A Dual-Polarization Switched Beam Patch Antenna Array for Millimeter-Wave Applications

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Abstract—Increasing data transfer rates for future millimeter-wave communications require high capacity radio channels and user-centric communication network environments. Therefore, special requirements, such as wide-angle beam steering, wide operational frequency bandwidth and simultaneous operation with orthogonal polarizations are essential for telecommunication antennas. In this paper, we present a compact patch antenna array with a Butler matrix feed network for the frequency band of 26–31.4 GHz. The 16-element planar array has been designed to operate with two linear orthogonal polarizations and provide $\pm 42^\circ$ beam switching. To ensure wideband operation, a novel combination of planar couplers, crossovers and phase shifters is designed to form the Butler matrix. A new phase shifter topology is used which is based on a combination of open-short stubs and Shiffman phase shifter. The design is fabricated on a low-cost multi-layer board with a size of 120x70x1.62 mm³. The size of the feeding network, which is implemented on a single board, is 76x23x0.1 mm³. Experimental measurements of return loss, mutual coupling and radiation patterns confirm simulated results.

Index Terms—Millimeter-wave antenna array, wideband antenna array, Butler matrix.

I. INTRODUCTION

Future wireless telecommunication standards are expected to address the increasing demand in bandwidth (BW) and higher data transfer rates and will operate at millimeter-wave (mm-wave) frequency bands. For example, future Local Multipoint Distribution Service (LMDS) standard promises to cover the following bands: 27.5–28.35, 29.1–29.25, and 31–31.3 GHz [1], therefore, operational BW covering all the bands is more than 13%. Also, the telecommunication standards require adaptive antenna arrays that are capable of wide-angle beam steering [2, 3] to provide optimal speeds for devices, thereby creating a user-centric environment. If these arrays can operate with two orthogonal linear polarizations, the channel capacity can be enhanced and thus the spectral efficiency of communication links can be boosted.

Designing an antenna that operates with two orthogonal linear polarizations and provides beam-steering with a BW of more than 13% is quite challenging. For beam steering, there are two popular choices, (1) an active antenna array, and (2) a switched-beam antenna array. Although active components for millimeter-wave adaptive antenna arrays are commercially available [4, 5], the use of switched-beam antennas based on a beamforming network (BFN), such as Butler matrix (BM) or Rotman lens, can be a cheaper alternative at the cost of a larger size.

In recent years, some designs of dual-polarized antenna arrays with BM BFN were reported in [6–8]. Those designs consist of two separate BM BFN for each polarization. Therefore, a single polarized antenna array with BM BFN can be extended to a dual-

polarization mode through the second separate BFN and radiating elements that support dual-polarization mode. Some designs of mm-wave antenna arrays based on single layer BM BFN were reported recently in [9–16]. A comparison of important parameters such as practical implementation of BFN, impedance BW, polarization, electrical size, efficiency and beam switching angle range has been

TABLE I
COMPARISON OF BUTLER MATRIX BASED ANTENNA ARRAYS

Paper	BFN type	Pol	Size, λ	f_0 , GHz	BW, %	Eff, %	SLL, dB	BSR
[6]	M	DL	NA	10	6	51	6	$\pm 38^\circ$
[7]	M	DC	NA	7.83	15	NA	8	$\pm 40^\circ$
[8]	M	DC	NA	1.58 2.2	8 9	NA NA	6.5	$\pm 45^\circ$ $\pm 42^\circ$
[9]	M	L	NA	43	NA	NA	7	$\pm 31^\circ$
[10]	M	L	2.4x2	62.4	8.5	NA	8	$\pm 40^\circ$
[11]	M	L	NA	60.5	11.5	NA	5	$\pm 40^\circ$
[12]	M	L	2.75x2.5	61	13.5	NA	7	$\pm 40^\circ$
[13]	M	L	2x2	60.2	11.6	NA	7	$\pm 40^\circ$
[14]	SIW	L	NA	61	11.5	NA	8	$\pm 41^\circ$
[15]	SIW	L	6x2.5	30	13.3	NA	7.5	$\pm 51^\circ$
[16]	M	L	2.5x5	27.9	2	NA	6.5	$\pm 32^\circ$
This work	M	DL	2.4x 1.9	28.7	19	36	8	$\pm 42^\circ$

BFN type - microstrip (M) or Substrate integrated waveguide (SIW) implementation of BM BFN; Pol - operating polarizations: dual linear (DL), dual circular (DC), or linear (L); Size - overall size of BFN in fractions of center operating wavelength (λ); f_0 - central operation frequency; Eff - efficiency; and beam switching angles range (BSR).

done in Table I. From the comparison, we can see that antenna arrays with BM BFN published in [7, 12, 15] can cover 13% impedance BW, but all those BFN designs are characterized by huge amplitude (>4 dB) and phase ($>20^\circ$) imbalances, as well as low isolation (<12 dB) between output ports of the BFN inside the operational impedance BW. Such imbalances and isolation lead to asymmetrical beams with increased SLL, thus making the steering angle range narrower. Therefore, we are interested not only to get wide operational impedance BW, but also to get stable radiation patterns at all operation frequencies.

In this paper, we present a novel switched-beam planar patch antenna array with dual polarization for future mm-wave applications. The beam switching is achieved through the use of a wideband BM. The wideband operation covering 26–31.4 GHz is achieved by using novel wideband crossovers, couplers and phase shifters along with dual-fed superstrate loaded radiating patches. The planar patch antenna array consists of 16 elements (in 4 x 4 arrangement). Such arrangement is chosen to get the highest antenna area efficiency for a massive multiple-input multiple-output (MIMO) system [17]. Good agreement is achieved between the simulated and measured results. The antenna array demonstrates beam switching for angles of $\pm 42^\circ$ with a maximum gain of 12 dBi and 19% BW. The obtained 8 dB SLL is comparable with the published designs and is sufficient for most mobile telecommunication system applications [3].

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II. ELEMENTS OF THE ANTENNA ARRAY

A typical scheme of a 4-beam Butler matrix contains four 3-dB quadrature directional couplers, two crossovers, two 45° and two 0° phase shifters. The purpose is to feed the neighboring antenna elements with equal amplitudes and equally varying phase shifts. Different phase shifts ($\pm 45^\circ$ and $\pm 135^\circ$) appear when different ports are excited, and that causes the beam to switch in different directions.

Many papers have previously been published on the topic of Butler matrix, however most of these papers present multilayer designs (for example, [18]) to increase the BW of the feeding network. In this work, one of the targets is to make 5G base station antenna array which is low-cost and that can be easily of manufactured. Due to this reason, we have employed a single layer planar design for Butler matrix feed network.

A. The Design of The Wideband Quadrature 3-dB Directional Coupler

Some designs of planar wideband 3-dB quadrature directional couplers have been previously published in [19, 20]. The common disadvantage of such designs is the large size and the complicated fabrication required to realize them. One of the simplest design of a 3-dB coupler, which can offer wide BW, is an elliptic-patch quadrature-hybrid coupler (see Fig. 1a) [21, 22]. The advantages include wide BW, small size and simple fabrication in a single-layer implementation.

Although the design has been published previously, there are no recommendations on the proper selection of both the ports' positions and the eccentricity of the ellipse to get the widest BW. We have

Fig. 1. (a) the wideband quadrature 3-dB directional coupler design and (b) the wideband crossover design.

performed a parametric optimization of the 3-dB quadrature elliptic coupler using simulations of the design in ANSYS Electromagnetic Suite V.17.2. The main conclusions of the optimization are summarized as follows.

1. Resonance frequency f of the coupler is given by the equation [23] for an elliptical patch antenna:

$$f = \frac{3 \times 10^8 p}{a \sqrt{\epsilon}}, \quad (1)$$

where p is an empirical constant ranging from 0.56 to 0.6; ϵ is the permittivity of the coupler's substrate; a is the length of the major axis of an ellipse.

2. When the eccentricity of an ellipse is $e=1$ (the case of a disk), only 10% impedance BW can be achieved. To get 20-25% BW, the eccentricity should be chosen from the range of $e=0.6 \dots 0.8$.

3. When the coupler is fed through a $50\text{-}\Omega$ microstrip line, the feed point should be located on a radial line of the ellipse which is oriented at an azimuthal angle of $\Phi=45^\circ$ relative to the major axis of the ellipse. In the case of $50\text{-}\Omega$ feeding, the impedance BW is less than 15%. To increase the impedance BW up to 25%, the impedance of a microstrip feeding line should be decreased to $20\text{-}\Omega$. With decreasing value of the impedance, the azimuthal angle Φ should be increased to values of $\Phi=55 \dots 65^\circ$.

We optimized the design for the band of 26-31 GHz. The optimized design is based on the use of a Rogers Ultralam 3850 laminate, which has a permittivity of 2.9, loss tangent of 0.0025, and a thickness of 0.1 mm. We choose the parameters $p=0.59$ and $a=3.83$ mm in (1) to get resonance frequency of 27.1 GHz. According to the recommendations mentioned above, we choose an eccentricity of $e=0.793$ and a port angle of $\Phi=60^\circ$. The coupler is fed by a $20\text{-}\Omega$ microstrip line with a width of $W=0.84$ mm and a length of $L=0.756$ mm. However, to connect this coupler with the rest of the elements

Fig. 2. The S-parameters of the coupler.

in the Butler matrix, the width of this feed line must be changed to achieve $50\text{-}\Omega$ characteristic impedance. This can be clearly seen in Fig. 1, where narrowing width ($W_s=0.25$ mm) of the feed line ensures this transition from $20\text{-}\Omega$ to $50\text{-}\Omega$. A 3-dB quadrature directional coupler with these parameters was simulated, and the simulated results demonstrate a phase imbalance of less than $\pm 4^\circ$, an amplitude imbalance of less than 0.4 dB, and an isolation that exceeds 14 dBs in the frequency range of 26-31.5 GHz (Fig. 2).

B. The Design of The Wideband Crossover

A few designs of planar wideband crossovers, based on Lange coupler [24], have previously been demonstrated in [22, 25, 26]. However, these designs are difficult to fabricate because of the required electrical connections between different sections of the Lange coupler. A simpler design from fabrication perspective is a square patch with circular slots (see Fig. 1b) [27]. The wide operational band of the crossover is based on excitation of two resonances inside the square cavity. The first resonance is the resonance of modes TM_{100} and TM_{010} of the square patch. The second one is obtained by the circular slots.

The design has been optimized for a Rogers Ultralam 3850 laminate with a thickness of 0.1 mm for the band of 26-31 GHz. The length L_c is critical for the first resonance, and from (4) we calculate it to be 3.3 mm to get the first resonance at 26.7 GHz. The radius of the slot R_c and the widths $W1=W2$ help in achieving the second resonance. From simulations, $R_c=1.3$ mm and $W1=W2=0.15$ mm have been obtained as the best values to place the second resonance close to the first resonance. A crossover with these dimensions has been simulated. The simulated results (Fig. 3) show that the crossover has an isolation of more than 15 dB and insertion loss of less than 1.5 dB within the range of 26-31.5 GHz. Due to the

Fig. 3. The S-parameters of the crossover.

influence of the slot, we can see a slight shift ($\sim 3\%$) of the first resonance from the calculated value. It can be shown that in the absence of the slot, formula (4) gives very precise value of the resonance frequency of a square crossover. It is also observed in Fig. 3, that isolated ports (2 and 4) do not have ideal isolation. This is due to some minor reflected currents going into those isolated ports. Nonetheless, the isolation is below -15 dBs for the entire band of interest.

C. The Design of The Wideband Phase Shifter

As per the requirement of Butler matrix feeding network, phase shifters must provide stable (with low phase imbalance) 45° and 0° phase shifts as compared to the signal coming from the wideband crossover. Many previous phase shifter designs use a straight transmission line as a reference. However, in the proposed Butler matrix implementation, the square crossover with slots has a group delay that is quite different from that of a straight transmission line. The ratio of the group delay of the crossover to the group delay of a $50\text{-}\Omega$ microstrip line is shown in Fig. 4 (curve 1). From this figure, it can be seen that the group delay of the crossover is more than twice of that of the microstrip line. Therefore, well-known wideband planar designs of Shiffman phase shifter [28], Lange coupler based

Fig. 4. The group delay ratio.

phase shifters [29, 30], or T-shaped open stub [31] and open-short stub phase shifters [32] cannot be used in the proposed Butler matrix design as they provide unit group delay ratio, whereas this particular crossover design requires 2.5 group delay ratio. Therefore, a new design is required to achieve the required group delay ratio and subsequently a stable phase shift over the entire BW.

In order to resolve the above-mentioned issue and achieve a low phase imbalance for the frequency range of 26-31 GHz, we have come up with a planar design of phase shifter which is a combination of series connected elements. An open-short phase shifter (Fig. 5a) can create stable phase shift (i.e. unit ratio of group delays) as compared with a $50\text{-}\Omega$ microstrip line in the frequency band of more than 20%, as it has been shown in [32]. Series connection of two such phase shifters gives us the ratio of group delays equal to 2. Additional group delay to get the required 2.35-2.6 ratio is possible through a Shiffman phase shifter (Fig. 5b), whose group delay ratio (compared with a $50\text{-}\Omega$ microstrip line) is shown in Fig. 4 (curve 2). Thus a novel combination of two open-short-stub phase shifter elements with the Shiffman phase shifter (Fig. 5c) provides the

Fig. 5. (a) the open-short-stub phase shifter element geometry; (b) the Shiffman phase shifter geometry; (c) the overall proposed wideband phase shifter geometry; (d) the single element patch antenna element geometry.

required group delay ratio. This design allows to get a stable phase shift with low phase imbalance when the square patch crossover is a reference line.

Following conclusions can be deduced from the design process.

1. The width $W3$ of the transmission line in the open-short stub phase shifter must have a characteristic impedance of $25\text{-}35\text{ }\Omega$, provided the interconnecting lines have a characteristic impedance of $50\text{-}\Omega$ (represented by Ws in Fig. 5a). For the proposed design, $W3 = 0.52\text{mm}$ which corresponds to $30\text{-}\Omega$.
2. The width of the transmission line $W4$ in the Shiffman phase shifter is the same as the width $W3$. The gap width $W5$ should be 4-6 times smaller than the width $W4$. For the proposed design, $W5 = 0.1\text{mm}$.
3. The widths Wo and Wt of the open and shorted stubs respectively must adhere to an impedance ratio of $\sim 3:1$ to maximize the operational BW and to reduce the phase variation below 10%. For the proposed design, an impedance ratio of 2.7:1 has been utilized, whereas Wo and Wt correspond to 62 and $23\text{-}\Omega$ respectively.
4. The lengths Lo and Lt of the open and shorted stubs have been

calculated by the formula $\lambda / (8\sqrt{\epsilon})$, where λ is the center wavelength of the operational frequency band, ϵ is the permittivity of the phase shifter's substrate.

5. The height Hps of the Shiffman phase shifter is critical to achieve the required nominal of the phase shift.

The final dimensions of 0° and 45° phase shifter elements are listed in Table II. Phase shifter with the optimized dimensions has been simulated and the results are shown in Fig. 6. It can be seen that the phase shifter has an insertion loss of less than 0.6 dB for the entire range of 26-31.5 GHz. Phase shift of the wideband phase shifter has been calculated by subtracting the phase of S_{13} (ports for the

TABLE II
DIMENSIONS (IN MM) OF 0° AND 45° PHASE SHIFTERS

Phase shift	Lps	Hps	Wo	Lo	Wt	Lt	W3	D
0°	4.7	4.11	0.18	0.84	0.75	0.74	0.52	0.2
45°	4.5	4.3	0.18	0.775	0.7	0.775	0.52	0.2

crossover) from phase of S_{56} (ports for the phase shifter). From Fig. 6 we can see that phase deviation of the phase shifter is less than $\pm 5^\circ$ from the required 0° and 45° phase shifts for the entire frequency range of 26-31.5 GHz. In the following section, with simulated and

Fig. 6. The S-parameters of the phase shifter.

measured results for a complete design of antenna array, we will show that the $\pm 5^\circ$ phase deviation does not affect the BSR and the SLL.

D. The Design of The Wideband Radiating Element

For this particular design, we need a wide band antenna element that can support two orthogonal linear polarizations over the frequency range of 26-31.5 GHz and have a broad radiation pattern. One possible antenna that satisfies the requirements on dual-polarization operational mode as well as broad radiation pattern is a rectangular or circular patch. But, such patch antennas can typically provide less than 7% BW [23]. In the proposed patch design (a square patch antenna with semi-circular cut-outs) [33], as shown in Fig. 5d, can fulfill the large BW requirement. The patch antenna is in the form of a square plate of length $L_p=2.65$ mm that has angled circular cutouts of the radii $R_p=0.95$ mm. The radiating element and the superstrate have been implemented on a Rogers RO4533 laminate, which has a permittivity of 3.3, loss tangent of 0.0025, and a thickness of 0.76 mm. The patch is excited by two metallic pins with diameters of 0.416 mm, which are located at distances of 0.78 mm from the center of the patch (along the X and Y-axes), through circular holes with a diameter of 1 mm in the metallic ground plane.

The simulated results for this patch antenna are shown in Fig. 7. Due to symmetry of the patch in ZX and ZY-planes, $|S_{11}|=|S_{22}|$ and $|S_{12}|=|S_{21}|$. From Fig. 7, we can see that the antenna is well matched for the desired frequency range and the isolation between the two ports is better than 12 dB.

III. RESULTS AND DISCUSSIONS

Fig. 7. The S-parameters of the patch element.

A. The Design of The Antenna Array

A 16 elements antenna array has been designed which requires three boards (Fig. 8). The 1st board, based on Ultralam 3850 laminate, contains the feeding network, the 2nd board has the planar 16 element (4x4) radiating patches and the 3rd is the superstrate board. Both, the second and third board, are based on Rogers RO4533 laminate. Each board has a size of 70x120 mm². The patch inter-element spacing is 5.5 mm (0.57 wavelength for the frequency of 31.4 GHz) in both X and Y-directions. The electrical contact

between the feeding network and the radiating patches is created by metallized vias between the 1st and the 2nd board.

For simultaneous operation with two orthogonal linear polarizations, we feed the radiating patches by two identical independent Butler matrices (Fig. 9 (a)). Each Butler matrix has four ports to create four beams. Each output microstrip line from the

Fig. 8. (a) the boards' view of 16-elements antenna array; (b) PCB stack up view ($h_1=h_2=0.76$ mm, $h_3=0.1$ mm).

Butler matrix feeds a 4-element linear patch array via a power divider (Fig. 9 (b)) with non-uniform amplitude distribution. The normalized amplitudes at the outputs of the power divider are: $A_2=A_3=1$, $A_1=A_4=0.6$. The non-uniform amplitude distribution is used to provide low SLL in the radiation pattern in the plane that is orthogonal to the plane of beam scanning. When ports 1-4 are excited, the antenna array operates with main polarization of the electrical field in the Y-direction. When ports 5-8 are excited, the main polarization of electrical field is in the X-direction.

B. Simulated and Measured Results

A prototype of the antenna array (Fig. 10) has been fabricated and measured in a far field chamber. The fabrication process includes drilling vias in the 1st and the 2nd boards with subsequent metallization of the vias. The final manufacturing is performed by bonding all the three boards using Rogers 4450F bondply with a

Fig. 9. (a) the feeding network of the dual-polarized antenna array; (b) the power divider.

Fig. 10. Photo of the fabricated prototype: (a) layer of radiating patches; (b) layer of the feeding network.

thickness of 0.1 mm as prepreg. The prototype is fed by mini-SMP connectors. During the measurements, all unused connectors are loaded by 50- Ω terminations.

Simulated and measured S-parameters (S_{nm} ; $n, m=1, 2, \dots, 8$) of the array are presented in Fig. 11. From Fig. 11 (a), we can see that the reflection coefficients for ports 1-4 are less than -10 dB for the frequencies of 26-31.4 GHz (the BW is 19%). The same BW we obtain for ports 5-8. Fig. 11 (b) show isolation between the ports of Butler matrix 1 (note that $|S_{12}|=|S_{21}|$, $|S_{13}|=|S_{31}|$, $|S_{14}|=|S_{41}|$). From this figure, we can see that isolation is more than 15 dB for the working frequency band. Measurements show that isolation between ports 2-3, 2-4 and 3-4 is also more than 15 dB. Measured curves of isolation

Fig. 11. (a) reflection coefficients for ports 1-4; (b) isolation between ports 1-2, 1-3, 1-4; (c) isolation between ports 1-8, 2-7, 3-6, 4-5.

between ports of Butler matrix 2 are similar to the curves of isolation between ports of Butler matrix 1, therefore they are not shown in a separate figure. Fig. 11 (c) shows isolation between ports of the Butler matrixes 1 and 2. Isolation between ports 1-8, 2-7, 3-6 and 4-5 is better than 25 dB in the working frequency band. For other ports, isolation between two Butler matrixes is more than 30 dB.

Fig. 12 shows the radiation patterns of the antenna array in the XZ-plane at frequencies of 26, 28.5, and 31 GHz when different ports are excited. The angle θ in the figure is measured from the Z-axis. From Fig. 12 we can see that the directions of the main beam are -42° (-42°), -13° (-13°), 13° (13°) and 42° (42°) when port 2 (6), 4 (8), 1 (5), and 3 (7) are excited, respectively. Maximum gain is 12 and 9 dBi for the beams with $\pm 13^\circ$ and $\pm 42^\circ$ main directions, respectively. The SLL in XZ-plane is -10 and -8 dB for the beams with $\pm 13^\circ$ and $\pm 42^\circ$ main directions, respectively. Due to non-uniform amplitude power dividers of the feeding network, the SLL in the YZ-plane is at least 15 dB. Measurements show that the cross-polarisation levels are 16 dB lower as compared to the maximum of the co-polarised patterns, for all ports inside the operating frequency band of 26-31.4 GHz.

From the simulated and measured results we can see that the proposed design of the antenna array demonstrates wide impedance BW as well as wide BW for stable beam steering.

Additional investigation has been performed on the efficiency of the array. This investigation shows that the efficiency is in the range of 29-36% inside the operational frequency band. It is worth mentioning here that the efficiency of a single radiating element is more than 90% inside the operational band [33]. Thus, the microstrip feed network losses (~ 4 dB) at such high frequencies are responsible for such relatively low efficiency.

Note that BFN has been realized in microstrip line implementation that leads to its potential radiation and interference with electronics placed behind the antenna array. To avoid the interference issue and

Fig. 12. Radiation patterns of the antenna at the frequency of (a, b) 26 GHz; (c, d) 28.5 GHz; (e, f) 31 GHz.

get high isolation, the BFN can be implemented as a stripline where an additional board with ground plane is added to the bottom of the board 1. Investigation of a stripline based BFN shows that it can provide the same BW and coupling characteristics as its microstrip counterpart, without distorting the radiation pattern.

IV. CONCLUSIONS

In this paper, we propose the design of a 16-element antenna array covering the bands 26-31.4 GHz, which operates with two orthogonal linear polarizations and provides $\pm 42^\circ$ beam switching. The antenna array is based on a Butler matrix feed network and wideband patch radiating elements. The Butler matrix design is implemented by planar wideband crossovers, couplers and phase shifters. The proposed design provides better operating bandwidth among the published switched-beam Butler matrix based antenna array designs. The construction is fabricated by three boards of low-cost Rogers laminates.

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