



An ultra-compact diplexer based on simple microstrip coupled lines for GSM and wideband wireless applications

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Abstract

In this work, a microstrip diplexer with $0.004 \lambda g^2$ (10.1×23 mm) overall size is designed, analyzed and fabricated. The proposed diplexer has the smallest size compared to the previously reported microstrip diplexers. The proposed diplexer has a simple and novel structure, wide flat channels and very low $|S_{11}|_{dB}$. An innovative microstrip structure based on thin coupled lines is used to design of the proposed diplexer. Since in a simple structure the possibility of manufacturing errors reduces, having a simple structure is one of its advantages. Another advantage of this diplexer is two low $|S_{11}|_{dB}$ of 0.17 and 0.14 dB at the lower and upper channels. The operational frequencies of our diplexer are tuned to work at 0.9 GHz and 1.8 GHz for GSM application. It has the privileges of very compact size, simple structure, small $|S_{11}|_{dB}$, two wide fractional bandwidths (FBWs) of 21 and 24.3% and acceptable $|S_{11}|_{dB}$ and isolation. Due to its two wide FBWs, the presented diplexer is suitable for broadband communication systems. We have fabricated and measured the introduced diplexer to verify the design methodology and simulation results. The obtained results of the diplexer measurement confirm the simulation.

Keywords Ultra-compact · LC circuit · Microstrip diplexer · GSM · Fractional bandwidth

1 Introduction

Due to being planar, low weight, low cost and easy to fabricate, compact microstrip devices, *i.e.* couplers, filters, diplexers and multiplexers, have been very much needed

by modern high-speed wireless communication systems [1–5]. To segregate signals and multiplex them, compact microstrip diplexers are the key devices. Recently, achieving a compact high-performance microstrip diplexer has been a challenge for designers. Accordingly, some microstrip structures are introduced to access dual-band bandpass-bandpass diplexers [4–23]. A common disadvantage of the designed diplexers in [1–4] and [6–20] is occupying large implementation areas. On the other hand, some communication links with high data rates are forced to use a wide fractional bandwidth (FBW). Nevertheless, the proposed diplexers in [1–4] and [6–21] could not increase FBWs significantly which is another common problem of them. Different types of coupled structures are used to design the microstrip diplexers in [4, 6–10]. Coupled meandrous cells in [4], stub-loaded coupled lines in [6], mixed electromagnetic coupling in [7], four coupled loops with step impedance cells in [8], coupled spiral cells in [9] and coupled E-shape structures in [10] are used. In [4], the second channel is narrow and cannot be used practically. Also, we can see the problem of low selectivity at the second channel of the designed diplexer in [6]. The

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reported structures in [7–10] have undesired $|S_{11}|_{dB}$. Using interdigital cells, the problem of high losses is solved in [11], but it could not improve the sharpness of its passbands edges. In [12], a diplexer with large $|S_{11}|_{dB}$ is proposed using ring resonator. To improve the bandwidths, a microstrip diplexer is presented in [13]. To design this diplexer, coupled step impedance cells are used which increase its FBWs up to 16.8%. Using half- and quarter-wavelength resonators, a novel diplexer with two narrow channels and high $|S_{11}|_{dB}$ is presented in [15]. Dual-mode ring filters in [16], coupled triangular microstrip structure in [17] and open rings in [18] are used for designing microstrip diplexers. In [19], engraved patches and triangle structures are used. Using computational intelligence, a microstrip multiplexer is obtained in [20]. In [20], a compact microstrip diplexer is presented for broadband wireless applications. In [21], a small microstrip diplexer is designed using coupling meandrous open-loops. In [24], a compact microstrip diplexer consists of a BPF and a low-pass filter (LPF) is introduced. In [25], a compact microstrip lowpass-bandpass diplexer with high isolation is presented. A high-selectivity microstrip diplexer is reported in [26] to work at 1.496 GHz and 2.759 GHz. A microstrip diplexer with close channels is designed in [27] for L-band applications. A high-isolation microstrip diplexer with high $|S_{11}|_{dB}$ is presented in [29] for 5G applications. The proposed diplexers in [26–29] have large sizes.

In this work, thin microstrip lines are coupled for designing a compact wideband microstrip diplexer. Our proposed diplexer is designed for GSM applications i.e. 0.9 GHz and 1.8 GHz. Duo to the use of only thin microstrip lines, the proposed diplexer has a completely new structure. But in the previous structures, in addition to thin lines patch or stepped impedance cells are used to create some essential capacitors. Meanwhile, by using a perfect mathematical design method this new structure can reduce the overall size and improve the performance simultaneously. The presented diplexer is well miniaturized so that it occupies only an overall size of $0.004 \lambda_g^2$, which is the smallest area compared with the previously reported diplexers. Moreover, it has desired $|S_{11}|_{dB}$ at the upper and lower channels. The other parameters of our diplexer such as frequency selectivity, $|S_{11}|_{dB}$ and isolation are suitable. Meanwhile, several TZs exist in its stopband. The paper structure is presented as follows: Firstly, an innovative resonator is presented and analyzed. Secondly, two

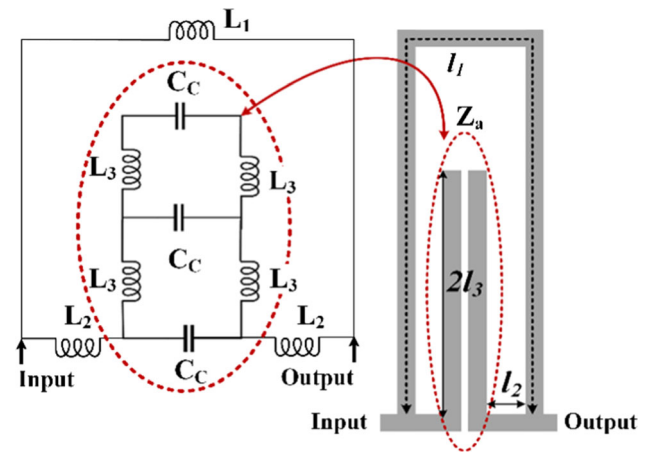


Fig. 1 Presented resonator and the approximated coupled line LC circuit

microstrip BPFs are designed using the analyzed resonator. Thirdly, a microstrip diplexer is created by integrating the BPFs. Fourthly, our diplexer is simulated using ADS software. Finally, the fabricated diplexer is measured. Also, our proposed design is compared with the other previously reported diplexers to show its features.

2 Designing method

Our proposed resonator consists of coupled thin lines. It includes a pair of coupled lines, which are connected to a thin rectangular ring as shown in Fig. 1. The LC circuit model of each thin line is an inductor while the coupling between lines creates several small capacitors. Due to having the microstrip cells with the capacitance and inductance features, our proposed structure can create a passband. An equivalent LC circuit (approximated) of the coupled lines with an input impedance of Z_a is also shown in Fig. 1. The capacitor C_c is created due to the coupling effect, while the inductor L_3 is the equivalent of the thin line with the physical length l_3 . In the exact model, we have to divide the line with the physical length l_3 into a large number of small lines, where there are several coupling capacitors. However, using the approximated model gives sufficient information about the resonator action and makes the calculations easier. The inductors L_1 and L_2 are the equivalent of the lines with l_1 and l_2 lengths, respectively. The capacitors created at the corners have a little impact at frequency $f < 10$ GHz. Hence, we can eliminate them and

show the line with the physical length l_1 only by an equivalent inductor.

Our resonator is symmetric; hence we can perform the even and odd modes analysis to get information about the proposed resonator. Accordingly, at ω (which is an angular frequency), the impedance from the input port to the output port (Z) is obtained as follows:

$$Z = \frac{j\omega L_1(2j\omega L_2 + Z_a)}{2j\omega L_2 + Z_a + j\omega L_1} \tag{1}$$

where

$$Z_a = \frac{\left(\frac{\left(\frac{1}{j\omega C_C} + 2j\omega L_3 \right) \times \frac{1}{j\omega C_C}}{\left(\frac{1}{j\omega C_C} + 2j\omega L_3 \right) + \frac{1}{j\omega C_C}} + 2j\omega L_3 \right) \times \frac{1}{j\omega C_C}}{\left(\frac{\left(\frac{1}{j\omega C_C} + 2j\omega L_3 \right) \times \frac{1}{j\omega C_C}}{\left(\frac{1}{j\omega C_C} + 2j\omega L_3 \right) + \frac{1}{j\omega C_C}} + 2j\omega L_3 \right) + \frac{1}{j\omega C_C}} \Rightarrow \tag{2}$$

$$Z_a = -j \frac{1 - 6\omega^2 L_3 C_C + 4\omega^4 L_3^2 C_C^2}{\omega C_C (3 - 8\omega^2 L_3 C_C + 4\omega^4 L_3^2 C_C^2)}$$

In the above equations, the inductances L_1 and L_2 are the equivalents of the transition lines with the physical lengths l_1 and l_2 respectively (depicted in Fig. 1). The capacitor produced by the coupling structure is in fF, hence it is very small. The inductors and angular frequency are in nH and GHz, respectively. Accordingly, we can rewrite (2) by applying this approximation as follows:

$$\begin{cases} 1 \rangle - 6\omega^2 L_3 C_C + 4\omega^4 L_3^2 C_C^2 \\ 3 \rangle - 8\omega^2 L_3 C_C + 4\omega^4 L_3^2 C_C^2 \end{cases} \Rightarrow Z_a \approx \frac{1}{j\omega C_C} \tag{3}$$

Now, by substituting Z_a in Z of (1), the impedance Z can be calculated as:

$$Z = j\omega \frac{L_1(1 - 2\omega^2 L_2 C_C)}{1 - \omega^2 C_C(L_1 + L_2)} \tag{4}$$

For the odd mode, the numerator of (4) must be zero. On the other hand, the denominator of (4) should be zero for the even mode analysis. Hence, the angular resonance frequencies of the odd and even modes are obtained as follows:

Odd mode:

$$1 - 2\omega_0^2 L_2 C_C = 0 \Rightarrow \omega_o = \sqrt{\frac{1}{2C_C L_2}} \tag{5}$$

Even mode:

$$1 - \omega_e^2 C_C(L_1 + L_2) \Rightarrow \omega_e = \frac{1}{\sqrt{C_C(L_1 + L_2)}} \rightarrow \omega_e \approx \frac{1}{\sqrt{C_C L_1}} \tag{6}$$

In Eq. (6), ω_e and ω_o are the even mode and the odd mode resonance frequencies respectively. As presented in Fig. 1, the stub with the physical length l_2 is very small so we can assume that $L_2 \ll L_1$. Therefore, ω_o will be $(2C_C L_2)^{-0.5}$. Hence, we can say that $\omega_e \ll \omega_o$ where ω_e is the main and the ω_o is a harmonic which should be attenuated. On the other hand, we can tune the values of C_C and L_1 . According to (5) and (6) and the above discussion, ω_e is strongly affected by the stub with the physical length l_1 and the value of C_C , as well. By decreasing the space between the coupled lines we can increase C_C . Therefore, we can reduce the length l_1 which saves the size significantly. Having some information about how the resonator works helps us to reduce dimensions and adjust the operational frequency simultaneously. Using the analyzed resonator, two bandpass filters (BPF1 and BPF2) are designed, as shown in Fig. 2(a), (b). Each BPF includes two pairs of the proposed resonator which are coupled. The dimensions of both BPFs are in millimeter. The resonance frequency of each BPF is tuned by adjusting the most important physical dimensions in accordance with the formula of the even mode analysis, as presented in (6). As mentioned in the resonator analysis, the gaps between coupled lines are very small which leads to increase the values of coupling capacitors. Therefore, we could decrease the overall dimensions in accordance with Eq. (6).

The S-parameters of BPF1 and BPF2 are shown in Fig. 2(c), (d), respectively. BPF1 works at $f_{o1} = 930$ MHz with two low return and $|S_{11}|_{dB}$ of 19 and 0.09 dB, respectively. The fractional bandwidth (FBW) of this filter is 22.6%. It could attenuate the harmonics up to 2.6 GHz which is $3.1 \times f_{o1}$. Therefore, it suppressed the 1st, 2nd and 3rd harmonics. As shown in Fig. 2(d), BPF2 resonates at $f_{o2} = 1.78$ GHz with 0.14 dB $|S_{11}|_{dB}$ and 16 dB return loss. The -3dB-passband of BPF2 is from 1.5 GHz to 1.94 GHz with 25.7% FBW. Since BPF1 and BPF2 work at 930 MHz and 1.78 GHz, they are suitable for GSM applications. As shown in Fig. 2(d), it attenuates the undesired harmonics up to 5.4 GHz ($3 \times f_{o2}$) so that they are attenuated from the first up to 3rd harmonics. Moreover, both filters have reasonable frequency selectivity. Figure 3(a), (b) show the surface current density distributions of BPF1 and BPF2, respectively. From these figures it is clear that the lengths T and D have significant effects on the

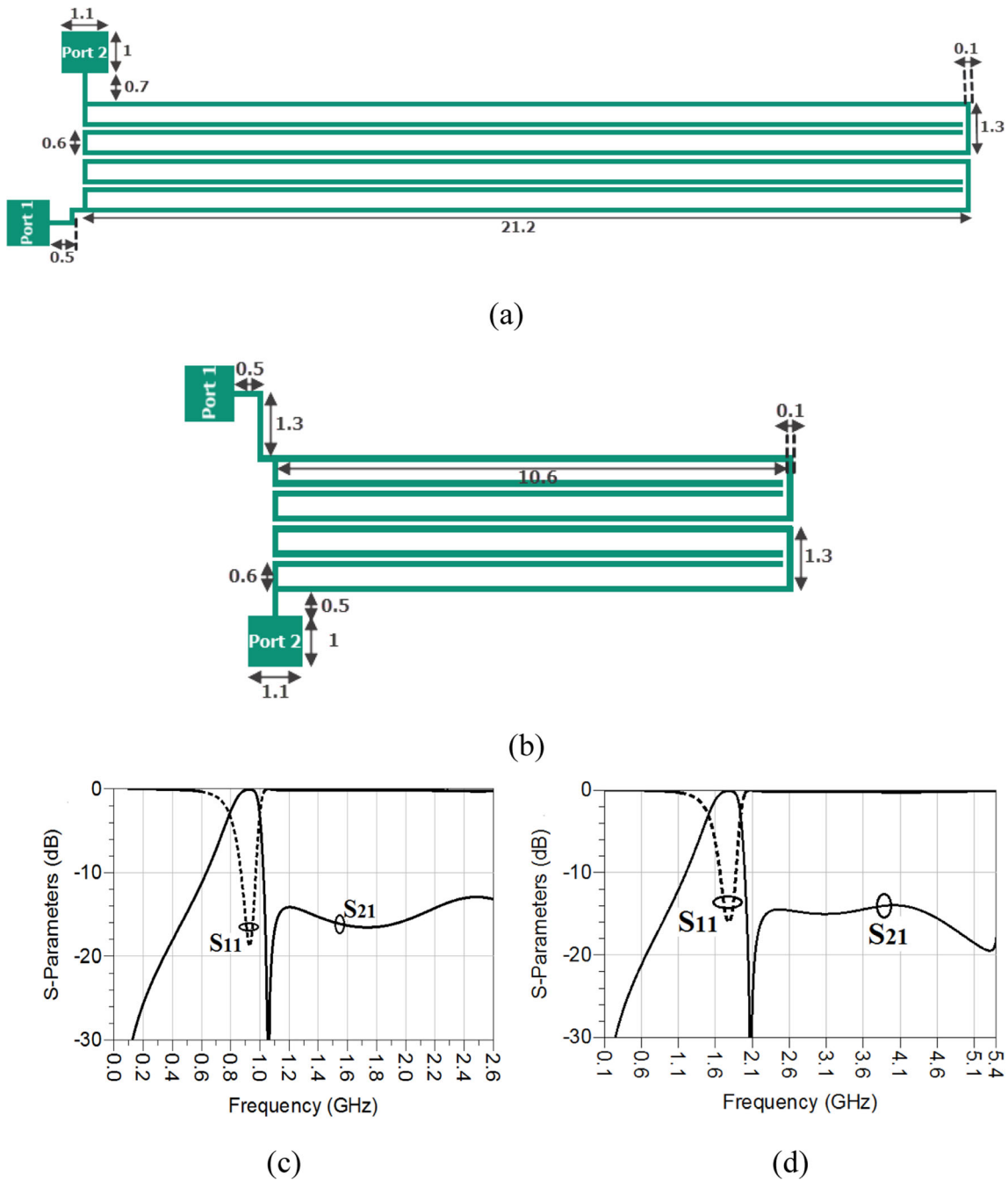


Fig. 2 **a** BPF1 layout (unit:mm), **b** BPF2 layout (unit:mm), **c** S-parameters of BPF1, and **d** S-parameters of BPF2

S-parameters of BPF1 and BPF2, respectively. In Fig. 3(c), (d), the S-parameters of BPF1 and BPF2 are shown as functions of the lengths T and D . As presented in these figures, the BPF1 and BPF2 frequency responses are moved to the lower frequencies by raising the lengths T and D .

Using the proposed BPF1 and BPF2 a novel microstrip diplexer with a compact size is proposed. Figure 4 shows the layout configuration of our diplexer. The presented diplexer is composed of the introduced BPFs with the dimensions exactly equal to the designed BPFs. The BPF with larger dimensions creates the lower channel. The

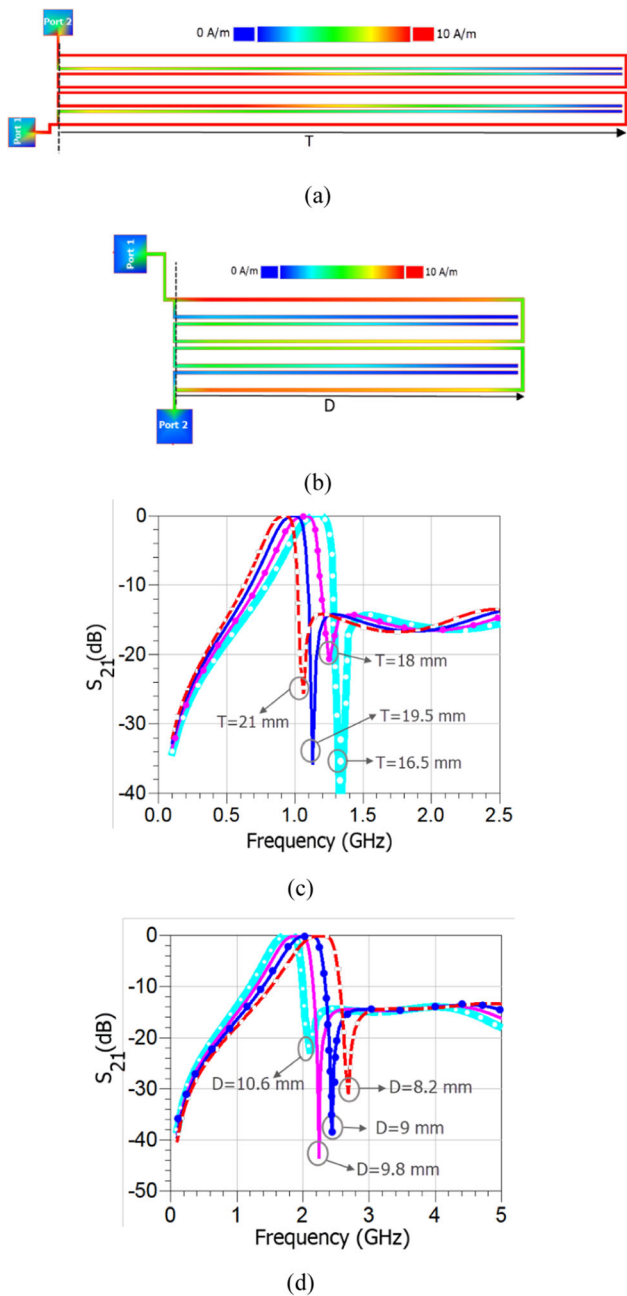


Fig. 3 **a** Distribution of current density for BPF1, **b** distribution of current density for BPF2, **c** S_{21} of BPF1 as a function of T , **d** S_{21} of BPF2 as a function of $h D$

BPFs are connected directly without an extra matching circuit which leads to save the size significantly.

3 Results and discussion

The introduced diplexer is designed and simulated using ADS software. It is fabricated on a Rogers RT/duroid5880 substrate with the following specifications:

$$\epsilon_r = 2.22$$

$$h = 0.7874 \text{ mm}$$

$$\text{Loss tangent} = 0.0009$$

For the measurement, we used an Agilent N5230A network analyzer. Our diplexer is an ultra-compact structure with an overall dimension of $0.044 \lambda_g \times 0.1 \lambda_g$ ($10.1 \times 23 \text{ mm}$), where λ_g is the guided wavelength calculated at the lower resonance frequency. Figure 5(a) presents the measured and simulated transfer parameters. The effective dielectric constant is $\epsilon_{re} = 1.9$ and $\lambda_g = 0.004 \text{ mm}$ at the first resonance frequency. Our presented structure resonates at 0.9 and 1.8 GHz with two desired $|S_{11}|_{dB}$ of 0.17 and 0.14 dB, respectively. Another privilege of the presented diplexer is its wide FBW i.e. 21% and 24.3% for the first and second channels, respectively. S_{11} and S_{23} are plotted in Fig. 5(b). From Fig. 5(b) it can be seen that S_{11} at the first and second resonance frequencies are 16.1 dB and 17 dB, respectively, while S_{23} is better than 20.5 dB. Figure 5(b) shows the fabricated microstrip diplexer. There are two transmission zeros (TZs) at the stopband with the levels better than -40 dB. These TZs are located near 1 GHz and 2.1 GHz, which are created in the stopband by S_{21} and S_{31} respectively. To prove the superiority of our diplexer, we compared it with the previous diplexers. In Table 1, the comparison results are shown where f_{o1} and f_{o2} are the lower and upper resonance frequencies, respectively and FBW_1 and FBW_2 are the FBWs at the lower and upper passbands, respectively. As shown in Table 1, the previous work in [14] have the minimum $|S_{11}|_{dB}$. Also, the proposed diplexer in [22] has the widest FBW%. Compared with the previous diplexers, the main advantage of our diplexer is

Fig. 4 Presented diplexer

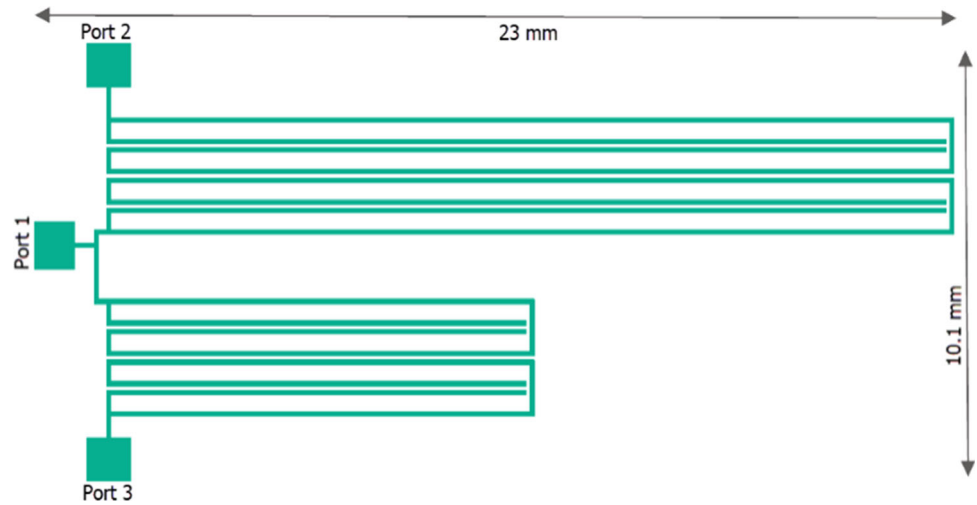
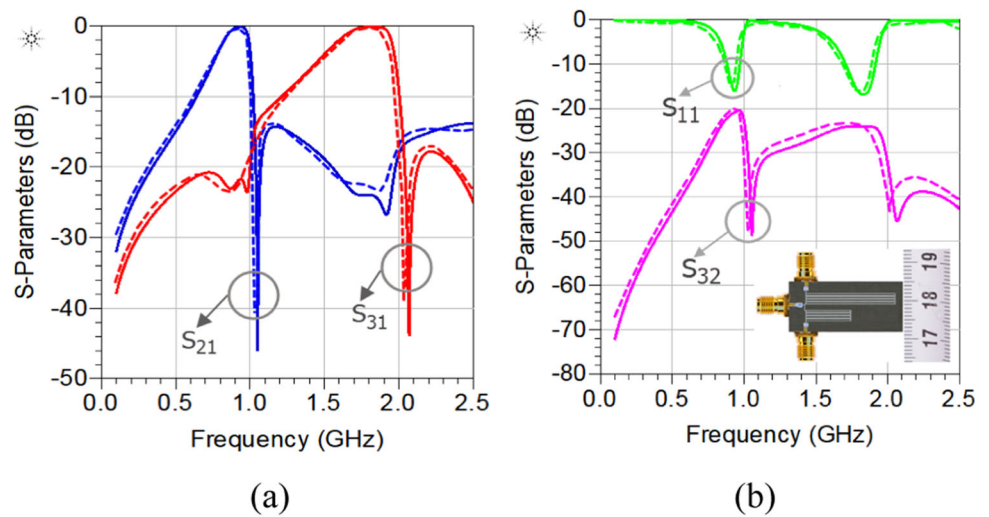


Fig. 5 Simulated and measured results: **a** transfer parameters, **b** S_{11} and S_{32} and the fabricated diplexer



its compact size. Our diplexer is 13.5 times smaller than the proposed diplexer in [14]. Moreover, the introduced diplexer in [22] has the minimum size. But our diplexer has lower $|S_{11}|_{dB}$ and it is two times smaller than the designed diplexer in [22]. From Table 1, it is clear that our designed structure is the most compact diplexer. Meanwhile, our diplexer has the widest FBWs, comes after the diplexer in [22], and has low $|S_{11}|_{dB}$ at the first and second passbands.

4 Conclusion

We designed and analyzed a very compact microstrip diplexer based on a novel and simple structure. The designing method started by analyzing a proposed resonator, which gave us general data about resonator action. These data helped us to optimize the proposed diplexer for better performance and smaller size, easily and efficiently.

Table 1 Our diplexer in comparison with the other previous designs

References	f_{o1}, f_{o2} (GHz)	$ S_{11} _{dB}$	FBW ₁ , FBW ₂	Size (λ_g^2)
This work	0.9, 1.8	0.17, 0.14	21, 24.3%	0.004
[1]	1.5, 2.1	0.7, 0.3	–	0.052
[2]	2.84, 4.08	0.7, 0.9	1.41, 2.2	–
[3]	16, 18.3	1.3, 1.36	2.5, 2.3	–
[4]	2.4, 2.7	0.18, 0.39	–	0.075
[6]	2.3, 4	0.2, 0.4	–	0.09*
[7]	2.4, 3.5	1.43, 1.59	–	0.282
[8]	2.6, 6	0.6, 0.9	–	0.076*
[9]	3.6, 5.2	1.35, 1.31	8.2, 7.6%	0.05
[10]	1.9, 2.1	1.2, 1.5	3.5, 3.2%	0.136
[11]	1.8, 2.4	0.4, 0.4	–	0.095
[12]	1.1, 1.3	1.83, 1.52	8, 9.2%	0.705
[13]	2.3, 2.7	1, 0.9	6.1, 5.8%	0.127
[14]	1.6, 2.1	0.10, 0.16	16.8, 11%	0.054
[15]	1.8, 2.4	2.2, 2.1	2.8, 1.9%	0.064
[16]	1.7, 1.8	2.1, 2.1	5, 5%	0.07
[17]	1, 1.3	0.21, 0.21	4.6, 4.6%	0.018
[18]	1.05, 1.87	1.4, 3.4	6.1, 4%	0.089
[19]	0.7, 1.8	0.17, 0.30	12.5, 4%	0.026
[20]	1.67, 1.88	0.43, 0.35	2, 2%	0.036
[21]	1.6, 3	0.45, 0.60	34.3, 38.6%	0.0081
[30]	2.1, 5.1	0.3, 0.4	14, 4%	0.083
[31]	3.7–4.2, 7.25–7.75	1.6, 2.6	–	–
[32]	0.895, 2.2	0.075, 0.14	31, 5.45%	0.008

*Approximated values

Then, using the analyzed resonator, two bandpass filters (BPFs) were designed. Finally, a microstrip diplexer with high-performance was created by integrating the proposed BPFs. The presented diplexer operates at GSM frequencies with an overall size of 232.3 mm² (0.004 λ_g^2). Moreover, having two wide fractional bandwidths makes it suitable for wideband applications. It has two low $|S_{11}|_{dB}$ of 0.17/0.14 dB, hence, it is suitable for energy harvesting.

Data availability Data sharing not applicable to this article as no datasets were generated or analysed during the current study.

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