 . To extract the slot admittance, the procedure in [25] has been utilized. A single slot is simulated with the wave guide being terminated by a short circuit at a distance of $3\lambda_g/4$ from the slot center and with the port plane being at $\lambda_g/2$ before the slot center. In this manner the input admittance at the port is equal to the admittance of the slot $Y_{in} = Y_{slot}$. Using this model, the slot length was varied at different offsets and each time the slot length that induces resonance at 13 GHz and slot resonant conductance (normalized to the waveguide characteristic admittance) G_0/Y_0 are recorded. The results for the slot conductance at resonance and slot resonant length at 13 GHz are depicted in Fig. 5. The available LTCC panel space ($48 \times 14 \text{ mm}$) could accommodate a maximum of three slots in addition to the

Fig. 8: Simulated beam steering. (a) E- plane radiation pattern (b) Beam angle versus magnetization (c) Gain over scan angle.

feeding structures. For a three-shunt-slots design, each slot should have resonant conductance equal to one-third of SIW's characteristic admittance ($\frac{G_0}{Y_0} = 0.33$). Using the shown characteristics curves of a single slot in Fig. 5, the resonant-length and offset were determined to be 4.87 mm and $100 \mu\text{m}$ respectively.

Following this step, two three-slot sub-arrays were integrated with a phase shift section that is a simple extension of their feeding SIW. It is worth mentioning that a slightly larger SIW width is used for the antenna sub-array ($a = 4.8 \text{ mm}$) than SIW for the phase shifter ($a = 4.5 \text{ mm}$). This small change in width has minor effects on the propagation and it helps relax the dimensions of the slot offset ($x \approx 0.21a$).

Fig. 9: Anti-symmetric biasing of a SIW using external coils and yokes.

Fig. 10: Embedded windings for biasing individual phase shifters.

To feed the sub-arrays, a tapered microstrip to SIW transition has been designed and a 2-way SIW power divider that uses inductive vias for matching the impedance has been employed as depicted in Fig. 6. The whole structure has been optimized in CST microwave studio for operation at 13 GHz. Slight adjustments were required for the slot antenna dimensions obtained previously to account for mutual coupling effects. The simulated reflection coefficient for the optimized structure is illustrated in Fig. 7 showing that the array is matched better than -12 dB at 13 GHz and with a 10 dB impedance bandwidth extending from 12.88 to 13.1 GHz. The array exhibits a simulated gain of 5.1 dBi and an efficiency of 41%. As mentioned earlier, the relatively low efficiency and gain are due to the high permittivity of the ferrite substrate. However, it is expected that gain and efficiency numbers can be increased if more complex fabrication is used. For instance, creating an air cavity or perforating the substrate inside the SIW's antenna array portion is expected to reduce the effective permittivity and enhance radiation.

In this configuration, the antenna array beam scans in the E-plane towards positive and negative elevation angles depending on which phase shifter is being activated. A beam scanning range of ± 45 degrees has been demonstrated in simulations by driving each phase shifter from the unbiased state ($M = 0$) to the saturated state ($M = M_s$) as depicted in Fig. 8(a,b). The antenna remains matched when either phase shifter is biased with very little variation in S_{11} . As expected the gain drops

Fig. 11: Magnetostatic simulation results. Normalized Magnetization distribution $|M_y(x,y)/\max[M_y(x,y)]|$ over SIW cross section at bias current of 1 ampere. (a) external biasing, (b) embedded biasing and (c) average magnetization versus bias current.

when the array is steered from broadside. Steering the beam up to ± 30 degrees is accompanied by slow drop in gain but after ± 30 degrees the gain drops more quickly with the scan angle as seen in Fig. 8(c).

D. Biasing Circuitry

In order to magnetize the individual phase shifters, a magnetic field having an anti-symmetric configuration needs to be generated. In this work, we adopted two different approaches to generate such a configuration. In the first approach, an external coil system is used to bias the phase shifters. An 80-turn coil is wound around high permeability U-shaped iron

Fig. 12: Fabricated prototype.

yokes which are fitted above each phase shifter. This configuration helps guide the flux in anti-symmetric fashion inside the guide as demonstrated in Fig. 9. The second approach is by using windings embedded in the Ferrite LTCC package. The windings consist of 2 solenoid-like coils occupying the five layers beneath the SIW. The solenoids are symmetrically-placed with respect to the SIW mid-plane with the current in the solenoids circulating in opposite sense to generate the required anti-symmetric bias magnetic field profile as illustrated in Fig. 10. Each solenoid has 3 turns in each layer, the largest turn has a length of 8.25 mm and a width of 2.7 mm. Windings' dc pads are placed in the bottommost layer for dc supply connection. It is worth noting that here the dc windings are isolated from the microwave signal path by the SIW ground plane in layer five therefore there is no need for adding microwave chokes or dc block. The bias circuits are simulated separately using a magnetostatic solver (CST EM studio). Simulation results are depicted in Fig. 11.

As can be seen in Fig. 11(a,b), in both scenarios the magnetization is anti-symmetric (same in each half of the SIW but with opposite direction). However, the magnetization is not uniform within each half of the SIW. The average magnetization level over the SIW volume is calculated for each current excitation and the results are depicted in Fig. 11(c). As it is clear from the figure, in spite of having less total number of turns, the embedded windings can magnetize the structure to higher level with the same amount of dc current compared to the external biasing method. This is due to the demagnetization effects that occur due to the air gap existing between the external yoke and the ferrite substrate (assumed to be 100 μm). It is also worth mentioning that the bias magnetization fields are highly localized around their specific phase shifter. For the present phase shifter separation of 3.5 mm, magnetizing one phase shifter to saturation (4000 Gauss) will magnetize the neighboring phase shifter to 4.4 Gauss (0.11%). This would result in a very small induced phase shift of about 0.2 degrees in neighboring phase shifter.

III. PROTOTYPE MEASUREMENTS RESULTS AND DISCUSSION

An array prototype, shown in Fig. 12, has been fabricated at VTT-electronics Finland. The prototype has been prepared

Fig. 13: Measured and post-measurement simulation of the array reflection coefficient.

Fig. 14: Simulated and measured E (xy) plane (up) and H (yz) plane (down) at 13.2 GHz.

by firing at a peak temperature of 875 degree Celsius for 30 minutes.

A. Measurements Results in Unbiased State

An SMA connector has been mounted on the antenna and the input reflection coefficient has been measured using Agilent PNA-L 2873 network analyzer and the results are presented in Fig. 13. This measurement revealed that the antenna exhibits a return loss better than 10 dB over the frequency range from about 13 to 13.3 GHz, with a center fre-

Fig. 15: Prototype under characterization with biasing circuitry.

quency of about 13.15 GHz. This measured center-frequency is slightly higher than the intended 13 GHz predicted by simulation. It is suspected that this deviation is due to using a higher permittivity value than the actual one of the material. A post measurement simulation has been carried out by varying the permittivity within the material tolerance range ($\epsilon_r = 14.6 \pm 0.6$), good match was obtained by using a permittivity of 14.0 as can be seen in Fig. 13.

The antenna radiation performance has also been characterized using SATIMO Star-Lab system. The antenna shows a maximum measured gain of 4.9 dBi which is very close to the simulated one of 5.1 dBi at 13.2 GHz at no-bias. The simulated and measured E (xy) and H (yz) plane patterns are displayed in Fig. 14. As it is clear from the figure, the measured radiation pattern are in good agreement with the simulated one.

B. Measurements in Biased States

In order to test the array using the external yoke systems, we fabricated an acrylic alignment-sleeve to keep the yokes in position over the phase shifters as shown in Fig. 15. About 80 turns were wound on each yoke. Using this configuration, the beam could be steered up to ± 28 degrees in E-plane by incrementing the current to maximum value of 1.75 A. The measured patterns are shown in Fig. 16(a). Some distortion of the pattern could be observed (relative to the unbiased case) and is likely to be because of the wiring and bias circuitry placed around the antenna. Also, some asymmetry is observed when scanning to positive and negative angles in the external windings case. This can be caused by a misalignment error in the placement of the external coils. A gain drop of about 1 dB occurs when scanned to 28 degrees. Increasing the bias current beyond 1.75 A was not possible as it caused melting of the insulation of the copper wires and short circuiting the turns. Also, it is worth mentioning that the input impedance of the array showed very little variation when the phase shifters are biased. The measurements results of the array input reflection coefficient when one phase shifter is being biased are displayed in Fig.17. As can be seen little variation can be observed and the input of the array remains satisfactorily matched to 50 Ohm over the frequency from about 12.98 GHz to 13.38 GHz.

Fig. 16: Measured E-plane pattern at maximum scanning using (a) external windings and $I=1.75$ A (b) embedded windings and $I=0.2$ A.

By using the embedded windings to bias the phase shifter, a maximum beam scanning of nearly ± 19 degrees could be obtained at a 200 mA of applied current as depicted in Fig. 16(b). It is worth noting that the embedded windings resistance was much higher than the external one (16 ohm compared to 0.5 ohm for the external coil). This high resistance limited the application of higher currents for biasing. At higher currents, the windings started to heat up appreciably over the course of the radiation pattern measurements resulting in a distorted pattern.

A summary of scanning results for both internal and external biasing techniques is given in table I. From the table, it can be observed that the embedded windings biasing approach can be used to magnetize the structure to a higher level for a given current compared to the external biasing approach. This is evident by observing that the embedded windings achieve about 20 degrees of scanning at 200 mA whereas the external windings achieve this value at 850 mA. Compared to simulations, it is noticed that measured scanning angles for a given current are generally higher. It is believed that this deviation from simulation predictions are coming from underestimating the generated magnetization from each current value. This underestimation can be due to inaccuracy of the BH-curve used for modeling the material and/or an inaccuracy in estimating the air-gap size in case of external windings. All inclusive, the simulations predictions are found to be fairly close to actual

Fig. 17: Measured reflection coefficient under different bias currents.

TABLE I: Summary of results for internal and external biasing.

I [A]	Embedded Windings			External Windings		
	Sim. M (M/M_s)	Θ_{sim}	Θ_{mes}	Sim. M (M/M_s)	Θ_{sim}	Θ_{mes}
0.2	1082 G (27%)	14.2	19	541 G (14%)	7.2	-
0.85	2890 G (72%)	30.2	-	1527 G (38%)	18.7	20
1.75	4000 G (100%)	45	-	1716 G (43%)	21	28

Θ_{sim} : Simulated Scan Angle, Θ_{mes} : Measured Scan Angle., G: Gauss

measurements and the deviations have small impact on the overall performance of either case.

IV. FUTURE FERRITE LTCC PHASED ARRAYS

The high permittivity and moderate loss tangent of Ferrite LTCC ($\epsilon_r = 14.6$, $\tan \delta = 0.005$) limits the antenna gain and efficiency. To improve the antenna performance, it is suggested that conventional low-loss LTCC material such as Dupont 9k7 ($\epsilon_r = 7.1$, $\tan \delta = 0.0009$) be used for antennas realization, while the phase shifters can still be implemented in the ferrite multilayer technology. This will improve the efficiency of the antenna considerably while maintaining the beam steering capability of the phased antenna array. However, this requires further research work on the fabrication side to enable monolithic integration of conventional and ferrite LTCC tape systems into a single package.

Another performance limitation for Ferrite LTCC based phased arrays comes from the self-heating of the embedded bias windings. This heating sets a limit on the allowable bias current and hence the realizable phase shift and scanning range. To mitigate this effect, the embedded windings need to be designed for minimum resistance while meeting the desired magnetization level. Additionally, in-package cooling mechanisms to dissipate heat generated from internal bias windings must be considered in future designs. One suggested way is to incorporate microfluidic channels within the LTCC package through continuous embedded cavities. Through these channels a liquid coolant can be pumped to diffuse the generated heat. With the internal windings cooled, more current can

be applied resulting in higher phase shift and larger scanning range for a phased array.

In addition to cooling mechanism, improving the remanent magnetization of the ferrite material can also help reduce the generated heat as well as the consumed power. If the material has sufficiently high remanent-to-saturated magnetization ratio (M_r/M_s), the bias current need not be continuously applied to maintain the material magnetized. Instead, the current need only be pulsed for a short duration (typically few hundred microseconds) to switch the magnetization level inside the material [22], [26]. This mode of operation is usually referred to as “latching” and is feasible to use when the material has near-square hysteresis ($M_r/M_s > 0.75$). Near-square hysteresis microwave ferrites are commercially available in bulk substrate form [27]. The development of similar LTCC process-compatible versions of those, currently being researched [28], can resolve self-heating problems as well as greatly reduce device power consumption.

V. CONCLUSION

In this work, a new implementation of SIW-based phased antenna arrays is proposed. Unlike previously demonstrated SIW phased arrays that use a SIW fixed beam-forming network (ex: butter matrix) or mountable switching elements on the SIW to switch the beam, the phase shifting is implemented by using tunable ferrite-filled SIW phase shifters. The proposed approach has the advantages of allowing for continuous beam scanning, avoiding the use of surface mount components and providing monolithic integration of the antennas with phase shifters in SIW technology. A demonstrator two-element slotted-SIW phased antenna array prototype has been designed fabricated and measured. For testing the array, two approaches were used for providing the biasing magnetic field for the phase shifters. The first approach relies on coil systems implemented using circuitry outside the Ferrite LTCC package and the second approach used windings that are realized within the package. A maximum beam steering of ± 28 degrees could be measured using the external biasing approach and a winding current of 1.75 A, while a maximum scanning of about ± 19 degrees is measured using the internal bias circuitry and a current of 0.2 A. It is important to mention that although the internal winding is found to be more efficient in magnetizing the substrate with low currents, they generally suffer from larger windings resistance (nearly 20 times higher) than external windings of similar size. Consequently, greater power dissipation is observed in the case of the internal windings for the same current. Also, the large embedded winding resistance limits the amount of current that can be used without causing permanent damage to the conductive traces and it makes the substrate subject to self-heating effects. If embedded windings are to be used for biasing in future works, it is recommended to attempt to minimize the winding resistance as much as possible and incorporate heat dissipation techniques into the package. As a first attempt, the demonstrated prototype results show promise for using Ferrite LTCC as a platform for SIW-based phased antenna arrays.

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REFERENCES

- [1] K. W. P. A. M. Bozzi, L. Perregrini, "Current and Future Research Trends in Substrate Integrated Waveguide Technology," *Radioengineering*, vol. 18, no. 2, pp. 201–209, 2009.
- [2] Y. J. Cheng, W. Hong, K. Wu, Z. Q. Kuai, C. Yu, J. X. Chen, J. Zhou, and H. J. Tang, "Substrate Integrated Waveguide (SIW) Rotman Lens and Its Ka-Band Multibeam Array Antenna Applications," *IEEE Trans. Antennas*, vol. 56, no. 8, pp. 2504–2513, Aug 2008.
- [3] A. Guntupalli and K. Wu, "Phase-Steered Fixed Beams in One and Two-Dimensional Scan Space for Substrate-Integrated Radar and Radio Systems at 60 GHz," *IET Microwaves, Antennas and Propagation*, vol. 8, no. 11, pp. 819–828, August 2014.
- [4] C.-C. Chang, R.-H. Lee, and T.-Y. Shih, "Design of a Beam Switching/Steering Butler Matrix for Phased Array System," *IEEE Trans. Antennas*, vol. 58, no. 2, pp. 367–374, Feb 2010.
- [5] L. Wu, A. Farrall, and P. Young, "Substrate Integrated Waveguide Switched Beam Antenna," *IEEE Trans. Antennas*, vol. 63, no. 5, pp. 2301–2305, May 2015.
- [6] P. Loghmannia, M. Kamyab, M. Ranjbar Nikkiah, and R. Rezaiesarlak, "Miniaturized Low-Cost Phased-Array Antenna Using SIW Slot Elements," *IEEE Antennas Wireless Propag. Lett.*, vol. 11, pp. 1434–1437, 2012.
- [7] X. Yang, J. Wu, Y. Gao, T. Nan, Z. Zhou, S. Beguhn, M. Liu, and N. Sun, "Compact and low loss phase shifter with low bias field using partially magnetized ferrite," *IEEE Trans. Magn.*, vol. 49, no. 7, pp. 3882–3885, July 2013.
- [8] W. Hui, J. Bell, M. Iskander, and J. Lee, "Low-Cost Microstrip-Line-Based Ferrite Phase Shifter Design for Phased Array Antenna Applications," *IEEE Antennas Wireless Propag. Lett.*, vol. 6, pp. 86–89, 2007.
- [9] H. Kishimoto, T. Ueda, and Y. Kado, "Experimental demonstration of nonreciprocal phase-shift composite right/left-handed transmission lines using a ferrite-rod-embedded substrate," *IEEE Trans. Magn.*, vol. 47, no. 10, pp. 3724–3727, Oct 2011.
- [10] L. Carignan, T. Kodera, D. Menard, and C. Caloz, "Moldable polymer/ferrite composite and application to an integrated cpw tunable phase shifter," *IEEE Microw. Wireless Compon. Lett.*, vol. 19, no. 4, pp. 206–208, April 2009.
- [11] A. Abuelma'atti, J. Zafar, I. Khairuddin, A. Gibson, A. Haigh, and I. Morgan, "Variable toroidal ferrite phase shifter," *Microwaves, Antennas Propagation, IET*, vol. 3, no. 2, pp. 242–249, March 2009.
- [12] W. Che, E.-N. Yung, and K. Wu, "Millimeter-wave ferrite phase shifter in substrate integrated waveguide (siw)," in *Antennas and Propagation Society International Symposium, 2003. IEEE*, vol. 4, June 2003, pp. 887–890 vol.4.
- [13] S. Adhikari, A. Ghiotto, S. Hemour, and K. Wu, "Tunable non-reciprocal ferrite loaded siw phase shifter," in *Microwave Symposium Digest (IMS), 2013 IEEE MTT-S International*, June 2013, pp. 1–3.
- [14] A. Ghiotto, S. Adhikari, and K. Wu, "Ferrite-loaded substrate integrated waveguide switch," *Microwave and Wireless Components Letters, IEEE*, vol. 22, no. 3, pp. 120–122, March 2012.
- [15] F. Ghaffar, J. Bray, and A. Shamim, "Theory and Design of a Tunable Antenna on a Partially Magnetized Ferrite LTCC Substrate," *IEEE Trans. Antennas*, vol. 62, no. 3, pp. 1238–1245, March 2014.
- [16] F. Abdul Ghaffar and A. Shamim, "A Partially Magnetized Ferrite LTCC Based SIW Phase Shifter for Phased Array Applications," *IEEE Trans. Magn.*, vol. PP, no. 99, pp. 1–1, 2015.
- [17] A. Shamim, J. Bray, N. Hojjat, and L. Roy, "Ferrite LTCC-Based Antennas for Tunable SoP Applications," *IEEE Trans. Compon. Packag. Manuf. Technol.*, vol. 1, no. 7, pp. 999–1006, 2011.
- [18] E. Arabi, F. Ghaffar, and A. Shamim, "Tunable Bandpass Filter Based on Partially Magnetized Ferrite LTCC With Embedded Windings for SoP Applications," *IEEE Microw. Compon. Lett.*, vol. 25, no. 1, pp. 16–18, Jan 2015.
- [19] A. Nafe and A. Shamim, "An integrable siw phase shifter in a partially magnetized ferrite ltcc package," *IEEE Transactions on Microwave Theory and Techniques*, vol. 63, no. 7, pp. 2264–2274, July 2015.

- [20] Electroscience thick-film materials and ceramic tapes. [Online]. Available: <http://www.electroscience.com/pdf/40012.pdf>
- [21] J. J. Green and F. Sandy, "Microwave Characterization of Partially Magnetized Ferrites," *IEEE Trans. Microw. Theory Techn.*, vol. 22, no. 6, pp. 641–645, 1974.
- [22] S. K. K. . B. Bhat, *Microwave and Millimeter Wave Phase Shifters*. Dedham, Massachusetts: Artech House Inc, 1991, vol. I, pp. 7–8, 115–116, 185.
- [23] J. Bray and L. Roy, "Development of a Millimeter-Wave Ferrite-Filled Antisymmetrically Biased Rectangular Waveguide Phase Shifter Embedded in Low-Temperature Cofired Ceramic," *IEEE Trans. Microw. Theory Techn.*, vol. 52, no. 7, pp. 1732–1739, 2004.
- [24] R. C. Johnson, *Antenna Engineering Handbook*, 3rd ed. McGraw-Hill Professional, 1992.
- [25] J. Massen, M. Frei, W. Menzel, and U. Möller, "A 79 GHz SiGe Short-Range Radar Sensor for Automotive Applications," *International Journal of Microwave and Wireless Technologies*, vol. 5, pp. 5–14, 2 2013.
- [26] V. G. Harris, "Modern microwave ferrites," *IEEE Trans. Magn.*, vol. 48, no. 3, pp. 1075–1104, 2012.
- [27] Trans-tech, inc. [Online]. Available: <http://www.trans-techinc.com/products.asp>
- [28] T. Zhou, H. Zhang, C. Liu, L. Jin, F. Xu, Y. Liao, N. Jia, Y. Wang, G. Gan, H. Su, and L. Jia, "Li_{0.43}Zn_{0.27}Ti_{0.13}Fe_{2.17}O₄ ferrite ceramics for LTCC Technology," *Ceramics International*, vol. 42, no. 14, pp. 16 198 – 16 204, 2016.

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