Dual-Band Power Divider Based on Semiloop Stepped-Impedance Resonators

Stephane Avrillon, *Student Member, IEEE*, Ismael Pele, Anne Chousseaud, and Serge Toutain

***Abstract*—This paper presents a power-divider architecture whose characteristics are adjustable independently at two frequencies. The circuits uses a single semiloop stepped-impedance resonator filter coupled to the input and output lines. A specific methodology has been developed to design such a circuit.**

***Index Terms*—Dual-band matching, power divider, stepped-impedance resonator (SIR).**

I. INTRODUCTION

THE different standards across the frequency spectrum in the mobile communication arena have led researchers to develop more and more versatile components. These components have to address the requirements relative to the multifrequency and bandwidths.

One component of interest is the power divider [1]–[4]. This component can be used in special microwave functions, such as a splitter in antenna beam-forming systems [3], [4]. Little or no interest has been focused in this area of research.

It is believed that it will generate a set of new investigations into multiband beam-forming techniques.

A requirement to achieve multiband beam forming is a perfect knowledge of the divider frequency response in and out of the bands of interest. This involves a precise understanding of the filtering and matching effects of the structure used as a power divider. This requirement plus the need for versatility to weigh the array's elements with any kind of amplitude and phase within a reduced size has lead us to investigate a new type of structure based on the stepped-impedance resonator (SIR).

We have mainly focused on a dual-band investigation, though the possibility to extrapolate to over two bands is obvious and under study. A first step has been a feasibility study of such a power divider using single-pole resonating filters. Though the filtering function has been limited to one resonator, this gives a good idea and understanding of the behavior of more complex structures using several resonators.

In order to have a compact structure, a semiloop SIR filter has been synthesized by modifying classic single SIR filters [5]. The power-divider structure is then derived from combination of these filters. The frequency response of the filter directly shapes the power divider's response.

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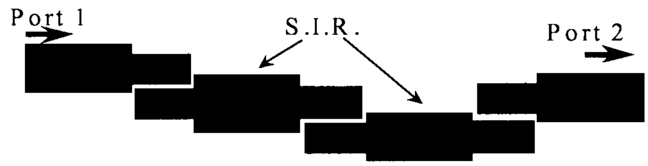

Fig. 1. Microstrip SIR filter example.

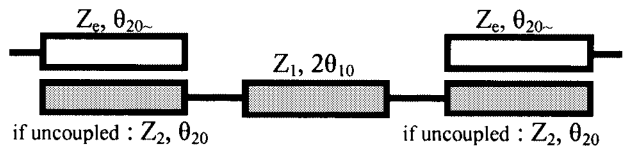

Fig. 2. Classic single SIR with input/output coupled lines.

II. DUAL-BAND MATCHING BASED ON SIR

Resonator filters are commonly used in microwave filtering functions. SIR filters present the interesting property of having several geometric degrees of freedom to set several resonant frequency values and to control the out-of-band frequency response.

The input impedance is matched and the output transmitted signal is maximized at resonant frequencies. These frequencies and corresponding bandwidths are not linked ones to the others and can be set independently by choosing appropriate geometric parameters.

A. Description of a Single SIR Filter

SIR filters are built with several line sections of nonconstant widths coupled together, as shown in Fig. 1, in microstrip technology. SIR filter syntheses are usually based on a low-pass prototype filter model with impedance or admittance inverters and susceptance slope parameters [6], [8]. Usually, thanks to this method, length and width values are set for the whole structure with a chosen frequency response function (Chebyshev, Butterworth, etc.). In this paper, the aim is different: the filter is used to obtain two matching bands without particular shape constraints. Thus, in a first step, this study is limited to the case of a single SIR

B. Frequency Behavior of Single SIR Filter

Each of the resonator's geometry and dimensions can be determined easily and, in general, the corresponding frequencies are numerically determined. In the case of a symmetric SIR (Fig. 2), an analytic solution exists. The symmetric resonator that we used in this paper is constituted by three line sections (Z_2, θ_{20}), $(Z_1, 2\theta_{10})$, and (Z_2, θ_{20}). Z_i is the characteristic

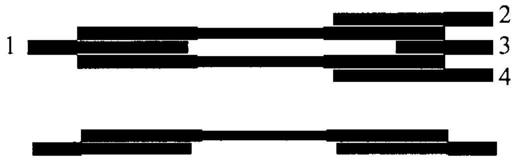

Fig. 3. Three-output port dual-band power divider and its corresponding single SIR filter.

impedance of the line i and Z_e is the even-mode impedance of coupled lines [Z_e and odd-mode impedance Z_o are linked with coupling coefficient k by (1)]. Z_c is the reference impedance and is set to 50 Ω. θ_{i0} is the electrical length of an L_i physical length line at the first resonant frequency f_0 and is linked to L_i, relative permittivity ε_r, light velocity c, and f_0 using (2) as follows:

$$\frac{Z_e}{Z_c} = \frac{Z_c}{Z_o} = \sqrt{\frac{1+k}{1-k}} \tag{1}$$

$$\theta_{i0} = \frac{2\pi L_i}{c\sqrt{\varepsilon_{\text{eff}}}} f_0. \tag{2}$$

To set up two frequencies (f_0 and f_1), we use the two resonant conditions (3) and (4), which depend only on SIR parameters [5]. We have introduced the parameter $K = Z_2/Z_1$ to simplify expressions. Actually, (Z_2, θ_{20}) line sections are coupled lines, and the electrical length of these coupled lines is $\theta_{20\sim}$. Though the approximation $\theta_{20\sim} = \theta_{20}$ introduces a shift in resonant frequency values, it really simplifies the problem as follows:

$$K = \tan\theta_{10} \cdot \tan\theta_{20} \tag{3}$$

$$K \cdot \tan\left(\frac{f_1}{f_0}\theta_{10}\right) + \tan\left(\frac{f_1}{f_0}\theta_{20}\right) = 0. \tag{4}$$

Bandwidth at f_0 can be controlled by spacing between coupling lines. However, contrary to resonant frequency determination, bandwidth is difficult to estimate without heavy calculation. In fact, the coupling coefficient k and coupling lengths influence the overall bandwidth. The smaller the spacing and longer the length, the wider the bandwidth.

C. Application to Dual-Band Power Divider

The transformation of a two-port filter into a power divider can be realized by coupling several output lines. To design a three-port power divider, two additional lines must be coupled on the output. To keep the structure symmetrical, a power divider has been designed, as shown in Fig. 3.

To start the basic design of the power divider, the values of spacings and widths obtained for the single SIR filter are kept. However, due to the configuration described in Fig. 3, we can see that ports 2 and 4 receive energy mainly from one resonator as port 3 receives from two resonators.

Port 3 coupling length must be reduced to balance the three outputs.

An optimization and simulation of this power divider has been performed with a Duroid substrate ($\varepsilon_r = 2.6$, loss tangent 0.001). Results are given in Fig. 4, showing an input matched at two frequencies. The return loss mainly reaches 30 and 25 dB at 0.9 and 1.96 GHz, respectively. The account for losses in the structure simulated gives fairly good transmission coefficient

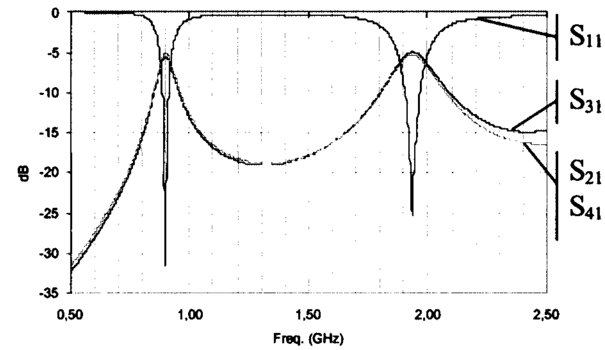

Fig. 4. Three-port dual-band power divider S-parameters.

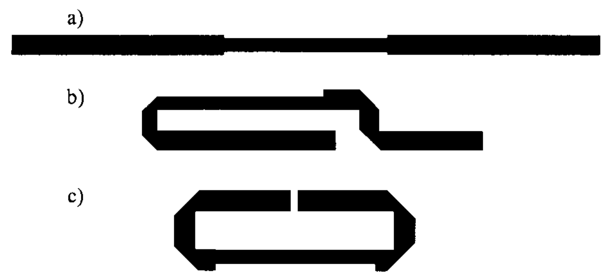

Fig. 5. Three types of SIRs. (a) Classic SIR. (b) Semiloop SIR. (c) Loop SIR.

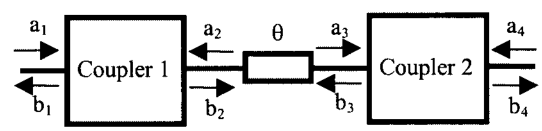

Fig. 6. Model for a single-resonator filter.

values. Theoretically, without any loss, the three-way power divider should present a 4.7-dB transmission loss compared to the 5 dB simulated in the actual structure.

This first approach of the dual-band power divider permits to validate the method and the kind of resonator chosen. The next step of the study is to have a compact structure keeping the same properties. A solution is to bend the SIR resonator with one constraint: to keep the same output configuration with three coupling ports. The solution is what we have termed a semiloop SIR.

III. Semiloop SIR Filters

A semiloop SIR must have the same geometry and dimensions as a classic SIR to keep the same resonant frequencies. Classic, semiloop, and loop resonators are presented in Fig. 5. It has to be noticed that the loop SIR has not been chosen because it is more difficult to make coupling with several lines at one of its extremity.

To have the same behavior as a classic SIR structure, it is necessary to keep resonant frequencies and bandwidths. For this last point, we have developed a model that permits comparison of the two kinds of structure bandwidths.

A. Global Model for Single-Resonator Filter

The frequency response of a single-resonator filter can be characterized using the model presented in Fig. 6. It consists of two couplers, of which S-matrices are given using (5a) and (5b), and a transmission line (electrical length θ), which can be simplified, in a first approach, in a phase shifter (6) as follows:

$$\text{coupler 1}: \begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} s_{11} & s_{12} \\ s_{21} & s_{22} \end{bmatrix} \cdot \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \tag{5a}$$

$$\text{coupler 2}: \begin{bmatrix} b_3 \\ b_4 \end{bmatrix} = \begin{bmatrix} s_{33} & s_{34} \\ s_{43} & s_{44} \end{bmatrix} \cdot \begin{bmatrix} a_3 \\ a_4 \end{bmatrix} \tag{5b}$$

$$\text{line}: \begin{bmatrix} a_2 \\ a_3 \end{bmatrix} = \begin{bmatrix} 0 & e^{-j\theta} \\ e^{-j\theta} & 0 \end{bmatrix} \cdot \begin{bmatrix} b_2 \\ b_3 \end{bmatrix}. \tag{6}$$

With this model, it is possible to find the frequency response S_{41} of the single-resonator filter. The result is given in (7) and (8) as follows:

$$S_{41} = \left.\frac{b_4}{a_1}\right|_{a_4=0} = \frac{s_{43}s_{21}e^{-j\theta}}{1 - s_{22}s_{33}e^{-j2\theta}} \tag{7}$$

and with $s_{ii} = |s_{ii}| \cdot e^{-j\varphi_{ii}}$

$$S_{41} = \frac{s_{43}s_{21}e^{-j\theta}}{1 - |s_{22}| \cdot |s_{33}| e^{-j\cdot(2\theta + \varphi_{22} + \varphi_{33})}}. \tag{8}$$

S_{41} has a bandpass response with a bandwidth generally quite small ($B < 10\%$), thus, all the S-parameters moduli of couplers can be considered as a constant in this range. By using new parameters (9), a simplified frequency response (11) is obtained as follows:

$$A = |s_{43}s_{21}| \quad b = |s_{22}| \cdot |s_{33}|$$

and

$$2\pi\frac{f}{f_0} = 2\theta + \varphi_{22} + \varphi_{33} \tag{9}$$

$$S_{41}(f) = \frac{A}{1 - b \cdot e^{-j2\pi(f/f_0)}}. \tag{10}$$

The linearization of the exponential term can be realized by using $df = f - f_0$

$$\begin{aligned} S_{41}(f) &= \frac{A}{1 - b \cdot e^{-j2\pi(df/f_0)}} \\ &= \frac{A}{1 - b \cdot \left(1 - j2\pi\frac{df}{f_0}\right)} \\ &= \frac{A}{1 - b + j2\pi b\frac{df}{f_0}} \end{aligned}$$

The new response can be identified using (11) as follows:

$$S'_{41}(f) = S_{21}^{\max} \cdot \frac{1}{1 + j\frac{df}{f_c}}. \tag{11}$$

It is a low-pass filter with well-known parameters presented in (12) and (13) as follows:

$$\text{Maximum magnitude: } s_{41}^{\max} = \frac{a}{1-b} \tag{12}$$

$$\text{Cutoff frequency: } f_c = \frac{1-b}{2\pi b} f_0. \tag{13}$$

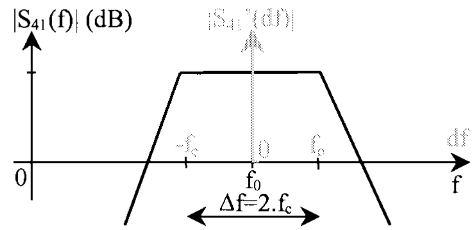

Fig. 7. Bandpass frequency response.

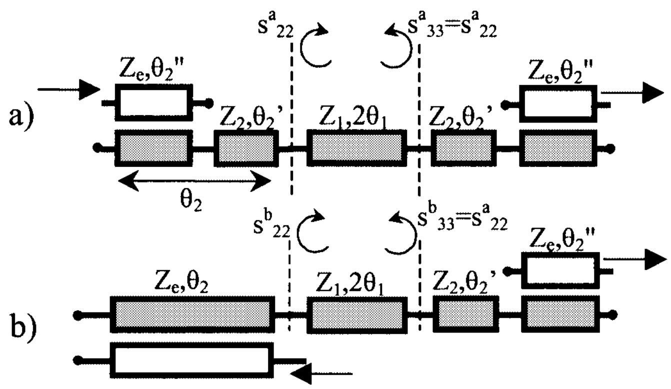

Fig. 8. Filter configurations. (a) Classic SIR filter. (b) Semiloop filter.

Assuming $\mathrm{df} \ll f_0$ and with the help of Fig. 7 showing the low-pass behavior around f_0, filtering parameters can be determined as follows in (14) and (15):

$$\text{Center frequency: } f_0$$

$$\text{Magnitude at } f_0: \frac{A}{1-b} \tag{14}$$

$$\text{Relative bandwidth at } f_0: 2\frac{f_c}{f_0} = \frac{1-b}{\pi b}. \tag{15}$$

Corresponding single-resonator parameters (16) and (17) are given using (9)

$$\text{Filter magnitude at } f_0: S_{41\,\max} = \frac{|s_{43}| \cdot |s_{21}|}{1 - |s_{22}| \cdot |s_{33}|} \tag{16}$$

$$\text{Filter bandwidth at } f_0: \frac{1 - |s_{22}| \cdot |s_{33}|}{\pi \cdot |s_{22}| \cdot |s_{33}|}. \tag{17}$$

Single-resonator filter bandwidth could be entirely defined by $|s_{22}| \cdot |s_{33}|$.

B. Bandwidth Equivalence Between Classic SIR Filters and Semiloop Ones

Fig. 8 presents two configurations of filters, i.e., the classic SIR filter [see Fig. 8(a)] and the semiloop filter [see Fig. 8(b)].

With the analogy between Figs. 6 and 8 and using (17), the two filter bandwidths, we have calculated B^a (18) for the classic SIR filter of Fig. 8(a) and B^b (19) for the semiloop one of Fig. 8(b). The equality between these two formulas gives a simple condition (20) on the return-loss parameters as follows:

$$B^a = \frac{1 - |s_{22}^a|^2}{\pi \cdot |s_{22}^a|^2} \tag{18}$$

$$B^b = \frac{1 - |s_{22}^a| \cdot |s_{22}^b|}{\pi \cdot |s_{22}^a| \cdot |s_{22}^b|} \tag{19}$$

$$|s_{22}^a| = |s_{22}^b|. \tag{20}$$

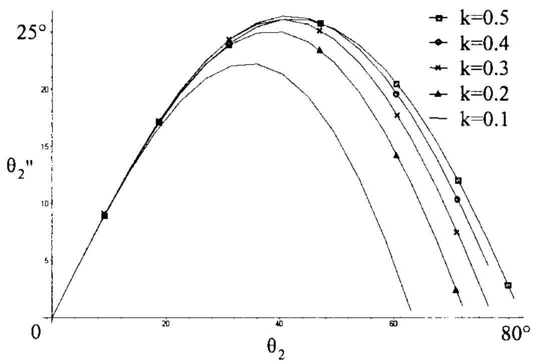

Fig. 9. Graph representing relation between the two configuration coupling lengths ($\epsilon_r = 2.6$).

The two parameters s_{22}^a and s_{22}^b have been calculated using the topology of Fig. 8 and using an accurate numerical model of coupling areas [7]. Equation (20) between $|s_{22}^a|$ and $|s_{22}^b|$ then gives conditions on θ_2 and θ_2'' according to the coupling coefficient k. An example of the result of this intricate solving is presented in Fig. 9 for a substrate of relative permittivity equal to 2.6. k is linked to even and odd impedances by (1).

Respecting these conditions, semiloop filters are more compact than classic SIR filters without changing resonant frequencies and bandwidths.

C. Semiloop Filter Design Process

The chosen filter structure is represented in Fig. 8(b). In this section, the different steps of a semiloop filter design are detailed.

Step 1) The first step is the determination of the input coupling electrical length θ_{20} at f_0. The larger θ_{20}, the smaller the bandwidth. At the same time, input coupled lines create a null in the frequency response at a frequency corresponding with $2\theta_2 = 180°$. This null must be kept distant from any resonant frequencies.

Step 2) The next step is how to obtain the coupling coefficient k. The larger k, the larger the bandwidth. However, k is limited by technological considerations, like minimum spacing between coupled lines. When k is set, Z_e and Z_o are fixed. The impedance Z_2 of a single line with the same width than coupled lines can also be calculated using a numerical model.

Step 3) With Fig. 9, θ_{20}'' is deduced with k and θ_{20}. Then, $\theta_{20}' = \theta_{20} - \theta_{20}''$. This chart must be plotted for the relative permittivity that corresponds to the realization.

Step 4) Finally, the resonator constituted by two Z_2 lines and one Z_1 line imposes the resonant frequencies. If two frequencies are set, θ_{10} and Z_1 are given using (3) and (4).

IV. Application to Dual-Band Power Divider

In this section, an example of design on Duroid ($\varepsilon_r = 2.6$) is given. The power divider works at 0.9 and 1.94 GHz.

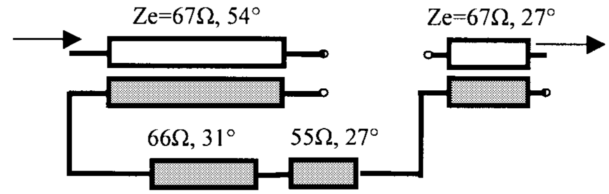

Fig. 10. Semiloop dual-band 0.9–1.94-GHz filter.

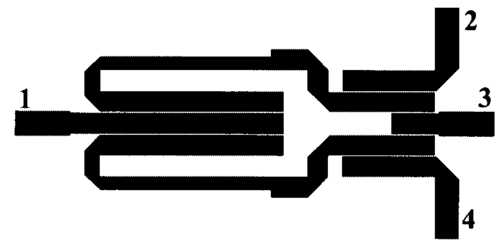

Fig. 11. Layout of the three-port semiloop-resonator power divider.

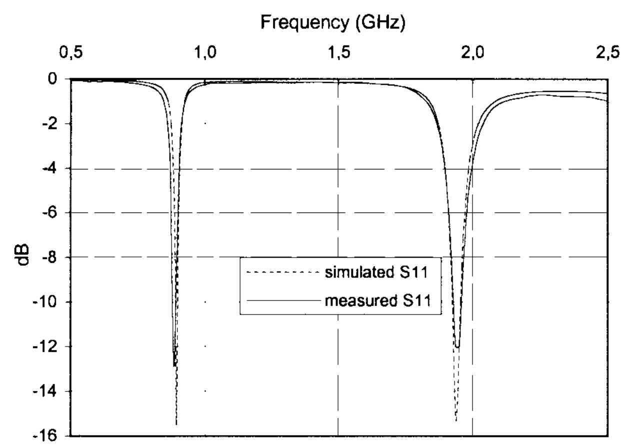

Fig. 12. Simulated and measured return loss.

A. Dual-Band Power-Divider Design

1) The zero introduced by the input coupler is set to 1.5 GHz leading to $\theta_{20} = 54°$.
2) In order to have the desired bandwidth ($B = 4.5\%$), k is set to 0.3 (0.2 mm between coupled lines), which is squaring with $Z_e = 67\ \Omega$. Thus, the coupled linewidth is 3.7 mm, which fits with a 55-Ω linewidth ($Z_2 = 55\ \Omega$).
3) Fig. 9 gives $\theta_{20}'' = 27°$ and, thus, $\theta_{20}' = 27°$.
4) Finally, θ_{20}, f_0, and f_1 values are introduced in (3) and (4) and give $K = 0.84$, $\theta_{10} = 31°$, and $Z_1 = Z_2/K = 66\ \Omega$.

The corresponding semiloop filter is given in Fig. 10.

The method used in Section II-C to transform this filter into a three-port power divider is applied and optimized with the use of circuit simulation software *Genesys* (Eagleware, Norcross, GA). It gives the final structure represented in Fig. 11.

B. Dual-Band Power-Divider Simulated and Measured Performances

The power divider has been simulated and realized. As seen in Figs. 12 and 13, measurements and simulation are in perfect agreement.

The device behavior presents two resonant frequencies (0.88 and 1.94 GHz) with expected bandwidths (4.5% at 0.88 GHz

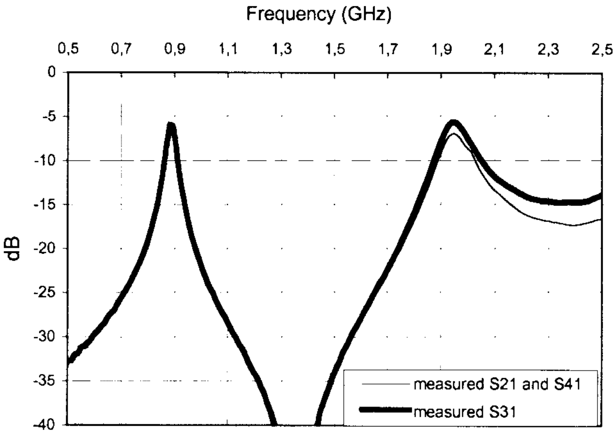

Fig. 13. Measured response.

and 7% at 1.94 GHz). The return loss at the input is approximately 13 dB at 0.9 and 1.94 GHz. Power levels on the three outputs are almost identical (Fig. 12) with a shift of 5 dB with respect to the theoretical 4.7 dB (third part of the input power in each output). However, at 1.94 GHz, outputs are not as well balanced ($S_{21} = S_{41} = -5.6$ dB and $S_{31} = -6.9$ dB).

The output return loss of ports 2 and 4 is 16 dB (in agreement with a classic SIR filter output), but for port 3, which has been added in the structure to have a three-way divider, the return loss is only 3 dB.

Isolations between outputs 2–4 are almost equivalent and equal to -10 dB.

V. Conclusion

This paper has presented a study of the feasibility of designing a new kind of dual-band power divider. We have developed a method using a classic SIR and then, in order to make the divider more compact, semiloop equivalent ones.

A dual-band power divider (0.9 and 1.94 GHz) has been designed. The design methodology, based on the couplings between microstrip lines and semiloop SIR resonators, allows the attainment of a compact structure. It allows also controlling each frequency, bandwidth, and output power level independently.

The next step of this study could be the development of more complex structures based on couplings between several resonators. More attention will be then be paid to the frequency response shape, particularly the filter order and introduction of transmission nulls, but compactness will constitute an inconvenience that may be solved by choosing an appropriate SIR ordering.

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