Dual-band filtering power divider with independent in-band power split ratio using a single quad-mode dielectric resonator

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Abstract
A design approach is presented based on insights into power division among modes for a dual-band filtering power divider using a single multimode dielectric resonator. The relationship between the energy of the input port under different modes and the corresponding external quality factor ($Q_e$) is analysed. Then, the power distribution among the in-band modes and between output ports can be all expressed by $Q_e$. Therefore, the synthesis of a coupling matrix to achieve independent power distribution in the two bands is simplified. For further verification, a quad-mode dielectric resonator with independently controllable modes is designed. Based on this, dual-band filtering power dividers with equal and unequal power split ratios (1:1 in the lower band and 2:1 in the higher band) are simulated and implemented. The simulation and measurement agree, proving the correctness of the approach.

1 | INTRODUCTION

With the rapid development of wireless communication technologies, multifunctional and multiband systems for processing a large amount of data are being widely investigated. The fusion design of a filtering power divider that has the two functions of frequency selection and power allocation in a single circuit can replace the traditional cascaded design of filter and divider. Because of the advantages of eliminating connection loss and miniaturising the system size, various filtering power dividers have been extensively investigated with equal or unequal power division [1–3] and out-of-phase signals [4] in the past few years. Among them, four typical design methods are included. The first one is to cascade the filter or filtering structure with a power divider [2], which results in a fairly large space. The second way is to replace the $\lambda/4$ transformers of the traditional Wilkinson power divider by coupling sections or filtering structures [3]. The third method is to incorporate the resonator or filtering structure within the quadrature coupler [5]. The last approach is based on the coupled resonator [6].

A dual-band operation is highly desirable in dual-channel concurrent wireless systems (satellite communication, for instance) [7–9], which makes the dual-band filtering power divider attractive. Various designs have been investigated based on a stub-loaded resonator [10], coupled line prototype [11], microstrip-to-slot transition structure [12], single multimode resonator [13], and so on. An even/odd mode and coupling matrix are important analytical methods. By neglecting stubs that have a slight effect on even-mode and odd-mode surface current distributions, equivalent circuits under the corresponding mode can be simplified [14]. The circuit design can also be facilitated by studying the relation between even-/odd-mode $S$-parameters and the generalized mixed-mode scattering matrix [15–17]. For a layout that is symmetric in orthogonal directions, two-level even-/odd-mode analysis is often applied [18]. To achieve an unequal power-dividing ratio, the
impedances of output ports will be asymmetric. Hence, the traditional even-/odd-mode method is unsuitable for an unequal case. Instead, a method based on the transmission line and circuit theories was proposed in Wu et al. [19]. Yet, the additional matching network at the outputs comes at the expense of circuit size. To the best of the authors’ knowledge, the dual-band filtering power divider with an independent power split ratio between two bands has not been reported. In Zhang et al. [18], the synthesis of an equivalent circuit is achieved by a high-order matrix, whereas optimization is not given. In Chi et al. [20], two coupling matrices are respectively applied to synthesise the lower and higher passbands, increasing computations compared with the analysis based on a single matrix. Most of these works use microstrip line and substrate integrated waveguide (SIW) technologies. Because of the relatively low quality factor \( (Q_a) \) of the resonator, the band selectivity is limited. The dielectric resonator (DR) with a high \( Q_a \) is suitable for multimode and narrow-band operations. However, its application in the dual-band filtering power divider is still rare.

A compact dual-band filtering power divider using a single quad-mode DR is proposed. Insight into power division among the modes is introduced into the derivation of the external quality factor \( (Q_e) \) under different modes. The recursive relation of \( Q_e \) exists only within the same passband. Therefore, the power split ratio in each band can be designed independently, indicating that the design procedure is simple. To verify the approach, prototypes with an equal and unequal power split ratio are designed and implemented. Good agreement between the simulated and measured results is obtained.

2 | THEORY OF FILTERING POWER DIVIDER BASED ON INSIGHT OF POWER DIVISION AMONG MODES

To facilitate the description of the power division concept among modes, we first take the second-order filtering power divider as an example. Figure 1a shows the topology for a filtering power divider based on cascaded identical single-mode resonators \( (R_1 \) and \( R_2) \), whereas Figure 1b shows the counterpart based on a single dual-mode resonator. \( S \) and \( L_n \) \( (n = 1 \) or \( 2) \) represent the source and loads, respectively. When the resonator is lossless, we define \( P^0 \) as the average dissipated power of the input port in Figure 1a and \( P_i \) \( (i = 1 \) or \( 2) \) as the average dissipated power of the input port in Figure 1b under each of the two modes of the resonator \( R_n \). Then, the \( Q_e \) of the input under different modes can be expressed as

\[
Q^e = \frac{Q_i}{W_S/P^0} \quad (1a)
\]

\[
Q^e_{ds} = \frac{Q_i}{W_S/P_i} \quad (1b)
\]

where \( Q^e \) is the \( Q_e \) of Port 1 in Figure 1a whereas \( Q^e_{ds} \) \( (i = 1 \) or \( 2) \) represents \( Q_e \) of Port 1 under the two modes in Figure 1b.

**FIGURE 1** Topologies for single-band filtering power divider. (a) Structure based on cascaded single-mode resonator. (b) Structure based on a single dual-mode resonator

\( W_S \) is energy stored in the resonant circuit. \( \omega_0 \) is the resonant frequency of \( R_1 \) (or \( R_2) \), whereas \( \omega_i \) \( (i = 1 \) or \( 2) \) represents the resonant frequencies of the dual modes in \( R_3 \). For narrow-band applications, there should be

\[
\omega_0 \approx \omega_i \quad (2)
\]

Then, by combining Equations (1) and (2), we obtain

\[
P_i/P^0 = Q^e_i/Q^e_{ds} \quad (3)
\]

Because of the equivalence of the two modes of \( R_3 \), energy through them should be equally divided, which means that \( P_i \) is half of \( P^0 \). Substituting the relationship into Equation (3), there is

\[
Q^e_i/Q^e_{ds} = 1/2 \quad (4)
\]

According to the relation between \( Q_e \) and the parameters in the coupling matrix, there should be

\[
M^e_{SI} = \frac{1}{\sqrt{FBW \cdot Q^e_e}} \quad (5a)
\]

\[
M^e_{SI} = \frac{1}{\sqrt{FBW \cdot Q^e_{ds}}} \quad (5b)
\]

where FBW represents the ripple fractional bandwidth. Coupling coefficients \( M^e_{SI} \) and \( M^e_{SI} \) correspond to the coupling schemes in Figure 1a,b, respectively.
By combining Equations (4) and (5), we obtain

\[ M_{S1}^b = M_{S2}^b = M_{S1}^e / \sqrt{2} \]  \hspace{1cm} (6)

Therefore, the topological transformation from Figure 1a to Figure 1b can be represented by the rotation transformation of the coupling matrix [21] as well as the analysis of the energy distribution between the modes.

The power division between the output ports, which is determined by the ratio of their \( Q_i \) [22], does not affect the power allocation between modes. For the common case of an equal split ratio, \( Q_{cd,2}^0 \) and \( Q_{cd,2}^2 \) should be twice that of \( Q_{e,5}^0 \) and \( Q_{e,5}^2 \), respectively. Then, we obtain \( Q_{cd,2}^0 = Q_{cd,2}^2 = 4 \ Q_e^0 \), which shows a two-level power distribution.

Figure 2 shows the topology for a dual-band filtering power divider. The applied resonator can be viewed as two parallel dual-mode resonators (black dotted box) shown in Figure 1b or a single quad-mode resonator (red dashed box). In particular, the result in Equation (4) is based on the condition of Equation (2), which means the power division between the modes exists only within a single band. Accordingly, the topology in Figure 2 can be regarded as two independent filtering power dividers shown in Figure 1b, which simplifies the design procedure. The lower band is formed by the previous two modes (Modes 1 and 2), whereas the higher band is constructed by the latter two modes (Modes 3 and 4).

Referring to the design approach for a single-band filtering power divider based on a dual-mode resonator, the design procedure of a second-order filtering power divider with dual-band Chebyshev response can be summarised as follows.

**Step 1** According to the specification, calculate \( Q_{e,1} \) and \( Q_{e,3} \) required by the two lowpass prototypes corresponding to the responses of the two bands, respectively.

**Step 2** (First-level power distribution): Replace the cascaded single-mode resonators in the lowpass prototypes with two parallel dual-mode resonators or single quad-mode resonator. Calculate \( Q_{e,j}^j (j = 1, 2, 3, 4) \) of the input under each mode according to Equation (7):

\[ Q_{e,j}^j = 2Q_{e,1} \text{ and } Q_{e,j}^{j+2} = 2Q_{e,2} \]  \hspace{1cm} (7)

**Step 3** (Second-level power distribution): Construct the relationship between \( Q_{e,j}^j \) of the output ports under each mode and \( Q_{e,j}^j \) according to Equation (8) while keeping the \( Q_e \) (\( Q_{e,1} \) and \( Q_{e,2} \)) of the total output required for each passband constant.

\[ Q_{e,j}^j = \frac{\alpha_1 + \alpha_2}{\alpha_n} Q_{e,5} \]  \hspace{1cm} (8a)

\[ Q_{e,j}^{j+2} = \frac{\beta_1 + \beta_2}{\beta_n} Q_{e,5} \]  \hspace{1cm} (8b)

where \( \alpha_1 : \alpha_2 \) and \( \beta_1 : \beta_2 \) represent the power split ratios of the lower and higher bands, respectively.

**Step 4** By combining Equations (7) and (8), the value of \( Q_{e,j}^j \) can easily be calculated:

\[ Q_{e,j}^j = \frac{\alpha_1 + \alpha_2}{\alpha_n} \cdot 2Q_{e,1} \]  \hspace{1cm} (9a)

\[ Q_{e,j}^{j+2} = \frac{\beta_1 + \beta_2}{\beta_n} \cdot 2Q_{e,2} \]  \hspace{1cm} (9b)

**Step 5** Adjust the coupling coefficients to meet the requirements of the \( Q_e \) (\( Q_{e,1} \) and \( Q_{e,2} \)) for the band construction.

In the case of \( \alpha_1 = \alpha_2 \text{ and } \beta_1 = \beta_2 \), we make \( Q_{e,5} = 2Q_{e}^0 \) (\( Q_{e,1} = Q_{e,3} = Q_{e}^0 \)) to construct two identical passbands. Then, there should be \( Q_{e,j}^j = 4Q_{e}^0 (j = 1, 2, 3, 4) \) after two-level power division. To simplify these analyses of the power distribution among modes, only the main couplings are considered in Figure 2. When the cross-couplings between the source and the loads (S-L, coupling) are considered, a coupling matrix to produce a dual-band filtering power divider with equal split ratio is
Self-resonance values $M_{jj}$ on the primary diagonal of the coupling matrix that represent the frequency offsets can be obtained by

$$M_{jj} = \frac{f_0^2 - f_j^2}{\Delta f \cdot f_j}$$

where $f_0$ and $\Delta f$ are the centre frequency and ripple bandwidth of a passband, $f_j$ is the resonant frequency of the corresponding mode. For the lower band, the resonant frequencies of Modes 1 and 2 are located at either side of the centre frequency. Therefore, there must be $M_{11} > 0$ and $M_{22} < 0$. Similarly, for the higher band, the signs of $M_{33}$ and $M_{44}$ are also opposite.

Figure 3a shows the corresponding theoretical response with four transmission zeros (TZ). The two TZs between the passbands are generated by the in-band modes. Specifically, the second TZ is generated by Modes 1 and 2. In the two transmission paths, the signs of $M_{11n}$ and $M_{22n}$ $(n = 1, 2)$ are opposite. There should be a pair of out-of-phase signals with equal amplitude, which cancel each other to generate the TZ. Similarly, the third TZ corresponds to Modes 3 and 4. The first and fourth TZs are generated by $S-L_n$ coupling. Figure 3b shows the theoretical responses of the dual-band filtering power divider with different values of $S-L_n$ couplings. The distance between the second and third TZs increases with an increase in the absolute value of $M_{51n}$.

In the case of $\alpha_1 : \alpha_2 = 1 : 2$ and $\beta_1 : \beta_2 = 2 : 1$, there should be $Q_{L_1}^e = Q_{L_2}^e = 6Q_2^e$ and $Q_{L_1}^{e^2} = Q_{L_2}^{e^2} = 3Q_2^e$ after two-level power division. The coupling matrix can be expressed as whose theoretical response is shown in Figure 3c. The generation mechanism of TZs is the same as the case with an equal split ratio.

These analyses reveal a two-level power distribution in the dual-band filtering power divider based on a multimode resonator. Power distribution among in-band modes and between output ports can be all expressed by $Q_n$, which makes it easy to achieve the independent power split ratio in the two bands.

### 3 | Dual-Band Filtering Power Divider Based on a Quad-Mode Dielectric Resonator

For validation and illustration, a dual-band filtering power divider is designed based on the topology in Figure 2. Figure 4 shows the electric-field distribution of the applied quad-mode DR with a silver layer inserted in the middle. The DR has two pairs of degenerate modes (TE$_{112}$ and TE$_{111}$ mode pairs). Each mode pair resonates at the same frequency. Among the four modes, TE$_{112}$ and TE$_{312}$ modes are caused by the silver layer (electric wall), which is an
important design approach to transform a dual-mode DR into a quad-mode DR. Figure 5 shows the resonant frequencies of TE₁₁₂ and TE₁₁₁ mode pairs with different side lengths ($b_1$) of the silver layer. As the value of $b_1$ increases, the resonant frequency of TE₁₁₂ mode pair rapidly decreases while the resonant frequency of TE₁₁₁ mode pair remains unchanged. The required centre frequencies of the two bands can be selected accordingly. Figure 6 shows the configuration of the dual-band filtering power divider using the quad-mode DR. The corner-cut is a kind of perturbation that is often used to separate the frequencies of the two orthogonal modes in a square resonator. After corner cutting, the TE₁₁₂ mode pair is transformed to Modes 1 and 2, whereas the TE₁₁₁ mode pair corresponds to Modes 3 and 4. Similar to the case in Figure 3, the lower band is formed by Modes 1 and 2, whereas the higher band is constructed by Modes 3 and 4. The permittivity ($\varepsilon_r$) and loss tangent (tan$\delta$) of the employed DR are 38 and 1.5 $\times$ $10^{-4}$. In this design, the centre frequencies of the two bands are 2.18 and 2.68 GHz with the 0.08-dB-ripple FBW of 1% and 0.8%, respectively. Figure 7 shows the simulated transmission responses excited by weak coupling with different corner cuts of the DR ($c_1$) and the silver layer ($c_2$). The bandwidths of the two passbands can both be increased by increasing $c_1$, whereas the bandwidth of the lower band can be adjusted by $c_2$ alone. Therefore, the required bandwidths of the two bands can be designed independently.

The $Q_e$ of the four modes are mainly controlled by the length and height of the feeding probe. Figure 8 shows the $Q_e$ versus probe length $d$ with different $d_0$ in the original and corner-cut square DR cavities. Because the $Q_e$ of the modes resonating at the same frequency cannot be extracted respectively, the obtained $Q_{e1}$ is the $Q_e$ under the combined effects of the TE₁₁₂ mode pair, whereas $Q_{e2}$ corresponds to the TE₁₁₁ mode pair. $Q_{e1}$ and $Q_{e2}$ are independent of each other. Meanwhile, corner-cut $c_2$ of the silver plane has a great influence only on the coupling coefficient of the lower band. The independence of the coupling coefficient makes it easy to meet the needs of $Q_e$ to construct the two bands. In addition, the equal/unequal power split ratio in the two bands can easily be achieved. Because of the approximation in Equation (2) and the limitation of extraction precision, the extracted $Q_e$ of the two modes for building the same band are not the same, as shown in Figure 8b. According to the analyses in Section 2, $Q_{e1}^{1+2}$ and $Q_{e3}^{1+2}$ should be twice $Q_{e1}$ and $Q_{e2}$, respectively. Figure 8 shows that the offset of $Q_e$ in this region is acceptable. To reduce the calculation for the matrix synthesis, the curves in Figure 8a are used to estimate the extracted results in Figure 8b. According to the design specification, there should be $Q_{e1} = 79.9$ and $Q_{e2} = 131.9$, which corresponds to $Q_{e1}^{1+2} = 2Q_{e1} = 159.8$ and $Q_{e3}^{1+2} = 2Q_{e2} = 263.8$. To meet the needs of the input for $Q_e$, $d_{01}$ = 6.2 mm and $d_1$ = 14.3 mm are fixed.

For the case of equal power division in the two bands, the appropriate values of $d_{02}$ ($d_0$) and $d_3$ ($d_4$) are respectively 5.5 and 14.3 mm to meet the requirement of $2Q_{e1}$ and $2Q_{e2}$. In this case, $Q_{e1}^{4+5}$ and $Q_{e2}^{4+5}$ should be $4Q_{e1}$ and $4Q_{e2}$, respectively, according to Equation (9). After adjusting the coupling coefficients to match the value of $Q_e$, the coupling matrix can be expressed as:

**FIGURE 3** Theoretical responses of dual-band filtering power dividers with different power split ratios. (a) Equal split ratio in two bands. (b) Equal split ratio with different values of S-Ln couplings. (c) Split ratio of 1:2 in the lower band and 2:1 in the higher band.
The theoretical response and simulated result match each other in the passbands, as shown in Figure 9. The dimensions of the dual-band filtering power divider can be determined as: \( A = 32 \text{ mm}, \ H = 27 \text{ mm}, \ b_1 = 12 \text{ mm}, \ b_2 = 16 \text{ mm}, \ c_1 = 3.2 \text{ mm}, \ c_2 = 0.5 \text{ mm}, \ b_1 = 4 \text{ mm}, \ b_2 = 15 \text{ mm}, \ d_{01} = 6.2 \text{ mm}, \ d_{02} = d_{03} = 5.5 \text{ mm}, \) and \( d_1 = d_2 = d_3 = 14.3 \text{ mm}. \) Figure 10 shows the simulated responses of the prototype with different values of \( b_1. \) As expected, the distance between the two passbands decreases with the decrease of \( b_1. \) When the silver layer is small, the couplings between the ports and \( \text{TE}_{112} \) mode pair will become too weak to form a passband. In the case of \( b_1 = 0, \) the electric wall disappears, which makes the \( \text{TE}_{112} \) mode pair no longer exist. Accordingly, the two bands cannot be joined into a wide single one with a fourth-order filtering response by reducing the value of \( b_1. \)

To obtain the power split ratio of 1:1 in the lower band and 2:1 in the higher band, \( Q_{d1}^2 \) and \( Q_{d2}^2 \) are both equal to \( 4Q_{e1} \) whereas \( Q_{d1}^2 = 3Q_{e1} \) and \( Q_{d2}^2 = 6Q_{e1} \) according to Equation (9). The probe lengths of Ports 2 and 3 need to be updated accordingly, whereas the other parameters remain unchanged. The theoretical and simulated responses are demonstrated in Figure 11. Figure 12 is a photograph of the implemented circuit. Figure 13 illustrates the measured and simulated S-parameters of the dual-band filtering power divider with equal and unequal split ratios in the two bands. Figure 14 shows the measured and simulated amplitude imbalance and phase difference of the two prototypes. In the case of equal power distribution, the amplitude imbalance (\( S_{21} - S_{31} \)) and phase differences (\( \angle S_{21} - \angle S_{31} \)) are respectively within 0.45 dB and 180° ± 5.3° in the two bands, showing good consistency. In the case of the unequal power distribution (split ratio of 1:1 in the lower band and 2:1 in the higher band), the amplitude imbalance is within 0.51 dB in the lower band. According to the ideal division ratio of 2:1 in the higher band, \( S_{21} \) should be \(-1.76 \text{ dB}, \) and \( S_{31} \) should be \(-4.77 \text{ dB} \) theoretically. The measured in-band insertion loss of \( S_{21} \) and \( S_{31} \) are respectively...
better than \((-1.76 - 0.55)\) dB and \((-4.77 - 0.21)\) dB in the higher band. The phase differences in the two bands are both within \(180^\circ \pm 4.8^\circ\). The slight differences between the measured and simulated responses can be mainly attributed to...
the tolerance of fabrication and assembly. Although it is the general situation that isolation is not ideal for power dividers with no resistance, these designs can be also applied in wireless transmitters. Comparisons with some previous filtering power dividers are provided in Table 1. Compared with the SIW designs, the proposed divider has the lowest insertion loss and narrowest bandwidth. Although the insertion loss in this work is slightly higher than the power dividers based on waveguide technology in Mohammed and Wang [25] and Lin et al. [26] owing to the loss of the Sub Miniature version A connector, the volume is much smaller. Moreover, the independent power division in the two bands can be realized in this work. Although a quad-mode DR is applied in Xu et al. [27], the design is suitable only for a single-band operation. Compared with the method in Chi et al. [20], there is no need to synthesise the two passbands separately in the proposed design approach. Owing to the insight of power division among modes, the recursive relation of coupling matrix parameters can be quickly obtained.

4 | CONCLUSION

The insight of power division among modes is introduced to design a dual-band filtering power divider based on a single multimode resonator. The required split ratio can be achieved by two-level power distribution. The first level occurs among modes whereas the second level occurs between output ports. To verify the proposed approach, a dual-band filtering power divider using a single quad-mode DR is constructed. Because of independence among the modes, the power split ratio in the two bands can be controlled independently. The simulated and measured results show good accordance, indicating the value of the approach for compact and narrower-band designs.
Figure 13: Measured and simulated results of quad-mode dual-band filtering power divider: (a) equal split ratio; (b) unequal split ratio of 1:1 in the lower band and 2:1 in the higher band.

Figure 14: Measured and simulated amplitude imbalance and phase difference of the prototypes. (a) Equal split ratio; (b) Unequal split ratio of 1:1 in the lower band and 2:1 in the higher band.

Table 1: Comparison of proposed power divider with other previous designs

| Reference | Technology | Frequency (GHz) | Insertion loss (dB) | 3 dB Fractional bandwidth (%) | Electrical size | Power-division ratio | Number of modes | Analysis method |
|-----------|------------|-----------------|---------------------|-------------------------------|----------------|----------------------|-----------------|-----------------|
| [20]      | SIW        | 28/39           | 1.9/2.2             | 3.6/3.1                       | $2.01\lambda_0 \times 1.45\lambda_0$ | Equal              | 4               | 2 coupling matrices |
| [23]      | SIW        | 11.8            | 1                   | 5.9                           | $1.0\lambda_0 \times 0.59\lambda_0$ | Equal/unequal     | 2               | Even-/odd-mode   |
| [24]      | SIW        | 5.5/8.3         | 0.9/1.5             | 14.5/9.6                      | $1.50\lambda_0 \times 0.62\lambda_0$ | Equal              | 2               | Circuit theory   |
| [25]      | Metal cavity | 10             | 0.7                 | 5.0                           | $3.9\lambda_0 \times 2.7\lambda_0 \times \frac{0.34\lambda_0}{0.34\lambda_0}$ | Equal              | 1               | 1 coupling matrix |
| [26]      | Metal cavity | 2.58/2.76      | 0.2/0.3             | 0.8/1.2                       | $1.38\lambda_0 \times 1.20\lambda_0 \times 0.69\lambda_0$ | Equal              | 3               | Mixed-mode $S$-parameters |
| [27]      | DR         | 3.52            | 0.5                 | 1.7                           | $0.39\lambda_0 \times 0.39\lambda_0 \times 0.26\lambda_0$ | Equal              | 4               | 1 coupling matrix |
| This work | DR         | 2.18/2.68       | 0.7/0.8             | 1.6/1.4                       | $0.23\lambda_0 \times 0.23\lambda_0 \times 0.26\lambda_0$ | Equal/unequal      | 4               | 1 coupling matrix |

Abbreviations: $\lambda_0$, wavelength in free space at the operating frequency or at $f_1$ of the dual-band operation; SIW, substrate integrated waveguide.
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