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Wideband Four-Way Filtering Power Divider With Sharp Selectivity and High Isolation Using Coshared Multi-Mode Resonators

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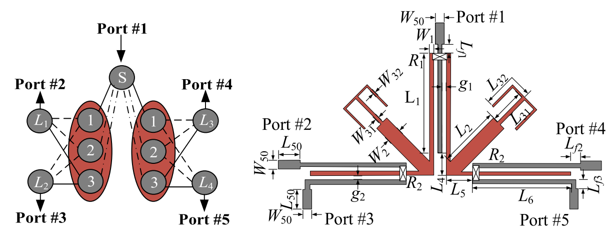
***Abstract*— This letter presents an alternative fusion-design of a wideband four-way filtering power divider (FPD). A coshared coupling topology is proposed by embedding one pair of multi-mode resonators (MMRs) and three isolated resistors, bringing out wide power division band, decent frequency selectivity, and high in-band isolation. In comparison with other reported ones, the presented four-way FPD stands out not only by its favorable operation performance but also flexible topology that can also be easily extended to other multi-way multi-mode FPD designs and competitive size with a reduced number of resonators. A prototype is implemented to validate the design concept and method. Simulated results coincide well with the measured ones, verifying the theoretical design equations.**

***Index Terms*— Coshared coupling topology, filtering power divider (FPD), isolation, multi-way.**

I. Introduction

POWER dividers (PDs) and filters are two essential passive components in modern wireless communication systems. They are usually cascaded together as a system which always results in bulky circuit size and high insertion loss (IL). In order to tackle such problems, the filtering PD (FPD), a multi-functional integrated device, which can not only provide power splitting/combining as a PD but also frequency band selection as a filter simultaneously, has been investigated.

Till now, much effort has been made on two-way FPD designs. However, to the best of the authors knowledge, only a few FPDs with multi-channel applications have been developed [1]–[4]. Among them, four-way FPDs in [1] and [2] are presented based on stub loaded coupled lines, which, however, exhibit either a narrowband (6%) [1] or relatively poor in-band isolation [2]. Another four-way reconfigurable FPD in [3] is developed with several coupled resonators ($N \geq 4$). But, additional input feedlines and multiple resonators result in large circuit size. For N order filtering response, the existing four-way FPDs always require $4 \times N$ resonators. In [4], a four-way FPD is also presented with unsatisfactory isolation based on dielectric resonators. While interesting results have been achieved in the above designs, it is still a challenge to build a wideband four-way FPD with decent operation performance, reduced number of resonators, but flexible design for the multi-way application.



(a) (b)

Fig. 1. Proposed wideband four-way FPD. (a) Designed coupling scheme. (b) Its configuration. (Substrate Rogers RO4003C: $\varepsilon_{\mathrm{re}} = 3.55$, $\tan\delta = 0.0027$, and $h = 0.508$ mm).

The main motivation of this letter is to introduce an alternative wideband four-way microstrip FPD fusion-design based on the designed coshared coupled multi-mode resonator (MMR) topology. In order to achieve wideband filtering response, the proposed four-way FPD is initially constructed by introducing a proper coupling topology between the input–output feed lines and a pair of MMRs. Meanwhile, satisfactory in-band isolation is achieved by reasonably introducing three resistors between the two input lines/sidelines of the resonators. For demonstration, a prototype wideband four-way FPD operating at $f_0 = 2.1$ GHz with a fractional bandwidth (FBW) of 48.0% is designed, fabricated, and tested. Both simulated and measured results are in good agreement validating the design concept.

II. Design and Analysis of the Proposed Wideband Four-Way Filtering Power Divider

The designed topology and its configuration of the proposed wideband four-way FPD are described in Fig. 1. It is basically composed of only a pair of coshared MMRs and three isolation resistors. As the output lines are symmetrically located at both sides of one arm of each MMR, the couplings between

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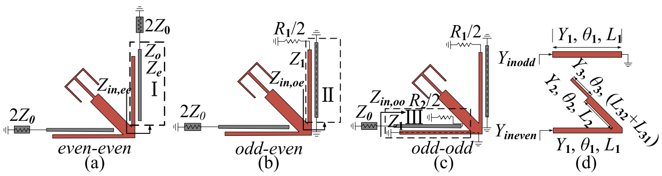

Fig. 2. Equivalent quarter circuit for (a) even–even excitations and (b) even–odd excitations. (c) Quarter circuit for odd–even excitations. (d) Equivalent circuit models of odd-/even-mode bisection of the designed MMR.

output lines and the resonator exhibit the same amplitude and in-phase property. Additionally, the resistor R_1 is added between the input lines, whereas two resistors R_2 are loaded between the output lines to achieve proper isolation. As the structure is symmetric, the even-/odd-mode analysis method is utilized to analyze it. The standard scattering matrix $[S_{\text{std}-\text{FPD}}]$ of the proposed four-way FPD in Fig. 1 can be calculated as [3]

$$S_{11} = S_{11ee} \tag{1a}$$

$$S_{21} = S_{31} = S_{41} = S_{51} = \frac{1}{\sqrt{2}} \cdot \frac{S_{21ee}}{\sqrt{2}} = \frac{S_{21ee}}{2} \tag{1b}$$

$$S_{24} = S_{25} = S_{34} = S_{35} = \frac{(S_{22ee} - S_{22oe})}{4} \tag{1c}$$

$$S_{23} = S_{45} = \frac{(S_{22ee} + S_{22oe} - 2S_{22oo})}{4} \tag{1d}$$

$$S_{22} = S_{33} = S_{44} = S_{55} = \frac{(S_{22ee} + S_{22oe} + 2S_{22oo})}{4}. \tag{1e}$$

As shown in Fig. 2(a), the quarter even–even mode circuit is utilized to specific filtering response of the proposed wideband four-way FPD, which is equivalent to a symmetrical two-port multi-mode filter. Therefore, the equivalent quarter circuit is designed with the same operating frequency and bandwidth as the four-way FPD. Herein, the employed resonator shown as Fig. 2 is adopted to achieve the desired wideband triple-mode response. Since it is symmetrical, the odd-/even-mode analysis method can be utilized to analyze its resonance property [5]. The equivalent circuit models of its half symmetrical bisections are depicted in Fig. 2(d). Accordingly, the input admittances of odd- and even-mode equivalent circuits under odd- and even-mode resonance conditions can be derived as

$$Y_{\text{in,odd}} = -jY_1 \cot\theta_1 = 0, \quad Y_{\text{in,even}} = Y_1 \frac{Y_L + jY_1 \tan\theta_1}{Y_1 + jY_L \tan\theta_1} = 0 \tag{2}$$

where $Y_L = j(Y_2/2) \cdot (Y_2 \tan\theta_2 + Y_3 \tan\theta_3)/(Y_2 - Y_3 \tan\theta_2 \tan\theta_3)$, $Y_{1/2/3}$ stands for the characteristic impedance of a microstrip line with the width W and $\theta_i = 2\pi f L_i \sqrt{\varepsilon_e}/c$, $(i = 1, 2, 3)$. C and ε_e denote the speed of light in the free space and effective dielectric constant, respectively. Three resonant frequencies denoted as f_{e1} and f_{o1},f_{e2} can be successively derived from the above equations with the corresponding odd- and even-mode resonance, which are readily grouped for the desired triple-mode filtering response. Besides, two inherent transmission zeroes (TZs) with the corresponding frequencies (f_{z1} and f_{z2}) can be derived from the input admittance relationship $Y_{\text{in,odd}} = Y_{\text{in,even}}$, i.e., $Y_2/Y_3 = \tan\theta_2 \tan\theta_3$. It can be verified and concluded that the TZs are controlled by θ_2 and θ_3 and drop down as the θ_2/θ_3 increase.

According to the prescribed specifications ($f_0 = 2.1$ GHz, ripple bandwidth of 0.63 GHz, the passband return loss (RL) of 21 dB), the targeted coupling matrix coefficients is synthesized with the analysis method in [6] as: $M_{S1} = M_{L1} = 0.4876$, $M_{S2} = -M_{L2} = 0.8374$, $M_{S3} = M_{L3} = 0.5031$, $M_{11} = 1.4580$, $M_{22} = 0.0289$, $M_{33} = -1.4711$. Based on (3), the required even-mode resonant frequencies (f_{e1},f_{e2}) and odd-mode resonant frequency f_o of the MMR can be derived as $f_{e1} = 1.69$ GHz, $f_o = 2.09$ GHz, and $f_{e2} = 2.61$ GHz. With the help of the derived resonant frequencies, the physical dimension of MMR can be determined

$$f_n = \frac{f_0}{2} \cdot \left[\frac{-\text{BW}}{f_0} M_{ii} + \sqrt{\left(\frac{\text{BW}}{f_0} M_{ii} \right)^2 + 4} \right], \quad i = 1, 2, 3; \; n = e1, o, e2. \tag{3}$$

The I/O couplings to even- and odd-modes can be characterized in terms of external quality factors $Q_{e\_n}(n = e1, o, e2)$, which are found from $Q_{e\_n} = f_0/(BW \cdot M_{si}^2)$ as: $Qe\_e1 = 14.02$, $Qe\_o = 4.75$, $Qe\_e2 = 13.17$. Based on these derived external quality factors, the width ($W_1 = 0.5$ mm) and gap ($g_1 = g_2 = 0.15$ mm) can be determined by extracting the I/O couplings from $Q_{e\_n} = \pi f_i \cdot \tau(f_i)$ through the group delays as in [7].

Till now, according to the above analysis, the two-port filtering circuit, i.e., the quarter equivalent even–even mode circuit has been determined by extracting the calculated coupling matrix coefficients. Therefore, the FPD can be initially built from the obtained equivalent mode circuit.

Once the filtering response is determined by the quarter even–even mode equivalent circuit, the port-to-port isolation and matching are mainly fulfilled by changing the resistances of the isolation resistors R_1 and R_2 of the odd–even- and odd–odd mode equivalent circuit. With Fig. 2(a) and (b) and the (1c), to achieve ideal port-to-port isolation between ports 2 and 4 (ports 3 and 5), the following condition should be satisfied as (4). Accordingly, the initial value of $R_1 = 450\ \Omega$ can be derived from (5). When R_1 is determined, for perfect matching at all output ports and satisfactory isolation between ports 2 and 3 (ports 4 and 5), the input impedance seen from the output ports of the equivalent circuits, as shown in Fig. 2(c) should satisfy the conditions as (6). Furtherly, $R_2 = 220\ \Omega$ can be initially calculated from (7). It is mentioned that $A1(2/3)$, $B1(2/3)$, $C1(2/3)$, and $D1(2/3)$ are the ABCD-matrix elements of the two-port network I/II/III derived as below.

For clearly clarifying the proposed four-way FPD design, an explicit design procedure is given as follows:

Firstly, according to the given specifications, determine the design parameters of the MMR with derived resonant frequencies from (3) in the targeted coupling matrix, which can be synthesized with [6]. Subsequently, determine the input–output coupling space (g) and feedline width (w) by extracting the group delays as described above. Then, build and optimize the four-way FPD. Next, determine initial resistances of the resistors (R_1 and R_2) to achieve favorable port matching and port-port isolation from (4)–(8), as shown at the bottom of the next page. Finally, execute the final optimization of the realized entire circuit layout relying on the electromagnetic (EM) simulator.

III. Implementation and Results

A demonstrator four-way FPD is implemented to verify the design concept. The final optimal layout parameters in Fig. 1 are determined as follows (Units: mm): $L_1 = 24.6$, $L_2 = 17.3$, $L_{31} = 9.7$, $L_{32} = 12.6$, $L_4 = 3$, $L_5 = 3$, $L_6 = 21.6$, $L_{50} = 5$, $L_{f1} = 2.2$, $L_{f2} = 2.2$, $L_{f3} = 2.2$,

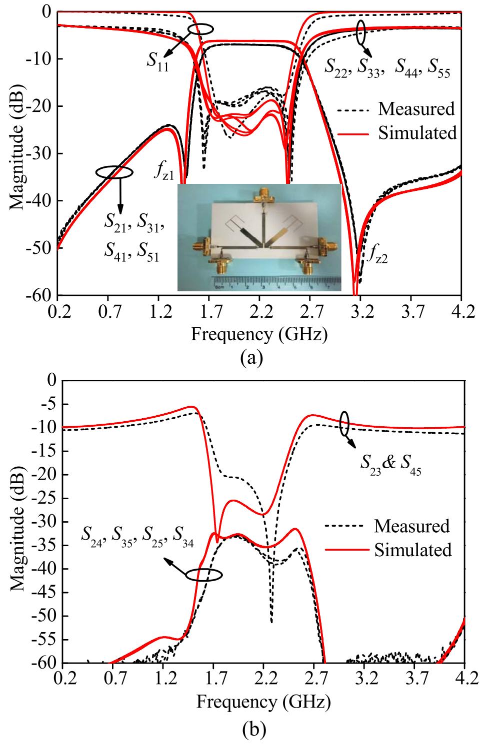

Fig. 3. Measured and simulated results of the wideband FPD. (a) Magnitudes of S_{11}, S_{21}, S_{31}, S_{22}, S_{33}, S_{44}, and S_{55}. (b) Magnitudes of S_{23}, S_{45}, S_{24}, S_{35}, S_{25}, and S_{34}.

$W_{50} = 1.18$, $W_1 = 0.48$, $W_2 = 3$, $W_{31} = 0.3$, $W_{32} = 0.15$, $g_1 = 0.16$, $g_2 = 0.17$, $R_1 = 460\ \Omega$, and $R_2 = 250\ \Omega$. The photograph of the fabricated four-way FPD is displayed in the insert Fig. 3. Its overall size is $0.63\lambda_g \times\ 0.33\lambda_g$, where λ_g is the guided wavelength at the center frequency. The circuit simulation and measurement were carried out by the EM simulator HFSS and the Agilent N5244A network analyzer, respectively. Fig. 3 plots the simulated and measured performances. As observed from the figure, the measured center frequency is 2.1 GHz with the 3-dB FBW of 47.6%. The corresponding measured in-band transmission or IL of each channel of the four-way PD is 7.0 dB, while the RL is better than 16.1 dB, respectively. Because of the resonance of the loaded stubs in the resonator, two TZs (TZ_1 and TZ_2) are introduced at the edges of the passband, which helps a lot to achieve high-frequency selectivity. Moreover, the measured isolation between the output ports is better than 20.2 dB within the passband. Slight discrepancy between the simulation and measurement are mainly due to the fabrication accuracy.

A detailed comparison of the proposal with other reported ones is provided in Table I. It can be observed that our proposed four-way FPD exhibits promising features in terms of compact size and good wideband performance. Actually, it can become more compact by folding the coshared resonators or loading the capacitors with them. It is worth mentioning that the size will exhibit much more competitiveness when the same resonators are applied to other works in Table I.

TABLE I
COMPARISONS WITH OTHER PREVIOUS WORKS

Refs.	FBW (%)	Topology/(Size)	Iso. (dB)	IL (dB)	Output RL (dB)	Input RL (dB)
[1]	6.0	Three resonators/ $(0.73\lambda_g \times 0.54\lambda_g)$	>17.5	7.1	16.7	19.0
[2]	56.5	Four folded parallel coupled lines/$(0.32\lambda_g \times 0.32\lambda_g)$	>13.0	6.8	15.0	10.0
[3]	34.4	Four varactor loaded resonators/ (Not given without varactors)	>19.6	7.1	15.5	14.2
[4]	1.7	Two coupled dielectric resonators$(0.93\lambda_g \times 0.93\lambda_g)$	>15	6.6	Not given	15.0
This work	**47.6**	**Co-shared a pair of multi-mode resonators/ $(0.63\lambda_g \times 0.33\lambda_g)$**	**>20.2**	**7.0**	**17.6**	**16.1**

IV. CONCLUSION

This letter has presented an alternative wideband four-way microstrip FPD design based on the developed coshared coupled MMR topology. The proposed method is flexible and can also be extended to other multi-way multi-mode FPD designs. With the illustrated analysis and design principle, a prototype FPD has demonstrated the design concept, exhibiting favorable operating performance. It is believed that the proposed design is promising for multi-way application in wireless communication systems.

$$Z_{in,ee} = Z_{in,oe} \Rightarrow \frac{2\ A_1 Z_0 + B_1}{2\ C_1 Z_0 + D_1} = \frac{A_2 R_1 + 2\ B_2}{C_2 R_1 + 2\ D_2} \tag{4}$$

$$R_1 = \frac{4\ Z_0 (A_2 D_2 - B_2 C_1) + 2\ B_1 D_2 - 2\ B_2 D_1}{2\ Z_0 (A_2 C_1 - A_1 C_2) + A_2 D_1 - B_1 C_2} \tag{5}$$

$$Z_{in,oo} = \frac{A_3 R_2 + 2\ B_3}{C_3 R_2 + 2\ D_3} = Z_0 \tag{6}$$

$$R_2 = \frac{2 (D_3 Z_0 - B_3)}{(A_3 - C_2 Z_0)} \tag{7}$$

$$\begin{bmatrix} A_1 & B_1 \\ C_1 & D_1 \end{bmatrix} = \begin{bmatrix} \dfrac{(Z_e + Z_o)\cos\theta}{Z_e - Z_o} & j\dfrac{(Z_e - Z_o)^2 - (Z_e + Z_o)^2\cos^2\theta}{2(Z_e - Z_o)\sin\theta} \\ \dfrac{j2\sin\theta}{Z_e - Z_o} & \dfrac{(Z_e + Z_o)\cos\theta}{Z_e - Z_o} \end{bmatrix} \begin{bmatrix} A_2 & B_2 \\ C_2 & D_2 \end{bmatrix} = \begin{bmatrix} \cos\theta & jZ_1\sin\theta \\ j\sin\theta/Z_1 & \cos\theta \end{bmatrix}$$

$$\begin{bmatrix} A_3 & B_3 \\ C_3 & D_3 \end{bmatrix} = \begin{bmatrix} \cos\theta & jZ_1\sin\theta \\ j\sin\theta/Z_1 & \cos\theta \end{bmatrix}. \tag{8}$$

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