Abstract  A reconfigurable bandstop microstrip line filter with wide continuously adjustable stopband is proposed in this letter. Dumbbell-shaped defected ground structure cells based on dual coupling slots are used to reduce the complexity of the filter and simplify the equivalent circuit modeling. Three varactors crossing the center coupling slots are utilized together with isolating capacitors to obtain the reconfigurability, of which the frequency range can be adjusted from 1.8 to 8.0 GHz. An equivalent-circuit model using coupled LC resonators is developed to firstly attain physical understanding of the filter’s working principle and secondly to assist the filter design. Three pairs of perpendicular stubs are introduced onto the microstrip line to improve the transmission and maintain signal integrity in the passband. The filter has been fabricated and experimentally characterized, with the good agreement observed between simulation and measurement validating the concept.

Keywords: reconfigurable filter, bandstop filter, defected ground structures, varactor-based

Classification: Microwave and millimeter-wave devices, circuits, and modules

1. Introduction

Bandstop filters (BSFs) [1, 2, 3, 4, 5, 6] are widely used in microwave and millimeter-wave circuits since they can suppress harmonics, spurious signals and out-of-band noise. In particular, in order to avoid interferences between multiple wireless systems in close proximity, wide-stopband band-pass filters are in high demand [7, 8, 9]. Reconfigurable microwave BSFs are developed to feature more functionalities than their conventional counterparts and can result in reduced system complexity. The desired characteristics for this type of reconfigurable filters include a large tuning range, a broad stopband, compact size and a sharp cutoff frequency response.

Many different approaches have been investigated to construct the reconfigurable BSFs with good characteristics for rejecting unwanted bands. The design of stopband filter based on the coupling relationship of microstrip line structure is one of the research focuses [10, 11, 12, 13, 14]. In [10], a frequency and bandwidth tunable BSF using substrate-integrated waveguide (SIW) resonators was proposed, which used a tunable coupling structure between the microstrip transmission line and the SIW resonator to obtain the bandwidth tuning capability. In [11], a reconfigurable bandstop filter using parallel coupled lines and two varactor diodes was proposed to realize a wide tunable range, of which the centre frequency could be tuned from 0.47 to 1.67 GHz. In [12], A varactor-based tunable bandstop filter was proposed, of which a dual-mode circuit is designed and the frequency tunability is achieved by using varactor diodes instead of the lumped capacitors in the circuit. In [13], a tunable constant bandwidth BSF was realized using two varactor-based resonators in a doublet configuration, where the coupling between the resonators and the main line was controlled by tuning the varactors’ capacitance. In [14], a three-pole tunable BSF using slotted ground structure was proposed, of which varactor was deployed to tune each resonator. The centre frequency of the filter was from 4.5 to 5.5 GHz, of which tuning range is 20% and fractional bandwidth is 11.6 and 15.9%, respectively. Various reconfigurable electromagnetic bandgap structure (EBG) [15, 16] and defected ground structures (DGS) have been proposed and applied for bandstop filters [15, 17, 18, 19]. In [15], a microstrip (MS) BSF is designed by embedding three metamaterial-based electromagnetic bandgap structure (MTM-EBG) in a MS line to produce a bandstop filtering response, and placing a dielectric plate directly on the surface of the MTM-EBGs to make it tunable. In [17], a compact reconfigurable bandstop resonator based on a modified dumbbell-shaped DGS and an embedded patch was proposed for coplanar waveguides, with the tunability of frequency range achieved by using varactors to change the resonant frequency of the DGS resonator. In [18], a compact reconfigurable bandstop resonator based on a modified dumbbell-shaped DGS and an embedded patch was proposed for coplanar waveguides, with the tunability of frequency range achieved by using varactors to change the resonant frequency of the DGS resonator. In [19], a tunable microstrip bandstop resonator with a C-shaped DGS was proposed, where varactors were embedded to the resonator cell and a 13% tuning range centered at 2.36 or 2.67 GHz is achieved. Some new technologies are used in the reconfigurable stopband filter [20, 21, 22, 23]. In [20], a tunable BSF based on a half-mode substrate integrated waveguide (HMSIW) was proposed, where reconfigurability was obtained by adjusting the bias voltage of multiple interdigitated microstrip resonators on liquid crystal material. However, the fractional bandwidth of only about 10% and a slow response time might limit its applications. In [22], a tunable frequency-agile BSF was proposed, which adopted two substrate-integrated evanescent-mode cavity resonators

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with piezoelectric actuators used to control the resonant frequency of each resonator. In all those cases, the targeted reconfigurable stopbands were relatively narrow, hence possible application of the proposed devices might not be suitable in scenarios that require a wider stopband.

In this letter, a concept of wideband DGS-based reconfigurable BSF for microstrip lines is proposed. The filter adopts three dual-coupling-slots dumbbell-shaped DGS cells to obtain frequency reconfigurability. When building a reconfigurable cell, varactor can be directly and conveniently welded to the dual-coupling-slots dumbbell-shaped DGS cell without changing its structure, which reduces the design difficulty and simplifies the equivalent circuit modeling. In order to effectively compensate the impedance variations due to the DGS structure underneath, a varactor can be directly and conveniently welded to the dual-coupling-slots dumbbell-shaped DGS cell without changing its structure, which reduces the design difficulty and simplifies the equivalent circuit modeling. In order to effectively compensate the impedance variations due to the DGS structure underneath, pairs of stubs with rounded edges are introduced onto the microstrip lines, which hence significantly improve the in-band transmission. An equivalent circuit model for the proposed filter is proposed to explain the operation principle, and to support the design process. As a demonstration, a BSF based on three varactor-loaded DGS cells is designed, fabricated and measured. The proposed filter exhibits a wide reconfigurable bandstop bandwidth within the 1.8 to 8.0 GHz range, leading to a fractional tuning range of around 75%.

2. Continuously reconfigurable low-pass filters

2.1 Design of filter structure

The configuration of the proposed reconfigurable BSF is shown in Fig. 1 and the dimensions are given in the caption, where dark gray represents metal microstrip line, white represents slot, light gray represents metal ground, and gray represents capacitor. Three DGS cells adopting coupled broad slots are embedded in the ground plane, and are placed symmetrically along the microstrip line. The application of dual-slotted DGS is used to solve the problem of introducing variable capacitance to the reconfigurable DGS cell, where conventional single slot DGS cell is not suitable for introducing varactor [14, 24]. Three varactors and DC-blocking capacitors crossing the slots are utilized to add tunability for the bandstop. Three pairs of round edges stubs with radius R are introduced on the microstrip line to smoothly compensate the impedance variations due to the DGS structure and thus improve the out-of-band transmission. In the process of simulating the filter, the thicknesses of air layers at top and bottom is set 75 mm.

2.2 Model of continuously tunable dual-slotted DGS cell

The working principle of dual-slotted DGS is similar to that of traditional single slot DGS. Take the left DGS cell in Fig. 1 as an example, there exist two equivalent gap capacitance $C_1$ and $C_2$ in the dual-slotted DGS cell, and one coupling capacitance $C_{M12}$ between the two gap capacitors $C_1$ and $C_2$ because of the short distance, as shown in Fig. 2.

To simplify the analysis, the total equivalent capacitance is expressed as $C_{D1}$. Therefore, the equivalent parameters $L_{D1}$ and $C_{D1}$ of the left DGS cell can be determined from equations (1) and (2) [25], where Z is the impedance of the microstrip line, $\omega_{c1}$ is 3 dB lower cutoff angular frequency and $\omega_{o1}$ is resonant angular frequency.

$$L_{D1} = \frac{2Z}{\omega_{c1}} \frac{\omega_{o1}^2 - \omega_{c1}^2}{\omega_{o1}^2} \quad (1)$$

$$C_{D1} = \frac{1}{2Z} \frac{\omega_{o1}^2 - \omega_{c1}^2}{\omega_{o1}^2} \quad (2)$$

In order to improve the filtering range of the filter, varactor is used in the DGS-based filter. Because the equivalent capacitance of varactor can be adjusted by loading different supply voltages on it, the resonance point of the varactor-based DGS cell can be reconstructed. The best location to place the varactors is the middle of each DGS cell where the strongest return currents would exist in the undisturbed microstrip line, as shown in Fig. 1(b). The chosen varactor ($C_v$) is a MA46H120 from MACOM and it has a wide 1:10 capacitance tuning ratio from 0.12 to 1.3 pF. A DC-blocking capacitor ($C_c$) in series with the varactor is needed for applying the reverse bias voltage from 18 to 0 V.

In the actual equivalent circuit modeling, variable capacitance and bypass capacitance can not be replaced by ideal
Capacitance, it needs to use RLC equivalent circuit model. The chosen varactor is modeled as a lumped element having a resistance \( R_v = 2 \Omega \), an inductance \( L_v = 0.05 \) nH and a variable capacitance \( C_v \) ranging from 0.12 to 1.3 pF, whereas the DC-blocking capacitor equivalently shows a capacitance \( C_s = 0.3 \Omega \) and an inductance \( L_s = 0.2 \) nH, as is shown in Fig. 3b.

As shown in Fig. 3, the varactor and DC-blocking capacitor are connected in series. Then the equivalent capacitance \( C_e \) equal to \( \frac{C_v C_s}{C_v + C_s} = C_v \). Since the \( C_v \) is much larger than \( C_s \), the equivalent capacitance \( C_e \approx C_v \). To simplify the equivalent circuit model, the equivalent resistance \( R_e \), capacitance \( C_e \) and inductance \( L_e \) for the in-series varactor and DC-blocking capacitor can be obtained as follows: \( R_e = R_v + R_s = 2.3 \Omega \), \( L_e = L_v + L_s = 0.25 \) nH and \( C_e = C_v \), as depicted in Fig. 3c.

Figure 4 shows the relationship between the resonant frequency of the left reconfigurable DGS cell and the different capacitance values of varactor. With the increase of variable capacitance, the resonant frequency decreases continuously. When the variable capacitance is 0.12 pF, the resonant frequency is 5.0 GHz. When the variable capacitance is 1.3 pF, the resonant frequency is 2.2 GHz.

![Varactor and bypass capacitor equivalent circuits.](image)

![Graph showing the relationship between resonant frequency and pad width d of dual-slotted DGS.](image)

2.3 Analysis of proposed reconfigurable DGS-based filter

The equivalent circuit of the proposed filter is shown in Fig. 5. The DGS cells can be modeled as three cascaded parallel LC resonator circuits which alter the path of return currents. The equivalent inductance \( L_{Di}(i = 1, 2, 3) \) comes from the additional path formed around the DGS for the return currents. The equivalent capacitance \( C_{Di}(i = 1, 2, 3) \) is formed over the coupling slots in the center of the DGS structure, accounting for the most significant displacement currents. At their resonance, the parallel LC resonators block the unwanted frequency components.

Since the distances between DGS cells are very small, there exist non-negligible mutual coupling inductances \( M_{ij}(i = 1, 2, j = 2, 3) \) between the DGS cells, which improve the filtering capability and hence enable the wide stopband. The mutual capacitances between adjacent DGS can be neglected as the center gaps are far enough from each other. For the mutual inductance, they can be estimated according to equation (3) [26].

\[
M_{ij} = -0.5 \frac{f_{oi} + f_{oj}}{f_{oi} f_{oj}} \times \frac{f_{oi}^2 - f_{oj}^2}{f_{oi}^2 + f_{oj}^2} \left( \frac{f_{oi}^2 - f_{oj}^2}{f_{oi}^2 + f_{oj}^2} \right)
\]

where \( f_{oi} \) and \( f_{oj} \) denotes the self resonance frequency of the isolated DGS cells while \( f_i \) and \( f_j \) stand for their two split resonance frequencies because of the coupling.

Figure 6 shows the full-wave simulated transmission coefficients of the proposed BSF, for which all the varactor diodes are controlled by the same dc-power supply. When all \( C_v \) are set to the featured values of 0.12, 0.3, 0.5 and 1.3 pF simultaneously, the -10 dB stopband frequency ranges are 3.0-7.0, 2.5-5.6, 2.2-4.7 and 1.7-3.0 GHz, respectively. This implies that, with only a single voltage supply to bias the three varactors, the filter is able to provide a satisfactory tunable bandstop performance. Advanced control of the filtering performance can be done through individual DGS manipulation if necessary. Figure 6 also shows the simulated transmission coefficients based on the equivalent circuit modeling. The agreement of full-wave simulation is satisfactory, suggesting that the equivalent circuit modeling can be used to estimate the filter’s performance effectively when varying the varactor capacitance.
Fig. 6 Full-wave and equivalent circuit simulated results of the proposed filter when $C_{v1} = C_{v2} = C_{v3} = 0.12, 0.3, 0.5$ and $1.3 \ \text{pF}$, for (a) to (d) respectively.

Fig. 7 Comparison simulated results of the proposed filter with or without the stubs on the microstrip line. (a) $C_{v} = 0.12 \ \text{pF}$, (b) $C_{v} = 1.3 \ \text{pF}$.

2.4 Optimization of the proposed reconfigurable filter

Because the ground under the microstrip line is etched to realize the DGS, the reference ground is changed, which increases the characteristic impedance of the microstrip line at this position [27, 28, 29]. It is obviously that the impedance of microstrip line can be reduced effectively by increasing the linewidth of microstrip line when other conditions remain unchanged. As a result, an efficient way to counterbalance the impedance change is to adopt suitable stub to the microstrip line at the position over the DGS cell. However, the length of the stub is limited by two factors. One is that a large size will be taken up if the length of the stub is too large, which is not suitable for application. The other factor is that coupling capacitance will appear between stubs when the length of stubs is too large. However, a proper introduction of stub can effectively reduce the impact of impedance mismatch.

As shown in Fig. 1, three perpendicular stubs centered onto the middle of the microstrip line are introduced to improve the transmission performance in the lower frequencies as well as the sharpness of the lower stop-band transition [30]. The rounded corners are utilized to smooth transitions and thus reduce reflection. As shown in Fig. 7, the lower-cut-off frequencies of the filter with stubs and without stubs at -10 dB are $2.9$ and $3.2 \ \text{GHz}$ respectively when all $C_{v} = 0.12 \ \text{pF}$. When all $C_{v} = 1.3 \ \text{pF}$, the lower-cut-off frequencies of the filter the with stubs and without stubs at -10 dB are $1.7$ and $1.8 \ \text{GHz}$ respectively. Moreover, the transmission loss of the proposed filter at low frequency is reduced by 1 or 1.2 dB with respect to the case without stubs when the varactor is $0.12$ or $1.3 \ \text{pF}$ respectively. As a result, the proposed filter with stubs achieves enhanced performance with improved impedance matching in the lower pass-band, which exhibits a nearly perfect transmission and a sharper transition to the stop-band. A slightly higher insertion loss is observed at the higher pass-band. However, it remains within an acceptable level.

The fractional tuning range (FTR) is a parameter of primary importance in reconfigurable filter, where it is defined as

$$FTR = 2 \cdot \frac{f_{c-high} - f_{c-low}}{f_{c-high} + f_{c-low}}$$

where $f_{c-high}$ is the center frequency of the bandwidth at high frequency, $f_{c-low}$ is the center frequency of the width at low frequency. As shown in Fig. 7, the center frequency of the bandwidth at high frequency and low frequency is $(3.0 + 7.0)/2 = 5.0 \ \text{GHz}$ and $(1.7 + 3.0)/2 = 2.35 \ \text{GHz}$ respectively, then the FTR of the filter is 72.1%.

3. Experimental results

A prototype filter has been fabricated on a $381-\mu \text{m}$ thick Rogers TMM3 substrate with a relative permittivity of 3.27 and a loss tangent of 0.002, as shown in Fig. 8. The overall dimensions of the BSF are $50 \times 45 \ \text{mm}^2$. A simple bias circuitry is added in the BSF, which provides a dc bias voltage for varactor control, as is shown in the Fig. 8(c). The bias circuitry includes an RF-blocking resistor $R_{b}$ (1 M$\Omega$) and an RF-choke inductance $L_{b}$ (100 nH) in series. In order to reduce the effect of wire, the resistance should be close to the pad where the varactor is located. The fabricated unit is measured using a vector network analyzer N5230A. A comparison between the simulated and measured reflection coefficients $|S_{11}|$ and transmission coefficients $|S_{21}|$ is shown in Fig. 9. When all $C_{v}$ are set to 1.3 pF, the stopband with rejection better than 10 dB ($|S_{21}| < -10 \ \text{dB}$) extends from 1.7 to $3.0 \ \text{GHz}$ in simulation whereas the measured values range from 1.8 to 3.2 GHz. Both the simulated $|S_{11}|$ parameters are around -1 dB. When all varactor capacitances $C_{v}$ are...
and largest tuning range in terms of stopband performance. It is noted that the proposed filter has the widest bandwidth, compared recently reported tunable filters is summarized in Table I. The simulation.

terial uncertainties can also introduce discrepancies from fabrication and manufacturing. First, the lumped RLC parameter of the varactor and DC-blocking capacitor are set as constant values in HFSS simulations. Second, practically they are frequency-dependent. While there exist capacitances between the soldered lumped components, they are neglected in simulation. Finally, tolerances introduced by the fabrication and material uncertainties can also introduce discrepancies from simulation.

A comparison between the proposed BSF and some typically recently reported tunable filters is summarized in Table I. It is noted that the proposed filter has the widest bandwidth, and largest tuning range in terms of stopband performance.

4. Conclusion

A continuously reconfigurable wideband DGS-based microstrip BSF adopting varactors has been presented. It uses dual coupling center-slots DGS to solve the problem of conventional DGS, which needs to alter the DGS pattern to load varactor and bypass capacitor. An equivalent circuit model for the varactor-loaded DGS has been developed and utilized for the filter analysis and design. By applying different bias voltages to the varactors, the frequency range (defined by the transmission loss not less than 10 dB) of the proposed filter is measured to be tunable between 1.8 to 8.0 GHz. To the authors’ best knowledge, this filter exhibits the largest tuning range and widest bandwidth in the stopband, among the reported bandstop filters in the open literature. All these findings indicate that the proposed reconfigurable filter is promising for high speed digital applications.

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Table I

Comparison among various reconfigurable bandstop filters

| Ref | Size (Å$_L$) | Tuning range (GHz) | $BW_{min}$ 10 dB (GHz) | $BW_{max}$ 10 dB (GHz) | FTR |
|-----|-------------|-------------------|------------------------|------------------------|-----|
| [10] | 0.28×0.56 | 2.8-3.4 | 0.02 | 0.1 | 19% |
| [11] | 0.23×0.53 | 2.5-5.5 | 0.07 | 1.4 | 19% |
| [12] | 0.17×0.21 | 1.2-1.7 | 0.2 | 1.2 | 19% |
| [13] | 0.18×0.19 | 0.6-0.99 | 0.11 | 0.25 | 19% |
| This work | 0.24×0.36 | 2.5-5.5 | 1.3 | 4.2 | 79% |

at 0.12 pF, the stopband with rejection better than 10 dB is 2.9 - 7.1 and 3.1 - 8.0 GHz for simulation and measurement respectively. The corresponding simulated return loss is around 0.5 dB and the measured value is about 1 dB.

Overall, Fig. 9 confirms that a wide stop band with tuning range of 75% centered at 4.0 GHz is obtained with the proposed reconfigurable filter. Measurements agree well with the simulations at the lower frequency range, while at the higher frequency range, the measured result shows an extended stop band. The discrepancy between the simulated and measured results is mainly caused by the following reasons. First, the lumped RLC parameter of the varactor and DC-blocking capacitor are set as constant values in HFSS while practically they are frequency-dependent. Second, there exist capacitances between the soldered lumped components and the ground, which are neglected in simulation. Finally, tolerances introduced by the fabrication and material uncertainties can also introduce discrepancies from simulation.

Fig. 9  Simulated and measured reflection and transmission coefficients of the proposed filter. (a) When $C_v = 1.3$ pF; (b) when $C_v = 0.12$ pF.
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