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Compact Microstrip Spurline Bandstop Filter with Defected Ground Structure (Dgs)

Abhijeet Kumar[1], Prity Mishra[2], Savita Kadian[3]
[1]M.Tech Scholar, CBS Group of Institutions, Jhajjar
[2]Asst. Prof. ECE Deptt. , Ambedkar Institute of Advanced Communication Technologies and Research, Delhi
[3]Asst. Prof. ECE Deptt.,CBS Group of Institutions, Jhajjar

ABSTRACT
This paper presents a new structure to implement compact narrowband high-rejection microstrip band-stop filter (BSF). This structure is the combination of two traditional BSFs: Spurline filter and BSF using defected ground structure (DGS). Due to inherently compact characteristics of both Spurline and interdigital capacitance (used as DGS), the proposed filter shows a better rejection performance than Spurline filter and open stub conventional BSF without increasing the circuit size. From, the proposed BSF has a rejection of better than 20dB and the maximum rejection level of 41dB.
KEYWORDS- Microstrip, Bandstop filter (BSF), Spurline, Defected ground structures (DGS)

I. INTRODUCTION

Filters are essential component in any communication system. They are used to select or confine RF or microwave signals within assigned spectral limit so as to share the limited Electromagnetic spectrum. Emerging application in wireless communication demand RF microwave filter with even more stringent requirement: small size, lighter weight, low cost along with better performance. Depending on requirement and specification, RF and microwave filters are realized on various transmission lines such as waveguide, coaxial line and microstrip line. Development of compact filter using resonators in microstrip configurations has been discussed.

II. SPURLINE BANDSTOP FILTER

Most conventional microstrip bandstop filters have open-circuited stubs and shunt stubs of a quarter wavelengths. The conventional microstrip bandstop filter may be designed using a conventional procedure. The microstrip bandstop filter using open circuited stubs consists of one main transmission line coupled to a half-wavelength resonator; it is generally electrically and magnetically coupled [5]. The resonant frequency depends on the half-wavelength resonator. However, the open-circuited stubs are large. The microstrip bandstop filter with the shunt stubs of a quarter wavelengths still must be larger than the quarter-wavelength resonator. Therefore, reducing the size of such filters is essential. In 1970, Alley proposed interdigital capacitors for use in lumped-element microwave integrated circuits [11]. The microstrip interdigital resonator is analogous to the planar interdigital capacitor, whose operating frequency can be changed by control of the capacitances. Hence, an interdigital resonator is described herein to realize a lumped element interdigital bandstop filter. The interdigital capacitance then can be referred to a periodic structure of transmission line. Moreover, the admittance matrix may be calculated from each pair of interdigital coupling lines by even- and odd-mode analysis.

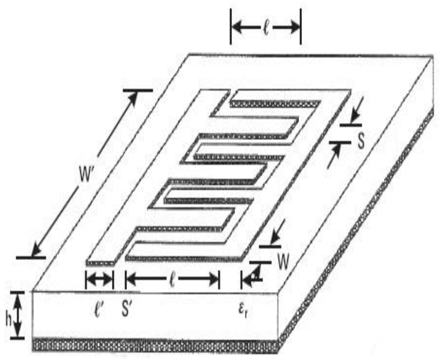

Fig.1 Structure of the novel bandstop filter with one microstrip interdigital resonator and tapped I/O.

The analyses were based on lossless microstrip coupled lines and lossy coupled microstrip lines. A more accurate characterization of these capacitors can be performed if the capacitor geometry is divided into basic microstrip sections such as the single microstrip line, coupled microstrip lines, open-end discontinuity, unsymmetrical gap, 90° bend, and T-junction discontinuities as shown in figure 2. This method provides an approximate solution, due to several assumptions in the grouping of subsections and does not include interaction effects between the basic microstrip sections. Figure.3 shows a simple EC model used to describe the characteristics of the interdigital capacitors shown in Figure1.

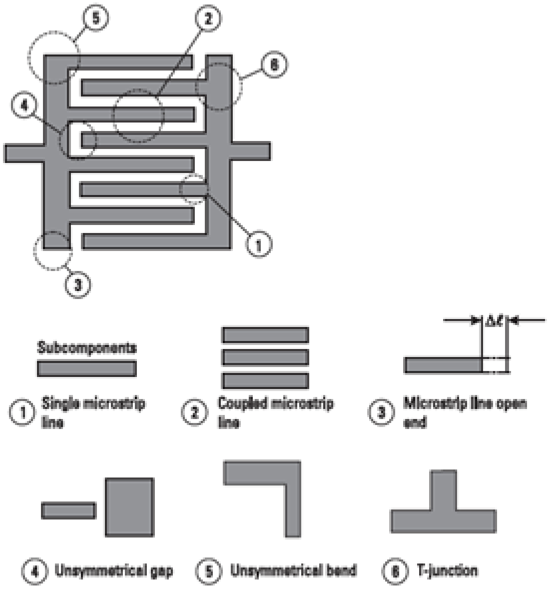

Fig.2 The interdigitated capacitor and its subcomponents.

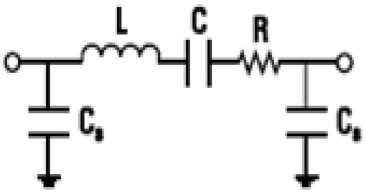

Fig.3 Lumped-element EC models of the interdigital capacitor low frequency

An approximate expression for the interdigital capacitance is given by

$$C = \frac{\epsilon_r + 1}{W'} \ell \left[(N-3)A_1 + A_2\right]$$

Where C is the capacitance per unit length along W', $A1$ (the interior) and $A2$ (the two exterior) are the capacitances per unit length of the fingers, N is the number of fingers, and the dimensions W' and l, are shown in Fig.1 and expressed in microns. For infinite substrate thickness (or no ground plane), $A1$ = 4.409×10^6 pF/mm and $A2$ = 9.92 ×10^6pF/mm. The total capacitance of an interdigital structure of length l, is expressed as

$$C = (\epsilon_r + 1)\ell\left[(N-3)A_1 + A_2\right] \quad \text{(pF)}$$

For a finite substrate, the effect of h must be included in $A1$ and $A2$. In the final design, usually $S = W$ and, l = λ /4. Approximate expressions for $A1$and $A2$ are obtained by curve fitting the data already given.

These expressions are as follows:

$$A_1 = 4.409 \tanh\left[0.55\left(\frac{h}{W}\right)^{0.45}\right] \times 10^{-6} \quad \text{(pF/}\mu\text{m)}$$

$$A_2 = 9.92 \tanh\left[0.52\left(\frac{h}{W}\right)^{0.5}\right] \times 10^{-6} \quad \text{(pF/}\mu\text{m)}$$

The series resistance of the interdigital capacitor is given by

$$R = \frac{4}{3}\frac{\ell}{WN} R_s$$

Where Rs is the sheet resistivity in ohms per square of the conductors used in the capacitors.
The effect of metal thickness t plays a secondary role in the calculation of capacitance. The Q of this capacitor is given by

$$Q_c = \frac{1}{\omega CR} = \frac{3WN}{\omega C 4 \ell R_s}$$

The capacitance Cs and inductance L are approximately calculated on the basis that for $S/h <<$ 1, magnetic field lines do not loop around each finger but around the cross section of the interdigital width W' (Fig.1). Under this assumption L and Cs are calculated from microstrip transmission-line theory using, as the length of the structure. Here Cs is half of the microstrip total shunt capacitance. Expressions for these quantities are given next:

$$L = \frac{Z_0\sqrt{\epsilon_{re}}}{c}\ell$$

$$C_s = \frac{1}{2}\frac{\sqrt{\epsilon_{re}}}{Z_0 c}\ell$$

Where Zo and εre are calculated using W' and h microstrip parameters and c=3×10^10 cm/s is the velocity of light in free space.

A general expression for the total series capacitance of an interdigital capacitor can also be written as

$$C = 2\epsilon_0\epsilon_{re}\frac{K(k)}{K'(k)}(N-1)\ell \quad \text{(F)}$$

$$= \frac{10^{-11}}{18\pi}\epsilon_{re}\frac{K(k)}{K'(k)}(N-1)\ell \times 10^{-4} \quad \text{(F)}$$

Or

$$C = \frac{\epsilon_{re}10^{-3}}{18\pi}\frac{K(k)}{K'(k)}(N-1)\ell \quad \text{(pF)}$$

Where, is in microns, N is the number of fingers, and εre is the effective dielectric constant of the microstrip line of width *W*. The ratio of complete elliptic integral of first kind *K* (*k*) and its complement *K* '(*k*) is given by

$$\frac{K(k)}{K'(k)} = \begin{cases} \frac{1}{\pi}\ln\left\{2\frac{1+\sqrt{k}}{1-\sqrt{k}}\right\} & \text{for } 0.707 \le k \le 1 \\ \frac{\pi}{\ln\left[2\frac{1+\sqrt{k'}}{1-\sqrt{k'}}\right]} & \text{for } 0 \le k \le 0.707 \end{cases}$$

And

$$k = \tan^2\left(\frac{a\pi}{4b}\right), \; a = W/2, \text{ and } b = (W+S)/2, \text{ and } k' = \sqrt{1-k^2}$$

III. SIMULATION RESULTS

In previous section we discussed the analysis and design equation of microstrip spurline and interdigital defected ground structure bandstop filter. In this section we will briefly see the software used for simulation of our structure and the best results.

Basically for verification of results of previous research paper and our structure's software stimulation, we used HFSS v10 for generating our structure's simulation and their respective comparison. HFSS is a high-performance full-wave electromagnetic (EM) field simulator for arbitrary 3D volumetric passive device modeling that takes advantage of the familiar Microsoft Windows graphical user interface. It integrates simulation, visualization, solid modeling, and automation in an easy-to-learn environment where solutions to your 3D EM problems are quickly and accurately obtained. Ansoft HFSS employs the Finite Element Method (FEM), adaptive meshing, and brilliant graphics to give you unparalleled performance and insight to all of your 3D EM problems. Ansoft HFSS can be used to calculate parameters such as S Parameters, Resonant Frequency, and Fields. The simulation technique used to calculate the full 3D electromagnetic field inside a structure is based on the finite element method. Although its implementation is largely transparent, a general understanding of the method is useful in making the most effective use of HFSS. It provides an overview of the finite element method and its implementation in HFSS. They also describe how modal S-parameters are computed from the simulated electric and magnetic fields and how they can be converted to "nodal" or "voltage" based pseudo-S-parameters used in circuit theory.

3.1 Structure design and result of spurline bandstop filter at 6 GHz

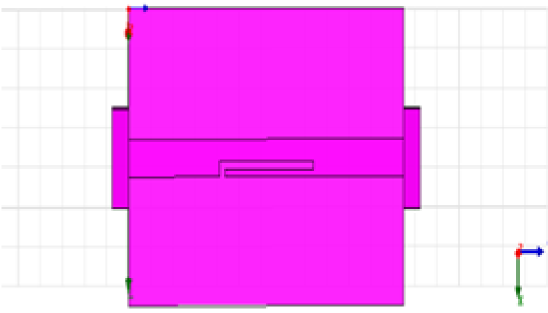

Fig.3 Structure design of spurline BSF at 6GHz

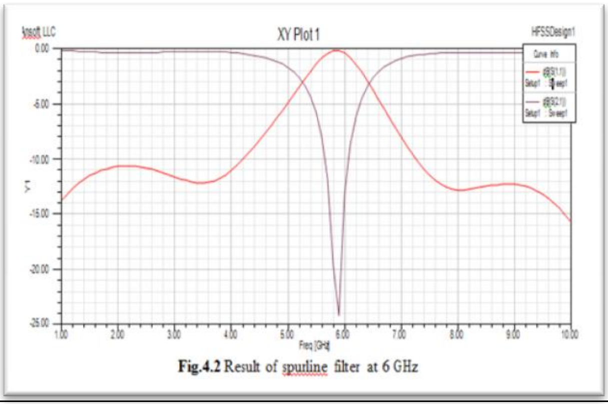

Fig.4 Result of spurline filter at 6 GHz

3.2 Structure design and result of interdigital capacitance bandstop filter at 6 GHz

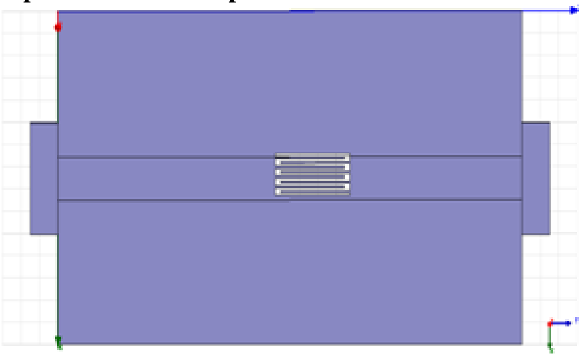

Fig.5 Structure design of interdigital capacitance bandstop filter at 6 GHz

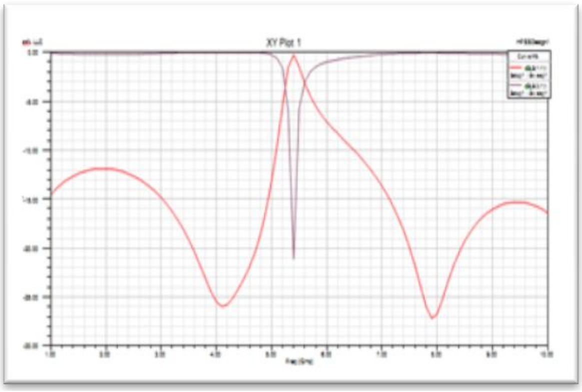

Fig.6 Result of interdigital capacitance bandstop filter at 6 GHz

3.3 Structure design and result of proposed structure at 6 GHz

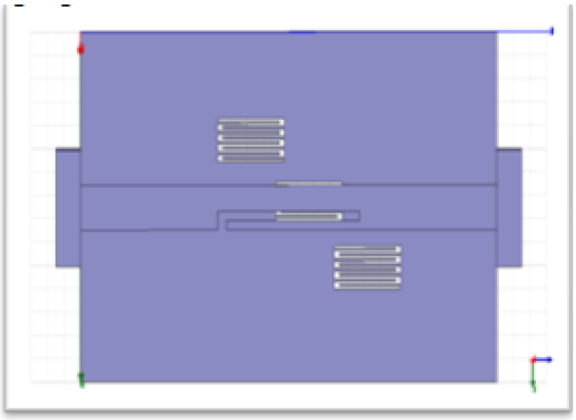

Fig.7 Structure design of proposed structure at 6 GHz

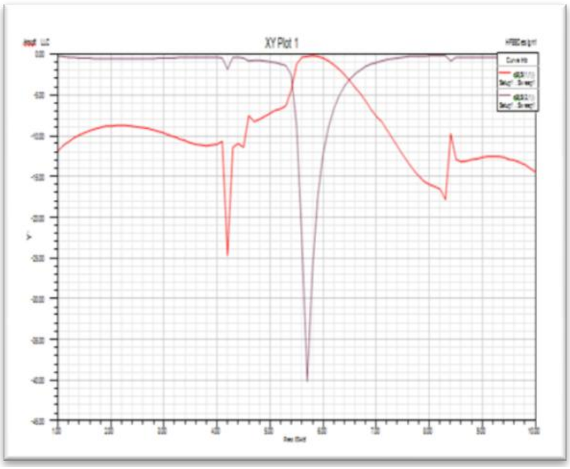

Fig.8 Result of Proposed structure at 6 GHz

IV. COMPARISON

Table below have shown the graph of comparison of Spurline bandstop filter, interdigital bandstop filter and proposed structure i.e. spurline with interdigital defected ground structure bandstop filter results.

Structure	Frequency	S21(dB)
Spurline	6 GHz	**-24**
Interdigital	6 GHz	**-20**
Proposed	6 GHz	**-40**

V. CONCLUSION

The aim of this paper is the analysis of Compact Microstrip Spurline Bandstop Filter and interdigital Defected Ground Structure (DGS) bandstop (notch) filter at designed frequency and its effect with and without diagonal coupling between interdigital structures. In this paper, lossless conditions are taken into account and the substrate used is RT duroid 3003 with dielectric constant 3 with the thickness of 30mil (0.762mm) and structure calculations have got simulated results with very good attenuation. Proposed structure consist of combination of spurline bandstop filter with interdigital Capacitance as defected ground structure (DGS) from which a sharpness of the roll-off response can be achieved, which is important in wireless communication systems. With proposed structure we can achieve sharp and steep rejection notch. We successfully reduced the radiation losses, dielectric losses, lowered the extra cost due to defected structure and also provide adjustable rejection band. We can achieve bandstop notch tunability at desired low frequency by changing the length of interdigital capacitance. Thus with reduced size and improved rejection notch and bandstop filtering performance and comparison to typical Spurline filter without DGS.

VI. FUTURE SCOPE

The work can be extended i.e. future scope of this is different shapes of spur line with different configurations can be further design and analyzed. Fabrication with active devices can be done to give tunability to the circuit to reduce the structure length.

VII. ACKNOWLEDGEMENT

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