COMPACT ULTRA-WIDEBAND BANDPASS FILTER WITH DEFECTED GROUND STRUCTURE

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Abstract—A compact planar microstrip ultra-wideband (UWB) bandpass filter is presented in this paper. The proposed UWB filter is realized by cascading a high pass filter (HPF) and a lowpass filter (LPF). HPF with short-circuited stubs is used to realize the lower stopband and a lowpass filter based on a defected ground array in the ground plane employed to attenuate the upper stopband. One such a bandpass filter is designed and simulated.

1. INTRODUCTION

In recent years, the ultra wideband technology has become quite interesting. One of the major components for a UWB communication system is UWB bandpass filter. The fractional bandwidth of such a BPF usually exceeds *100%* (3.1–10.6 GHz). Designing a BPF with wide bandwidth, compact size, low insertion loss and also wideband rejection is still a challenging task. Researchers have proposed a few new structures [1–4].

In this paper the UWB bandpass filter is realized by cascading a high pass filter and a LPF. The designed UWB bandpass filter has a good frequency performance in the passband and a wide rejection in the stopband. It also has a compact size. The HPF consists of short-circuited stubs. The designed LPF utilizes a non uniform DGS array with nine elements.

2. HIGH PASS FILTER DESIGN AND SIMULATION

A typical high pass filter prototype and also the layout and the dimensions of the utilized high pass filter of this paper are depicted in
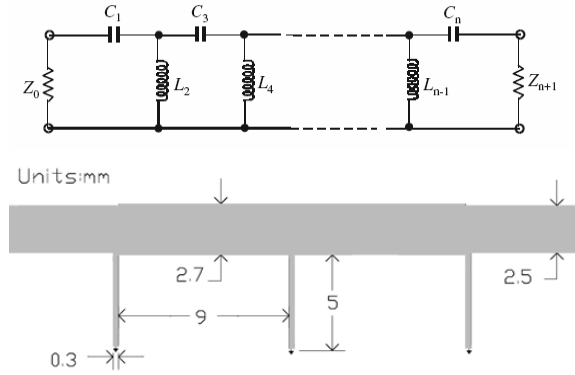
Figure 1. High pass prototype circuit and layout of the utilized HPF.

Fig. 1. The complete design procedures of such a HPF are addressed in [5, 6].

As we know this type of high pass filter shows a periodic response due to its distributed nature and it has a bandpass characteristic from DC up to $2f_0$ and a notch at $2f_0$.

So its centre frequency must be large enough in order not to disturb the passband of our BPF. Hence the major point of designing the HPF for the proposed UWB filter is choosing f_0 [5, 6]. This parameter determines the lengths of the short-circuited stubs and must be chosen in order to give a sufficient wide passband for our design. In this paper the stubs length are determined by choosing $f_0 = 10\,\text{GHz}$.

Simulated S-parameters of the HPF are shown in Fig. 2. S-parameters were simulated Using Ansoft Designer planar EM simulator. The substrate used for the simulation was *Rogers RT/duroid 5880* with the thickness of $h = 31\,\text{mils}$ and a dielectric constant of $\varepsilon_r = 2.2$.

3. LOWPASS FILTER DESIGN AND SIMULATION

An etched area on the ground plane disturbs the shield current and it can change the effective inductance and capacitance of a transmission line. It is one the most interesting areas of the LPF design and research [7, 9]. Microstrip line with defect ground structure pattern provides bandgap effect at certain frequency [8, 9].

An equivalent parallel RLC circuit corresponds to this bandgap effect, which can be used in filter design [7]. Since a DGS unit presents both the cutoff (lowpass) and the resonance (bandstop) characteristic, instead of traditional microstrip LPF designs, the microstrip low pass
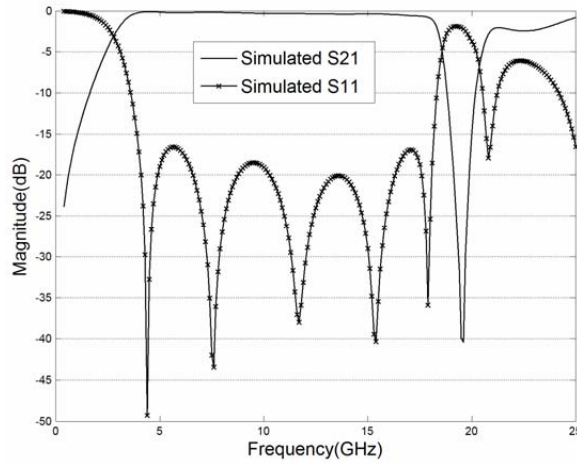
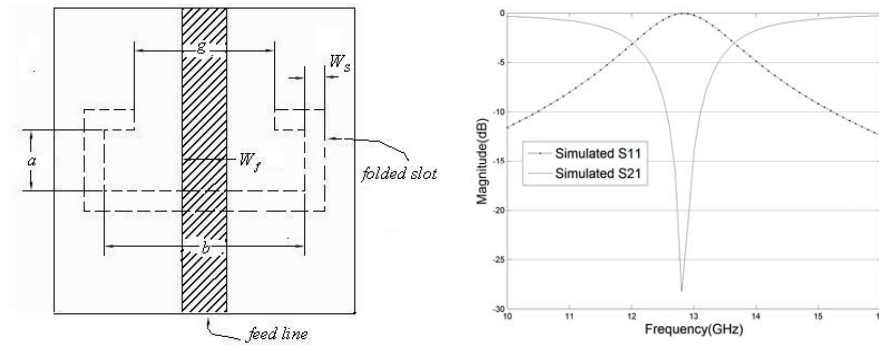**Figure 2.** Simulated S-parameters of the HPF.

Figure 3. DGS unit section with its simulated S-parameters. The dimensions $W_s = 10$ mils, $a = 80$ mils, $b = 220$ mils, $g = 100$ mils, $W_f = 2.5$ mm.

filter can be realized by cascading a specific number of such defects. The stopband widens in relation to the increased number of defects leading to a lowpass filter [7]. Fig. 3 shows a DGS unit and its simulated S-parameters. The substrate used for this simulation was *Rogers RT/duroid 5880* with the thickness of $h = 31$ mils and a dielectric constant of $\varepsilon_r = 2.2$.

The resonant frequency of the structure is determined by two factors shown in Fig. 3 (i.e., $a \times b$ and g). Variation of the resonant frequency versus these parameters (i.e., a, b and g) for different

substrates is investigated and the results for two dielectric constants, i.e., $\varepsilon_r = 2.2$ and $\varepsilon_r = 10.2$ with the thickness of $h = 20$ mils are shown in Fig. 4 and Fig. 5 respectively.

As we can see from Fig. 4 and Fig. 5, increasing the gap distance yields a higher resonant frequency and also a larger area of the T-shaped metal loading $(a \times b)$ causes a lower resonant frequency.

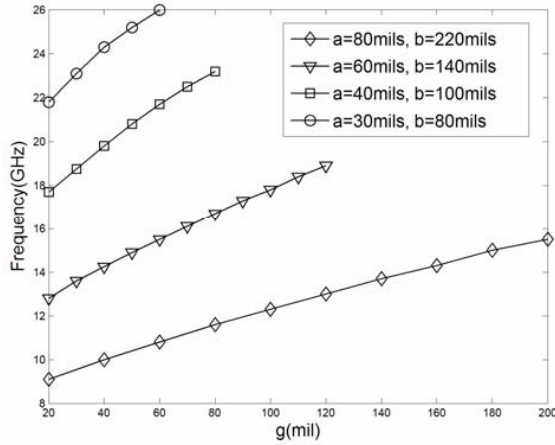It can be seen that, Fig. 4 and Fig. 5, cover the frequency range

Figure 4. Variations of resonant frequencies of the folded slot DGS versus a, b and g for $\varepsilon_r = 2.2$ and $W_S = 10$ mils.
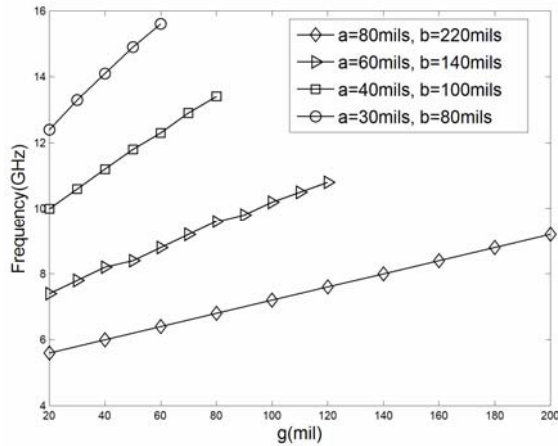
Figure 5. Variations of resonant frequencies of the folded slot DGS versus a, b and g for $\varepsilon_r = 10.2$ and $W_S = 10$ mils.

from 9 GHz to 26 GHz and 5.6 GHz to 16 GHz respectively. Because there is no considerable change on effective dielectric constant due to substrate thickness variations and so on the resonant frequency, the results of this effect are not included here.

The same procedure as above, has been made (for a substrate with $h = 31$ mils and a dielectric constant of $\varepsilon_r = 2.2$) to choose the dimensions of the DGS slots. Based on the resonant frequencies of various defects with different dimensions we adopted a non uniform DGS array with nine elements to realize a LPF.

Dimensions of the largest unit are chosen to fulfil the desired cutoff frequency. Here the desirable cutoff frequency is 10.6 GHz. Dimensions of the other sections have been chosen to cover the desired stopband. Note that each unit section adds a transmission zero at the stopband of the LPF and hence more sections provide more transmission zeroes and the resultant stopband widens. Since cascading unit sections perturbs the transmissions zeros of the folded slots, therefore with the use of Ansoft HFSS these dimensions have been optimized and adjusted to have the best performance in the stopband of the LPF. Finally the resultant scheme has been simulated using Ansoft Designer planar EM simulator for verification.

Periodicity of the DGS sections is $\lambda_g/4$ (λ_g is the guided-wavelength at the centre frequency in the stopband). Fig. 6 shows the schematic view of the proposed LPF and its simulated S-parameters. It can be seen from Fig. 6 all the ripples of return loss are lower than -14 dB in the passband. Also the proposed LPF has a wide -20 dB rejection in the stopband up to at least 22.7 GHz. The total length of the LPF is 28 mm.

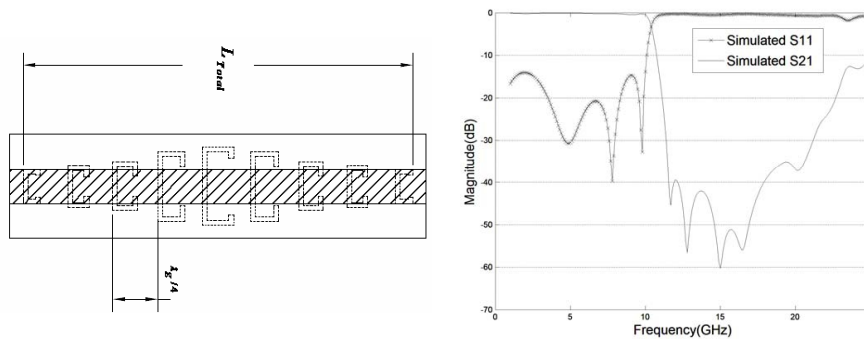At first no optimization is applied to the directly cascaded filter.

Figure 6. LPF with a non uniform DGS array with nine elements and its simulated S-parameters.
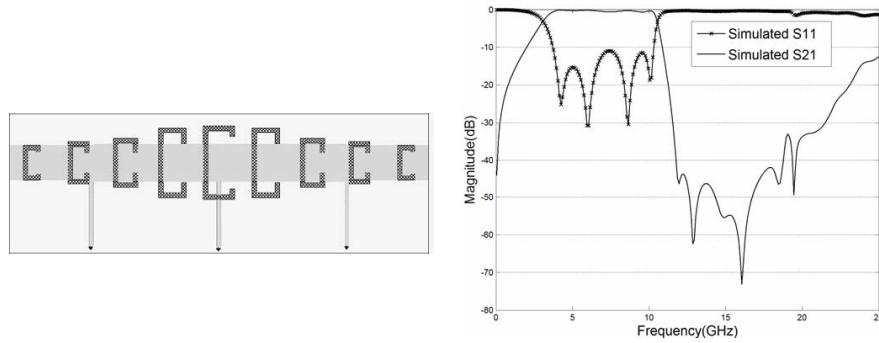
Figure 7. Proposed UWB bandpass filter together with its simulated *S*-parameters.

Simulations show that after cascading, the LPF and the HPF have a little influence on each others, i.e., the lower cutoff frequency of HPF had a 200 MHz shift in comparison with its counterpart without DGS sections in the ground plane. But the entire design shows a satisfactory bandpass characteristic over the specified frequency band. So the dimensions of the HPF had been adjusted to have a cutoff frequency of 2.9 GHz. Fig. 7 shows the layout of the proposed UWB bandpass filter together with its simulated *S*-parameters. It can be seen from Fig. 7 that the return loss in the pass band of the proposed UWB bandpass filter is better than -10 dB while the attenuation in the stopband is better than 20 dB up to 22.7 GHz. The total dimensions of this UWB BPF are 15 mm by 30 mm.

4. CONCLUSIONS

A new compact structure for microstrip bandpass filter suitable for UWB wireless communication is presented in this paper. The filter consists of a LPF and a HPF. The LPF is designed using DGS sections. Variations of the resonant frequency versus DGS parameters were investigated and the method to incorporate DGS units in designing the LPF was suggested. The HPF is realized by coupled short circuited stubs. The composite bandpass filter was simulated using EM simulators. Simulated *S*-parameters showed that the filter has a 3 dB bandwidth of 7.5 GHz (3.1–10.6 GHz) and the attenuation in the stopband is better than 20 dB up to 22.7 GHz.

ACKNOWLEDGMENT

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