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Published in:
IEEE Access

Published: 01/01/2018

Document Version:
Final Published version, also known as Publisher’s PDF, Publisher’s Final version or Version of Record

Publication record in CityU Scholars:
Go to record

Published version (DOI):
10.1109/ACCESS.2017.2759961

Publication details:
YE, XIAO. FENG., ZHENG, SHAO. YONG., PAN, YONG. MEI., HO, DEREK., & LONG, YUNLIANG. (2018). A New Class of Components for Simultaneous Power Splitting Over Microwave and Millimeter-Wave Frequency Bands. IEEE Access, 6, 146-158. [8059763]. https://doi.org/10.1109/ACCESS.2017.2759961

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A New Class of Components for Simultaneous Power Splitting Over Microwave and Millimeter-Wave Frequency Bands

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The work was supported in part by the National Natural Science Foundation of China under Grant 61671485, in part by Science and Technology Program of Guangzhou, China under Grant 201510010084, in part by the Fundamental Research Funds for the Central Universities under Grant 16lgzd04, and in part by the Guangdong Natural Science Foundation under Grant 2015A030312010.

Abstract: The co-existence of the microwave and millimeter-wave technologies becomes the inexorable trend of the future wireless communication systems. The corresponding components within the system are required to cover these two frequency bands simultaneously. But the existing dual-/multi-band components cannot satisfy this requirement. For the first time, a new class of components are proposed to provide simultaneous power splitting functions at microwave and millimeter-wave frequency bands. To achieve such a large frequency ratio, an effective feeding approach should be proposed to properly route the signal between the input/output ports and the two elements which operate at microwave and millimeter-wave frequency bands respectively. For universality, the popular microstrip line and substrate integrated waveguide (SIW) structures are utilized for the implementation of microwave and millimeter-wave elements, respectively. An aperture coupling mechanism which can suppress the high order mode of the microstrip structure and excite the $TE_{10}$ mode of the SIW structure at the same time is proposed. A simple transmission model is utilized to explain the working principle along with the theoretical analysis. Based on this novel feeding approach, four dual-/tri-band components which can achieve a large frequency ratio up to 33.3 were designed, fabricated, and measured. Besides the flexibility in operating frequency, the proposed structure can provide arbitrary coupling coefficients (3–10 dB) and even different functionalities at the two frequency bands, which had not been reported in the literature.

Index Terms: Dual band, triple band, large frequency ratio, millimeter-wave, substrate integrated waveguide (SIW), quadrature coupler, power divider, aperture coupling.

I. INTRODUCTION

To deal with the explosive growth of data traffic, emerging new services and scenarios, the millimeter-wave technology will be utilized in the next-generation wireless communication, while the existing microwave technologies will still play an important role. For example, the IEEE standard 802.11ad had already been proposed for the high speed wireless personal network (WPAN) by enabling the switching between 802.11 Wi-Fi networks operating at the microwave frequency bands and those operating at the 60 GHz WiGig bands. It can be expected that the future wireless communication system should be more adaptable to different environments and capable of switching between high-data-rate short-range wireless communications (millimeter wave technologies) and traditional wireless communications (microwave technologies). Thus, the coexistence of microwave and millimeter-wave technologies becomes the inexorable trend of the future wireless communication systems. This motivates studies of the front ends which can support microwave and millimeter wave technologies simultaneously. As the critical component to transmit/receive the signals within the system, the antenna should cover microwave and millimeter-wave frequency bands simultaneously. Two antenna structures were reported to achieve a large frequency ratio by cleverly combining two configurations [1], [2]. Unfortunately, two different input ports should be used for the two operating frequency bands.
This may become a problem when the signals with different operating frequencies should be transmitted through the same signal chain, which had been widely used in the existing wireless communication standards, such as 802.11 a/b/g. The same requirement also emerges for the implementation of different circuits within the front end. For the dual-band operation, the utilization of components with dual mode nature is the most straightforward approach. The stub-loaded resonators (SLR) had been widely used in the design of bandpass filters [3], [4], power dividers [5], [6], and couplers [7]–[10]. The coupled-line section can be also utilized to implement different dual-band power splitting components [11]–[13]. In [14], the composite right/left-handed (CRLH) transmission lines based on the folded substrate integrated waveguide (FSIW) was proposed to realize a dual-band quadrature coupler. Complementary split ring resonators (CSRRs) were loaded onto a patch quadrature coupler to provide arbitrary power division ratios over two arbitrary frequency bands [15]. Alternatively, the coplanar waveguide feeding mechanism was proposed to feed two elements simultaneously for independently controlled center frequencies and bandwidths [16], [17]. To further increase the system integration level, the relevant techniques have been extended for triple band operation [18]–[22]. However, the existing dual-/multiple-band structures can only provide a maximum frequency ratio up to 4, which is not large enough to cover the microwave and millimeter wave frequency bands simultaneously. Therefore, the realization of components which can support microwave and millimeter wave frequencies simultaneously remains a key problem for the next generation wireless communication systems.

In this paper, a new class of components which can cover the microwave and millimeter-wave frequency bands simultaneously are proposed. To combine two elements operating at different frequency bands within a device for dual band operation, a novel aperture coupling mechanism is proposed to implement the harmonic suppression of the microstrip lines and excitation of the TE_{10} mode of the substrate integrated waveguide (SIW) structures simultaneously. This paper is organized as follows. Section II gives the theoretical analysis on simultaneous harmonic suppression and excitation properties of an aperture. For demonstration, a simple transmission model is presented. Based on this mechanism, a dual-band 3-dB quadrature coupler operating at 0.9 and 30 GHz is proposed in Section III. To demonstrate the independent operations at microwave and millimeter-wave frequencies, a dual-band quadrature coupler is designed to provide 3 dB and 10 dB coupling coefficients at 0.9 and 30 GHz respectively. The dual-band quadrature coupler is further extended for triband operation by simply introducing a dual-band structure for the microwave frequency band, as presented in Section IV. In Section V, the function of the proposed circuit at millimeter wave frequency is changed to a three-port power divider, which is totally different from that at microwave frequency. Finally, a conclusion is drawn in Section VI.

II. ANALYSIS OF A NOVEL APERTURE COUPLING STRUCTURE FOR DUAL-BAND OPERATION

A. OPERATING PRINCIPLE

As stated in the introduction, the implementation of a dual-band component usually relies on the dual-mode nature of a structure resulting in a limited frequency ratio. To extend the frequency ratio significantly, the approach to feed two elements operating at two specific frequencies is the promising one. However, the realizable frequency ratio is limited by the requirement on the bandwidth of the feeding methodology and comparable sizes for the two elements, as reported in [16] and [17]. Moreover, it is also affected by the interaction between the two elements especially for the large frequency ratio case. For example, the operating frequency of the high-frequency element may locate near the harmonic frequency or high order mode frequency of the low-frequency one resulting in unwanted interaction, the implementation of dual-band operation becomes difficult. Therefore, an effective approach to independently feed two elements with a very large frequency ratio is desired for the realization of components which can support microwave and millimeter-wave frequencies simultaneously.

As the most common structures to implement the microwave and millimeter-wave elements, the microstrip line and SIW are considered respectively. However, there is no feeding mechanism reported to feed these two elements simultaneously. So a new mechanism to feed these two elements and suppress the interaction between the microstrip element and SIW element should be proposed. The aperture coupling is commonly used as the feeding structure in patch antennas and dielectric resonator antennas [23]–[25]. Similarly, it can be also utilized to feed the SIW element [26], [27]. Therefore, the simultaneous feeding of microstrip and SIW elements becomes possible. But the problem with the interaction between two elements remains unsolved. Because the frequency for the fundamental mode of the SIW element is much higher than that of the microstrip element, the SIW element does not affect the characteristics of the microstrip element under the fundamental mode. But the higher order operating mode of a microstrip element affects the characteristic of a SIW element. It should be properly suppressed to eliminate the interaction between two elements. Since the aperture under a microstrip line belongs to the family of patterned ground structures, it has the band-stop characteristics [28], [29]. And numerous low-pass filters have been implemented accordingly [30]–[32]. Therefore, the aperture coupled structure will be investigated in the following sections for the implementation of suppressing the unwanted characteristics of the microstrip element and exciting the SIW element simultaneously.

B. SUPPRESSION OF THE MILLIMETER-WAVE CHARACTERISTICS OF A MICROSTRIP STRUCTURE

To demonstrate the feasibility, a simple circuit model consisting of a 50 Ω microstrip line and a SIW structure is shown in
Two apertures are employed on the middle metal layer. Their locations are indicated using the parameter $P_s$. In this model, two pieces of the Rogers RT/Duroid 5880 dielectric substrate with a relative dielectric constant of 2.2 are used. The dimensions of the transmission model are as follows:

- $H_1 = H_2 = 0.254$ mm,
- $L_{siw} = 8.1$ mm,
- $W_{m} = 0.78$ mm,
- $W_{siw} = 6$ mm,
- $L_{slot} = 3.65$ mm,
- $W_{slot} = 0.15$ mm,
- $P_s = 0.65$ mm,
- $R_{via} = 0.15$ mm,
- $P_{via} = 0.5$ mm.

$R_{via}$ is the radius of the metal via, $P_{via}$ is the distance between two shorting vias. For the microstrip line with apertures, the aperture can be modeled by a parallel resonant circuit exhibiting the band-stop property [29]. It is worthy pointing out that the location of aperture does not affect the characteristics of the circuit at microwave frequency band. Fig. 2 shows the equivalent circuit of the simple model without SIW element.

The anti-resonance of the parallel $RLC$ circuit occurs when the electrical length of the slot is about $0.5\lambda_g$, where $\lambda_g$ is the guided wavelength at the operating frequency $f_0$ of the SIW element under TE$_{10}$ mode. The length of the apertures to maintain a good rejection for the microstrip line at $f_0$ can be obtained as:

$$L_{slot} = \frac{c_0}{2f_0} \cdot \sqrt{\frac{\varepsilon_{r1} + 1}{2}}$$

where $\varepsilon_{r1}$ is the relative dielectric constants of two pieces of the substrate used, $c_0$ is the speed of light in free space. The frequency response of this model is shown in Fig. 3.

As expected, a good rejection larger than 16 dB and return loss better than 1 dB have been achieved from 25 to 35 GHz. Therefore, it is confirmed that the high order mode of a microstrip element can be effectively suppressed by introducing the half wavelength aperture underneath the microstrip feed line.

The relationship between the width of the TE$_{10}$ mode SIW and operating frequency can be obtained as [35]:

$$W_{siw} = 4.32 \cdot \frac{R_{via}^2}{P_{via}} + 0.4 \cdot \frac{P_{via}}{W_{siw}} = \frac{c_0}{2f_c \cdot \sqrt{\varepsilon_r}}$$

where $f_c$ is the cutoff frequency of the TE$_{10}$ mode.
For further verification, the characteristics of the simple transmission model at the millimeter-wave frequency is evaluated using ANSYS HFSS. Fig. 5 shows the simulated E-field distribution within the SIW element at 30 GHz. To further show the effectiveness of the aperture in coupling signal to the SIW element, Fig. 6 illustrates the frequency responses of the transmission model with/without the SIW element. It can be observed that the millimeter-wave signal can be guided from port 1 to the SIW element, and further routed from the SIW element to port 2 through the two apertures respectively. A good transmission with the return loss better than 15 dB and the insertion loss better than 0.5 dB has been achieved from 28.9 to 32.4 GHz, indicating the effectiveness of the proposed aperture coupling mechanism.

III. THE DUAL-BAND QUADRATURE COUPLER CONFIGURATION TO COVER MICROWAVE AND MILLIMETER-WAVE FREQUENCY BANDS SIMULTANEOUSLY

A. THE DUAL-BAND QUADRATURE COUPLER CONFIGURATION

To demonstrate the design concept, a dual-band 3-dB quadrature coupler (prototype I) operating at 0.9 and 30 GHz is designed, firstly. Fig. 7(a) shows the configuration of the proposed coupler. A quadrature coupler based on the classical branch lines configuration is implemented on the top layer for the operation at microwave frequency. For miniaturization, the branch-line coupler is folded inwards like a flower. The electrical length of the fan-shaped microstrip line $L_4$ is set to be quarter-wavelength at the millimeter-wave frequency to introduce a wideband virtual short near the joint with the 50-ohm microstrip line [36]. Another quadrature coupler based on the SIW cruciform directional coupler configuration is implemented between the middle and bottom metal layers for the operation at the millimeter-wave frequency. The detailed design procedures can be found in [37]. Two metallic vias are arranged along the diagonal line of the SIW cruciform to control the coupling strength. In addition, four inductive vias in each SIW arms are necessary for...
good matching [38]. The novel aperture coupling approach described previously is applied on the middle metal layer. To enhance the rejection property of the aperture at the millimeter-wave frequency, two parallel open stubs are employed to act as the parallel capacitance [30]. The detailed parameter specification of the proposed coupler is given in Fig. 7(b).

To further demonstrate the flexibility of proposed configuration, the coupling coefficient of the coupler at 30 GHz is changed to 10 dB with other parameters fixed. Owing to the independent operations at microwave and millimeter-wave frequencies achieved by the proposed configuration, another dual-band coupler which exhibits coupling coefficients of 3 and 10 dB at 0.9 and 30 GHz respectively can be realized. Fig. 8 shows the configuration and parameter. The central loaded metallic via is utilized to adjust the coupling coefficient of a SIW cruciform directional coupler [39]. The 10 dB coupling coefficient can be easily obtained at Port 3 of the coupler by fine tuning the width of the inductive coupling windows of the SIW coupler ($W_s$) without affecting the characteristics of the circuit at the microwave frequency.

**B. DESIGN PROCEDURE AND EXPERIMENTAL RESULTS**

Owing to the simplicity of proposed feeding mechanism and the independent operations of the microwave and millimeter-wave elements, a simple design procedure can be summarized as follows.

1) According to the desired microwave operating frequency and specified substrate material, calculate the initial values for a classical branch line coupler. Fold the four branches for the miniaturization of circuit size.

2) Set the initial length of the aperture as one-half wavelength and the length of the fan-shaped microstrip as a quarter-wavelength at the target millimeter-wave operating frequency respectively. Choose a relatively narrow width for the apertures to minimize any potential radiation loss while transmitting the EM power from the microstrip line to the SIW element.

3) Adjust the dimensions of the apertures to obtain a good suppression at the target millimeter-wave operating frequency with the SIW structure removed.

4) Calculate the initial width ($W_{siw}$) of each TE$_{10}$ SIW arms using the equations (2). Set the initial length of each TE$_{10}$ SIW arms to be one wavelength at the desired millimeter-wave operating frequency.

5) Combine the SIW and microwave elements into a circuit, and fine tune the position of inductive posts and apertures for the desired coupling coefficient and a good mode matching.

According to the previous design procedure, two dual-band couplers operating at 0.9 and 30 GHz simultaneously are designed. Two pieces of substrate Rogers Duroid 5880 with a dielectric constant of 2.2 and thickness of 0.254 mm are used. The obtained dimensional parameters of prototypes I and II are given in Table 1 ($P_{via}$: the distance between two vias; $R_{via}$: the radius of the metal vias). It should be noted that the dimensions of the aperture in the proposed coupler are: $L_s = 2.8$ mm, and $W_s = 0.5$ mm, while the dimensions of aperture in the transmission model given in Section II are: $L_{slot} = 3.65$ mm, and $W_{slot} = 0.15$ mm. This disagreement is mainly caused by the difference in lengths. As the slot length decreases, the anti-resonant frequency increases slowly with the slot width fixed [29]. In other words, for a specified frequency to be suppressed, the shorter the length is, the wider the width should be. All of the measured data are obtained by Keysight N5247A PNA-X network analyzer.

**TABLE 1. Dimensions of the proposed prototype I and II.**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>$W_0$</th>
<th>$W_1$</th>
<th>$W_2$</th>
<th>$W_3$</th>
<th>$W_4$</th>
<th>$W_5$</th>
<th>$W_6$</th>
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<tbody>
<tr>
<td>Prototype I</td>
<td>0.8</td>
<td>1.35</td>
<td>0.8</td>
<td>0.4</td>
<td>5.05</td>
<td>0.5</td>
<td>3.17</td>
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<tr>
<td>Prototype II</td>
<td>0.8</td>
<td>1.35</td>
<td>0.8</td>
<td>0.4</td>
<td>2.13</td>
<td>0.5</td>
<td>3.22</td>
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<table>
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<tr>
<th>Parameter</th>
<th>$W_{siw}$</th>
<th>$L_1$</th>
<th>$L_2$</th>
<th>$L_3$</th>
<th>$L_4$</th>
<th>$L_5$</th>
<th>$L_6$</th>
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</thead>
<tbody>
<tr>
<td>Prototype I</td>
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<td>56</td>
<td>56</td>
<td>6.6</td>
<td>2.75</td>
<td>6.45</td>
<td>5.5</td>
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<tr>
<td>Prototype II</td>
<td>4.22</td>
<td>56</td>
<td>56</td>
<td>6.55</td>
<td>2.8</td>
<td>6.45</td>
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<tr>
<th>Parameter</th>
<th>$L_7$</th>
<th>$P_t$</th>
<th>$P_{via}$</th>
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<td>Prototype I</td>
<td>2.8</td>
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<tr>
<td>Prototype II</td>
<td>2.8</td>
<td>0.6</td>
<td>0.5</td>
<td>0.15</td>
</tr>
</tbody>
</table>

The simulated and measured frequency responses of the fabricated prototypes I and II are shown in Fig. 9 and 10.
FIGURE 9. The photographs and frequency response of the fabricated prototype I. (a) Photographs. (b) Amplitude response at the microwave frequency band. (c) Phase difference at the microwave frequency band. (d) Band-rejection characteristic of the apertures. (e) Amplitude response at the millimeter-wave frequency band. (f) Phase difference at the millimeter-wave frequency band.

together with the corresponding photographs, respectively. The phase difference between two output ports for all couplers is defined as $\Delta S_{21} - \Delta S_{31}$. The frequency response of the prototype I at microwave frequency is shown in Fig. 9(b). An excellent agreement between the simulation and measurement is observed. The measured $|S_{21}|$ and $|S_{31}|$ follow each other closely with the amplitude imbalance smaller than 0.6 dB and return loss better than 15 dB throughout the measured frequency range from 0.78 GHz to 0.96 GHz, which corresponds to a relative bandwidth of 20.7% centered at 0.87 GHz. The phase difference between the two output ports is $87.6 \pm 0.6^\circ$ over the same band, as depicted in Fig. 9(c). For illustration purposes, the millimeter-wave characteristics of the microstrip element with apertures applied are also investigated by removing the SIW element. A good rejection with $|S_{21}|$ lower than $-25$ dB and $|S_{11}|$ higher than $-1$ dB has been achieved from 27 to 33 GHz, as depicted in Fig. 9(d). The frequency response of the prototype I at the millimeter-wave frequency is shown in Fig. 9(e). The simulated $|S_{21}|$ and $|S_{31}|$ are both $-3.9 \pm 0.5$ dB with return loss ($|S_{11}|$) and isolation ($|S_{41}|$) better than 15 dB across the frequency range from 29.2 GHz to 30.6 GHz, which corresponds to a relative bandwidth of 4.7% centered at 30 GHz. The simulated phase difference between the two output ports is $92 \pm 3^\circ$ over the same band, as depicted in Fig. 9(f). A reasonable agreement between simulation and measurement can be observed. The measured $|S_{21}|$ and $|S_{31}|$ follow each other closely with the amplitude imbalance smaller than 1.5 dB throughout the measured frequency range from 29 GHz to 30.8 GHz, which corresponds to a relative bandwidth of 6.1% centered...
The photographs and frequency responses of the fabricated prototype II are shown in Fig. 10(b) and (c). The characteristics are almost the same as those of prototype I, which confirms the independence between the microwave and millimeter-wave operations achieved by the proposed aperture coupling mechanism. The frequency responses of the prototype II at millimeter-wave frequency are shown in Fig. 11(d) and (e). Within the operating frequency band from 29.3 GHz to 30.7 GHz, both the simulated return loss ($|S_{11}|$) and isolation ($|S_{41}|$) are better than 13 dB, the simulated coupling coefficient at Port 3 ($|S_{31}|$) is 10±0.5 dB. The simulated phase difference between the two output ports is 92±5° over the same band. A reasonable agreement between simulation and measurement can be also observed. The coupling coefficient ($|S_{31}|$) is measured to be 10.9±0.4 dB with return loss ($|S_{11}|$) and isolation ($|S_{41}|$) better than 15 dB from 29.2 GHz to 30.4 GHz. The measured phase difference between the two output ports is 93±6° over the same band.

The discrepancy between the simulation and measurement results of the fabricated circuit at millimeter-wave frequency band is mainly caused by the fabrication errors. The increase in insertion loss observed in the measurements is caused by the dielectric loss and leakage loss from the air gap between layers in the fabricated prototypes, which have not been accounted in the simulation.

To show the advantages of the proposed structure, Table 2 summarizes the comparison between the proposed dual-band couplers with other structures that can be found in the literature. First of all, the proposed configuration is found to
exhibit the largest frequency ratio up to 33.3, which cannot be implemented by the existing techniques including the stub-loaded resonators (SLR) [7]–[10], coupled-line [12], the composite right/left-handed (CRLH) transmission lines [14] and complementary split ring resonators (CSRRs) [15], coplanar waveguide feeding mechanism [16], [17]. It is worth mentioning that, most of the existing dual-band structures can only provide a limited frequency ratio up to 4. More importantly, the proposed structure is the first dual-band quadrature coupler structure to cover the microwave and millimeter-wave frequency bands simultaneously, which is a critical unresolved issue for the next generation wireless communication systems. In addition, this structure shows the independent operations at the two frequency bands and the high flexibility in controlling operating frequencies, which cannot be implemented by the existing structures [7]–[10], [12], [14]. Besides, the proposed structure exhibits the simplicity in structure to provide arbitrary coupling coefficients at two frequency bands compared to the structures reported in [12] and [15]. The number of design variable for the implementation of different coupling coefficients required in the proposed structure is only one.

### IV. THE TRI-BAND QUADRATURE COUPLER CONFIGURATION TO COVER MICROWAVE AND MILLIMETER-WAVE FREQUENCY BANDS SIMULTANEOUSLY

#### A. THE TRI-BAND QUADRATURE COUPLER CONFIGURATION

As a mature technology, the microwave technology had been widely utilized in the modern wireless communication systems. Thus there already exist several wireless communication standards operating at different microwave frequency bands. To meet the demand of the future wireless communication systems in supporting multiple standards whose operating frequencies are located at both microwave and millimeter-wave frequency bands, a tri-band quadrature coupler (prototype III) is proposed accordingly. Based on the stub loading approach reported in [9], a dual-band branch-line coupler is designed at 0.9 and 2.4 GHz for the dual-band operation firstly. Owing to the independent operation achieved by the proposed feeding approach, the dual-band

<table>
<thead>
<tr>
<th>TABLE 2. Comparison with other dual-band couplers.</th>
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<td>Frequency (GHz)</td>
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<tr>
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<tr>
<td>( f_0 )</td>
</tr>
<tr>
<td>[7]</td>
</tr>
<tr>
<td>[8]</td>
</tr>
<tr>
<td>[9]</td>
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<tr>
<td>[15]</td>
</tr>
<tr>
<td>This work</td>
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microwave coupler is further combined with the 3 dB SIW coupler proposed in Section III to realize the desired tri-band operation. Fig. 11(a) shows the configuration and parameter specification of the proposed tri-band coupler. The dual-band branch-line coupler operating at 0.9 and 2.4 GHz simultaneously is located on the top layer. In this configuration, the electrical lengths of all branch lines including the open stubs are set as quarter-wavelength at the center frequency of the two operating bands. The impedances of these branch lines/stubs can be calculated by the closed-form design formulas in [9].

The main difference between the proposed coupler and the structure in [9] is the cross coupling between the feed lines, which are used to split the attenuation pole located at the center frequency of the two operating bands into two poles, resulting in the good suppression and sharp roll off [40]. Under the dual-band coupler is the aperture etched on the middle metal layer. The remaining design procedures are similar to those described in Section III. B for prototypes I and II.

**B. EXPERIMENTAL RESULTS**

For experimental validation, the prototype III with tri-band characteristics is designed, fabricated and measured. The dimensional parameters of prototype III are given in Table 3. As can be observed, the dimensions of the SIW structure remain almost the same as those of the prototype I.

The simulated and measured frequency responses of the prototype III are depicted in Fig. 12 together with...
TABLE 3. Dimensions of the proposed prototype III.

<table>
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<tr>
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<td>0.5</td>
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<td></td>
<td>3.16</td>
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<th>Dimension</th>
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<th>$L_2$</th>
<th>$L_3$</th>
<th>$L_4$</th>
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<td>17.2</td>
<td>6.8</td>
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<td>30.8</td>
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the photographs. The frequency responses of the prototype III at microwave frequency are shown in Fig. 12(b) and (c). Similar to previous two designs, an excellent agreement between the simulation and measurement can be observed. A high suppression level better than 40 dB has been achieved between the 0.9 and 2.4 GHz frequency bands. As observed in Fig. 12(b), the measured insertion losses ($|S_{21}|$ and $|S_{31}|$) are $-3.8 \pm 0.5$ dB and $-4.4 \pm 0.5$ dB with return loss and isolation better than 15 dB within the two frequency bands (0.8-0.92 GHz and 2.3-2.43 GHz) respectively. The measured phase differences are $89 \pm 2^\circ$ and $90 \pm 8^\circ$ within the two operating frequency bands simultaneously, as depicted in Fig. 12(c). Phase ripples can be observed between 1.5 GHz and 2.0 GHz. This is caused by the very weak signals obtained at Ports 2 and 3. As given in Fig. 12(b), the magnitudes of $S_{21}$ and $S_{31}$ are smaller than $-30$ dB from 1.5 GHz to 1.9 GHz. Thus the corresponding results for phase characteristics are not accurate resulting in the unwanted ripples. The fabricated coupler occupies the relative bandwidths of 14% and 5.5% at 0.9 and 2.4 GHz band respectively.

The simulated frequency responses of the prototype III at millimeter-wave frequency are shown in Fig. 12(d) and (e). Within the operating frequency from 29.2 GHz to 30.8 GHz, the return loss ($|S_{11}|$) and isolation ($|S_{41}|$) are better than 15 dB, the $|S_{21}|$ and $|S_{31}|$ are $-4.6 \pm 0.8$ dB. The simulated phase difference between the two output ports is $89 \pm 5^\circ$ over the same band. A reasonable agreement between simulation and measurement can be observed. The measured $S_{21}$ and $S_{31}$ follow each other closely with the amplitude imbalance smaller than 1.4 dB with return loss ($|S_{11}|$) and isolation ($|S_{41}|$) better than 15 dB throughout the measured frequency range from 28.7 GHz to 30.5 GHz. The measured phase difference between the two output ports is $87 \pm 5^\circ$ over the same band. Similar to the prototypes I and II, this circuit demonstrates the high flexibility of the proposed configuration in operating frequencies. Besides, the frequency ratio is high as 33.3, which cannot be implemented by the existing structures [21], [22].

V. MICROWAVE COUPLER AND MILLIMETER-WAVE POWER DIVIDER APPLICATION

A. A NEW CONFIGURATION TO PROVIDE DIFFERENT POWER SPLITTING FUNCTIONALITIES OVER TWO FREQUENCY BANDS

Both the dual-band and triple-band configurations to cover the microwave and millimeter-wave frequency bands simultaneously had been proposed in the previous sections. Similar to all the previous dual-band and triple-band configurations, the functionalities for the two operating frequency bands should be the same. With the rapid development of modern wireless communication, the system is required to exhibit higher flexibility. However, the required circuit functionalities maybe different for the two frequency bands. For example, the in phase property is also widely demanded in different wireless communication systems [41], [42]. However, none of the existing dual-band configurations can achieve such a high flexibility. Based on the signal routing property of proposed aperture coupling structure, the desired circuit can be simply implemented by combining a microstrip 3-dB quadrature coupler with a SIW in-phase power divider. For ease of design, the T-junction SIW configuration is utilized to realize the in phase equal power division at the millimeter-wave frequency. The detailed design principle of the T-junction SIW power divider had been given in [43]. Based on the design procedures provided in Section III, the prototype IV can be implemented accordingly. Fig. 13 shows the configuration and parameter specification of the proposed prototype IV.

B. EXPERIMENTAL RESULTS

For validation, the prototype IV which exhibits the quadrature equal power division and in phase equal power division at 0.9 and 29 GHz respectively, are designed, fabricated and measured. It should be noticed that the objective operating frequency of prototypes I over millimeter-wave frequency
V. SIMULATION AND MEASUREMENT RESULTS

The frequency responses of the prototype IV at microwave frequency band is set as 29 GHz to show the flexibility in operating frequency of the proposed structure. The obtained dimensional parameters of prototype IV are given in Table 4. As can be observed, the dimensional parameters of the 3 dB microstrip coupler structure are almost the same as those for the proposed prototype I.

The simulated and measured frequency responses of the prototype IV are shown in Fig. 14 together with the photographs. The frequency responses of the prototype IV at microwave frequency are shown in Fig. 14(b), (c) and (d). It can be found that the frequency responses at microwave frequency band keep almost the same as those of prototype I, which again confirms the independence of the microwave and millimeter-wave operations achieved by the proposed approach.

![Photographs and frequency responses of the fabricated prototype IV.](image)

**TABLE 4. Dimensions of the proposed prototype IV.**

<table>
<thead>
<tr>
<th>Dimension</th>
<th>$w_0$</th>
<th>$w_1$</th>
<th>$w_2$</th>
<th>$w_3$</th>
<th>$w_4$</th>
<th>$w_5$</th>
<th>$w_{nw}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Value (mm)</td>
<td>0.8</td>
<td>1.35</td>
<td>0.8</td>
<td>0.4</td>
<td>1.25</td>
<td>0.5</td>
<td>0.4</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Dimension</th>
<th>$L_1$</th>
<th>$L_2$</th>
<th>$L_3$</th>
<th>$L_4$</th>
<th>$L_5$</th>
<th>$P_1$</th>
<th>$P_2$</th>
<th>$P_{wa}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Value (mm)</td>
<td>28</td>
<td>28</td>
<td>1.9</td>
<td>3</td>
<td>1.7</td>
<td>2.8</td>
<td>0.55</td>
<td>0.6</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Dimension</th>
<th>$R_{wa}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Value (mm)</td>
<td>0.2</td>
</tr>
</tbody>
</table>

The frequency responses of the prototype IV at millimeter-wave frequency are shown in Fig. 14(d). The simulated $|S_{21}|$ and $|S_{41}|$ are $-4.1 \pm 0.4$ dB with return loss ($|S_{11}|$) better than 15 dB throughout the simulated frequency range from 26.7 GHz to 31 GHz, which corresponds to a relative bandwidth of 14.9% centered at 28.9 GHz. The simulated and measured results are in good agreement. The measured $|S_{21}|$ and $|S_{41}|$ follows each other closely with the amplitude imbalance smaller than 1.4 dB throughout the frequency range from 26.7 GHz to 31 GHz, which corresponds to a relative bandwidth of 14.9% centered at 28.9 GHz. The measured $|S_{21}|$ and $|S_{41}|$ are both $-4.8 \pm 0.7$ dB with return loss ($|S_{11}|$) better than 15 dB from 26.7 GHz to 30.8 GHz. It should be pointed out that this circuit exhibits lower additional loss at the millimeter-wave frequency band compared to prototypes I and II. This is because the T-junction SIW configuration is less affected by the fabrication error owing to its simple structure.

VI. CONCLUSION

A new class of components which can provide simultaneous power division over the frequency bands with a very large ratio is proposed for the coexistence of microwave and millimeter-wave technologies required in the future communication systems. A novel aperture coupling mechanism is utilized to route the microwave and millimeter-wave signals for the independent operations of the two frequency bands. Besides the flexibility in operating frequency, arbitrary coupling coefficient can be also achieved. Furthermore, different circuit functionalities with different numbers of input/output ports had been implemented over two different frequency bands for the first time.
This paper has demonstrated the concept of implementing a dual-/tri-band component to cover microwave and millimeter-wave frequency bands simultaneously. This concept can be easily extended to other types of circuits such as filters, baluns and so on. This concept can also be considered for applications in antennas. The dual-/tri-band operation for the antenna can be easily implemented based on the proposed feeding mechanism. However, the radiation characteristics such as the direction of maximum radiation would need to be investigated further, as the two radiating elements may affect each other.

REFERENCES


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