A new SIDGS-based tunable BPF design method with controllable bandwidth

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SUMMARY This paper provides a new method to implement substrate integrated defected ground structure (SIDGS)-based bandpass filter (BPF) with adjustable frequency and controllable bandwidth. Compared with previous literature, this method implements a new SIDGS-like resonator capable of tunable frequency in the same plane as the slotted line using a varactor diode, increasing the design flexibility. In addition, the method solves the problem that the tunable BPF constituted by the SIDGS resonator cannot control the bandwidth by introducing a T-shaped non-resonant unit. The theoretical design method and the structural design are shown. Moreover, the configured structure is fabricated and measured to show the validity of the design method in this paper.

key words: Substrate integrated defected ground structure (SIDGS), tunable band-pass filter (BPF), controllable bandwidth.

1. Introduction

With the development of modern wireless systems, some current communication devices require miniaturization while also operating in multiple frequency bands. In order to satisfy the development needs of these communication devices, bandpass filters (BPFs) are constantly being updated and progressed. Therefore, there is a strong need for a BPF filter that can operate in multiple frequency bands while maintaining bandwidth and miniaturization.

SIDGS consists of slot lines and metal vias that surround the slots [1]. Compared with waveguides and SIWs [2][3], it has a smaller volume [4], and also has a lower radiation loss [5]. Compared with a microstrip line [6], it has lower radiation loss, a more robust ability to resist interference from external circuits [7], and also has the characteristic of the compact structure of a microstrip line [8][9]. Therefore, it can satisfy the requirements of miniaturization and low radiation loss at the same time. Regarding the realization of frequency adjustment, the more common method is to connect the resonator to a varactor diode [10][11][12] or MEMS to control the frequency [13][14], or to adjust the length of slot lines manually [15]. It is also possible to use a particular substrate, such as liquid crystal materials [16], which a voltage can adjust the dielectric constant ratio of the substrate to achieve frequency tunability. The method of electronic control used in [10][11][12][13][14] cannot be directly applied to SIDGS for structural reasons. In [17], a harmonic-controlled SIDGS resonator is proposed, which uses a varactor diode to indirectly control the equivalent inductance of the SIDGS to change the resonant frequency. Additionally, there is a method to adjust the resonant frequency by crossing a varactor diode on both sides of the slot of CSRR and DGS(similar to SIDGS in structure), which changes the equivalent capacitance in the equivalent circuit [18][19][20]. In the above method, all the adjustment circuits must pass through the substrate, and there is a problem in that the center frequency and bandwidth cannot be controlled independently.

In this paper, a new center frequency adjustment method and bandwidth adjustment method are proposed, which enables the center frequency and bandwidth to be adjusted independently, and the control circuit of the center frequency will be in the same plane of the SIDGS and outside the SIDGS, significantly improving the flexibility of the design. The principles of frequency control and bandwidth control by this method are analyzed in detail in Chapter 2. In Chapter 3, the practical application of the frequency adjustment circuit and the bandwidth adjustment circuit is discussed, as well as the effect on the external Q value after loading a varactor diode on the port. Next, Chapter 4 presents the design of a tunable BPF with constant bandwidth. Finally, conclusions are given in Chapter 5.

2. Adjustment methods of center frequency and coupling coefficient

2.1 An adjustment method of the center frequency

Fig. 1 shows an SIDGS resonator cell designed for frequency adjustment. The cell consists of two parts, a C-shaped slotted line and metal vias surrounding the slot forming the SIDGS resonator part (the part inside the red dotted line), and a frequency adjustment circuit part consisting of an I-shaped slotted line loaded with capacitors and grounded metal vias.
When we change the capacitance of capacitor C, the scattering parameters shown in Fig. 2 can be obtained. By substituting the scattering parameters obtained in Fig. 2 into Eq. 1 [21], we can obtain the equivalent impedance at the resonant frequency under different capacitances of about 50Ω above 3.1GHz[22]. This is because, on the whole, although the width of the slot line of SIDGS is the same, it is not a uniform transmission line, it does not have a constant characteristic impedance, but it can be regarded as a unit of periodic lumped elements. It is similar to complementary split rings resonators [23] in that the equivalent impedance keeps changing with frequency and resonates when it is equal to 50Ω. Therefore, its equivalent circuit is shown in Fig. 3, and it is equivalent to a transmission line with a characteristic impedance of 50Ω at the resonant frequency.

\[
Z_B = \frac{2b}{a-d+\sqrt{(a+d)^2-4}}
\]  

(1)

In Eq. 1, a, b, c, and d are ABCD parameters, which can be obtained from the simulation results. Where \(a = \frac{2521}{2521} \), \(b = (1+S_{11})(1+S_{22})-S_{12}S_{21} \), \(c = (1-S_{11})(1-S_{22})-S_{12}S_{21} \), \(d = \frac{2521}{2521} \), and the characteristic impedance of the source \(Z_0 = 50 \Omega \). \(S_{11}, S_{12}, S_{21}, \) and \(S_{22} \) are scattering parameters.

In Fig. 3, \(Z_1 \) and \(l_1 \) are the characteristic impedance and length of the 50Ω transmission line. \(Z_2 \) and \(l_2 \) are half of the characteristic impedance and length of SIDGS. \(Z_{in}(50\Omega) \) and \(Z_L(50\Omega) \) are input impedance and load impedance respectively.

According to Fig. 3, we can obtain Eq. 2 and Eq. 3.

\[
[A] = \begin{bmatrix}
\cos(\beta l_1) & jZ_1 \sin(\beta l_1) \\
\frac{j\sin(\beta l_1)}{Z_1} & \cos(\beta l_1)
\end{bmatrix}
\]  

(2)

\[
\beta_0 (x = 1 \text{ or } 2) \text{ in Eq. 2 and Eq. 3 is the phase constant.} \\
\omega_0 \text{ is the central angular frequency.}
\]

At the resonant frequency \(f_0 = \omega_0 / 2\pi \), \([A] = [B]\), and further, we can obtain Eq. 4. Since SIDGS is a half-wavelength resonator [24], we can obtain Eq. 5 by substituting \(\beta l_1 = \pi \) into Eq. 4.

\[
\cos(\beta l_1) = \cos(2\beta l_2) - 2\omega_0C \sin(2\beta l_2)
\]  

(4)

\[
C = \frac{2(\cos(2\beta l_2)+1)}{\omega_0Z_2\sin(2\beta l_2)}
\]  

(5)

In Eq. 5,

\[
\beta_2 = \frac{2\pi f_0}{\sqrt{\varepsilon_{reff}}}
\]  

(6)

In Eq. 6, \(\varepsilon_{reff} = 2.97 \) is the effective dielectric constant of SIDGS in Fig. 1.

In addition, since the equivalent resistance of SIDGS is 50Ω at the resonance frequency, Eq. 7 [25] can be listed according to Fig. 3, thus obtaining Eq. 8.

\[
Z_{in} = Z_2 + \frac{1}{\omega_0C} + \frac{1}{Z_2+1/\omega_0C+Z_L}
\]  

(7)

\[
Z_2 = \sqrt{\frac{Z_{in}^2 - \frac{2}{\omega_0C}}{\omega_0C}}
\]  

(8)

Substituting Eq. 5 into Eq. 8, the resonant frequency under different capacitances can be obtained. The calculation results were obtained using Matlab R2020b.
Taking the center frequency of 4.33GHz as an example, \( Z_2 = 49.98 - j1.34 \ \Omega \) and \( C = 0.55 + j0.014 \ \text{pF} \) are calculated. We will find that \( Z_2 \) in the calculation result is a complex value close to 50\( \Omega \), and \( C \) will also have a small imaginary part, which is considered to be the neglect of the influence from the I-shaped slot and the metal via. The final calculation result takes real part.

Comparing the calculated results with the model simulation results of HFSS 12.0, as shown in Fig. 4, we can find that the error is within 0.04 pF in the range of 4.09-4.86 GHz. However, the error generated becomes larger as the frequency decreases below 4.09 GHz. The error is considered to be the neglect of the influence from the I-slot when calculating the value of \( C \). According to the above results, the resonant frequency can be predicted by the method in a certain range, the above equivalent circuit is reasonable, and the loading capacitance in the same plane of the SIDGS is also proved feasible.

![Fig. 4 Calculation and simulation results of C at different frequencies.](image)

2.2 An adjustment method of the coupling coefficient

![Fig. 5 Topology Logic for SIDGS Bandwidth Controllable.](image)

Achieving independent control of the coupling between two SIDGS is an unsolved problem. SIDGS cannot be controlled by connecting two resonators with varactor diodes, similar to literatures [26][27] because of the structural reasons. In this paper, a T-type non-resonator is provided between two SIDGS. By loading the varactor diode on the T-type non-resonator, the independent control of the coupling between the SIDGS can be achieved, as shown in Fig. 5.

According to Fig. 5, due to the addition of this structure, a coupling path (R1-\( k_{n1}\)-N-\( k_{n2}\)-R2) through the non-resonator is added to the original coupling of the two SIDGS resonators. The coupling between the SIDGS and the non-resonator can be changed by changing the bias voltage on the varactor on the non-resonator so that the total coupling \( k \) between the two SIDGS can be changed.

![Fig. 6 Structure of the T-shaped non-resonator.](image)

In order to study the influence of the structure of the T-type non-resonator on the coupling, the T-type non-resonator with different \( L_{d1} \) and different capacitances was simulated and calculated. The structure of T-type non-resonator is shown in Fig. 6. Fig. 7 shows the simulation results when \( L_{d1} = 7 \text{mm} \) and \( C_x = 1 \text{pF} \).

![Fig. 7 An example of calculating coupling coefficient \( k \).](image)

When calculating \( k \), Eq. 9 is used [28].

\[
k = \frac{f_2^2-f_1^2}{f_2^2+f_1^2}
\]  

(9)

In Eq. 9, \( f_1 \) and \( f_2 \) are the lower resonant frequency and the higher resonant frequency of the two resonators coupled, respectively, as shown in Fig. 7. The results obtained from Eq. 9 are collated in Fig. 8 and Fig. 9. According to Fig. 8, we found that the longer the length of \( L_{d1} \), the greater the variation of \( k \) when adjusting \( C_x \), and the greater the tuning range. In the extreme case when the length of \( L_{d1} \) is 0 mm, adjusting \( C_x \) does not cause \( k \) to change. We do not find any pattern in Fig. 8, but when we fix the length of \( L_{d1} \) as in Fig. 9, we can see that \( k \) increases and converges to a certain value as \( C_x \) increases. And the longer the length of \( L_{d1} \), the more significant the change.
3. Design of a tunable BPF

3.1 Calculation of $Q_{ex}$ and $k$

To further verify the feasibility of the above method, a BPF with a frequency adjustment range of 3.7–4.2 GHz is designed, and a tunable BPF with a constant bandwidth of 100 MHz can be made by voltage control. The design specification of the tunable BPF is shown in Table 1. Since $Q_{ex}$ and $k$ change monotonically and continuously, and take the maximum and minimum values at 3.7 GHz and 4.2 GHz, only the results at 3.7 GHz and 4.2 GHz are calculated for reference [29].

<table>
<thead>
<tr>
<th>$f_0$ (GHz)</th>
<th>Characteristics</th>
<th>Order</th>
<th>BW (MHz)</th>
<th>RL (dB)</th>
<th>$Q_{ex}$</th>
<th>$k$</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.7</td>
<td>Chebyshev</td>
<td>2</td>
<td>100(0.05dB)</td>
<td>19.4</td>
<td>52.2</td>
<td>0.0191</td>
</tr>
<tr>
<td>4.2</td>
<td>Chebyshev</td>
<td>2</td>
<td>100(0.05dB)</td>
<td>19.4</td>
<td>59.3</td>
<td>0.0169</td>
</tr>
</tbody>
</table>

3.2 Design of frequency adjustment circuit

The reasonableness and validity of the method are verified in section 2.1, but the circuit can not be directly used for the actual design of tunable BPF. A varactor diode needs to be used in the actual design, and a bias voltage needs to be applied. Therefore, it is also necessary to modify the structure of the frequency adjustment circuit of the SIDGS.

The modified circuit is shown in Fig. 10.

Different $C_f$ cases are simulated, and the relationship between $C_f$ and $f_0$ is shown in Fig. 11. At 0.23–2.1 pF, $f_0$ can be varied from 3.67 to 4.86 GHz. This circuit can satisfy the design requirements.

Compared with Fig. 4, we find that its variation range is smaller. It is considered that the original I-shaped slot has changed, and two small copper sheet structures and a series-connected DC blocking capacitor $C_{cut}=12$ pF have been added, which makes the equivalent capacitance value of the circuit smaller. However, the changing trend has stayed the same, and the method is still valid.

3.3 Design of bandwidth adjustment circuit

According to Fig. 9, the required capacitance and adjustment range of $C_k$ is very small, and the varactor diode with the smallest capacitance that can be found is SMV-2201, with an adjustment range of 0.23–2.1 pF. Then the $f_0$ will be adjusted to 3.7 GHz and 4.2 GHz, respectively, and the coupling coefficients at different values of $C_k$ are calculated.

The measurement circuit of the adjusted T-type non-resonator is shown in Fig. 12.
The value of $k$ obtained from the circuit shown in Fig. 12 is collated in Fig. 13.

According to Fig. 13, when changing $C_1$, the change tendency of $k$.

3.4 Design of the Adjustment Circuit of $Q_{ex}$

Regarding the adjustment of the $Q_{ex}$, the method in [27] is adopted, and its structure is shown in Fig. 14. Among them, $C_q$ is the varactor diode of SMV-2201(range: 0.23-2.1pF), $C_{cut}$ is the chip capacitor of 6 pF, and $R_q$ is the chip resistor of 100 kΩ.

3.5 Simulation of Tunable BPF

According to sections 4.2, 4.3, and 4.4, a tunable BPF can be composed, as shown in Fig. 16.
The circuit simulation in Fig. 16 can be obtained in Fig. 17 and Fig. 18.

According to the simulation results, the circuit can be adjusted from 3.7 to 4.2 GHz, and the bandwidth can be controlled individually by controlling \( C_q \) and keeping the bandwidth at 100 MHz. In the lossy case, the IL is 1.6 dB for \( f_0=3.7 \) GHz, 1.61 dB for \( f_0=3.95 \) GHz, and 1.7 dB for \( f_0=4.2 \) GHz, to the maximum insertion loss.

The bandwidth tuning is mainly done by \( C_k \), when the bandwidth is 125 MHz, \( C_k=1.2 \) pF, \( C_f=0.37 \) pF, and \( C_q=0.81 \) pF. When the bandwidth is 150 MHz, \( C_k=1.42 \) pF, \( C_f=0.37 \) pF, and \( C_q=0.87 \) pF. When the bandwidth is 175 MHz, \( C_k=2.2 \) pF, \( C_f=0.39 \) pF, and \( C_q=0.81 \) pF.

**4. Measured results of Tunable BPF**

The tunable BPF circuit in Fig. 16 was worked out in practice. The actual circuit structure is shown in Fig. 20.
In Fig. 20, the substrate uses MEGTRON7 R-5785(N) with a thickness of 0.5 mm, a relative dielectric constant of 3.4, and a dielectric dissipation factor of 0.002. The circuit size is $32 \times 18 \text{mm}^2$. $C_f$, $C_k$, and $C_q$ all use MSV2201-040LF, $R_f$ and $R_q$ all use 667-ERJ-PA2F1003X, $C_{cut1}$ and $C_{cut2}$ use UMK-105-CH060DW, $C_{cut3}$ and $C_{cut4}$ use UMK-105-CH120JW, and the input of the black wire is the DC voltage $V_f$ that controls $C_f$, the input of the blue wire is the DC voltage $V_k$ that controls $C_k$, and the yellow wire is the DC voltage $V_q$ that controls $C_q$.

Fig. 21 and Fig. 22 show the measurement results at different frequencies when the BW is held constant. When $V_f = 6.6 \text{ V}$, $V_k = 7.6 \text{ V}$, $V_q = 23.5 \text{ V}$, $f_0 = 4.65 \text{ GHz}$, BW = 198 HMz, IL = 10.3 dB. When $V_f = 8.2 \text{ V}$, $V_k = 14.3 \text{ V}$, $V_q = 23.5 \text{ V}$, $f_0 = 4.75 \text{ GHz}$, BW = 201 HMz, IL = 6.8 dB. When $V_f = 17.6 \text{ V}$, $V_k = 18 \text{ V}$, $V_q = 21.3 \text{ V}$.
\[ f_0 = 4.85 \text{ GHz}, \quad BW = 200 \text{ HMz}, \quad IL = 4 \text{ dB}. \] All voltage values are readings from the constant voltage source. Comparing Fig. 18 and Fig. 21, it can be found that the measured results show a higher frequency, wider BW, larger IL, and tilted waveforms than the simulated results. The reasons are as follows: 1. During soldering, each component is not entirely asymmetrical, and there is an inevitable error in each component itself, which causes the frequency of the two resonators to be different, thus widening the BW and tilting the waveform. 2. Due to the larger BW, the required \( Q_{\text{en}} \) becomes smaller, and a gap of about 0.1mm between the two substrates, which makes \( Q_{\text{en}} \) larger, even if \( V_q \) is adjusted to 23.5 V (near the maximum voltage that the SMV2201 can withstand after voltage division). It cannot be reduced to the target value, causing IL to increase sharply.3. The ideal components are used in the simulation, but the influence of parasitic inductance and parasitic capacitance of various components at high frequencies, in reality, will become apparent, which results in higher insertion loss and a higher resonant frequency.

Fig. 23 and Fig. 24 show the measurement results under different bandwidths when \( f_0 \) is 4.8 GHz. When \( V_q \) changes from 20 V to 0 V, BW changes from 202 MHz to 276 MHz. Fig. 23 also has the same problem as Fig. 21. However, through Fig. 21, Fig. 22, Fig. 23 and Fig. 24, it can be demonstrated the amount of their variation decreases as the voltage increases that the adjustable method proposed in this paper is effective and can achieve independent control of frequency and bandwidth. If the manufacturing process is improved, a circuit close to the simulation result can be obtained. About the improvement methods: 1. In the case of conditions permitting, we can use such as low temperature co-fired ceramic substrate to make the circuit to reduce the structural error. 2. Use higher precision chip capacitors, resistors, and varactor diodes to reduce the degree of inconsistency in the center frequency of the resonator.

To further demonstrate the controllability of \( f_0 \) and BW, the relationship between \( V_q \) and \( f_0 \), and the relationship between \( V_q \) and \( \Delta f \) are shown in Fig. 25 and Fig. 26, respectively. \( \Delta f \) represents the bandwidth between the two resonant frequencies. \( f_0 \) becomes higher as \( V_q \) increases, and \( \Delta f \) decreases as \( V_q \) increases, but the amount of variation decreases as the voltage increases. Due to insufficient external Q, the IL will increase as the frequency becomes lower, resulting in the waveform becoming indistinguishable, and its maximum tuning range cannot be known. However, it can be determined that the maximum tuning range of \( f_0 \) contains 4.65-4.85 GHz, and the maximum tuning range of \( \Delta f \) contains 0.87-0.2 GHz, and the maximum insertion loss at 4.65 GHz is 10.3 dB.

Finally, a comparison with other tunable miniaturization BPFs is made in Table 2. The tunable center frequency and bandwidth are achieved while maintaining the miniaturization.

### 5. Conclusion

This paper proposes a new independently adjustable method of frequency and bandwidth based on SIDGS BPF and proves the method’s effectiveness through simulation and actual measurement.

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**Table 2. Comparison of this work with other tunable BPFs.**

<table>
<thead>
<tr>
<th>Items</th>
<th>Type</th>
<th>( f_0 ) (GHz)</th>
<th>Order</th>
<th>BW tuning</th>
<th>BW tuning range</th>
<th>Size ((\lambda_p \times \lambda_s))</th>
<th>IL (dB)</th>
</tr>
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<tr>
<td>[30]</td>
<td>DGS</td>
<td>2.87-3.63</td>
<td>2</td>
<td>no</td>
<td>-</td>
<td>0.19 x 0.25</td>
<td>3.6-2.4</td>
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<tr>
<td>[31]</td>
<td>HMDISIW</td>
<td>2.5-3.2</td>
<td>1</td>
<td>no</td>
<td>-</td>
<td>0.22 x 0.18</td>
<td>3-1</td>
</tr>
<tr>
<td>[32]</td>
<td>Microstrip line</td>
<td>1.27-1.6</td>
<td>2</td>
<td>yes</td>
<td><a href="mailto:67-107MHz@1.6GHz">67-107MHz@1.6GHz</a></td>
<td>0.16 x 0.11</td>
<td>3.35-1.6</td>
</tr>
<tr>
<td>[33]</td>
<td>SIW</td>
<td>1.24-1.8</td>
<td>3</td>
<td>yes</td>
<td><a href="mailto:75-187MHz@1.5GHz">75-187MHz@1.5GHz</a></td>
<td>0.83 x 0.30</td>
<td>4.2-8</td>
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<tr>
<td>[17]</td>
<td>SIDGS</td>
<td>1.75-2.7</td>
<td>2</td>
<td>no</td>
<td>-</td>
<td>0.07 x 0.08</td>
<td>2.7-1.75</td>
</tr>
<tr>
<td>This work: measurement</td>
<td>SIDGS</td>
<td>3.7-4.2</td>
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<td>1.9-1.6</td>
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<tr>
<td></td>
<td>SIDGS</td>
<td>4.65-4.85</td>
<td>2</td>
<td>yes</td>
<td><a href="mailto:202-276@4.8GHz">202-276@4.8GHz</a></td>
<td>-</td>
<td>4-10.3</td>
</tr>
</tbody>
</table>

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**References**


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