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PERFORMANCE ANALYSIS OF RESONATORS IN MICROSTRIP BANDPASS FILTERS

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Abstract: Filters are substantial microwave components. RF/Microwave filters can be implemented using transmission lines. In this paper microstrip bandpass filters had been designed for RF/microwave applications. Some novel techniques like implementing the open resonators, open split square loop resonators, and star shaped multi-mode resonators are implemented in designing the microstrip bandpass filters. Microstrip filters are used in this report to design bandpass filters because of their compact sizes. The goal of this thesis is to investigate on some planar microwave bandpass filters. In this simulation 5GHz Frequency used to find the Result and this simulation done in HFSS Software in the result gain for S11, S12,S21,S22 find and Coupling coefficients and external Q-factor of tunable BPF also find in the simulation.

Index Terms - **Resonators, Microstrip, Bandpass Filters, antenna, gain, HFSS**

I. INTRODUCTION

Microwave filters are an important component in front end receiver of microwave communication system. They are used for passing frequency components within a particular pass band and to reject the interfering signal outside of that operating frequency band. Their functionality also includes to reject the unwanted product from the output of the mixers and amplifiers, and to set the IF bandwidth of the IF receiver. Important parameters include cutoff frequency, insertion loss, out of the band attenuation rate measured in dB per decade of the frequency [1]. Filters with sharper cutoff frequency provide more rejection for out of the band signals. Insertion loss, measured in dB, is the amount of attenuation seen by signal through the pass band of the filter. Another important consideration is the size. Much of the front end circuitry today can be integrated monolithically form in IC's packages in the range of 800 MHz to 2 GHz. Now days however, It is not possible to construct Filters in IC's form for high performance, since inherent loss of the RF and microwave IC's leads to high insertion loss and high out of band attenuation rates. For this reason, Wireless systems today use individually "off chip" filter that are located outside of the circuit board, rather than IC packaging. Size requirement for most of the PCS (Personal Communication Systems) require filters to be compact in size and narrowband. The use of periodic structures combined with microwave designs has become very common during the last few years, although the basis of some of these structures were already defined some decades ago, such as the case of "soft/hard" surfaces. These surfaces are periodic repetitions of perfect electric and magnetic conductors (PEC/PMC) which produce an enhancement of the transmitted power in one of the directions (named "hard"), and a band gap in the orthogonal one (named "soft")[1]. However, although the implementation of the perfect electric conductor is easy at microwave frequencies with the use of metallic materials, the magnetic one has to be realized with artificial magnetic conductors (AMC) which limits the bandwidth of operation of these surfaces [2]. More recent investigations have established the limits in terms of bandwidth and operation frequency of these structures in their different implementations (corrugations and microstrip technology).Corrugations have been used to suppress surface wave propagation in predefined directions, for different purposes such as mutual coupling or back radiation reduction. Particularly, horizontal corrugations have been demonstrated to produce a significant size reduction defining ribbed electrical paths from the short-circuit to the open condition [3] and some examples of how to reduce its size have been proposed .However, when the size of the unit cells is reduced, the Q factor is increased, and as a consequence, the bandwidth of operation is narrower .A possible solution to increase the operating frequency range is to make the device reconfigurable. Thereby, when an external parameter is modified (i.e., an external voltage), the structure can work at different bands, achieving a frequency reconfigurable stop-band [4], which is a satisfactory solution to the problem of narrow bandwidth for many applications. On the other hand, for multifunctional devices and for the development of smart receivers [5] all parts of the device must be reconfigurable. In this case, reconfigurability should be also applied to the corrugations

II. Microwave Communication

The electromagnetic waves are the waves whose frequency ranges from 300 MHz - 300 GHz, these range of frequencies are referred as microwaves. The wavelength of this waves in free space is about 1 m – 1 mm. The electromagnetic spectrum is shown in figure 1.1, it demonstrates schematically the electromagnetic spectrum. Further some selected frequency spectrums are allocated into many frequency bands as betokened in Table 1.1. Frequency boundaries between RF and microwave are almost arbitrary. The boundary rely on the specific technologies established for the utilization of that particular frequency range. The applications that use RF/microwave frequency ranges are communications, remote sensing and many more[6].

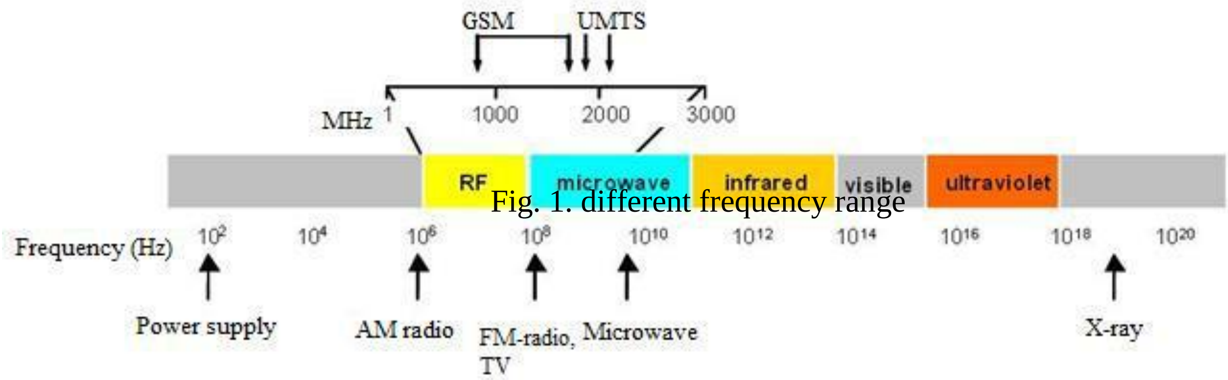


Figure 1. different frequency range

A filter is used to regulate the frequency response at a fixed point in the EM spectrum by providing low loss transmission at the preferred frequency band and high attenuation at remaining frequencies. Filters are extensively used in many applications like communications, remote sensing, radars etc. A filter is generally a two-port network.

III. Role of filters in microwave communication

Filters are essential in separating and sorting signals in communication systems. To cull or confine the RF/microwave signals within given spectral limits, filters are used. The role of filters in communication systems is to usually transmit and receive amplitude and/or phase modulated signals through a communication channel. To get rid of or suppress spurious frequencies from being transmitted or received in radio transmitters and receivers, filters are used. Evolving applications such as wireless communication remains to challenge RF/microwave filters with even more rigid requirements like smaller size, lighter weight and lower cost with better performance. Filters used in communication and radar applications, are implemented in different kinds of transmission lines comprising strip line, rectangular waveguide, and microstrip. Filters are also the integral part of multiplexers which are of major demand in the BroadBand wireless access communication systems[7].

IV. Microwave filters in cellular communication

Microwave filters are very important components in cellular systems where stringent filter specifications are required both on the mobile station and base station levels. All modern full duplex personal communications systems require transmit and receive filters for each transceiver unit at least at the base station level. Transmit filters must be very selective to prevent out of band inter-modulation interference to satisfy regulatory requirements as well as prevent adjacent channel interference. Acceptable levels of adjacent channel interference in TDMA second generation mobiles are specified in GSM ETSI standards as C/A > -9dB. In practice a C/A of -6 dB is used in the network design. Also, the transmit filters must have low insertion loss to satisfy efficiency requirements. A typical transmit filter contains a return loss of 20dB and passband insertion loss of 0.8 dB. It is obvious that the technology used in filter realization in base stations is significantly different from that used in handsets. Although the filter specifications in handsets are less stringent due to lower power handling (33dBm maximum transmit power), size requirements remain a challenging task. One of the main difficulties is parasitic or unwanted coupling that is caused by the close proximity of the resonators. Another application of filters is in cellular systems microwave links to connect base stations to BSC (base station controller) and then to the MSC (Mobile Switching Center). These are high-speed links with directive dish antennas. There are few licensed bands for transmission such as 8, 11, 18, 23, 24 and 38 GHz. The choice of the frequency band depends on spectrum availability, length of the hop and required link reliability. Filters for transmission systems are usually constructed using waveguide technology due to the high quality factors requirements and high power handling capabilities.

V. RELATED WORK

Despite the extensive literature in the field of microwave filters, several issues are still either not well understood or lack a systematic solution or accurate design procedure. For instance, one of the major difficulties with miniaturized and compact filters is parasitic coupling that can be in the order of the main coupling in compact filters. This makes it difficult to identify a sparse topology on which most of design and optimization methods are based. It is sometimes impossible to predict the behavior of the filter when using a conventional coupling topology based on the arrangement of physical resonators. Most importantly, the absence of a reliable circuit model that represents compact filters makes the optimization of this class of filters, within efficient and systematic techniques such as space mapping technique, impossible. The same argument holds for wideband filters that cannot be represented by the conventional low-pass prototypes that are based on a narrowband approximation.

One of the major difficulties with microwave filter design is the absence of a generic design technique to transform the low-pass prototype into physical dimensions except for few cases as in [5, 6]. A careful investigation of the available literature reveals that this is due to basing most of the design techniques unswervingly on the phenomenon of resonance. The dependence of the resonant frequencies of the resonators on the coupling strength (loading) is not systematically accounted for in the model. It is indeed shown that by using the phenomenon of propagation instead of resonance, it becomes straightforward to account for the piling of the resonances by the coupling components as in in-line direct-coupled resonator filters [3].

Despite the widespread use of dual-mode filters in satellite communications, the design and realization of this class of filters is not straightforward. The concept of dual-mode filters was anticipated by Atia and Williams in the 1970's [5-10]. In general, filters designed according to this theory require extensive optimization. Tuning elements are used as part of the CAD design as well as in compensating for inherent manufacturing errors. The resulting designs are very time-consuming, labor intensive, costly and at times extremely sensitive. A satisfactory solution to this problem is not known. The main goal of this thesis is to find new techniques for the design of bandpass microstrip filters. Detailed investigation of the relevance of equivalent circuits used to represent the filter response to the field theory is carried out in detail. The new view is exploited to formulate new design and implementation methods for microwave filters.

Hanyue Xu et.al (2021) this article presents novel designs of reconfigurable microwave filters based on the recently introduced concept of field programmable microwave substrate (FPMS). Using reconfigurable FPMS substrate, significant tuning freedom of the designed microwave filters can be achieved. It allows implementation of smart filter designs with tunable center frequencies, operational bandwidths, and filter orders. Moreover, different microwave filter types are also realizable on one FPMS board with simple dynamic control. As a proof-of-concept, design steps are described for three filter types, including waveguide bandpass filter, quarter-wave-coupled bandpass filter, and waveguide bandstop filter. A good match between the simulated and measured results is presented, showing good tuning range for the center frequency and bandwidth of the bandpass filters around 2 GHz. In addition, the reconfigurability of the design allows it to switch to bandstop filters, which is a complete change in topology from a single physical design. The merits of the proposed design are reflected in the realized bandpass filter, exhibiting center frequency tuning of more than 20% along with bandwidth variability spanning three times its minimum value. Finally, a significant size reduction compared with the designs using conventional technologies is also demonstrated. With its inherent flexibility, low cost, and high degree of integration, FPMS filtering is suited to a wide variety of RF applications.

HaiYang Wang et.a; (2020) Cavity filter is widely used in microwave receiver front-end subsystem for the advantage of its low insertion loss, excellent band selectivity, and relative high power handle characteristics. However, the short high power microwave pulse transmission response in cavity filter is seldom studied, especially under high power microwave (HPM) electromagnetic environment. By the Particle-In-Cell/Monte Carlo Collision simulation and high power tests, the transient microwave breakdown phenomenon of an S band comb-line cavity filter are discussed. The simulation and test results show that when the input microwave pulse width is 50ns, the microwave breakdown power threshold value is about 500W. And the filter output would be nearly suppressed while the input power above 3.5kW. The simulation data of different microwave pulse widths demonstrate that the breakdown power threshold is approximately inversely proportional to the microwave pulse width in case of the pulse width is lower than 100ns.

Rong Cong et.al (2019) A notch/bandpass microwave photonic filter (MPF) with tunable frequency based on a micro ring resonator (MRR) and a LiNbO3 phase modulator is proposed and experimentally demonstrated. By adjusting the polarization controller (PC) between the PM and the MRR, the MPF can switch between a bandpass filter and a notch filter. The frequency of the MPF can be tuned by changing the wavelength spacing between the optical carrier and the resonator wavelength of the MMR. A proof-of-concept experiment is carried out. The notch/bandpass filter with a rejection ratio of 27 dB/15 dB can be achieved during a frequency tuning range from 9 GHz to 18.5 GHz.

VI. PROPOSED DESIGN

The main motive of this research to designed Hairpin Short-ended Resonators. A symmetrical hairpin short-ended resonators with different connection of the capacitance to their middle part. The dashed line through the middle of the resonators will be divided into two equal upper and lower halves. Transmission matrix of the lower half of the resonators will be take [A′B′C′D′] and the transmission matrix of a distributed circuit of the same half we denote [a′b′c′d′]. Resonance in the symmetrical resonators under consideration .when the voltage in the middle of the resonator reaches the maximum or minimum values. The case is considered when a parallel or series variable capacitance is connected to the middle of these resonators. It is established that odd resonance frequencies of such resonators do not depend on the variable capacitance value and they are constant. Two traditional symmetrical open-end hairpin resonators, to which one or two variable capacitors are connected, are also considered. New resonance equations of the hairpin resonator with two capacitors have been obtained, where a segment of uniform transmission line is replaced by a segment of non-uniform transmission line Loop hairpin and combine resonators provide a wide tuning range and a wide rejection band for tunable BPFs. Therefore, the set of resonators used in tunable BPFs need to expanded. For this purpose, it is advisable to investigate two short-ended hairpin resonators and a parallel or series variable capacitance connected to their middle part.

The main feature of the designed given below

- Design: the dimensions and designs will be refer and implement on HFSS software
- Two or more will be designed (due to slow run of software HFSS, minimum two will be there)

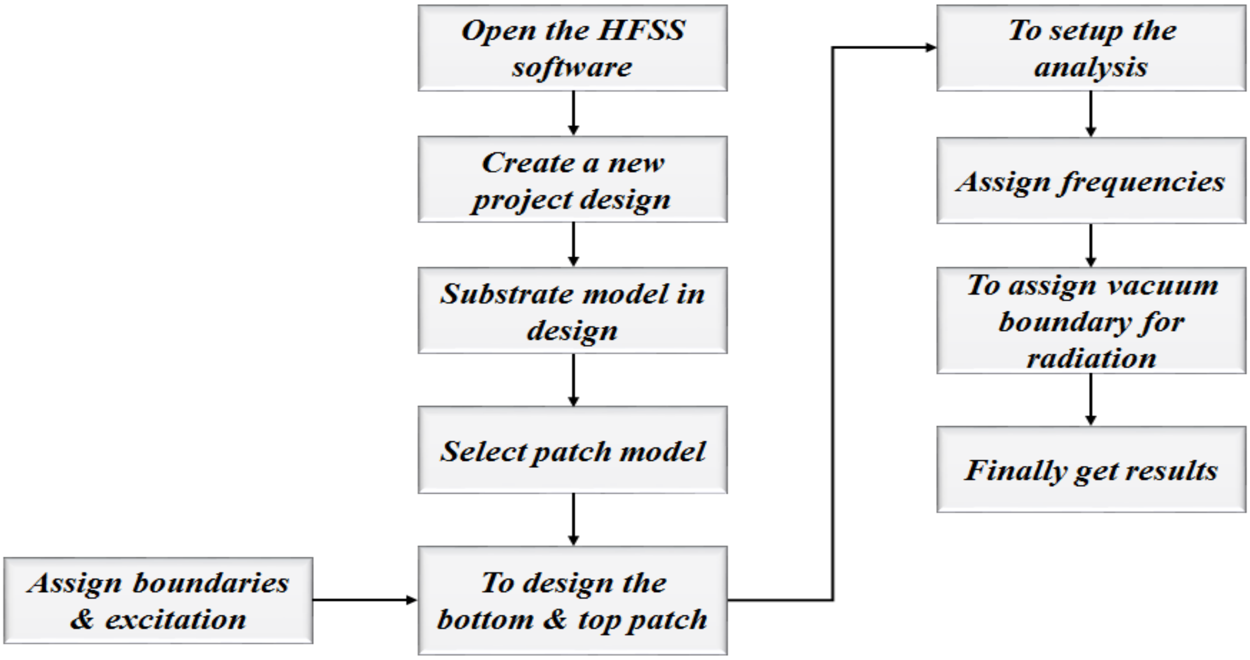


Fig. 2. Flow Diagram

VII. ANTENNA CALCULATION:

Step 1: Calculation of the Width (W)

$$W = \frac{c}{2f_o\sqrt{\frac{\varepsilon_r+1}{2}}} \qquad \text{Equation 1}$$

Step 2: Calculation of the Effective Dielectric Constant. This is based on the height, dielectric constant of the dielectric and the calculated width of the patch antenna.

$$\varepsilon_{eff} = \frac{\varepsilon_r+1}{2} + \frac{\varepsilon_r-1}{2}\left[1+12\frac{h}{W}\right]^{-1/2} \qquad \text{Equation 2}$$

Step 3: Calculation of the Effective length

$$L_{eff} = \frac{c}{2f_o\sqrt{\varepsilon_{eff}}} \qquad \text{Equation 3}$$

Step 4: Calculation of the length extension ΔL

$$\Delta L = 0.412h\frac{(\varepsilon_{eff}+0.3)(\frac{W}{h}+0.264)}{(\varepsilon_{eff}-0.258)(\frac{W}{h}+0.8)} \qquad \text{Equation 4}$$

Step 5: Calculation of actual length of the patch

$$L = L_{eff} - 2\,\Delta L \qquad \text{Equation 5}$$

Properties

Name	Value	Unit	Evaluated Value
Name	Infinite Sphere1		
Type	Infinite Sphere		
Start Theta	0	deg	0deg
Stop Theta	180	deg	180deg
Theta step	10	deg	10deg
Start Phi	0	deg	0deg
Stop Phi	360	deg	360deg
Phi Step	10	deg	10deg

Table 1 Properties

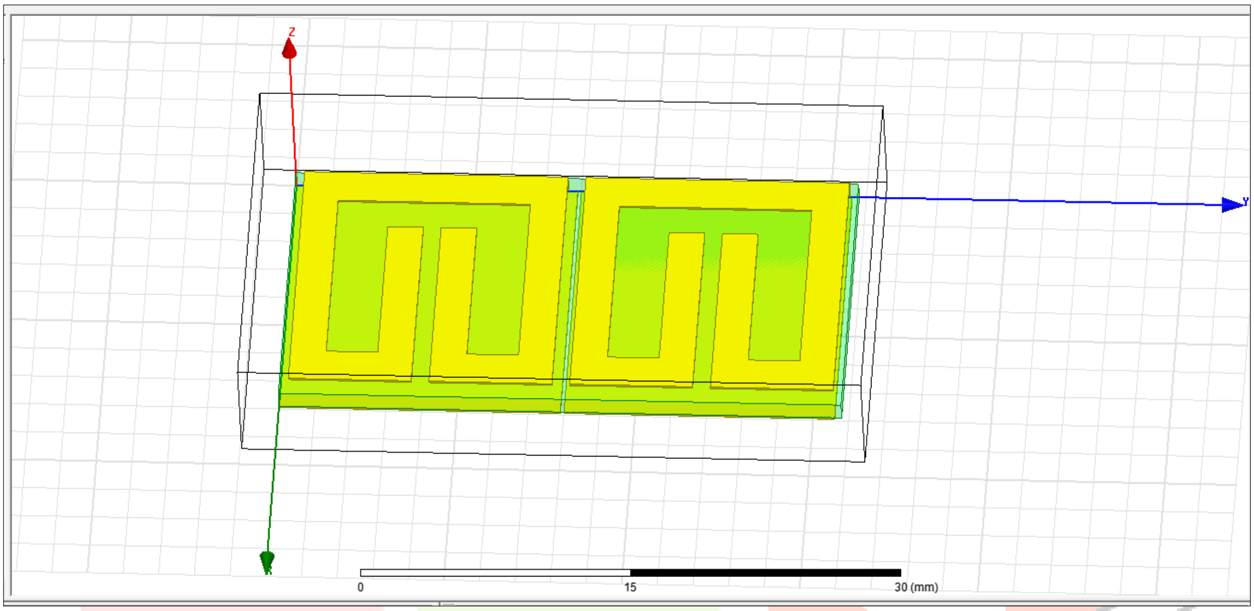


Fig. 3. Antenna plane patch

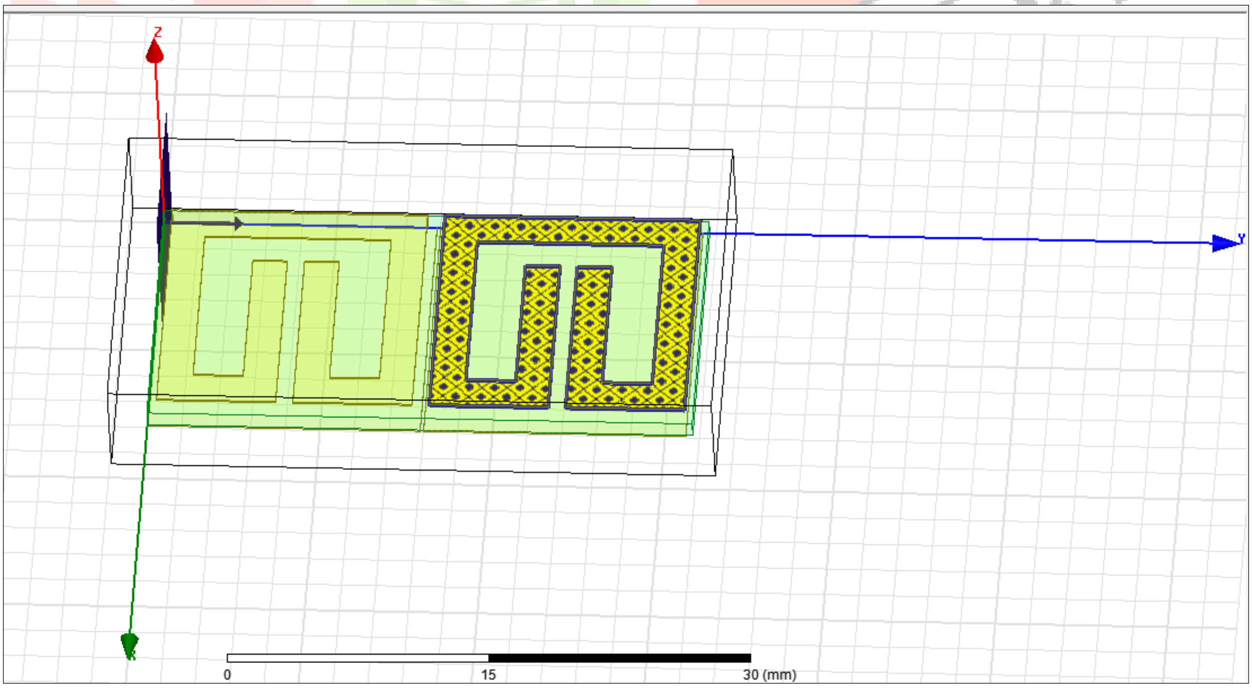


Fig. 4. Antenna plane patch E2

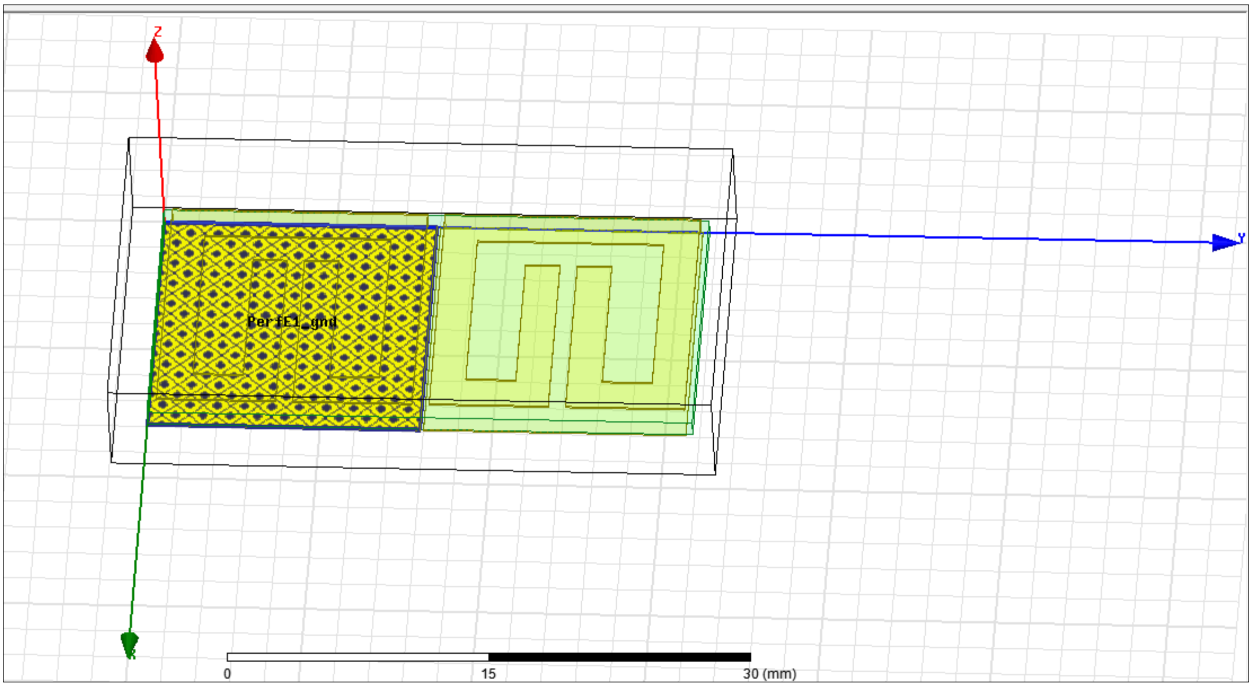


Fig. 5. Antenna Ground patch E1

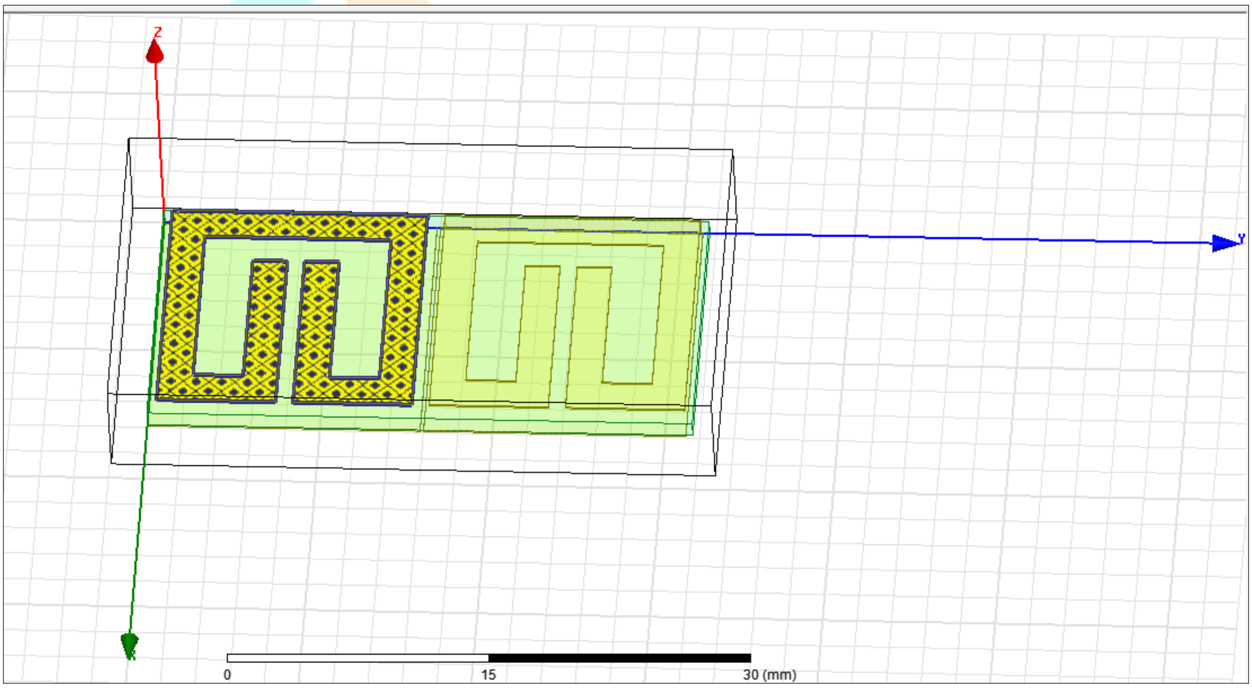


Fig. 6. Antenna plane patch E1

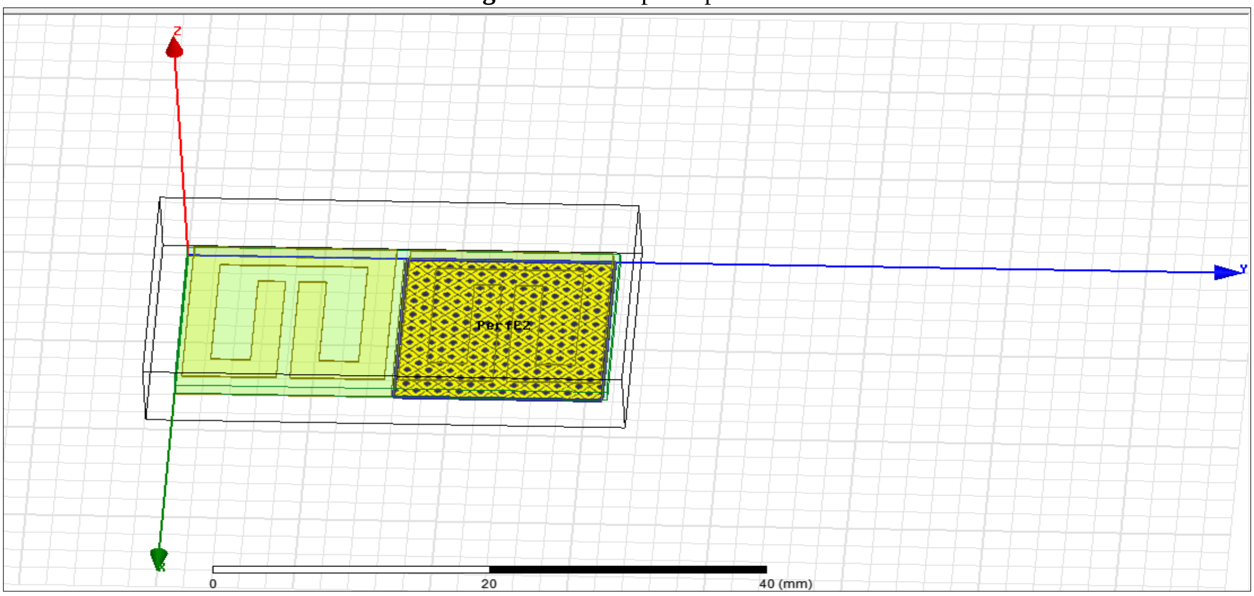


Fig. 7 Antenna Ground patch E2

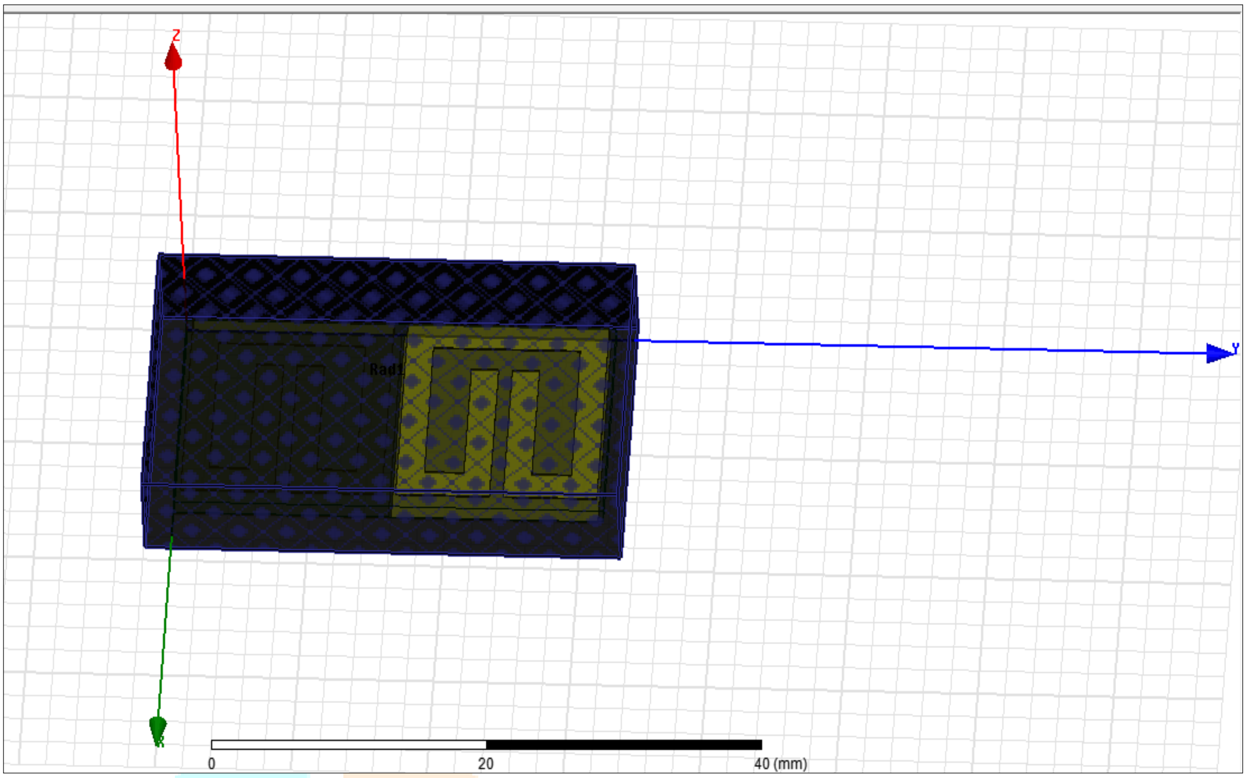


Fig. 8 Antenna Ground patch Radiation

S-parameters describe the input-output relationship between ports (or terminals) in an electrical system. For instance, if we have 2 ports (intelligently called Port 1 and Port 2), then S12 represents the power transferred from Port 2 to Port 1. S21 represents the power transferred from Port 1 to Port 2. In general, SNM represents the power transferred from Port M to Port N in a multi-port network. A port can be loosely defined as any place where we can deliver voltage and current. So, if we have a communication system with two radios (radio 1 and radio 2), then the radio terminals (which deliver power to the two antennas) would be the two ports. S11 then would be the reflected power radio 1 is trying to deliver to antenna 1. S22 would be the reflected power radio 2 is attempting to deliver to antenna 2. And S12 is the power from radio 2 that is delivered through antenna 1 to radio 1. Note that in general S-parameters are a function of frequency (i.e. vary with frequency).the following two-port network:

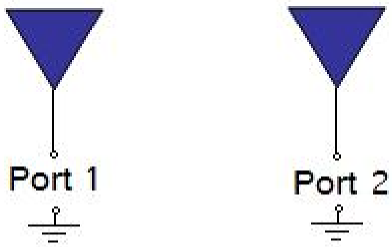


Fig. 9. two-port network of antenna elements

In the above Figure, S21 represents the power received at antenna 2 relative to the power input to antenna 1. For instance, S21=0 dB implies that all the power delivered to antenna 1 ends up at the terminals of antenna 2. If S21=-10 dB, then if 1 Watt (or 0 dB) is delivered to antenna 1, then -10 dB (0.1 Watts) of power is received at antenna 2.

If an amplifier exists in the circuitry, then S21 can show gain (i.e. S21 > 0 dB). This means that for 1 W of power delivered to Port 1, more than 1 W of power is received at Port 2. as a final consideration, since the unit cell has two branches, it is possible to modify only one of these capacitances with a single varactor diode. In that situation, there is a variation of the stop band, but the notch is maintained at the initial band, i.e., obtaining a dual band filter or an increased bandwidth if the two bands are very close to each other. S11 represents how much power is reflected from the antenna, and hence is known as the reflection coefficient (sometimes written as gamma: or return loss. If S11=0 dB, then all the power is reflected from the antenna and nothing is radiated. If S11=-10 dB, this implies that if 3 dB of power is delivered to the antenna, -7 dB is the reflected power. The remainder of the power was "accepted by" or delivered to the antenna. This accepted power is either radiated or absorbed as losses within the antenna. Since antennas are typically designed to be low loss, ideally the majority of the power delivered to the antenna is radiated.

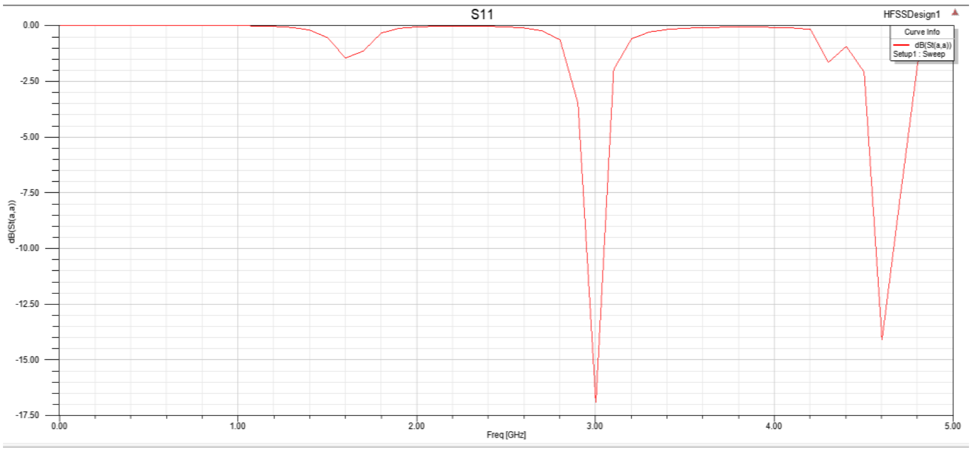


Fig. 10: S11=-10 dB, antenna radiates best at 2.5 GHz

The above figure implies that the antenna radiates best at 2.5 GHz, where S11=-10 dB. S11 is the input port voltage reflection coefficient. S12 is the reverse voltage gain. S21 is the forward voltage gain. S22 is the output port voltage reflection coefficient. The S-parameter matrix can be used to determine reflection coefficients and transmission gains from both sides of a two port network.

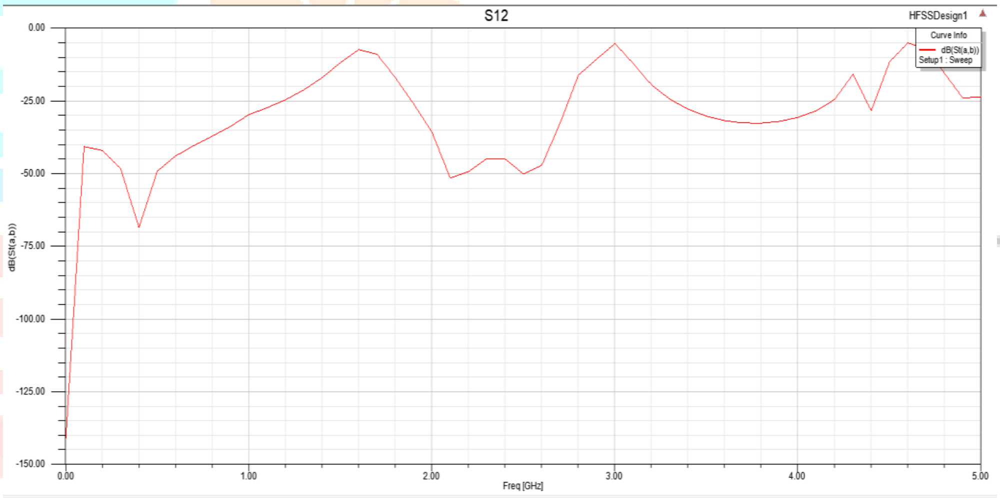


Fig. 11. S11=-10 dB, antenna radiates best at 2.5 GHz

The above figure implies that the antenna radiates best at 2.5 GHz, where S11=-10 dB

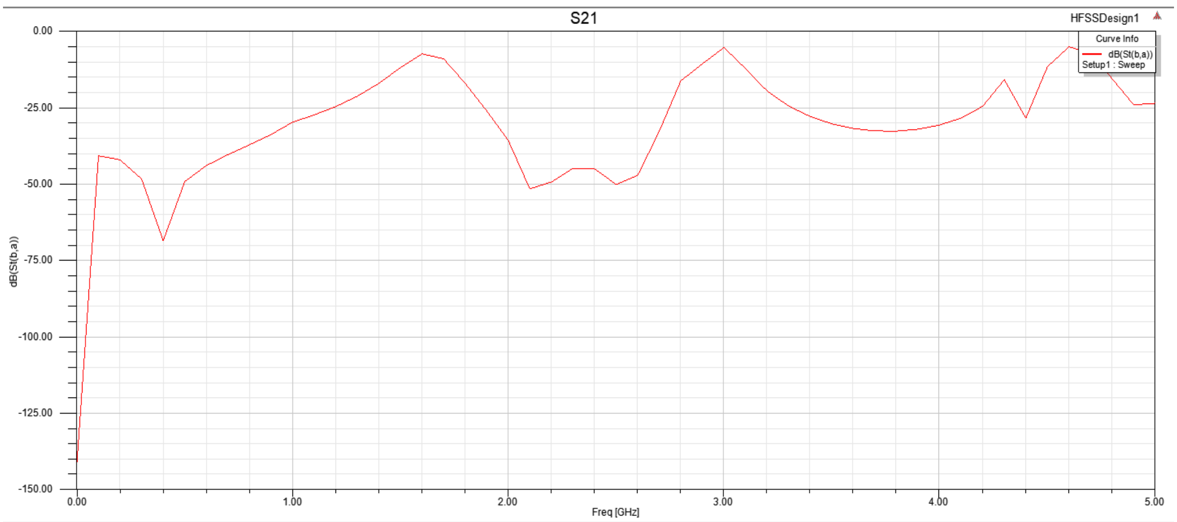


Fig.12. S11=-10 dB, antenna radiates best at 2.5 GHz

The above figure implies that the antenna radiates best at 2.5 GHz, where S11=-10dB. S21 is the forward transmission (from port 1 to port 2), S12 the reverse transmission (from port 2 to port 1) and S22 the output reflection coefficient.

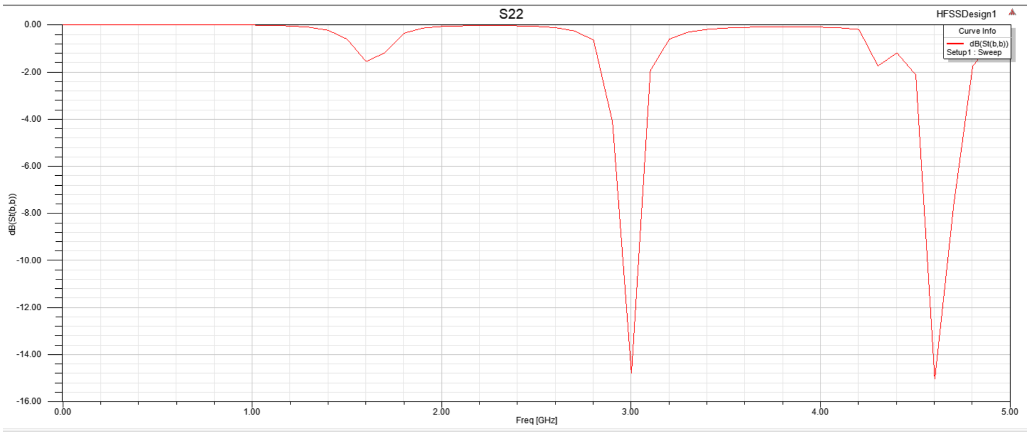


Fig. 13 S11=-50 dB, antenna radiates best at 2.5 GHz

The above figure implies that the antenna radiates best at 2.5 GHz, where S11=-50 dB

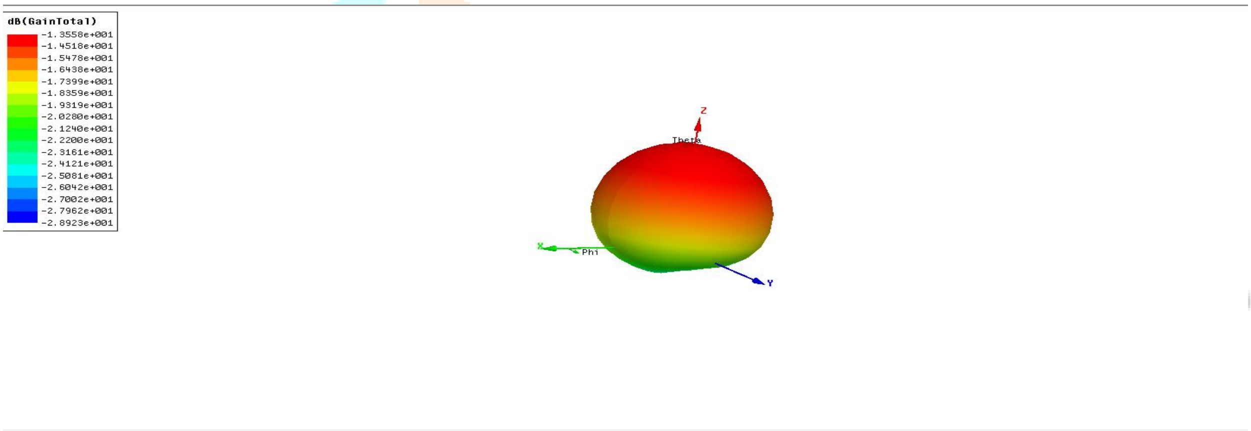


Fig. 14 . Radiation pattern

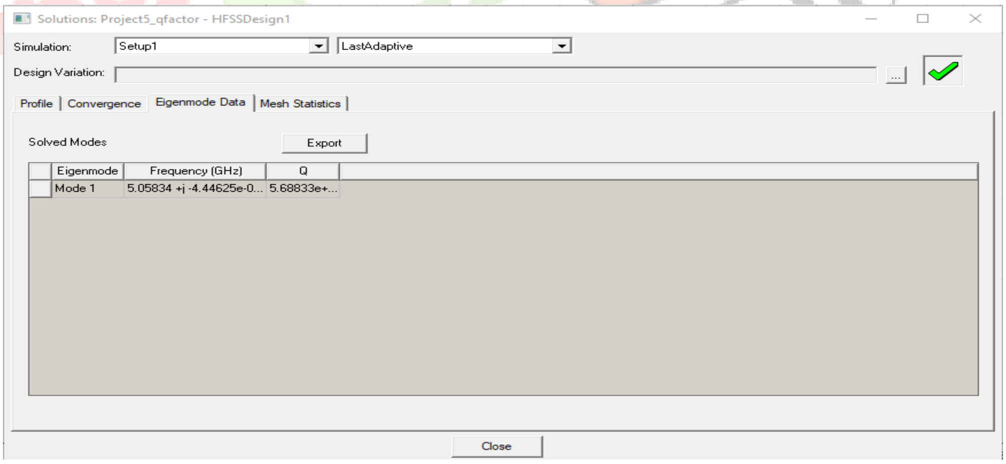
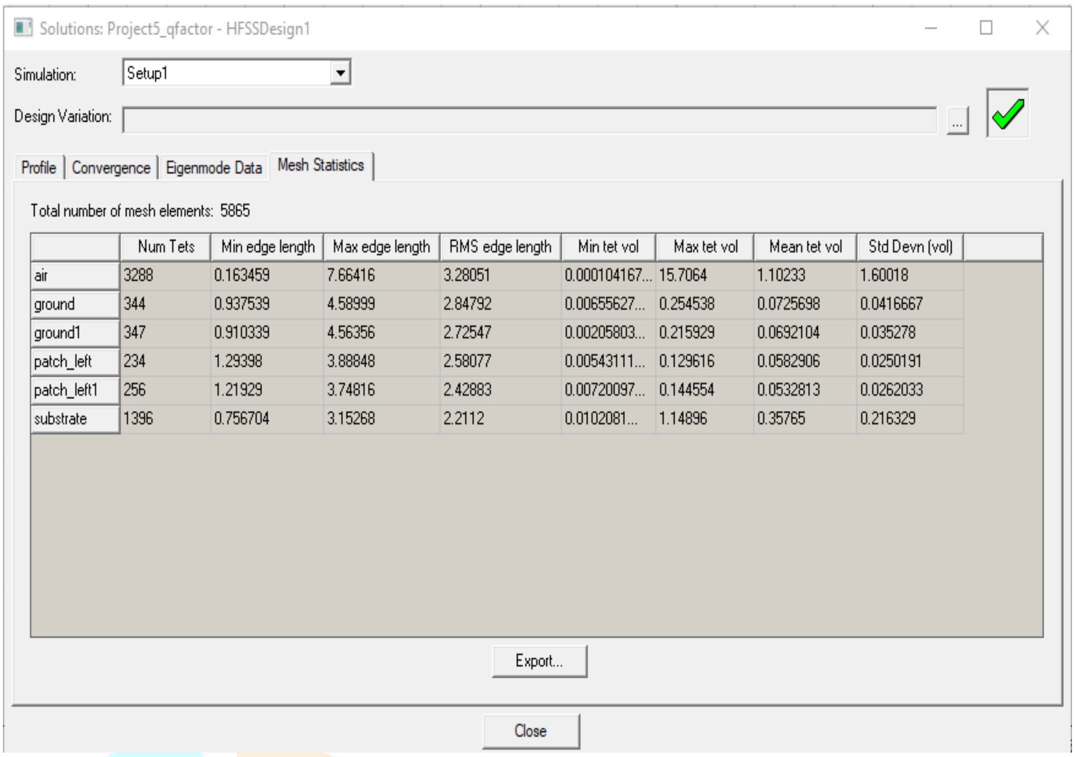


Fig. 15. Q Factor Value



	Num Tets	Min edge length	Max edge length	RMS edge length	Min tet vol	Max tet vol	Mean tet vol	Std Devn (vol)
air	3288	0.163459	7.66416	3.28051	0.000104167...	15.7064	1.10233	1.60018
ground	344	0.937539	4.58999	2.84792	0.00655627...	0.254538	0.0725698	0.0416667
ground1	347	0.910339	4.56356	2.72547	0.00205803...	0.215929	0.0692104	0.035278
patch_left	234	1.29398	3.88848	2.58077	0.00543111...	0.129616	0.0582906	0.0250191
patch_left1	256	1.21929	3.74816	2.42883	0.00720097...	0.144554	0.0532813	0.0262033
substrate	1396	0.756704	3.15268	2.2112	0.0102081...	1.14896	0.35765	0.216329

Fig. 16 Mesh Statics of Q Factor

VIII. CONCLUSION

The possibilities of reconfigure ability that the miniaturized periodic structures defined in have, with the use of varactor diodes. These unit cells can then be used to implement reconfigurable planar soft surfaces (corrugations). The varactor diodes are an appropriate choice for a range of corrugation geometries, their inclusion is not always possible for some classical geometries. The following results will be show in this simulation part. This work investigates how to develop compact reconfigurable planar corrugations. The proposed structure has a unit cell of a planar horizontal microstrip corrugation with a modified geometry that resembles a hairpin resonator shape. simulations and measured results are in good agreement as a consequence of the rigorous description of the included in the simulations.

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