Novel Dual-Band Beam-Scanning/Switching Network Based on a Hybrid Coupler with Synchronously Tuned Phase Differences

PEI-LING CHI, (Senior Member, IEEE), CHUN-PIN CHIEN, ANRONG CHEN, AND TAO YANG, (Senior Member, IEEE)

1Department of Electrical and Computer Engineering, National Yang Ming Chiao Tung University, Hsinchu 30010, Taiwan
2School of Electronic Science and Engineering, University of Electronic Science and Technology of China, Chengdu 610054, China

Corresponding author: Pei-Ling Chi (e-mail: peilingchi@nycu.edu.tw)

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ABSTRACT This paper proposes a novel feeding network that enables the antenna array to perform dual-band continuous beam scanning and beam switching between forward and backward directions. Specifically, the switchable beam is controlled by the network input and the beam in each direction is continuously steerable at either frequency. To support a wide range of beam-scanning angles, a novel dual-band coupler with synchronously tuned differential phases is applied to the feeding network, together with a pair of phase shifters. The operating principles and design equations of the dual-band beam-scanning/switching network, coupler, and phase shifters will be given in detail. As a proof-of-concept, a 0.9-GHz/1.8-GHz beam-scanning/switching network was prototyped alongside a 1×4 linear array. Experimental results show that the array can scan and switch its main beam from -24° to -42° in the backward direction and from 24° to 34° in the forward direction at 0.9 GHz, while the beam scanning is from -12° to -24° and from 12° to 22° at 1.8 GHz.

INDEX TERMS Beam-scanning network, beam-switching network, coupler, dual-band, hybrid, phase difference, phase shifter, tunable, varactor.

I. INTRODUCTION

The beam-scanning or beam-switching antenna array in nowadays wireless communications has gained ever-increasing interest due to its adaptive capability to increase the signal to noise ratio and thus improve channel capacity in spite of the existing multipath fading and various sources of electromagnetic interference. To produce a scanned beam at a frequency, the composite right/left-handed leaky-wave antenna provides a solution to the fixed-frequency beam scanning by electronically tuning the antenna circuit parameters [1], [2]. On the other hand, the beam-scanning array is generally driven by a feeding network, where the 360° controllable phase progression between antenna ports is theoretically necessary to scan the entire spatial range. While a greater phase shifting range will extend beam coverage of the antenna system, designing the feeding network becomes more challenging as a result. Thus, a practical solution is to take account of size, complexity, cost, and the achievable performance of a beam-scanning network. In [3], an additional beam-steering network was used to form more beams by appropriately weighting the excitation ratio between two beam ports of the Butler matrix. Tunable phase shifters were incorporated into a conventional 8-way Butler matrix to contribute to a small steering range around the original (switched) beam [4]. For an acceptable quantization loss, the 1-bit or 2-bit PIN-controlled reflectarrays [5]-[9] were realized to alleviate design complexity and element loss necessary for implementing continuous phase shift. Moreover, the array driven by the beam-switching network is an alternative solution to providing multiple characteristic beams toward different angles. The Butler matrix, for example, is a popular choice and its extended capabilities have been extensively studied in the literature [3], [4], [10]-[17]. By replacing quadrature couplers with nonstandard-phase-difference couplers [10], [16], the switched beam can be engineered to other angles. In [14] and [15], the filtering
Butler matrix is inherited from the filtering 180° hybrid coupler. To carry out dual-band operation, the Butler matrix was built with dual-band components [12], [13]. Besides, the single-ended-to-balanced power divider can feed the 1x4 differential array to produce two switched beams [17].

It is noted that, in the literature, the beam-scanning/switching networks are mainly designed with single-band scheme and the array feeding network is usually on a basis of beam-switching operational mechanism with reduced spatial resolution (by discrete beams). To lend itself to multiple operating standards and to cover the spatial range in a continuous manner, this work proposes a novel, single-layer, and microstrip feeding network that allows beam-scanning and switching at dual frequencies of interest. Specifically, a dual-band hybrid coupler with two equal and tunable differential phases is the building block of the dual-band feeding network, as shown in Fig. 1. It will be shown that the proposed coupler can achieve a wide range of tunable phase differences at both frequencies, which will contribute to an extended beam-scanning range. To enable the proposed beam-scanning/switching network to be utilized at two arbitrary frequencies, a detailed coupler study of flexible beam-scanning/switching networks are mainly designed with the basis of beam-switching operational mechanism with reduced spatial resolution (by discrete beams).

![Figure 1. Schematic representation of proposed dual-band beam-scanning/switching network.](image)

II. PROPOSED DUAL-BAND BEAM-SCANNING/SWITCHING NETWORK

A. SCHEMATIC DIAGRAM AND OPERATING PRINCIPLES

Fig. 1 shows the schematic diagram of proposed dual-band beam-scanning/switching network, consisting mainly of three identical dual-band phase tunable couplers in two-stage configuration. Two kinds of tunable phase shifters (PS) 1 and 2 are used in-between stages to connect couplers and the connection is controlled by two single-pole double-throw (SPDT) switches. Input ports 1 and 2 are used to switch beam direction; ports 3-6 are the antenna ports of the network and each will be connected to 1x4 antenna elements. Note that to reduce integration complexity, each individual component is designed to match 50 Ω and thus when passing through it, only the transmission phase is considered. To ensure isolation between ports 1 and 2 and between either two of ports 3 to 6 of the network, the internal ports with port numbering 1’ and 2’ or 3’ and 4’ are the isolated pair of the proposed tunable coupler and this port numbering convention will be kept consistent with that used in Section III. For the excitation from coupler port 1’ or 2’, the resulting phase difference φ1 or φ2, respectively, of the tunable coupler is defined as

$$\phi_1 = \angle S_{34} - \angle S_{35},$$  
$$\phi_2 = \angle S_{36} - \angle S_{35}.$$  

This feeding network has two operating modes. When port 1 is the input port, the SPDTs 1 and 2 are switched to PS 1 and PS 2, respectively, i.e., to the upper paths for both SPDTs, and thus the phase differences between adjacent outputs Δφ34, Δφ56, and Δφ36 are given by

$$\Delta \phi_{34} = \angle S_{34} - \angle S_{35} = \phi_1,$$  
$$\Delta \phi_{56} = \angle S_{56} - \angle S_{51} = \phi_1,$$  

where Δφ34 and Δφ56 are the introduced phase shifts of PS 1 and PS 2, respectively. Thus, a constant phase progression will require the phase shifts Δφ34 and Δφ56 to satisfy:

$$\phi_1 = \Delta \phi_{35} - \Delta \phi_{34}.  \quad (3)$$

On the other hand, when port 2 is the input port, the SPDTs 1 and 2 are both switched to the lower paths, PS 2 and PS 1, respectively. The phase progression Δφ45, Δφ56, or Δφ34 is

$$\Delta \phi_{45} = \angle S_{45} - \angle S_{41} = \phi_2,$$  
$$\Delta \phi_{56} = \angle S_{56} - \angle S_{52} = \phi_2,$$  

Similarly, the required phase relation can be found as

$$\phi_2 = \Delta \phi_{54} - \Delta \phi_{52}.  \quad (4)$$

From (3) and (5), it can be shown that, if the differential phase Δφ34=Δφ56 between PS 1 and PS 2 can be synchronously tuned with the coupler’s phase difference φ2, the main beam of the antenna array can be continuously steered; furthermore, since the phase progression between antenna ports has the form of φi or -φi, the antenna array has greater potential to scan its main beam in either forward and backward direction by simply switching the excitation port. The latter is an important advantage for simplifying the coupler design. The structural analysis and design of the proposed coupler will be comprehensively detailed in Section III.
B. VARACTOR-TUNED LUMPED-ELEMENT BASED PHASE SHIFTERS

To support dual-band beam steering, the phase shifters PS 1 and PS 2 are also necessary to provide tunable phase shifts at both frequencies of interest. To simplify circuit complexity, the lumped-element based lowpass network is paired with a highpass network to implement the tunable PS 1 and PS 2, respectively, as shown in Fig. 2. In the lowpass network, each unit cell consists of two series inductors \( L_i \) and one shunt varactor \( C_i \) in the T-type configuration, and thus the incremental phase shift of a unit cell is tuned by the varactor \( C_i \). Similarly, each series varactor \( C_h \) in the \( \pi \)-type highpass unit cell is used to engineer the total phase shift \( \Delta \phi_{p2} \). The transmission \( (ABCD) \) matrix \([T]_l,h\) of the lowpass or highpass unit cell can be derived as follows:

\[
[T]_l = \begin{bmatrix} 1 - \omega^2 L_i C_i & j2\omega L_i - j\omega^2 L_i^2 C_i \\ j\omega C_i & 1 - \omega^2 L_i C_i \end{bmatrix}, \quad (6a)
\]

\[
[T]_h = \begin{bmatrix} 1 - \frac{1}{\omega^2 L_h C_h} & j\omega C_h \\ \frac{j\omega L_h}{\omega^2 L_h C_h} & 1 - \frac{1}{\omega^2 L_h C_h} \end{bmatrix}. \quad (6b)
\]

where \( \omega \) is the angular frequency. Thus, the overall phase shift \( \Delta \phi_{p1} \) or \( \Delta \phi_{p2} \) is equal to the sum of the phase shift of each corresponding unit cell or the incremental phase shift multiplied by the number of unit cells. By conversion between \( ABCD \) and \( S \) network parameters, the \( \Delta \phi_{p1}, \Delta \phi_{p2}, \) and the impedance matching conditions are given by

\[
\Delta \phi_{p1} = -N \cdot \tan^{-1} \frac{\omega L_i + \omega Z_0 C_i}{Z_0 - \frac{1}{\omega^2 L_i^2 C_i} - \frac{2\omega L_i}{Z_0}}, \quad (7a)
\]

\[
\Delta \phi_{p2} = -N \cdot \tan^{-1} \frac{\omega^2 L_h C_h + \omega Z_0}{Z_0 - \frac{1}{\omega^2 L_h^2 C_h} - \frac{2\omega L_h}{Z_0}}, \quad (7b)
\]

\[
\frac{L_i}{C_i} = \frac{L_h}{C_h} = Z_0, \quad (7c)
\]

where \( N \) is the number of unit cells in a network and \( Z_0=50 \) \( \Omega \). The required differential phase between lowpass and highpass networks, as dictated by (3) and (5), can therefore be obtained by simultaneously controlling \( C_i \) and \( C_h \) in PS 1 and PS 2, respectively.

To validate the applicability of the paired lowpass and highpass networks in the proposed dual-band beam-scanning network, the varactor-tuned lumped-element based PS 1 and PS 2 were prototyped for dual operating frequencies at 0.9 GHz and 1.8 GHz, and the measured/simulated magnitude and phase responses are shown in Fig. 3 (b) and (c). Here, the PS 1 and PS 2 were built on a RT/Duroid 5880 substrate of thickness 0.508 mm and of dielectric constant \( \varepsilon_r=2.2 \). The phase shifts \( \Delta \phi_{p1} \) and \( \Delta \phi_{p2} \) are tuned by \( C_1, C_2 \) and \( C_3, C_4 \), respectively, using varactors MA46H202 (0.7 pF–7 pF).
the prescribed phase differences at dual frequencies, the required inductances and capacitances can be preliminarily decided by (7). Furthermore, to maintain good impedance matching across a wide range of phase differences, the fixed inductances \(L_1\) to \(L_4\) were properly optimized. In this design, \(N=3\) and the inductances \(L_1=2.2\) nH, \(L_2=4.3\) nH, \(L_3=15\) nH, and \(L_4=7.2\) nH. Measured and simulated results are in good agreement. In Fig. 3 (b), the measured phase difference between two paths was synchronously controlled at 0.9/1.8 GHz and continuously tuned from 60° to 120°. In Fig. 3 (c), the measured return loss is greater than 15 dB at either frequency for both PS 1 and PS 2, thus ensuring good power transmission when applied to the proposed feeding network.

III. PROPOSED DUAL-BAND HYBRID COUPLER WITH SYNCRONISUMLY TUNED PHASE DIFFERENCES

As aforementioned, the proper operation of the proposed dual-band beam-scanning/switching network needs a well-performed coupler that enables synchronous phase engineering at two flexible frequencies. To this end, a varactor-based dual-band hybrid coupler with two synchronously tunable phase differences is proposed for the first time to offer an adaptive solution without the need of incorporating all functional hardware with substantial increase in complexity, size, and loss.

Electronically tunable couplers with tunable operating frequency [18], [19], tunable power-dividing ratio [20]-[22], or both [23]-[25] were extensively reported. Owing to the adaptability to providing flexible phase compensation in a beamformer, such as in the switched-beam Butler and Nolen matrices, and the contribution to optimum performance of reflectometers [26], [27], a coupler with tunable phase difference between through and coupled ports has received increasing attention and several tunable or switchable solutions [28]-[33] were proposed in the recent literature. In [29], the switchable coupler can realize three discrete phase differences. To extend the phase coverage, varactor-tuned couplers were studied to fulfill the continuously controllable differential phase at a fixed frequency [30], [32]. The frequency agility was also demonstrated in addition to the phase control [28], [31], [33]. To allow for a greater degree of flexibility and applicability in the existing and emerging multiband applications, however, the simultaneous dual-band and continuous phase engineering is increasingly useful and is not explored till now.

A. COUPLER ANALYSIS

The schematic of proposed tunable coupler with dual-frequency phase control is shown in Fig. 4. This coupler consists of two host microstrip lines of characteristic impedance \(Z_1\) and electrical length \(\theta_1\), two horizontal microstrip lines of characteristic impedance \(Z_2\) and electrical length \(\theta_2\), and two groups of stub networks with the respective input impedances of \(jZ_{q1}\) or \(jZ_{q2}\). Note that to carry out tunable phase differences, varactor diodes will be used in the stub networks. When port 1 is the input port, port 2 is the isolated port while ports 3 and 4 will be two outputs. Therefore, to ensure ideal impedance match and port isolation, this proposed coupler has to hold the condition in (8) at either operating frequency:

\[
S_{11} = S_{22} = 0, S_{21} = 0.
\]  

The phase difference \(\phi_1\) and \(\phi_2\), and power division ratio \(K\) are defined as

\[
\phi_1 = \angle S_{41} - \angle S_{31},  \tag{9a}
\]

\[
\phi_2 = \angle S_{32} - \angle S_{42},  \tag{9b}
\]

\[
K = \frac{|S_{41}|^2}{|S_{31}|^2}.  \tag{10}
\]

where \(\phi_1\) and \(\phi_2\) are the seen phase differences between output ports 3 and 4 with the input ports 1 and 2, respectively. It can be shown that \(\phi_1\) and \(\phi_2\) are supplementary angles, i.e., \(\phi_1 + \phi_2 = 180^\circ\), and thus only \(\phi_1\) will be used in the mathematical derivations that follow. As seen, this coupler is symmetric with respect to the plane of symmetry PP’ and thus the even-odd mode decomposition technique is used for coupler analysis. By applying the open-circuited and short-circuited boundary conditions along the PP’ plane, the
resulting half circuits are shown in Fig. 4 (b) and (c). The transmission \( (ABCD) \) matrix elements of the even-mode or odd-mode half circuit are given as

\[
\begin{align*}
A_{e,o} &= \cos \theta_1 + \frac{1}{Z_{2}} \sin \theta_1 \left( \frac{1}{j \frac{Z_1}{Z_2}} + Y_1^{e.o} \right), \\
B_{e,o} &= j \frac{Z_2}{Z_1} \sin \theta_2, \\
C_{e,o} &= \cos \theta_2 \left( \frac{1}{j \frac{Z_1}{Z_2}} + \frac{1}{Z_2} + \frac{j}{Z_2} Y_1^{e.o} \right) + j \sin \theta_2 \left( \frac{1}{Z_2} + \frac{1}{j \frac{Z_1}{Z_2}} + Y_1^{e.o} \right), \\
D_{e,o} &= \cos \theta_2 + j \frac{Z_2}{Z_1} \sin \theta_2 \left( \frac{1}{j \frac{Z_1}{Z_2}} + Y_1^{e.o} \right),
\end{align*}
\]

where \( Y_1^{e.o} \) is the input admittance seen toward the bisected host line under even- or odd-mode excitation, and they are given by:

\[
Y_1^{e} = j \tan \frac{\theta_1}{Z_1}, \quad Y_1^{o} = -j \coth \frac{\theta_1}{Z_1}.
\]

From (16a), it is noted that the phase difference of this coupler can be tuned by controlling input impedances of the stub networks. For the coupler designed with equal power division ratios \((K_1=K_2=K)\) and equal phase differences \((\phi_1=\phi_2=\phi)\) at dual frequencies \(f_i\) and \(f_2\) \((f_1 < f_2)\), the electrical length \(\theta_1\) of these microstrip lines can be solved from (16b),

\[
\theta_1 = \frac{\pi}{1+\left(2 f_i / f_1 \right)} = \frac{\pi}{1+M},
\]

where \(M\) is frequency ratio and the smallest value of \(\theta_1\) is chosen for implementation. Thus, the required impedance \(Z_2\) will then be determined by (16b) for the given \(f_1, f_2\), \(K = 1\), and \(\phi\). By using (11) and (14), the results in (18) yield:

\[
\begin{align*}
\frac{1}{Z_{s1}} + \frac{1}{Z_{s2}} &= -2 \left( \frac{1}{Z_{s1} \tan \theta_1} + \frac{1}{Z_{s2} \tan \theta_2} \right), \\
\frac{1}{Z_{s1} - Z_{s2}} &= \left\{ \begin{array}{ll}
-2 & \frac{1}{Z_{1} \sin^2 \theta_1} = \frac{1}{Z_{2} \sin^2 \theta_2} = \frac{1}{Z_0} \; \phi_1 \leq 90^\circ, \\
2 & \frac{1}{Z_{1} \sin^2 \theta_1} = \frac{1}{Z_{2} \sin^2 \theta_2} = \frac{1}{Z_0}, \; 90^\circ < \phi_1 < 180^\circ
\end{array} \right.
\end{align*}
\]

With (16a) and (18b), the following result can be derived:

\[
Z_{1}^2 \sin^2 \theta_1 = \left( \frac{1}{Z_{2} \sin^2 \theta_1} + \frac{1}{Z_{2} \sin^2 \theta_2} \right)^{-1}.
\]

Again, to realize equal \(K\) and \(\phi\) in dual bands with (17), the microstrip length \(\theta_1\) in (19) is given by

\[
\theta_1 = \frac{\pi}{1+\left(2 f_i / f_1 \right)} = \frac{\pi}{1+M},
\]

and the characteristic impedance \(Z_1\) can readily be obtained by (19). Finally, the input impedances of the stub networks at \(f_1\) and \(f_2\), i.e., \(Z_{s1}(f_i)\), \(Z_{s2}(f_i)\), \(Z_{s1}(f_2)\), and \(Z_{s2}(f_2)\), can be deduced from (18) and they are shown at the bottom of this page. By inspection of (21), it is seen that \(Z_{s1}(f_1)=Z_{s2}(f_1)\) and \(Z_{s1}(f_2)=Z_{s2}(f_2)\).

Given two operational frequencies \(f_1\) and \(f_2\), power division ratio \(K=1\), and the phase difference \(\phi\), the microstrip impedance \(Z_1\) or \(Z_2\) and electrical length \(\theta_1\) or \(\theta_2\) can be computed by (16b), (17), (19), and (20); moreover, the required \(Z_{s1}\) and \(Z_{s2}\) at each frequency can be computed by (21). Note that \(\theta_1\) and \(\theta_2\) are decided by \(f_1\) and \(f_2\) only. On the other hand, the calculated characteristic impedances \(Z_1\) and \(Z_2\) of the microstrip lines and the stub’s input impedance \(Z_{s1}\) against phase difference \(\phi\) are illustrated in Fig. 5 for a wide range of frequency ratio \(M\). Note that, due to the fact of
The value of distribution about 90° the required structure is proposed and detailed in Section III-B. To accommodate a wide range of stub’s input impedances at two flexible frequencies, a highly tunable stub with the loaded microstrip stub is of characteristic impedance $Z_1$ and of electrical length $\theta_1$ away from the input end and the other varactor capacitance $C_2$ is loaded at the open end. The microstrip stub is of characteristic impedance $Z_s$ and of electrical length $\theta_1 + \theta_2$. By transmission line theory, the input impedance $jZ_{st}$ is given by

$$jZ_{st} = Z_s Z_{in} + jZ_{tan} \tan \theta_{s1}, \quad i = 1, 2 \quad (22a)$$

$$Z_{in} = Z_s + j\frac{1}{\omega C_1}, \quad (22b)$$

This means that the capacitive load $C_1$, i.e., equal to the parallel effective impedance of the varactor capacitance $C_1$ and the terminated stub (by $C_2$). Eq. 7 shows the calculated $Z_{st}$ as a function of $C_2$ for $C_1$, 0.37, 1, and 1.5 pF. When $\theta_1$ is small (<10°) and $C_1/C_2$ are in the range of 0.3 pF to 0.8 pF, the $Z_{in}(f_2)$ is mainly controlled by $C_2$. Thus, to carry out the required $Z_{in}(f_1)$ and $Z_{in}(f_2)$, the tuning capacitance $C_2$ is essentially determined by $Z_{in}(f_2)$ and $C_1$ is then additionally decided by $Z_{in}(f_1)$. This property will simplify the control of tuning varactors and will be exploited in the fabricated prototype. Besides, $C_1$ can be used to enlarge the tunable range of $Z_{st}(f_2)$ to adapt to the required variation across phase difference $\phi$, and thus the resulting capacitance tuning range (of $C_1$) will be larger than that of $C_2$, as observed in simulation and measurement.

For an arbitrary differential phase $\phi$, five electrical parameters ($Z_1$, $\theta_1$, $\theta_2$, $C_1$, and $C_2$) are used to realize $Z_{st}(f_1)$ and $Z_{st}(f_2)$. To maximize the tuning range, it is important to properly determine the non-tunable parameters $Z_1$, $\theta_1$, and $\theta_2$ in accordance with the available varactor capacitances. Since both of $Z_{st}(f_1)$ and $Z_{st}(f_2)$ increase with $C_1/C_2$ and thus $Z_1$, $\theta_1$, and $\theta_2$ can be solved by using the minimum $Z_s(f_1)$ and $Z_s(f_2)$.
capacitance $C_{\text{min}}$ in place of $C_1$ and $C_2$ to realize the minimum $Z_{\text{st}}(f_1)$ and $Z_{\text{st}}(f_2)$ in the attainable or required phase tuning range. As a result, larger $Z_{\text{st}}(f_1)$ and $Z_{\text{st}}(f_2)$, corresponding to other differential phases, can be easily fulfilled with available diode capacitances. Note that to allow parasitic effects, $C_{\text{min}}$ is chosen slightly larger than the minimum available diode capacitance provided by the manufacturer’s data sheet. According to (22), Fig. 8 shows the calculated $\theta_1$, $\theta_2$, and $Z_i$ for $Z_{\text{st}}(f_1)=-84.99$ $\Omega$ and $Z_{\text{st}}(f_2)=23.60$ $\Omega$. Here, only the value ranging from 0° to 90° is concerned for $\theta_1$ or $\theta_2$. Two important conclusions can be drawn. For a given $C_{\text{min}}$, there may exist up to two solutions of $(\theta_1, \theta_2)$ for a $Z_i$. For example, $(\theta_1, \theta_2)= (17.5^\circ, 11.7^\circ)$ (sol. 1) or $(5.4^\circ, 28.2^\circ)$ (sol. 2) for $C_{\text{min}}=0.4$ pF and $Z_i=140$ $\Omega$. The optimal choice will depend on the potential for realizing a maximal tuning range and will be exemplified in Section III-C. Furthermore, it is noted that as $C_{\text{min}}$ increases, the required $Z_i$ increases. Thus, the impedance $Z_i$ can be used to accommodate the practical minimum capacitance influenced by the parasitic effects or manufacturing tolerance.

### C. EXAMPLES AND DESIGN GUIDELINES

As illustrated in Fig. 5 (a), the ideal impedance $Z_i$ or $Z_2$ of the host microstrip lines varies with $\phi_i$. In order to comply with this requirement, additional tuning varactors are generally necessary for host lines, thus substantially increasing structural and design complexity. If a smaller phase tuning range is allowed or further optimization is applied to the tunable stubs, four simple microstrip lines can be conveniently used given that a reasonable approximation to the ideal impedances is fulfilled. Because the impedance $Z_i$ or $Z_m$ is symmetric with respect to $\phi_i=90^\circ$ and a quadrature coupler, i.e., $\phi_i=90^\circ$, is widely used in many components and networks, the values of $Z_{1,\phi_i=90^\circ}$ and $Z_{2,\phi_i=90^\circ}$ are chosen to realize $Z_i$ and $Z_m$, respectively. To examine the resulting impact and estimate the achievable tuning range, the reflection coefficient of a microstrip line of characteristic impedance $Z_{\phi_i=90^\circ}$ and length $\theta_{\phi_i}$ with the reference impedance $Z_{\phi_i}$ was calculated in Fig. 9. Note that $Z_{\phi_i}$ represents the ideal impedance required for the phase difference $\phi_i$ and it can be $Z_i$ or $Z_2$ in Fig. 5 (a). The length $\theta_{\phi_i}$ is a function of the frequency ratio $M$ only and can be obtained by (17) or (20). Here, a normalized impedance $Z_i$ ($i=1, 2$), defined as the impedance ratio of the characteristic impedance $Z_{i,\phi_i}$ to $Z_{1,\phi_i=90^\circ}$, is derived based on (16b) and (19):

$$Z_i^2 = \frac{Z_{1,\phi_i=90^\circ}^2}{Z_{Z_i}} = \sin^2 \phi_i,$$

where it is shown that $Z_i = Z_2 (= Z_{\phi_i})$ and thus in Fig. 9 (b), the calculated results are plotted against $Z_{\phi_i}$ for a wide range of frequency ratio $M$. When $\phi_i=90^\circ$, $Z_{\phi_i}=1$ and this corresponds to perfect match in Fig. 9 (b) regardless of $M$ (or $\theta_{\phi_i}$). As $\phi_i$ varies from 90°, $Z_{\phi_i}$ becomes smaller and the degree of impedance mismatch increases, implying that a feasible tuning range can be preliminarily estimated according to the acceptable return loss of this study. For example, for $M=2$, the range for greater than 15-dB return loss is from $Z_{\phi_i}=0.81$ to $Z_{\phi_i}=1$, corresponding to the phase
tuning range of 54°–126°, and the phase tuning range for 20-dB return loss is from about 63° to 117° for \( Z_f \geq 0.89 \). As \( M \) becomes larger, the attainable tuning range is improved as a result of a smaller \( \theta_f \). Nevertheless, the maximum feasible \( M \) will be limited by the characteristic impedance \( Z_s \), which increases to ~150 \( \Omega \) for \( M=8.24 \) at \( \phi_f=90^\circ \) and approaches to upper bound of the realizable line impedance based on standard PCB manufacturing process.

As mentioned in Section III-B, a proper choice of the stub’s parameters \( Z_s, \theta_1, \) and \( \theta_2 \) is crucial to the achievable phase tuning range. Based on the results in Fig. 8, Fig. 10 illustrates the calculated varactor capacitances \( C_1 \) and \( C_2 \) at \( \phi_f=50^\circ, 70^\circ, 90^\circ, 110^\circ, \) and \( 130^\circ \) for three sets of \( (Z_s, \theta_1, \theta_2) \). Note that each curve illustrates all combinations of \( C_1 \) and \( C_2 \) to realize \( \phi_f \) at \( f_1 \) or \( f_2 \) and thus, the intersection point of the same \( \phi_f \)-curves represents the desired \( C_1 \) and \( C_2 \) for realizing \( \phi_f \) in dual bands. Also note that the black dashed line in each figure is used to indicate those required capacitances across different phase differences. When \( \theta_1 \) is very small, for example, 5.4° in Fig. 10 (c), it is seen that the phase tuning from \( \phi_f=50^\circ \) to \( 130^\circ \) will result in a small variation of \( C_2=0.4 \) pF to \( C_2=0.56 \) pF, implying that precise capacitance control is requisite across different phase differences and will considerably increase implementation difficulty. On the other hand, Fig. 10 (b) shows the result when the value \( \theta_1 \) is comparable to \( \theta_2 \). In this case, a relatively irregular variation in either \( C_1 \) or \( C_2 \) between neighboring differential phases is obtained and may result in some impractical values (\( C_2<0.3 \) pF at \( \phi_f=70^\circ \) and \( 80^\circ \) given that 0.3 pF is the minimum available diode capacitance) and thus a reduced phase tunable range. With \( \theta_1=9.5^\circ \) and \( \theta_2=24.6^\circ \) in Fig. 10 (a), the phase differences can be synchronously tuned by a steady increase in tuning capacitances \( C_1 \) and \( C_2 \), thus contributing to an extended tuning range.

The design procedure for the proposed dual-band coupler with synchronously tuned phase differences is summarized as follows.

Step 1) For given two frequencies \( f_1 \) and \( f_2 \) (=\( Mf_1 \)), use (17) and (20) to compute electrical lengths \( \theta_1 \) and \( \theta_2 \) of the host lines. The characteristic impedance \( Z_s \) is solved by (16b) with \( K=1 \) and \( \phi_f=90^\circ \), followed by the determination of \( Z_s \) with (19).

Step 2) Use (21) to compute the stubs’ input impedances \( Z_{s1} (f_1), Z_{s2} (f_2), Z_s (f_1), \) and \( Z_s (f_2) \) with respect to \( \phi_f \).

Step 3) Use (22) to determine the parameters \( \theta_{1s}, \theta_{2s}, \) and \( Z_s \) of the varactor-loaded stub with a given \( C_{\text{min}} \) and the minimum \( Z_{s1} (f_1) \) and \( Z_{s2} (f_2) \) in the phase tuning range. The attainable phase tuning range can be estimated by the results in Fig. 9 (b). In practice, the value \( Z_s \) is limited to the highest realizable line impedance and from the parametric study, it is seen that \( Z_s=10^\Omega \) is a proper choice.

Step 4) With the chosen \( \theta_{1s}, \theta_{2s}, \) and \( Z_s \) from Step 3), the required varactor capacitances \( C_1 \) and \( C_2 \) for each differential phase \( \phi_f \) can be calculated by the results from Step 2) and (22).

Step 5) Choose commercially available tuning varactors to cover the required capacitances across the tuning range.

**FIGURE 10.** Calculated varactor capacitances \( C_1 \) and \( C_2 \) required for \( \phi_f=50^\circ, 70^\circ, 90^\circ, 110^\circ, \) and \( 130^\circ \). (a) \( (Z_s, \theta_{1s}, \theta_{2s})=(140 \Omega, 9.5^\circ, 24.6^\circ) \) for \( C_{\text{min}}=0.37 \) pF. (b) \( (Z_s, \theta_{1s}, \theta_{2s})=(140 \Omega, 13.7^\circ, 19^\circ) \) for \( C_{\text{min}}=0.37 \) pF. (c) \( (Z_s, \theta_{1s}, \theta_{2s})=(140 \Omega, 5.4^\circ, 28.2^\circ) \) for \( C_{\text{min}}=0.4 \) pF. All electrical lengths of \( \theta_{1s} \) and \( \theta_{2s} \) are evaluated at 0.9 GHz.

**D. MEASURED AND SIMULATED S-PARAMETERS OF THE FABRICATED COUPLER**

A hybrid coupler with operating frequencies at 0.9 GHz and 1.8 GHz, i.e., \( M=2 \), was experimentally developed to perform synchronously tuned phase differences. Fig. 11 shows the physical layout and photograph. The physical
dimensions of the fabricated coupler are given in Table 1. The fabricated coupler was built on a RT/Duroid 5880 substrate of thickness 0.508 mm. Loaded with the tuning varactors $C_1$ and $C_2$, two microstrip stubs on the left side are tuned by control voltages $V_1$ and $V_2$, respectively, to realize tunable input impedance $jZ_1$; the voltages $V_3$ and $V_4$ are used to control varactors $C_3$ and $C_4$, respectively, in right-sided microstrip stubs of tunable input impedance $jZ_2$. The DC block capacitance $C_{dc}$ of 20 pF and the bias resistance $R_{rf}$ of 100 KΩ were used. In addition, four DC blocks were connected to each port to prevent the flow of DC frequencies to RF signals and prevent damage to the network analyzer. The surface-mount GaAs tuning varactors MA46H201 (0.3 pF) were used to implement $C_3$ and $C_4$, as predicted in Section III-C for $Z_r=140$ Ω, $\theta_1=9.5^\circ$, and $\theta_2=24.6^\circ$. Figs. 12-14 show measured and simulated $S$-parameters of the fabricated prototype with dual-frequency phase differences of 60°, 90°, and 120°, respectively. As seen, measured results agree well with the simulated counterparts, and dual-band equal power division and equal/synchronous phase control are experimentally validated. The bias voltages $V_1 - V_4$ for each tuning state are given and it is seen that the voltages $V_1/V_2$ and $V_3/V_4$ monotonically increase and decrease, respectively, with increasing $\phi$, as expected from the result in Fig. 10 (a). In the phase tuning range from 60°

| $W_1$ | $W_2$ | $W_3$ | $W_4$ | $L_1$ |
|------|------|------|------|------|
| 1.57 | 2    | 1.2  | 0.2  | 4    |
| $L_3$ | $L_4$ | $L_5$ | $L_6$ |
| 40.3 | 39.5 | 17.15 | 2.6  | 44.1 |

via-hole diameter: 0.5 mm.

| $S$-parameters and (b) $\angle S_{41}$ $/\angle S_{43}$ |
|---|---|---|---|---|
| Sim. | Meas. | Sim. | Meas. |
| $S_{11}$ | $S_{21}$ | $S_{31}$ | $S_{41}$ |
| $S_{12}$ | $S_{22}$ | $S_{32}$ | $S_{42}$ |
| $S_{13}$ | $S_{23}$ | $S_{33}$ | $S_{43}$ |
| $S_{14}$ | $S_{24}$ | $S_{34}$ | $S_{44}$ |

FIGURE 11. Proposed dual-band hybrid coupler with synchronously tuned phase differences. (a) Physical layout. (b) Photograph of fabricated prototype.

FIGURE 12. Measured and simulated results of the fabricated prototype. The control voltages $V=2.82$ V, $V_2=3.2$ V, $V_3=20$ V, and $V_4=20$ V. The simulated junction capacitances $C_1=1.414$ pF, $C_2=0.649$ pF, $C_3=0.547$ pF, and $C_4=0.324$ pF. At 0.9/1.8 GHz, the measured power difference and phase difference are -0.22/0.29 dB and 59.4°/60.6°, respectively. (a) $S$-parameters and (b) $\angle S_{41}$ $/\angle S_{43}$.
to 120°, the measured return loss $|S_{11}|$ and port isolation $|S_{22}|$ at two operational frequencies are greater than 15.4 dB and greater than 17.5 dB, respectively. At $\phi=60^\circ$, the measured $|S_{11}|$ and $|S_{22}|$ at 0.9/1.8 GHz are -3.5/-3.9 dB and -3.7/-3.6 dB, respectively. Thus, the measured coupler loss, including conductor, dielectric, and varactor losses, is about 0.64/0.78 dB. At $\phi=90^\circ$, the measured $|S_{11}|$ and $|S_{22}|$ at 0.9/1.8 GHz are -3.2/-3.6 dB and -3.5/-3.6 dB, respectively, leading to the measured coupler loss of about 0.4/0.6 dB. At the upper end of phase tuning range, i.e., $\phi=120^\circ$, the measured $|S_{11}|$ and $|S_{22}|$ at 0.9/1.8 GHz are -3.5/-3.9 dB and -3.2/-3.7 dB, respectively, leading to the measured coupler loss of about 0.36/0.8 dB. The magnitude imbalance, i.e., $|S_{11}|-|S_{22}|$, is less than 0.32 dB, except the measured imbalance of 0.53 dB at $\phi=70^\circ$. As shown in Fig. 15, the differential phases at dual frequencies can be synchronously and continuously tuned from 60° to 120° and meanwhile, they can be well controlled with phase error less than 2.5° across the entire tuning range. Note that the achieved phase tuning range (60°-120°) agrees very well with the prediction result in Fig. 9 (b) for $M=2$ and 15-dB return loss.

Table 2 summarizes measured bandwidths of several key performances. Note that the phase bandwidth is calculated as the fractional bandwidth with less than 5° phase deviation.
from each measured/desired $\phi_i$. It is shown that the obtained bandwidth at the first frequency is larger than the corresponding bandwidth at the second frequency, as a result of the larger wavelength in the first band. Moreover, the minimum bandwidth comes from either end of the tuning range, i.e., 60° or 120°. In other words, the asymmetry of coupler structure to realize a nonstandard differential phase (other than 90°) will lead to bandwidth reduction. The linearity of proposed tunable coupler was also studied and for the entire phase tuning range, the measured IIP$_3$ is from 23.6 dBm to 36.3 dBm at 0.9 GHz and is from 23.7 dBm to 32.9 dBm at 1.8 GHz.

IV. MEASURED AND SIMULATED RESULTS OF THE DUAL-BAND FEEDING NETWORK AND THE BEAM-SCANNING/SWITCHING RADIATION PATTERNS

In Section III, the measured phase tuning range of the fabricated coupler is from 60° to 120° for $\phi_i = (\angle S_{41} - \angle S_{31})$ at 0.9/1.8 GHz and it can also be experimentally shown that the corresponding $\phi_i = (\angle S_{32} - \angle S_{42})$ ranges from 120° to 60°. Thus, based on the previous discussions in Section II-A, the continuous beam scanning in the switchable forward or backward direction can be carried out with the proposed dual-band feeding network. Figs. 16 and 17 show the photograph and measured/simulated $S$-parameters of the fabricated dual-band beam-scanning/switching network, respectively. This network was again developed on a RT/Duroid 5880 substrate of thickness 0.508 mm. Between neighboring antenna ports, the measured phase differences of 60°, 90°, and 120° for port 1 as the input are illustrated in Fig. 17 (a). In Fig. 17 (b), the measured return loss $|S_{11}|$ is greater than 20 dB at 0.9/1.8 GHz and the measured insertion loss $|S_{31}|$, $|S_{41}|$, $|S_{51}|$, or $|S_{61}|$ is about 8.2 dB on average. The amplitude imbalance between output ports is less than 1.5 dB. Note that the Analog Devices HMC284A SPDT switch was used in the dual-band feeding network and at these frequencies, the measured insertion loss is about 1 dB. In Fig. 17 (c), the measured isolation between antenna ports is greater than 15 dB for all tuning states. Based on this beam-scanning/switching network, a 1×4 antenna array was further developed for integration in the proposed dual-band beam-scanning/switching module. Fig. 18 shows the measurement setup for far-field radiation patterns in the $xy$ (horizontal)-plane, i.e., the plane with this antenna array. The array is connected to the network by cables and its radiation patterns were measured in the anechoic chamber. Fig. 19 shows the measured beam-scanning radiation patterns when the coupler’s phase difference was tuned to $\phi_i = 60°$, 90°, and 120° and the input was switched between ports 1 and 2 of the module. The fabricated 1×4 array has measured return loss greater than 15 dB at 0.9/1.8 GHz and the inter-element spacing is 130 mm, corresponding to 0.39$d_0$ at 0.9 GHz (or

![FIGURE 16. Photograph of the fabricated dual-band beam-scanning/switching network.](image1)

![FIGURE 17. (a) Measured and simulated phase differences $\Delta \phi$ between adjacent antenna ports. (b) Measured and simulated return/insertion losses for $\Delta \phi = 90°$. (c) Measured and simulated isolation between antenna ports for $\Delta \phi = 90°$.](image2)
0.78\lambda_0 at 1.8 GHz). When the module is fed through port 1, it is seen that the main beam is steered from 336° (-24°) to 318° (-42°) with the measured peak gains of 1.3-3.2 dBi at 0.9 GHz. At the same frequency, when feeding the module through port 2, the main beam is steered from 24° to 34° with the measured peak gains of 1-3.3 dBi. Moreover, at 1.8 GHz, the main beam is steered from 348° (-12°) to 336° (-24°) with measured gains of 5.8-6.8 dBi and from 12° to 22° with measured gains of 4.7-6.6 dBi for input excitation at port 1 and port 2, respectively. Thus, continuous beam scanning in the forward or backward direction is validated by experimental results obtained at 0.9 GHz and 1.8 GHz, and the slight asymmetry in beam-scanning angles in two opposite directions is mainly attributed to the measurement error and the output amplitude/phase imbalance of the fabricated dual-band beam-scanning/switching network.

In Table 3, careful comparisons are made between this fabricated prototype and the prior art in terms of the beam-scanning/switching technique, the number of applicable operating band(s), the experimentally achieved phase progression $\Delta \phi$ between antenna ports, beam-scanning range, and the array gain. By exploiting the proposed dual-band coupler with synchronously tuned phase differences, it is seen that the proposed beam-scanning/switching network features the remarkable advantage of flexible dual-frequency applications (and beam scanning at two fixed frequencies) whereas the prior work is mainly developed with single-frequency scheme. Moreover, in spite of its simple configuration, i.e., consisting of merely passive tunable couplers and phase shifters, this proposed beam-scanning/switching network provides a cost-effective solution to continuous beam-scanning, the feeding network based, and several-GHz applications without extra complex and active control circuitry [3]. Also note that the practically attainable beam-scanning range is dependent on the antenna’s element pattern. Especially, at 0.9 GHz the beam-scanning range is reduced because the main beam resulted from $\Delta \phi=120°$ is shifted towards boresight by the directional element pattern.

V. CONCLUSION

A novel and simple dual-band beam-scanning/switching network is proposed to contribute to dual-band beam scanning in both of the forward and backward directions. To support the beam-scanning/switching capabilities at two flexible frequencies, for the first time, a varactor-tuned microstrip coupler with continuously and synchronously tunable phase differences is proposed for this network realization. As a proof-of-concept, a 0.9/1.8-GHz beam-scanning/switching network and the array module were prototyped and achieved the measured beam-scanning ranges of -24° to -42° or 24° to 34° at 0.9 GHz, and -12° to -24° or 12° to 22° at 1.8 GHz.

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TABLE 3. Comparison between the fabricated beam-scanning/switching module and the prior art

| Technique | Freq. (GHz) | 'Diff. phase Δϕ(°) | Array BSR(°) | Array gain (dBi) |
|-----------|------------|-------------------|-------------|-----------------|
| Dual-band coupler with synchronously tuned phase differences | 0.9 | $P_1: +60$ to $+120$  $P_2: -120$ to $-60$ | $P_1: -24$ to $-42$  $P_2: +24$ to $+34$ | $P_1: 1.3-3.2$  $P_2: 1.3-3.3$ |
| CRLH leaky-wave antenna | 5 | N.A. | -37 to $+32$ | 4.4-5.7 |
| CMOS 4×4 ‘BM + beam steering network | 25 | $+45$, $-45$, $+135$, $-135$ | $-90$ to $+90$ | X |
| 1-bit reflect-array | 13.5 | N.A. | 0 to $+40$ | 15.7-16.5 |
| Dual-band modified BM | 2.4 | $+90$, $-90$ | 2 switched beams | 9 |
| $24$ | $-25$, $+155$, $-115$, $+65$ | 4 switched beams | X |
| Modified 2×4 BM | 5 | $+90$, $-90$ | 2 switched beams | 12.5 |

1 'Diff. phase between antenna ports. 'Beam-scanning range. 'Port 1 or port 2 as the input. 'Butler matrix. 'Predicted results based on the measured single-channel S-parameters.

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PEI-LING CHI (S’08–M’11–SM’17) received the B.S. and M.S. degrees in communication engineering from National Chiao Tung University (NCTU), Hsinchu, Taiwan, in 2004 and 2006, respectively, and the Ph. D. degree in electrical engineering from the University of California at Los Angeles (UCLA), in 2011. Since 2011, she has been with the National Chiao Tung University, as an Assistant Professor of electrical and computer engineering, where she is currently an Associate Professor. She holds several U.S. and international patents in the area of the left-handed metamaterials and has co-authored over 80 journal and conference papers. Her research interests include the analysis and design of the left-handed metamaterial circuits, design of microwave/mm-Wave components and integrated systems, and development of terahertz antennas and communications. She is an Associate Editor of the IEEE MICROWAVE AND WIRELESS COMPONENTS LETTERS.

CHUN-PIN CHIEN received the B.S. degree in electronic engineering from Chang Gung University, Taoyuan, Taiwan, in 2017, and is currently working toward the M.S. degree in the Institute of Communications Engineering, National Chiao Tung University (NCTU), Hsinchu, Taiwan. His research interests include the design of microwave circuits.

ANRONG CHEN was born in Nanping, Fujian, China. He is currently pursuing the Master degree in electromagnetic field and microwave techniques at the University of Electronic Science and Technology of China (UESTC), Chengdu, China. His current research interests include reconfigurable filters, multibeam antennas.

TAO YANG (S’09–M’11–SM’15) received the B.Eng. and Ph.D. degrees from the University of Electronic Science and Technology of China (UESTC), Chengdu, China, in 2005 and 2011, respectively. He is currently a professor with University of Electronic Science and Technology of China. From February 2016 to September 2017, he was a R&D IC Engineer with Broadcom Ltd., San Jose, CA, USA. From April 2014 to February 2016, he was with Qualcomm Inc, San Diego, CA. From October 2012 to April 2014, he was with the Department of Electrical and Computer Engineering, University of California at San Diego (UCSD). From August 2008 to September 2010, he was a Visiting Scholar with the Electrical Engineering Department, University of California at Los Angeles (UCLA). His research includes Ka-band circuit designs such as Ka-band frequency synthesizers and transceivers; miniaturized passive microwave and millimeter-wave components such as filters, diplexers, triplexers, and baluns; broadband microstrip antennas and leaky-wave antennas; metamaterial-based microwave circuits; design and development of RF passive components for highly integrated RF integrated circuits in deep sub-micrometer CMOS and silicon-on-insulator (SOI) technologies, and development of high Q FBAR resonators and filters.