Modified Dual Band Gysel Power Divider With Isolation Bandwidth Improvement

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Abstract—**This paper presents a new design of a dual band power divider with high power handling capability and isolation bandwidth improvement, which contains six branch line sections, two grounded resistors, one resistor between output ports, an extension line at the input, and a single stub line at the end. Exact closed-form equations are derived for the proposed structure using even-mode and odd-mode analysis. In this design, flexibility in choosing two grounded resistors enables us to determine the frequency ratio range, bandwidth of the circuit, and the level of power handling capability. In addition, line admittance analysis shows that the proposed structure can support the acceptable frequency ratio range. For verification, the proposed structure is simulated, fabricated and measured using microstrip lines. Simulation and measurement results show wide isolation bandwidth, while maintaining high power handling capability over the Wilkinson power divider.**

Keywords-component; dual band, isolation bandwidth, power divider

I. INTRODUCTION

Power dividers are widely used in the design of radar and communication systems. Among various power dividers, the Gysel power divider is suitable for high power applications, because of the using two grounded resistors [1]. In the Gysel power divider, there is a direct path to the external ground for heat sinking, while in the Wilkinson power divider, an internal isolation resistor is embedded, and there is not any path for generated heat sinking [2].

Recent years, several Gysel power dividers have been designed for dual band operations. In [3], a dual band Gysel power divider using an open and a short stub is proposed. The extended structure of [3] with unequal power dividing ratio and arbitrary termination resistance is developed in [4]. The major drawback of the reported power dividers in [3] and [4] is their limited isolation bandwidth. A dual band Gysel power divider using microstrip/slotline phase inverter with the features of the wide frequency ratio range and broadband operation is suggested in [5]. However, the complex structure of ground in the phase inverter design makes the fabrication process more rigorous and more expensive. Another design of dual band Gysel power divider using coupled lines is employed in [6]. In this design, using coupled lines increases the size of the circuit and the difficulty of fabrication process. In [7], a dual band out of phase Gysel power divider with double-sided parallel strip lines (DSPSLs) is employed; however the cost of multi-layer fabrication process is high. Another structure for the dual band Gysel power divider using an extension line at the input and a stub line at the opposite end is reported in [8]. Moreover, in [9], a dual band Gysel power divider using a dual band transformer is proposed. But structures in [8] and [9] suffer from the limited isolation bandwidth, too.

In this paper we propose a dual band power divider with high power handling capability, overcoming the mentioned problems. The proposed structure has the following features: 1) flexibility in choosing two load resistors to determine the level of power handling capability, frequency ratio range, and isolation bandwidth; 2) wide isolation bandwidth (73.33%); 3) achievable structure without the difficulty in fabrication process; 4) using meandering lines to reduce the occupied area. This power divider is based on the reported power divider in [8]. In our structure, for enhancing the isolation bandwidth, one isolation resistor between output ports is embedded. Using this method enables us to choose the load resistors with more flexibility for determining the level of power handling capability, frequency ratio range, and bandwidth of the circuit. In section II, explicit closed-form equations are obtained by using even-mode and odd-mode analysis. In addition, line admittance analysis is performed to determine the frequency ratio range. For verification, simulation and measurement results of the fabricated power divider are reported in section III. Finally, the paper is concluded in section IV.

II. DESIGN PROCEDURE AND ANALYSIS

Fig. 1 depicts the structure of the proposed dual band power divider, in which all admittance values are normalized to the source admittance ($y_0 = 1$). This power divider consists of two load resistors with normalized conductance g for better power handling capability, one resistor between output ports with normalized conductance g_0 for enhancing the isolation bandwidth, extension line y_1 at the input, and single stub line y_s between the two branch line sections with the admittance of y_4 for the dual band operation. All the transmission lines in the proposed power divider have the same electrical length θ_0. In

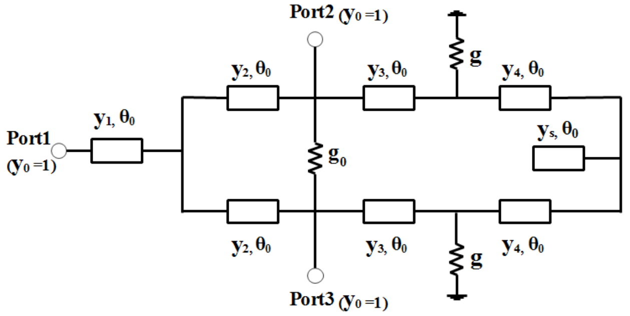

Fig. 1. Proposed structure of the dual band power divider

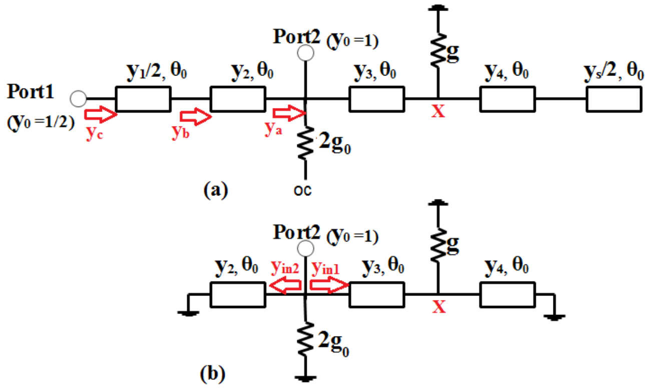

Fig. 2. Equivalent circuit of the proposed power divider. (a) even-mode half circuit. (b) odd-mode half circuit.

this section, we use even-mode and odd-mode analysis to determine the parameters of the proposed structure.

Fig. 2(a) shows the equivalent even-mode half circuit of the proposed power divider under the even-mode excitation. In this mode, the admittance of port 1 is halved and the conductance g_0 becomes $2g_0$. This conductance can be omitted, because no current flows through it. For having matched ports with lossless transmission lines, power must not be dissipated in the conductance of g_0 so shorting node X to ground is required. In other words, the admittance of cascaded stubs y_4 and $y_s/2$ lines must be infinity. Thus, this condition can be expressed as (1)

$$y_4 \frac{j(\frac{y_s}{2})\tan\theta_0 + jy_4\tan\theta_0}{y_4 + j(j(\frac{y_s}{2})\tan\theta_0)\tan\theta_0} \rightarrow \infty. \quad (1)$$

Under the lossless condition (1), we can ignore g, y_4 and y_s in the even-mode analysis, and y_3 is terminated by the virtual ground. So, according to Fig. 2(a), we have

$$y_a = 1 - jy_3\cot\theta_0 \quad (2)$$

$$y_b = y_2 \frac{y_a + jy_2\tan\theta_0}{y_2 + jy_a\tan\theta_0} \quad (3)$$

$$y_c = \frac{y_1}{2}\frac{y_b + j(y_1/2)\tan\theta_0}{y_1/2 + jy_b\tan\theta_0} \quad (4)$$

It is noted that port 2 is matched with y_0=1.

In the odd-mode analysis, because of the virtual ground along the symmetry plane, y_1 and y_s is not regarded. By consideration of matching ports and perfect isolation of the power divider, the odd-mode half circuit equations are expressed as (5)-(7)

$$y_{in1} + y_{in2} + 2g_0 = 1 \quad (5)$$

$$y_{in1} = y_3 \frac{(-jy_4\cot\theta_0 + g) + j(y_3)\tan\theta_0}{y_3 + j(-jy_4\cot\theta_0 + g)\tan\theta_0} \quad (6)$$

$$y_{in2} = -jy_2\cot\theta_0 \quad (7)$$

Suppose that the parameters y_1 and g are known variables. After a few manipulations, from (1)-(7) the unknown variables y_2, y_3, y_4, y_s, and g_0 can be determined simultaneously as

$$y_2 = \frac{2y_1\tan\theta_0 - 2y_1^3\tan\theta_0 \pm\sqrt{\Delta_1}}{4(y_1^2-1)\tan\theta_0 + 2y_1^2\tan^3\theta_0 - 2y_1^2\cot\theta_0} \quad (8)$$

$$y_3 = \frac{y_1y_2 + (2y_2^2 - y_1^2)\tan^2\theta_0}{y_1 + 2y_2} \quad (9)$$

$$y_4 = \frac{(4y_2y_3 + 2y_3^2)\cot\theta_0 - 2y_3^2\tan\theta_0 \pm\sqrt{\Delta_2}}{-4(y_3+y_2)\cot\theta_0} \quad (10)$$

$$g_0 = \frac{y_3 + y_4 - g(y_3+y_2)}{2(y_3+y_4)} \quad (11)$$

$$y_s = 2y_4\cot^2\theta_0 \quad (12)$$

where

$$\Delta_1 = (2y_1^6 - 4y_1^4)\tan^2\theta_0 + (4y_1^6 - 2y_1^4)\tan^4\theta_0 + 2y_1^6\tan^6\theta_0 - 2y_1^4 \quad (13)$$

$$\Delta_2 = (2y_3^2\tan\theta_0 - 2y_3^2\cot\theta - 4y_2y_3\cot\theta)^2 + 16(y_2+y_3)(y_3^3 - g^2y_3 - g^2y_2 - y_2y_3^2\cot^2\theta) \quad (14)$$

The electrical lengths of all transmission lines are chosen as follows for dual band characteristics at two bands of interest f_1 and f_2 [10]

$$\theta(f_1) = \frac{\pi}{1+f_2/f_1}, \qquad \theta(f_2) = \frac{\pi}{1+f_1/f_2} \quad (15)$$

For fabrication process, we must take into account the admittance range constraints (e.g., 0.37-2.5), which is obtained by choosing y_1 and g properly. Fig. 3 shows the variation of normalized admittance versus frequency ratio for three different values of g. In this figure, y_1 is selected to 0.9 for having better frequency ratio range. According to Fig. 3, when g = 0.666 the frequency ratio range would be between 1.83 and 2.3 GHz, for g = 0.5, the upper bound of frequency ratio range increases to 2.44 (because of increasing g_0) and the lower bound increases to 1.87. If g =0.333 the lower bound increases to 1.89 and the upper bound remains unchanged. Thus, we can see that when g = 0.5, the frequency ratio range is wider. On the other hand, based on the analysis method in [11], lower

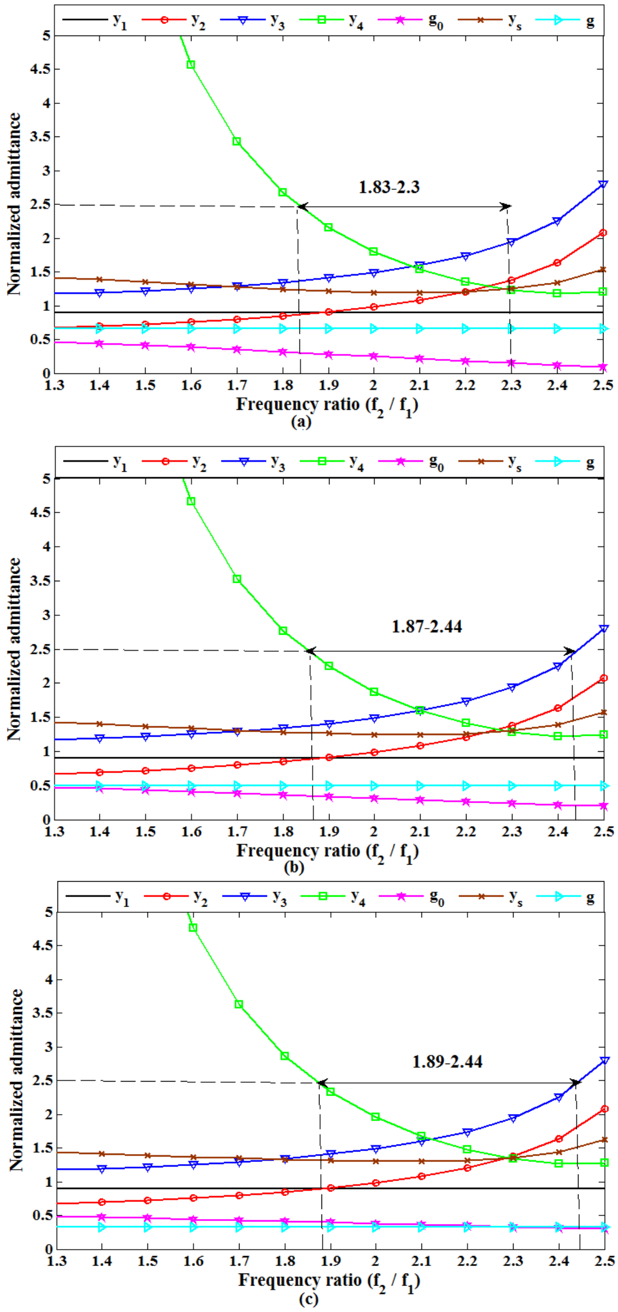

Fig. 3. Variation of normalized admittances versus frequency ratio for y_1=0.9, (a) g = 0.666 (b) g = 0.5 (c) g = 0.333

values for conductance g result in wider bandwidth but lower power handling capability. In fact, there is a trade-off between the bandwidth and power handling capability. In our design, if we choose g = 0.666, then the proposed power divider willhave more power handling capability and narrower bandwidth. In contrast, if we choose g = 0.333, then the proposed power divider will have less power handling capability and wider bandwidth. Therefore, if the bandwidth, the frequency ratio range, and the power handling capability are considered simultaneously, the appropriate solution will be obtained for g = 0.5.

III. SIMULATION AND MEASUREMENT RESULTS

For verification of the theoretical analysis, a dual band power divider using microstrip lines is fabricated at two bands of interest, f_1 = 1 GHz and f_2 = 2 GHz. The circuit is fabricated on dielectric substrate (Rogers RO4003C) with dielectric constant of 3.38, thickness of 0.813 mm and loss tangent of 0.0027. Before the fabrication of power divider, we must take into account the parasitic effects of lumped components and junction discontinuities by performing electromagnetic (EM) simulations. For this purpose, Agilent's ADS EM simulator is utilized. In addition, LineCalc is used for obtaining the line widths. It is worth mentioning that the fine tuning of center frequencies is carried out by slightly adjustment of microstrip lengths for having better response. Furthermore, meandering lines are used in the layout of the proposed structure to reduce the size of the circuit. The measurement results of the fabricated power divider are collected by using a vector network analyzer (ROHDE & SCHORZ ZVA40) over the frequency range from 0.5 to 2.5 GHz.

Making use of (8)-(14), assuming y_1 = 0.9 and g =0.5, and calculating electrical length of $\theta_0 = 60^\circ$ at f_1 = 1 GHz by using (16), we have: y_2 = 0.984, y_3 = 1.487, y_4 = 1.84, g_0 = 0.314, and y_s = 1.226. For convenience, standard value of conductance (or resistor) g_0 is selected (g_0 = 0.3125). However, in practical realization, external high power resistors can be used. A photograph of the fabricated power divider is shown in Fig. 4. The occupied area of the structure is $0.53\lambda_g*0.30\lambda_g$, where λ_g is the guided wavelength at the center frequency ((f_1+f_2)/2). Fig. 5 indicates the simulation with ideal elements, EM simulation, and the measurement results of the fabricated power divider. The measurement results of the first band at 1 GHz are $|S_{11}|$ = -25 dB, $|S_{22}|$ = -24.5 dB, $|S_{23}|$ = -27 dB, and $|S_{21}|$ = -3.36 dB. The second band results at 2 GHz are $|S_{11}|$ = -21.5 dB, $|S_{22}|$ = -24.3 dB, $|S_{23}|$ = -27 dB, and $|S_{21}|$ = -3.98 dB. There are slight discrepancies between simulation and measurement results because of the fabrication tolerances, parasitic effects of resistors, and the loss of connectors. The input return loss bandwidth around 1 GHz and 2 GHz are 110 and 135 MHz, respectively, with the reference of -15 dB. The output return loss bandwidth around 1 GHz and 2 GHz are 120 and 115 MHz, respectively, with the reference of -15 dB. The isolation is better than 25 dB over the wide frequency range from 0.95 GHz to 2.05 GHz or a fractional bandwidth of 73.33%.

In the proposed power divider, using the grounded resistors, generated heat of the circuit is transferred with a direct path to the external ground. Therefore this power divider's power handling capability depends on the grounded resistors. In the Wilkinson power divider, there is only one isolation resistor between its internal straps and there is not any path for heat sinking. So, the proposed power divider has high power handling capability over the Wilkinson divider. It

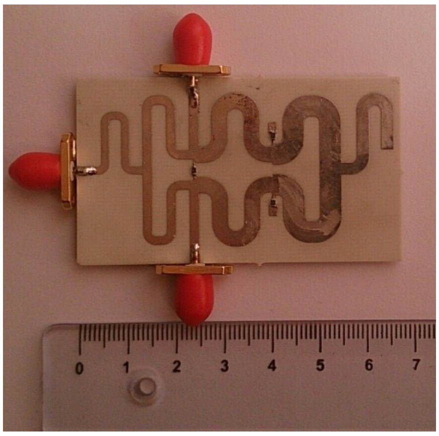

Fig. 4. Photograph of the fabricated dual band power divider.

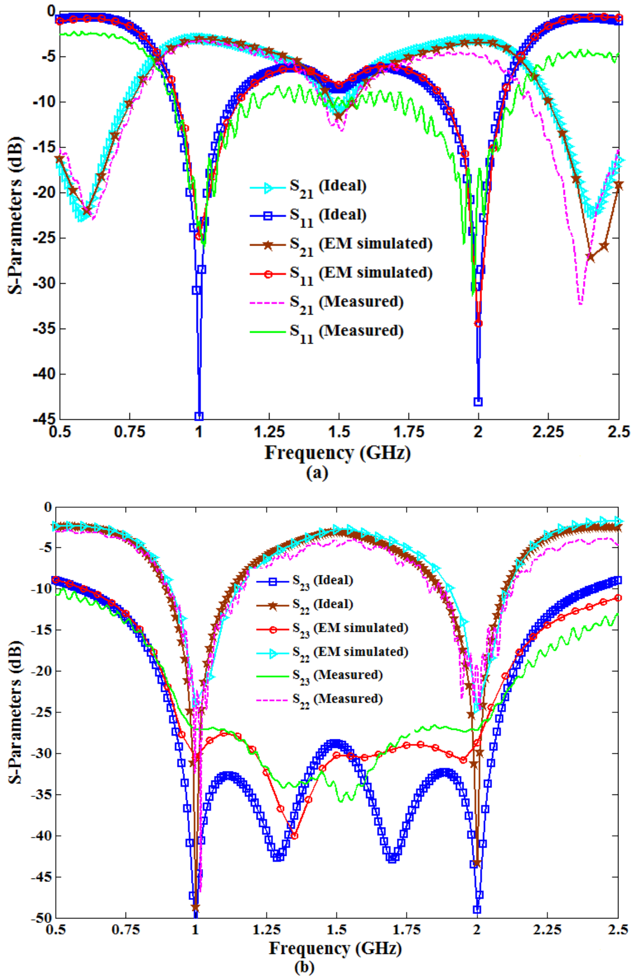

Fig. 5. S-parameters of the dual band power divider. (a) input return loss and insertion loss. (b) output return loss and isolation.

must be noted that according to [12], the meandering lines can jeopardize the power handling capability, because of the creation of parasitic elements, but for compensating this problem, the effect of curved lines in the whole length of each line are exactly calculated and the lengths are slightly changed. So, high power handling capability with this compensation is guaranteed.

IV. CONCLUSIOND

In this paper, a dual band power divider with high power handling capability and isolation bandwidth improvement has been presented. For designing this structure, derivation of accurate equations has been performed. Moreover by considering the admittance constraints and selecting proper values for both the admittance of the extended line at the input and the conductance of grounded resistors, the optimum design is obtained. Furthermore, for verification of the structure, one prototype of the power divider using microstrip lines is fabricated and measurement results show good agreement with the simulation results. Finally, these results show that this power divider with the features of wide isolation bandwidth, acceptable frequency band ratio, and high power handling capability is suitable for dual band applications.

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