A new 1 × 4 phase distribution network (PDN) featuring a fully controllable progressive phase shift between outputs is proposed for continuous beam-scanning arrays. The PDN is made up of two parts: a modified 2 × 4 Butler matrix integrated with four phase shifters (PSs), and a tunable power divider (TPD) whose power division ratio can be controlled over a wide tuning range. The synthesis equations show that the relative phase shift between PDN outputs can be fully controlled by the TPD and embedded PSs without using an external single-pole quad-throw (SP4T) switch. The proposed PDN is demonstrated at 2.4 GHz as the feed network of a 4-element linear array. The experimental results display a fully controllable progressive phase shift (from $-180^\circ$ to $180^\circ$) between PDN outputs over a 20% bandwidth with good performance of matching, power division, and relative phase shifts. A spatial coverage of 116° with the feature of continuous beam scanning and negligible dc power consumption is achieved. Benefitting from the single-input topology, a planar 16-element phased array for 2D beam scanning is then realized by simply stacking and cascading five instead of eight PDN modules. It removes the high-cost and bulky SP16T switch. Experimental results demonstrate the uniqueness of the proposed designs.

INDEX TERMS Antenna arrays, beam-steering, Butler matrix, phase shifters, tunable circuit and devices.

I. INTRODUCTION

Beam-forming networks (BFNs), an essential part in beam switching and beam scanning phased arrays, have become the core subsystem in fifth-generation (5G) communication systems and beyond [1]. Unlike the feed network for continuous beam scanning, the BFNs for beam switching arrays, or equivalently the multi-beam arrays, mostly rely on passive components without using the costly and lossy commercial phase shifters. They become a good trade-off solution in terms of cost, insertion loss, and beam agility. While the spatial resolution is compromised when compared to their continuous scanning counterparts, the overall cost and loss are significantly reduced. In general, 4-16 beams can be achieved by a multi-beam phased array. Among the well-known multi-beam arrays, the Butler matrix (BM) [2] offers various advantages in terms of the operating bandwidth, loss, and simplicity when compared with other solutions such as the Blass matrix [3], Nolen matrix [4], and Rotman len [5]. An $N \times N$ Butler matrix ($N = 2^n$ with $n = 1, 2, 3 \ldots$) can typically generate $N$ orthogonal beams [6]–[8], which set the limitation in spatial resolution. To improve the beam agility, a high-order Butler matrix can be implemented at the expense of circuit complexity and insertion loss.

Various studies have been conducted to alleviate the design complexity of a Butler matrix, specifically the constraints imposed by the crossovers [9]–[21]. Multi-layered printed circuit board (PCB) [9]–[11], low-temperature co-fired ceramic (LTCC) [12], and CMOS technologies [13], [14] are used to realize crossovers with compact footprints. Rearranging the network deployment as a planar array can reduce the number of crossovers [15], [16]. Without additional fixed phase shifters, hybrid couplers with arbitrary phase differences are utilized to realize Butler matrices with alternative beam sets as in [17] and [18]. Nonetheless, the number of beams in all aforementioned complexity-reduced BFNs is still limited to $N$. To augment the beam set, embedding reconfigurable components into

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a conventional Butler matrix is a viable and cost-effective solution [19]–[21]. The beam controllability of a common 4th-order Butler matrix can be greatly boosted by factors of two, three, and four by using embedded switched-line type phase shifters (PSs) [19], discrete reconfigurable couplers [20], and phase reconfigurable synthesized transmission lines (PRSTLs) [21], respectively. Better spatial resolution with a low beam-cross level is achieved to fulfill seamless coverage.

From another point of view, the Nth-order Butler matrix still requires an additional single-pole multiple-throw (SPNT) switch to select the excitation port from the common input. It makes the system not fully integrated and costly. Also, the SPNT component could be non-planar and bulky in size. An active Butler-matrix-based receiver with a common input is developed as an alternative solution showing continuous beam-scanning property [22]. It is fulfilled by the subsector beam-steering method, whose weighting factor controller is a narrowband solution with considerably high dc power consumption. Also, it experiences amplitude imbalance during scanning, which can be verified from (9) of [22].

To fill up the gap, a novel phase distribution network (PDN), fed by a single common input and featuring four outputs (1 × 4) with a fully controllable progressive phase shift and equal power splitting, is proposed in this paper for beam-scanning arrays. The proposed 1 × 4 PDN consists of a modified 2 × 4 Butler matrix and a tunable power divider (TPD). Without using an external single-throw four-way (SP4T) switch, the inputs of the modified Butler matrix are integrated into and controlled by the TPD. The tunable progressive phase shift is fulfilled by the TPD along with the PSs embedded in the 2 × 4 Butler matrix. As a proof-of-concept, the 1 × 4 PDN is first designed as the feed network of a 4-element linear array at 2.4 GHz. Over a 20% relative bandwidth, the measured circuit responses exhibit a fully controllable progressive phase shift from −180° to 180° with good performances in terms of the matching, power division, and relative phase shift. The main beam can be continuously scanned over a spatial coverage of 116° with negligible dc power consumption. Benefiting from the single-input topology, a planar 16-element phased array for 2D beam-scanning is then realized with a reduced number of PDN modules. It thereby dramatically reduces the complexity of a scanning antenna system since the high-cost and bulky SP16T switch is removed, and the number of modules is reduced from eight (4 vertical plus 4 horizontal ones) [23]–[24] to five (one vertical plus 4 horizontal ones).

This paper is organized as follows. The operating principle of the proposed 1 × 4 PDN is discussed in Sec. II, followed by the circuit implementations of the building blocks in Sec. III. Experimental validation, including the S-parameters of the feed network as well as the radiation characteristics of the array patterns, are provided in Sec. IV along with a discussion on the power handling capability. Finally, the 2-D beam-scanning array is demonstrated in Sec. V.

II. OPERATING PRINCIPLES

Fig. 1 shows the schematic diagram of the proposed 1 × 4 PDN. The modified 2 × 4 BM is originated from the conventional topology of a 4th-order one with modifications to reduce the structural complexity. Specifically, the crossover at the output side is removed, and the number of inputs is reduced to a half by replacing the pair of 90° hybrids with two-way equal power dividers (2-way PDs). The inputs of the 2 × 4 BM are co-excited by the TPD whose power ratio between signals s1 and s2 can be adjusted over a very wide range in either the in-phase mode (Σ-mode) or out-of-phase mode (Δ-mode). The two signals are processed by the 2 × 4 BM so as to achieve the desired progressive phase shift (Δφ) with equal power splitting at the output ports.

Let us assume a1 = 1 represents the normalized input signal and the TPD is operated in the Σ-mode to split the input signal into two in-phase signals s1 and s2 with a power division ratio (PDR) k²

\[
PDR = k^2 = \frac{s_1^2}{s_2^2} \quad \text{and} \quad s_1^2 + s_2^2 = 1, \tag{1}
\]

where 0 ≤ k² ≤ ∞ (i.e., −∞ dB to ∞ dB). The signals s1 and s2 are then equally split by a pair of 2-way PDs and their phases are shifted by groups of phase shifters PSx and PSy. The signals s1x, s2x, s1y, and s2y entering the 90° hybrid couplers are

\[
s_{1x} = \frac{s_1}{\sqrt{2}} e^{-j\phi_x} \quad \text{and} \quad s_{2x} = \frac{s_2}{\sqrt{2}} e^{-j\phi_x}, \tag{2a}
\]

\[
s_{1y} = \frac{s_1}{\sqrt{2}} e^{-j\phi_y} \quad \text{and} \quad s_{2y} = \frac{s_2}{\sqrt{2}} e^{-j\phi_y}. \tag{2b}
\]

ϕx and ϕy are the phase shifts due to PSx and PSy, respectively. The outputs b2, b3, b4, and b5, by manipulating s1x, s2x, s1y, and s2y using 90° hybrid couplers, are given as

\[
b_2 = \frac{s_{1x} e^{-j\pi/2} + s_{2x} e^{-j\pi}}{2} = -\frac{e^{-j\phi_x}}{2} (s_1 + s_2), \tag{3a}
\]

\[
b_3 = \frac{s_{1x} e^{-j\pi} + s_{2x} e^{-j\pi/2}}{2} = -\frac{e^{-j\phi_y}}{2} (s_1 + js_2), \tag{3b}
\]

\[
b_4 = \frac{s_{1y} e^{-j\pi/2} + s_{2y} e^{-j\pi}}{2} = -\frac{e^{-j\phi_y}}{2} (js_1 + s_2), \tag{3c}
\]
practical building blocks. Accordingly, a wideband solution is evolved from the designs in [25]–[27]. In this work it is realized by varactor-based components including a reconfigurable directional coupler and an absorptive SPDT (single-pole-double-throw) switch, as depicted in Fig. 3. The reconfigurable directional coupler, with PSs as the tuning element, is evolved from the designs in [25]–[27]. Without using an active tunable amplifier or attenuator as in [22], it is capable of splitting the input signal $a_1$ into $s_1$ and $s_2$ with a wide PDR tuning range and switchable phase difference ($\angle s_1 - \angle s_2 = 0^\circ$ or $180^\circ$). Also, to overcome the narrowband limitation in [25] and [26], the new topology in Fig. 3 adopts a wideband PS to be detailed in the following subsection.

### A. WIDEBAND VARACTOR-BASED PS

The circuit schematic of the wideband PS is depicted in Fig. 4(a). It is made up of a cascade connection of a left-handed (LH) and a right-handed (RH) phase reconfigurable synthesized transmission lines (PRSTLs) [21]. In [25] and [26], the original PRSTL was designed as a tunable PS between two bounded states ($\theta_{L1}$ and $\theta_{L2}$, or $\theta_{R1}$ and $\theta_{R2}$). By applying ABCD matrices to the left- and right-handed PRSTLs separately, their relative phase shifts between bounded states, $\Delta \theta_L = \theta_{L1} - \theta_{L2}$ (LH) and $\Delta \theta_R = \theta_{R1} - \theta_{R2}$ (RH), are derived as

\[
\Delta \theta_L = \cos^{-1}\left(1 - \frac{1}{\omega^2 L_L C_v1}\right) - \cos^{-1}\left(1 - \frac{1}{\omega^2 L_L C_v2}\right),
\]

\[
\Delta \theta_R = \cos^{-1}\left(1 - \frac{1}{\omega^2 L_R C_v1}\right) - \cos^{-1}\left(1 - \frac{1}{\omega^2 L_R C_v2}\right).
\]

$C_v(i = 1, 2)$ is the value of the varactor diode in the corresponding state $\theta_{L(R)i}$. Here, $\theta_{L(R)i}$ are all positive and $\theta_{L1} < \theta_{L2}$ and $\theta_{R1} < \theta_{R2}$. Using (6), the relative phase shifts of selected examples of pure RH and LH PRSTLs are plotted in Fig. 4(b). Clearly, the slopes of both curves are nonzero over the band of interest, suggesting the narrowband nature of selected examples of pure RH and LH PRSTLs are plotted.

### III. DESIGN AND IMPLEMENTATION OF THE TPD

The proposed PDN was developed on a 1.524-mm RO4003C substrate ($\varepsilon_r = 3.55$ and $\tan\delta = 0.0027$). The center frequency for demonstration is chosen as 2.4 GHz; nonetheless, another frequency covering the sub-6G 5G mobile systems can be selected using the same design concept.

The TPD is responsible for adjusting the power ratio and phase difference between $s_1$ and $s_2$ (Fig. 1) to attain wide PDR tuning range in both $\Sigma$- and $\Delta$-modes. To keep design simplicity with low cost and low dc power consumption ($P_{dc}$), the tuning element is evolved from the designs in [25]–[27]. Without using an active tunable amplifier or attenuator as in [22], it is capable of splitting the input signal $a_1$ into $s_1$ and $s_2$ with a very wide PDR tuning range and switchable phase difference ($\angle s_1 - \angle s_2 = 0^\circ$ or $180^\circ$). Also, to overcome the narrowband limitation in [25] and [26], the new topology in Fig. 3 adopts a wideband PS to be detailed in the following subsection.

![FIGURE 2. Output signals ($b_2, b_3, b_4,$ and $b_5$) in the complex plane when two signals $s_1$ and $s_2$ are excited: (a) in-phase and (b) out-of-phase.](image)

\[
\Delta \phi = \angle b_5 - \angle b_3 = \angle b_5 - \angle b_4 = \tan^{-1}\left(k^{-1}\right) - \tan^{-1}\left(k\right) = \frac{\pi}{2} - 2\tan^{-1}\left(k\right).
\]

It is clear from (4) that the phase difference $\Delta \phi$ can be controlled by the ratio $k$ between $s_1$ and $s_2$. If the two signals are in-phase excited ($\Sigma$-mode), $\Delta \phi$ is adjustable between $-\pi/2$ and $\pi/2$. The complement set of $\Delta \phi$, from $-\pi$ to $-\pi/2$ and from $\pi/2$ to $\pi$, can be fulfilled if the TPD is operated in the $\Delta$-mode such that $s_1$ and $s_2$ are excited with a $180^\circ$ phase difference. The result is depicted in Fig. 2(b).

Furthermore, according to (3) the magnitude of the output signal ($b_2$ to $b_5$) is always kept equal to 0.5 (or $-6$ dB), which implies the input signal $a_1 = 1$ is equally divided into four parts despite of the selection of $k^2$. This feature makes the proposed concept outperforms the subsector beam-steering method in [22], whose amplitude imbalance is significant upon steering the main beam.

Finally, a constant progressive phase shift $\Delta \phi$ between adjacent PDN outputs can be guaranteed by setting the relative phase shift of the embedded phase shifters $\Phi_1$ and $\Phi_2$ as

\[
\Phi_y - \Phi_x = 2\Delta \phi
\]

such that $\angle b_3 - \angle b_5 = \Phi_y - \Phi_x - \Delta \phi = \Delta \phi$.

In other words, by properly manipulating the TPD along with embedded PSs, a fully controllable progressive phase shift $\Delta \phi$ can be achieved with equal output amplitude for continuous beam-steering. Note that (1)-(5) are frequency independent and the bandwidth of the PDN is limited by practical building blocks. Accordingly, a wideband solution is highly demanded.

![FIGURE 3. Block diagram of the tunable power divider.](image)
of the original design. A quantitative evaluation is derived at $f_0$ as

$$\frac{\partial (\Delta \theta_L)}{\partial \omega} \bigg|_{\omega_0} = \frac{2}{\omega_0} \left[ \tan \left( \frac{\theta_{L1}}{2} \right) - \tan \left( \frac{\theta_{L2}}{2} \right) \right], \quad (7a)$$

$$\frac{\partial (\Delta \theta_R)}{\partial \omega} \bigg|_{\omega_0} = -\frac{2}{\omega_0} \left[ \tan \left( \frac{\theta_{R1}}{2} \right) - \tan \left( \frac{\theta_{R2}}{2} \right) \right]. \quad (7b)$$

The slopes reveal an opposite trend of variations by the negative sign in (7b). Accordingly, a cascade connection of the two units gives the opportunity of obtaining a flat response over the band of interest (i.e. with zero-slope condition). In addition, from (7) the slope is exactly equal to zero when $\theta_{L1} = \theta_{R1}$ and $\theta_{L2} = \theta_{R2}$, which in turn suggests that the LH and RH PRSTLs should be designed to have the same phase tuning range. The relative phase shift of the cascaded unit is also plotted in Fig. 4(b). A flattened response over the band is observed, therefore significantly widening the error tolerance bandwidth ($\delta$) and hence operating bandwidth when compared with conventional loaded-line PSs using LH and/or RH structures [25], [26], [28], [29].

As a demonstration, a wideband 90° PS, serving as the tuning element of the reconfigurable coupler in the TPD, was designed in Fig. 5(a) by a cascade connection of a LH and a RH PRSTLs. Each PRSTL has a 45° phase tuning range. By applying the design procedure in [21] with $f_0 = 2.4$ GHz, the calculated component values are $L_L = 3.59$ nH, $L_R = 3.06$ nH, $C_{v1} = 0.89$ pF, and $C_{v2} = 1.98$ pF. The PS is implemented using microstrip technology with varactor diodes SMV1405 from Skywork. The choking inductor (RFC) and blocking capacitor ($C_b$) are 33 nH and 15 pF from Murata Manufacturing, respectively. Only one bias voltage $V$ is required since all varactors are controlled at the same time in accordance with (7) to maximize the bandwidth.

The measured electrical responses of the proposed PS are depicted in Figs. 5(b) and (c), with a photo of the fabricated sample shown as the inset. Prior to the measurement, the capacitance versus voltage (C-V) curve of the varactor SMV1405 was firstly obtained as a look-up table. The bias voltages are then determined based on the C-V curve of the varactor diode, and the required capacitance value in the simulation. The bias voltages listed in Fig. 5(c) correspond to the capacitance values of 0.64 pF (20 V), 0.68 pF (16.7 V), 0.74 pF (13.1 V), 0.81 pF (10.4 V), 0.89 pF (8.3 V), 0.99 pF (7.1 V), and 1.12 pF (5.8 V). The wideband phase response is achieved while maintaining good matching and insertion loss. Specifically at 2.4 GHz, the measured $|S_{11}|$ is lower than -20 dB, and measured $|S_{21}|$ is -0.8 dB in average. Over the bandwidth from 2.1–2.7 GHz, an average insertion loss of 1 dB with a maximum phase error of $\leq 12^\circ$ is observed.

FIGURE 4. (a) Circuit schematic of the new varactor-based PS and (b) the relative phase shifts of selected examples of RH, LH and proposed (cascaded) PRSTL PSs.

FIGURE 5. The proposed tunable PS: (a) layout, (b) measured reflection and transmission coefficients, and (c) simulated and measured relative phase shifts. Inset: photo of the fabricated prototype.
FIGURE 6. The proposed tunable PS: (a) layout, (b) measured reflection and transmission coefficients, and (c) simulated and measured relative phase shifts. Inset: photo of the fabricated prototype.

The simulated and measured relative phase shifts in Fig. 5(c) also agree with one another very well. In addition to provide wider bandwidth than its conventional loaded-line counterparts, the proposed varactor-based PS also yields a simple structure with compact size when compared with other tunable PSs such as the vector-sum PS [30], the reflection-type PS [31], and the liquid crystal-based PS [32].

B. RECONFIGURABLE COUPLER

The tunable PSs are inserted between two 90° six-finger Lange couplers to fulfill the reconfigurable coupler in Fig. 3. The layout is shown in Fig. 6(a). The two inputs of the coupler are selected by an absorptive SPDT, while the coupler outputs are the outputs of the TPD. During the reconfiguration, one input is selected (excited) while the other one is terminated by a matched (absorptive) load.

The simulated and measured S-parameters of the coupler with \( k^2 = -10 \, \text{dB} \) (\( \Delta \)-mode) are depicted in Fig. 6(b) as an example; a photo of the fabricated sample is shown as the inset. The measured bias voltages, selected from the C-V curve, are \( V_A = 11.2 \, \text{V} \) and \( V_B = 20 \, \text{V} \). Good agreement between simulated and measured results is observed. Over a bandwidth from 2.1–2.7 GHz, the insertion loss is read as 1 dB with a good matching (\( |S_{11}| \leq -17 \, \text{dB} \)) and isolation (\( |S_{31}| \leq -14 \, \text{dB} \)); the variation of PDR and phase error are \( \leq 1.5 \, \text{dB} \) and \( 11^\circ \), respectively. Such a wide bandwidth cannot be feasible by previous techniques [25]–[26] while the design in [27] is a high-cost and high-loss solution. Fig. 6(c) further illustrates a very wide PDR tuning range with a flat response over the bandwidth of interest. It is worth mentioning that in the extreme case, say \( k^2 \geq 20 \, \text{dB} \) or \( k^2 \leq -20 \, \text{dB} \), the coupler serves either as a crossover (\( |S_{31}| = |S_{42}| = 1 \)) or as a pair of isolated lines (\( |S_{31}| = |S_{43}| = 1 \)), suggesting that the remaining ports are satisfactorily isolated.
FIGURE 8. The proposed PDN: (a) photo of the fabricated prototype; electrical responses of selected discrete states: (b) measured input and output matching; (c) simulated and measured transmission coefficients, and (d) simulated and measured progressive phase shifts.

C. ABSORPTIVE SPDT SWITCH

The absorptive SPDT is responsible for selecting one input of the reconfigurable coupler as the excitation while terminating the other one to avoid reflections. As available commercial SPDTs cannot provide adequate absorption as requested, the switchable synthesized line in [33], inherently serving as an on/off RF switch, is extended to fulfill the design. The layout of the absorptive SPDT is shown in the left of Fig. 7(a). It consists of two pairs of switches A and B and a 50-Ω resistor inserted between the pair of switches B to bridge the outputs (ports 2 and 3).

The operating principle is depicted in the right side of Fig. 7(a). The corresponding circuit schematics of switches A and B are depicted in Figs. 7(b) and (c), respectively. On one hand, when a switch is operated in the on-state, the series tank \( L_2 C_{sw} \) \((i = 1, 2)\) is off-resonance in the capacitive region to form a right-handed network to mimic a uniform 50-Ω transmission line. On the other hand, in the off-state, the series tank at the center of a switch should resonate and create a virtual short circuit. By resonating \( L_1 \) and \( C_1 \) at the same time, the switch A assembles an open circuit when looking into either side, whereas the switch B creates an open circuit at the left input terminal and a virtual short circuit at the right one. The inductor \( L_s \) in switch B guarantees good matching in the on-state such that only one universal bias voltage is needed for both switches. The design procedure of switches A and B has been detailed in [33].

The electrical responses of the SPDT when operated in state 1 \( (|S_{21}| = 0 \text{ and } |S_{31}| = 1) \) are plotted in Fig. 7(d); the bias voltages for Switches A and B are 25 V and 3 V, respectively, in the measurement. Good input and output matching \( (|S_{11}|, |S_{22}|, \text{ and } |S_{33}| \leq -15 \text{ dB}) \) and isolation \( (|S_{32}| \leq -26 \text{ dB}) \) are observed over the frequency range of 2.1–2.7 GHz. The average insertion loss in the active path is 0.8 dB while the signal suppression in the inactive path is up to 40 dB at the center frequency and higher than 16 dB over the band. Similar responses are obtained in state 2 \( (|S_{21}| = 1 \text{ and } |S_{31}| = 0) \), but not shown here to avoid redundancy.

IV. LINEAR 1 × 4 BEAM-SCANNING ARRAY

A. LAYOUT AND CIRCUIT RESPONSES

The TPD, together with Wilkinson power dividers (WPDs), six-finger Lange couplers, and a crossover by microstrip-to-conductor-backed coplanar waveguide transition [34] are integrated together to realize the proposed 1 × 4 PDN,
as shown in Fig. 8 (a). Apart from the TPD, two groups of additional phase shifters \( PS_x \) and \( PS_y \), as suggested by (5), are required to achieve a fully reconfigurable progressive phase shift \( \Delta \phi \) at the outputs. The relative phase difference between \( PS_x \) and \( PS_y \) should range from \(-180^\circ\) to \(180^\circ\), which can be realized by a pair of tunable PSs each with a phase tuning range of \(180^\circ\). Accordingly, a cascade connection of two \(90^\circ\) PSs presented in Sec. III.A is adopted to fulfill \( PS_x \) and \( PS_y \). The measured responses show a good matching (\(|S_{11}| \leq -10 \text{ dB}\)) with an average insertion loss of 1.7 dB and a phase error less than \(16^\circ\) over the bandwidth from 2.1–2.7 GHz.

The experimental validation starts from the circuit responses (S-parameters), followed by the radiation patterns of a 4-element linear quasi-Yagi array fed by the PDN. During the measurement, the reconfiguration was completed by controlling the bias voltages of the TPD and PSs without using SP4T switches as in previously reported counterparts. Since all varactor diodes are reversely biased, the reversed current is of the order of nA and the dc power consumption \( P_{dc} \) is negligible (21 \( \mu \)W).

Simulated and measured responses of the PDN in discrete states are selected and plotted in Figs. 8(b)–(d) for demonstration. In general, the agreement between simulated and measured results is good. The measured input and output return losses are better than 11 dB while an average transmission coefficient of \(-10 \text{ dB}\) is observed over a 20% FBW (2.16 – 2.64 GHz). The amplitude imbalance is \(\pm 1.4 \text{ dB}\) while the maximum phase error is \(9.3^\circ\) at 2.4 GHz and \(18^\circ\) over the bandwidth. As will be presented shortly, although the phase error becomes worsened in some states, the deviation of the main beam direction from the calculated one is less than \(1.5^\circ\).

B. RADIATION CHARACTERISTICS
A 4-element 0.5 \( \lambda_0 \)-spaced quasi-Yagi array fed by the proposed 1 \( \times \) 4 PDN is adopted to validate the radiation characteristics. The measurement was performed in a spherical anechoic chamber with NSI-2000 software from Nearfield Systems Inc. The antennas were connected to the PDN by
TABLE 1. Comparison of various butler-matrix-based BFNs.

| Ref. | Architecture (Process) | FBW (%) | No. of inputs | No. of beams | Control range of $\Delta \phi$ | Max. amplitude error (dB) | Max. phase error | Trans. gain/loss (dB) | Array peak gain (dB) | 3-dB spatial coverage | $P_{ae}$ (mW) | No. of modules required for 2D BFN |
|------|------------------------|---------|---------------|--------------|-------------------------------|---------------------------|-------------------|-------------------|-------------------|---------------------|-------------|----------------------------------|
| [6]  | BM ($N = 4$) (PCB)     | N.A     | 4             | 4            | $\pm 135^\circ$               | $\pm 7.5^\circ$           | $\pm 1.5$        | -2.6$^a$          | 6.4               | 132$^a$             | 0           | 2$N$ (8)                          |
| [8]  | BM ($N = 4$) (IPD)     | 8       | 4             | 4            | $\pm 135^\circ$               | $\pm 14^\circ$            | $\pm 1.1$        | -3.5$^a$          | N.A               | N.A                 | 0           | 2$N$ (8)                          |
| [14] | BM ($N = 8$) (CMOS)    | 18.2    | 8             | 8            | $\pm 157.5^\circ$             | N.A                       | N.A               | -5.5$^a$          | 4.8               | 98$^b$              | 0           | 2$N$ (16)                         |
| [15] | BM ($N = 8$) (PCB)     | 33      | 8             | 8            | $\pm 180^\circ$               | $\pm 10^\circ$            | $\pm 0.5$        | -1$^a$            | N.A               | 74$^c$              | 0           | 2$N$ (16)                         |
| [19] | BM ($N = 4$) + 1-bit PS (PCB + SMD switches) | N.A | 4             | 8            | $\pm 157.5^\circ$             | N.A                       | $\pm 2.3$        | -1.8$^a$          | N.A               | N.A$^c$             | N.A         | 2$N$ (8)                          |
| [20] | BM ($N = 4$) + switchable coupler (PCB + PIN diodes) | 30 | 4             | 12           | $\pm 150^\circ$               | $\pm 12^\circ$            | $\pm 1.7$        | -1.8$^a$          | N.A               | N.A$^c$             | 120         | 2$N$ (8)                          |
| [21] | BM ($N = 4$) + 2-bit PS (PCB + varactors) | 4.2 | 4             | 16           | $\pm 180^\circ$               | $\pm 20^\circ$            | $\pm 1.2$        | -1.7$^a$          | 10                | 118$^e$             | 0.008       | 2$N$ (8)                          |
| [22] | BM ($N = 4$) + sub-sector modules (CMOS) | 4 | 1 | Cont. | $\pm 180^\circ$ | $\pm 15^\circ$ | $\pm 3.8^e$ | 15$^e$ | N.A$^e$ | N.A$^e$ | 30 | $N + 1$ (5) |
| This work | Modified BM ($N = 4$) + TPD (PCB + varactors) | 20 | 1 | Cont. | $\pm 180^\circ$ | $\pm 18^\circ$ | $\pm 1.4$ | 4 | 8.2 | 116$^e$ | 0.021 | $N + 1$ (5) |

$^a$: Calculated from synthetic equations.  
$^b$: Calculated values from the normalized radiation pattern.  
$^c$: Radiation patterns are calculated based on measured S-parameters with ideal point sources, making it no fair way to comparison.  
$^e$: Using active device.  
Cont.: Continuously scanning.  
$^f$: PCB: printed circuit board; IPD: integrated passive device; SMD: surface mounted device.

coaxial cables, whose responses were calibrated in advance. The peak gain of single antenna is 5.8 dBi.

The measured results are plotted in Fig. 9(a). Good agreement is observed. The progressive phase shift between elements can be controlled to steer the beam continuously but of the sake of presentation, a $\Delta \phi = 22.5^\circ$ step was selected. Taking the element pattern into account, the main beam is steered to $\theta_a = 0^\circ, \pm 7.5^\circ, \pm 15^\circ, \pm 22.5^\circ, \pm 30^\circ, \pm 36^\circ, \pm 44^\circ, \pm 54^\circ$, and $\pm 60^\circ$. The measured peak gain varies from 4.3 dBi ($\Delta \phi = 180^\circ, \theta_a = \pm 60^\circ$) to 8.2 dBi ($\Delta \phi = -22.5^\circ, \theta_a = 7.5^\circ$); their corresponding 3-dB beamwidths (HPBW) in the plane of scanning ($H$-plane) are 42$^\circ$ and 25$^\circ$, respectively.

Given the loss of the PDN ($IL$) is 4 dB and the radiation efficiency of the quasi-Yagi array ($\eta_a$) is 95%, the overall efficiency can be calculated as

$$\eta = 10^{-IL(\text{dB})/10} \times \eta_a = 10^{-4/10} \times \eta_a = 37.8\%.$$ (8)

Apparenty, it is predominated by the transmission loss of the PDN.

Fig. 9(b) further depicts the measured radiation patterns of usable beams at different frequencies in rectangular form. Beams at large scanning angles are removed due to high sidelobe level ($SLL \geq -6$ dB). The gain envelopes are estimated from the selective usable beams in Fig. 9(a) with a red dash curve, and the 3-dB spatial coverage is defined by the two outermost angles at which the measured gain is 3 dB below the overall peak gain [21]. The maximum 3-dB spatial coverage is 116$^\circ$ at 2.4 GHz, while a narrower coverage is observed at the band edges (92$^\circ$ and 94$^\circ$). Within the usable scanning angle, the array peak gain is at least 7.8 dBi over the bandwidth.

Table 1 compares the proposed PDN with state-of-the arts BFNs based on Butler matrix (BM). The data, if not given, were estimated from the figures in the literature. From the comparison table, the proposed PDN can provide a fully controllable progressive phase shift ($\Delta \phi \in [-180^\circ, 180^\circ]$) to continuously scan the main beam direction of a phased array. Accordingly, beam controllability is greatly boosted when compared with previous designs. The measured array peak gain and equivalent 3-dB spatial coverage are also remarkable. Although the proposed PDN suffers from a higher loss, it eliminates the need for the bulky SPNT switch, which is not possible in conventional beam-switching counterparts. The amplitude and phase errors are in the same order of magnitude of previous BFNs, as well. To compensate for the loss, adding an additional amplifier at the common input would be a viable solution at the expense of additional dc power consumption.

Alternatively, utilizing a varactor diode with a higher $Q$-factor...


### Table 2. Power handling capability.

| Component          | $P_{1dB_{	ext{gain}}}$ (dBm) |
|--------------------|-------------------------------|
| 90° PS             | 23.3                          |
| 180° PS            | 25.7                          |
| Absorptive SPDT    | 25.8                          |
| PDN                | 25.8                          |

can further reduce the loss of all tuning element, while still keeping the feature of low power consumption. For example, by selecting the MA46H071 varactor from M/A-COM, from the datasheet the estimated transmission loss shows an improvement of 1.3 dB when compared to the current design using SMV1405. In addition, benefiting from the single-input topology, the proposed PDN can be easily extended to a planar 16-element phased array for 2D beam-scanning with a reduced number of modules, which will be demonstrated in the next section.

Also from Table 1, the proposed PDN and the design in [22] are the only two BFNs capable of achieving full beam-scanning capability with a single input. Nonetheless, the sub-sector technique in [22] is very narrow in bandwidth (1 GHz at a center frequency of 25 GHz), and requires high dc power consumption (30 mW) despite of its transmission gain. In contrast, the proposed design shows a respectable FBW which is five times wider than that in [22] (20%), and a negligible power consumption (21 $\mu$W) which is less than one thousandths of that in [22]. Moreover, it can be proved from the theoretical analysis that the amplitude imbalance is an inevitable drawback of the PND in [22] as long as the main beam is pointed to a direction other than the four conventional beams of a BM. Referring to (9) of [22], the amplitude difference can theoretically reach up to 7.6 dB for the equal weighting cases ($w_i = w_j$ for $1 \leq i, j \leq 4$ and $i \neq j$). In contrast, our design mitigates the undesired amplitude variation during beam steering, which further highlights the advantages of the proposed $1 \times 4$ PDN.

### C. POWER HANDLING CAPABILITY

The individual building blocks, as well as the PDN, were tested under large power injection. All tests were performed when the varactors are operated with their largest capacitance values. The data were recorded at 2.4 GHz and are listed in Table 2. The power handling capability of the PDN is limited by the input $P_{1dB}$ of the absorptive SPDT switch, which is 25.8 dBm. Taking into account the loss of the feed network (4 dB), the output of the PDN would be 21.8 dBm.

### V. PLANAR 16-ELEMENT BEAM-SCANNING ARRAY

A planar 16-element ($4 \times 4$) phased array for 2D beam-scanning applications is built by stacking/cascading the PDNs as depicted in the left plot of Fig. 10. During the integration, four PDNs were installed horizontally ($H$) and the remaining one was placed vertically to combine four inputs of the horizontal PDNs; the main beam direction is hence scanned two dimensionally. Compared with a conventional 2D BFN, as depicted in the right side of Fig. 10, the required module is reduced from eight ($2N$ with $N = 4$) to five ($N + 1$) [23], [24]. This is achieved even without a costly and bulky nonplanar SP16T switch.

The experimental setup of the proposed 2D scanning phased array is shown in Fig. 11(a), while the measured 3D patterns at 2.4 GHz are depicted in Fig. 11(b). Each radiation pattern is labeled with its corresponding main beam direction ($\phi_b$, $\theta_b$) and measured peak gain. From Fig. 11(b), the main beam can be continuously scanned with full 360° coverage in the $\phi$ orientation, and with a maximum steering angle of 63.5° in the $\theta$ direction. A considerably wide 3D spatial coverage for beam scanning is hence achieved. The degradation of $SLL$
again limits the array from further widening the scanning range. The HPBW of the main beam varied from 25° to 45° in two principle cuts; the measured peak gain varies from 5.8 dBi to 8.3 dBi within the coverage.

VI. CONCLUSION

A new Butler-matrix-based PDN, capable of providing full controllable progressive phase shift between radiating elements, has been presented and successfully demonstrated. Validated by synthesis equations, the proposed PDN is constructed by a TPD and a modified 2 × 4 Butler matrix whose inputs are simultaneously excited by the TPD to fully manipulate the output phases for feeding a beam-scanning array. All tuning elements embedded in the PDN are realized by low-cost RH/LH synthesized transmission lines in planar form with negligible dc power consumption. Benefiting from the newly proposed 1 × 4 PDN, a linear 4-element array and a planar 16-element array, both having continuous beam-scanning capability over a considerably wide angular range, have been realized and experimentally verified. When compared with its conventional counterparts, the proposed design provides much better beam resolution while getting rid of the costly non-planar SPVTs at the input port. Also, the required number of modules to fulfill 2D scanning is reduced. Meanwhile, when compared with the existing work in [22] having a similar 1 × 4 topology, the new architecture features much wider operating bandwidth and extremely low dc power consumption. It avoids the gain variation issue previously noticed when steering the main beam, as well. The advantage and uniqueness of the design concept have been clearly manifested. The main limitation of the proposed design is the power dissipation (parasitic loss). It can be eased by using a varactor with a higher Q-factor, or by introducing an additional amplifier at the common input. In a future work, relaxing the control of grouped phase shifters PS_v,γ for power tapering can further reduce the sidelobe level. Also, developing an 8th-order PDN by modularizing the existing building blocks for widening the spatial coverage would be possible at the expense of additional transmission loss. Using the current phase shifting components, the estimated extra loss of the 1 × 8 PDN is 1.8 dB.

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