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Narrow-Band Microwave Filter Using High Q Groove Gap Waveguide Resonators with Manufacturing Flexibility and no Sidewalls

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***Abstract*— The paper presents a new type of narrow band filter with good electrical performance and manufacturing flexibility based on the newly introduced groove gap waveguide technology. The designed 3rd order and 5th order filter work at Ku band with 1% fractional bandwidth. These filter structures are manufactured with an allowable gap between two metal blocks, in such a way that there is neither requirement to electrical contact nor alignment between the blocks. This is a major manufacturing advantage compared to normal rectangular waveguide filters. The measured results of the manufactured filters show reasonably good agreement with the full-wave simulated results, without any tuning or adjustments.**

***Index Terms*—Band-pass filters, Chebyshev response, Coupling Coefficient, External Q factor, Rectangular Waveguide filters, Gap Waveguide Technology.**

I. Introduction

MICROWAVE filters are integral parts of any communication system. For full duplex systems, these filters are normally realized in rectangular waveguide technology due to large power handling capacity and low losses. Narrow band-pass filters with low insertion loss and steep roll-off requires cascading of loosely coupled high Q waveguide cavity resonators [1-3]. Ideally, the Q factor for conventional waveguide cavity is very high, but when realized as filters and manufactured in two blocks, it is difficult to achieve the theoretical high Q value at high frequencies, and spurious resonances may appear near the pass-band. The reasons are usually due to finite conductivity and field leakage through imperfections and minor gaps between the two metal blocks (originating due to manufacturing flaws or metal deformation due to thermal expansion). Also, poor electrical contact between the conductors is one of the most common sources of passive intermodulation (PIM) [4], which is considered as a hidden threat in many microwave applications. To remove the tiny gaps between the two split-blocks, very good electrical contact must be achieved as well as good alignment of the two blocks. Also high quality surface finishing over the whole metal contact area is required for good mechanical assembly. Contacting surfaces must be protected from corrosion and oxidation over the entire lifecycle of the product. These strict mechanical requirements lead to very high precision metal machining technique which increases the cost of manufacturing and cause much delay in production chain, thus neither very suitable for high volume nor batch production.

To address the above-mentioned issue, a high Q resonator and band-pass filter with chebyshev response are presented in this work. The filter is designed for Ku band. Both the resonator and the filter are based on the newly proposed low loss gap waveguide technology presented in [5] and later experimentally demonstrated in [6-7]. The basic principle of operation is studied and explained in terms of a plane wave spectrum in [8] and in terms of classical modal field expansions in [9]. The articles [5]-[9] treat the ridge gap waveguide but the gap waveguide technology can also be used to realize groove gap waveguides and microstrip gap waveguides [10]. This technology can also be used for packaging of conventional microstrip circuits [11]. All the three types of gap waveguide transmission lines can be realized in two separate metal plates with a small gap between them, thus without any requirement for conductive contact between the two plates [12]. The present filters are realized using groove gap waveguides. The groove gap waveguides are in parallel work shown to have a Q-factor that is about 80% of that of the Q of a rectangular waveguide of the same cavity size [18], and, it has significant higher Q than the ridge gap waveguide.

The intended propagation mode of a groove gap waveguide is very similar to that of TE_{10} mode of rectangular waveguide [13]. The bottom metal plate consists of a groove with a texture of pins on both sides and the upper metal plate is a smooth plate, as shown in figure 1. Unlike other metal waveguides, there is not at all need for electrical contact

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between the lower and upper metal plates. The upper metal plate is placed intentionally at a distance smaller than λ/4 over the top of pin surface. The pin surface acts as an artificial magnetic conductor (AMC), and this stops the propagation of wave in all directions between the two plates within a stopband [5], and this stopband is larger for smaller gap between the pins and the upper plate [14].

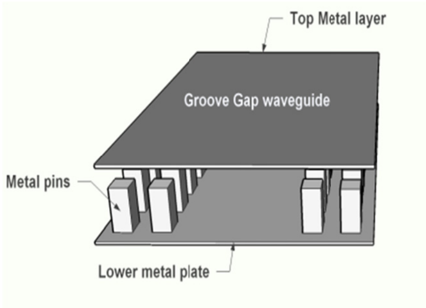

Fig.1. Geometry of groove gap waveguide

As a result, the electromagnetic field decays exponentially in the pin region. After two rows of pins, the field decay is about -45 dB. This field is so weak that the leakage can be ignored even without using metal sidewalls outside two pin rows. This leads us to build waveguide cavities without any requirement of electrical contact and alignment between the lower and upper metal plates. Consequently, the high precision machining required for manufacturing standard waveguide cavity filters can be avoided and cost of production can be minimized. The high Q groove gap waveguide resonator and the band-pass filter presented in this paper are based on this principle of fast decay of electromagnetic field in the pin region.

The aim of this paper is first to explain the design of the groove gap waveguide resonator. Then, we explain how to obtain the coupling coefficients between the cavities and how to design narrow band 3rd order and 5th order band-pass filters. At this point, the scope of the paper is to demonstrate the capability and manufacturing flexibility of the proposed groove gap waveguide filter only.

II. High Q Groove Gap Waveguide Resonator

High-Q resonator is one of the most important circuit components in many microwave applications such as narrow band-pass filters and low phase noise oscillators. Usually, the unloaded Q of dielectric free practical waveguide resonator can be expressed in the following form:

$$\frac{1}{Q_{unloaded}} = \frac{1}{Q_{metal}} + \frac{1}{Q_{leakage}} \qquad (1)$$

Thus, the losses that determine the unloaded Q of the resonator are composed of two main factors: metal loss and leakage loss. The metal losses are usually related to the volume of the resonator and conductivity of the material used to build the resonator. The leakage losses are related to the escaping of energy through the tiny gaps formed due to mechanical imperfections of the metal split block sections. Since the two factors are in parallel, the lowest Q value dominates in determining the unloaded Q of the resonator. Ideally, there are no losses due to leakage, but in practice leakage losses may become dominant. As a result, the leakage losses due to mechanical imperfections can cause a significant overall degradation of the Q-factor. Also, this leakage can shift the resonant frequency of the resonator and cause undesired extra ripples on the filter response. The frequency shift may not be a problem for wide-band filters, but for narrow band filters it can become critical and is difficult to predict. The unwanted frequency drift can only be found after manufacturing and testing, and therefore normally requires mechanical tuning screws to be adjusted after manufacturing. Also for high power applications, heat generated inside the cavity walls must be transferred to the environment in order to minimize temperature drift. Although not studied in the present work, in groove gap waveguide filter, it is expected that the thermal effects will be much smaller than in rectangular waveguide filters because the gap waveguide structure is open and has larger surfaces, one of which having a smooth plane surface.

The groove gap waveguide technology makes it possible to control and almost eliminate the leakage of EM energy through the gap between two metal blocks or plates. By carefully designing the square pins of the AMC surface, we can stop the field leakage over an octave bandwidth or more [14]. It has been observed in previous studies of the gap waveguide that after two consecutive rows of pins, the field leakage goes down up to a level of -45 dB over significant bandwidth [5]. The total E-field distribution within the resonator (at resonant frequency) also shows the same and is presented in fig.3(a).This motivates the design of the high Q resonator with much more manufacturing flexibility than normal rectangular waveguide. The metal walls on each side of the resonator do not play a significant role and can be removed unless needed for mechanical support. On the other hand, the two rows of pins that stop the field propagation on all sides make it flexible from a mechanical point of view.

Fig.2 shows the half wavelength groove gap waveguide resonator. As mentioned in the introduction, the groove gap waveguide has a lower metal plate with square pins forming a groove between them. The detailed design procedure of the pin surface to achieve a desired stopband is described in [14]. The pins are about λ/4 in height. The upper metal plate is a smooth plate that is placed at a distance of 1 mm from the top of the pin surface. It is chosen to have two rows of pins on each side of the resonator to make the field decay up to a level of about -45 dB. The pins have a dimension of 1×1 mm^2 and a period of 3 mm. The width of the groove is 15.8 mm that is similar to the width of standard Ku band rectangular waveguide. The total length of the groove is about 15.6 mm. The simulated Q value for this type of resonator with conductivity of aluminum is about 4605, which is lower than the rectangular waveguide resonator at that frequency. However, the groove gap resonator simulation includes the losses due to the current around the pins and the field leakage loss (whatever leakage is present after two rows of pins). In comparison, a short circuited standard rectangular waveguide cavity has a simulated Q of 5444 at 14GHz, which does not include losses due to an imperfect mechanical assembly. In a real case, the practical waveguide cavity has about 30% to 40% lower Q than the simulated value due to the imperfect mechanical assembly, surface roughness [17-18].Q factor of a practically realized groove gap resonator and a rectangular waveguide resonator is compared more in details in [19]. The simulated frequency response of the resonator is shown in Fig. 3(b).

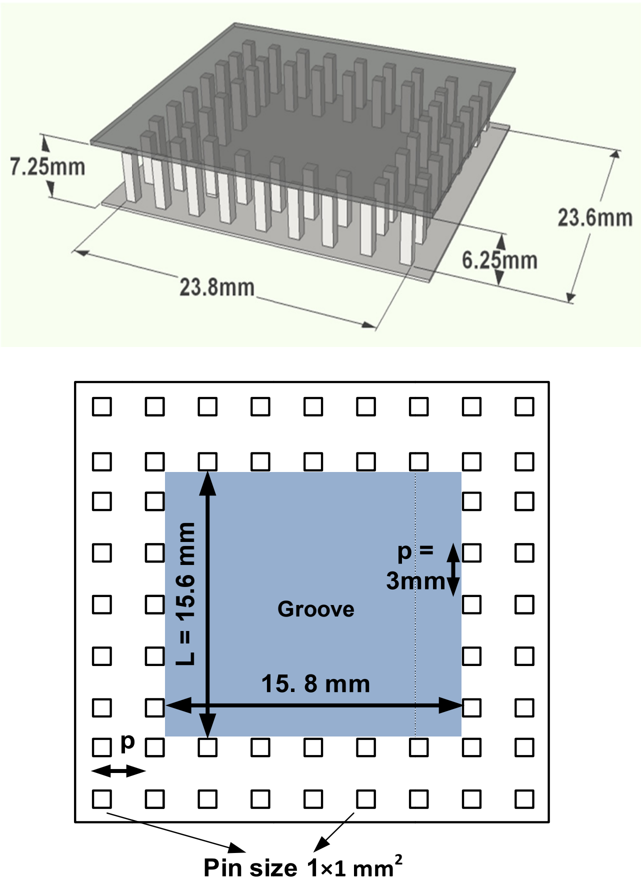

Fig.2. Geometry of the resonator, perspective view and top view

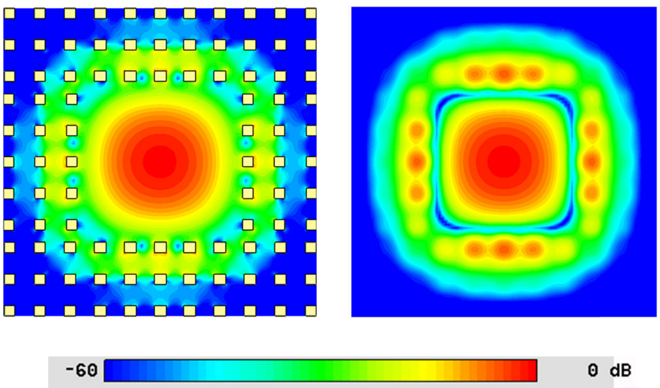

Fig. 3(a) Plots of total E- Field amplitude for the resonator with 3 rows of pin around it. The left graph shows the field at a surface that is cutting the pins in the middle, and the right one at a surface along the air gap between the top of the pins and the upper surface.

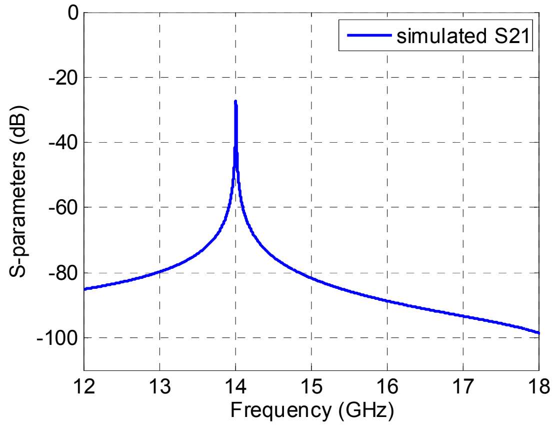

Fig.3(b). Simulated S_{21} for the Groove Gap waveguide resonator

III. The Coupling Coefficient And External Q

Any narrow-band, lumped element or distributed band-pass filter can be designed if three variables are determined beforehand [15]. These are synchronous tuning frequency of each resonator, the coupling between the resonators 'κ' and the singly loaded Q or external 'Q_{ex}' of the first and last resonators.

A. Coupling coefficient, κ:

The coupling coefficient is calculated for a pair of groove gap waveguide resonators placed side by side separated by two rows of pins. To obtain the required loose coupling for narrow band-pass filtering, the parameter *'s'* shown in Fig.4(a) is varied keeping all the other parameters of the groove gap waveguide resonator similar to that mentioned in section II. Complete full wave simulations are done to check how the parameter *s* influences the coupling. The effect of varying this parameter *s* is also shown in Fig. 4(b) for two values of *s*.

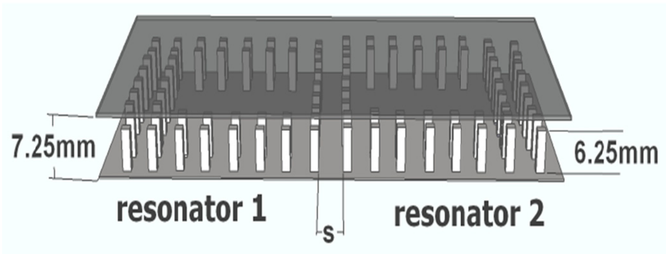

Fig.4(a). Two coupled groove gap waveguide resonators without metal sidewalls. The sidewalls are anyway not needed.

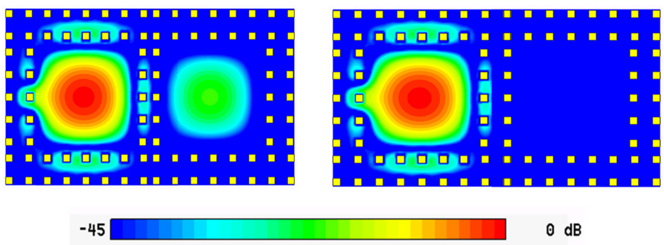

Fig. 4(b). Plot of total E field amplitude showing the effect of parameter *s* on coupling, for s=1.5 mm (left graph) and s=3.25 mm (right graph).

When this type of paired coupled resonator structure is simulated as a two port network, two distinct resonant peaks (closely placed with respect to each other) in S_{21} response are observed for each value of *'s'*. Usually, the stronger the coupling, the wider is the separation between the two frequency peaks and vice versa. The coupling coefficient is calculated from the frequencies of these resonant peaks by using this equation, which can be found in both [15] and [16]

$$\kappa = \pm \frac{f_{high}^2 - f_{low}^2}{f_{high}^2 + f_{low}^2} \quad (2)$$

The sign of the coupling is significant only in case of cross coupled resonator filter, and, therefore, it is not taken into account in our case. The obtained coupling coefficient for different *'s'* parameter values are obtained and plotted in Fig.5. As shown in Fig.5, quite low values of coupling coefficient can be achieved by the above mentioned arrangement of side by side resonators. While designing the 3rd order and 5th order filters, the plot in Fig.5 is used to determine the separation between consecutive resonators.

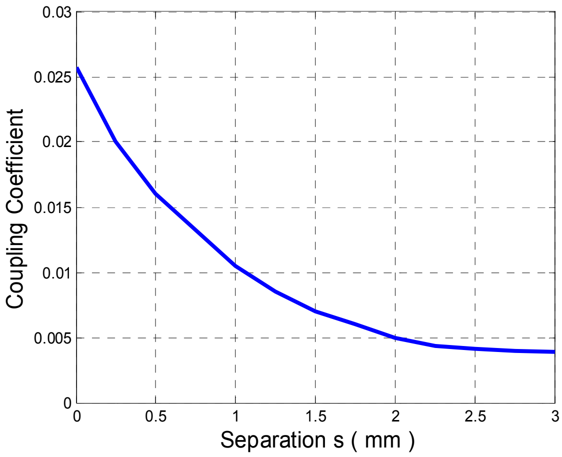

Fig.5. Coupling Coefficient, κ vs separation, *s*

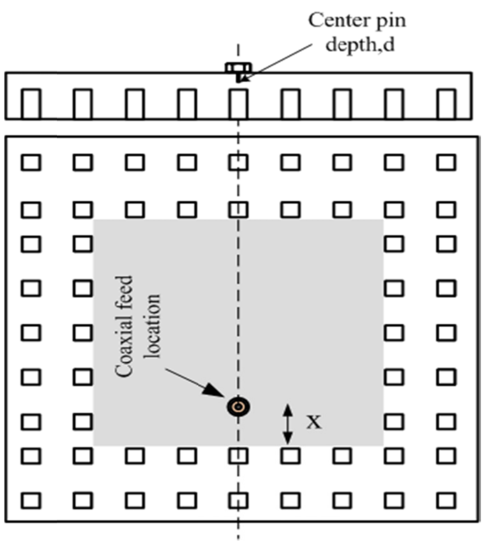

Fig.6. Location of coaxial feeding

B. External quality factor and excitation of the groove gap resonator:

The groove gap waveguide resonators used for the present filter design are excited with coaxial SMA connectors in order to simplify measurements with the help of available Vector Network Analyzer.

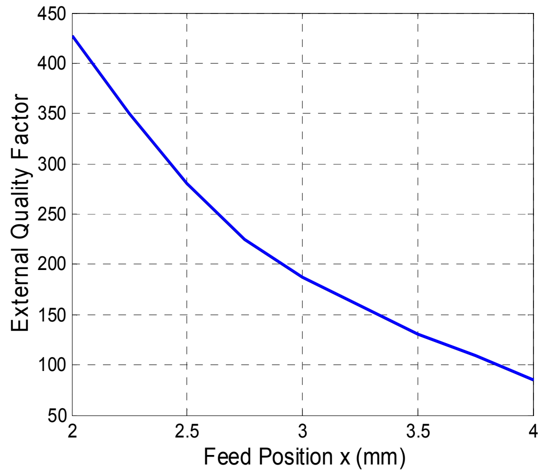

Fig.7. The simulated external quality factor

The typical I/O coupling structure of the proposed filter thus consists of one of the groove gap waveguide resonators and 50 ohm coaxial feeding. The coaxial probe is located in a hole in the top metal plate of the resonator. The position of the coaxial feed point *'x'* and the depth *'d'* of the center conductor of the coaxial probe are controlled to achieve the required coupling or the external quality factor. This is shown in Fig.6. To check the level of the achievable external quality factor, the parameter *'x'* is varied from 2 mm to 4 mm and *'d'* is kept at 2 mm. The simulated external quality factor is shown in Fig.7

IV. Band-pass Filter Design

The design of the microwave filter normally starts by using a low-pass prototype network regardless of the eventual practical realizations in waveguide or other transmission media. The synthesis of lumped element low pass prototype networks later enables the design of filters with transfer function such as chebyshev, elliptical function etc. In this work, the design of the band-pass filter starts with low-pass prototype network for chebyshev response. A 3rd order and 5th order chebyshev filters for Ku band are designed. The filters are designed for a 0.1dB equal ripple response with a pass band return loss of 16.4 dB. The fractional bandwidth Δf is chosen to be 1% around the center frequency $f_o = 14\ GHz$.

A. 3rd order filter design:

Now, to design a 3rd order filter, Chebyshev low pass prototype table in [15] is used. The low-pass prototype element values are

$$g_0 = g_4 = 1;\ g_1 = g_3 = 1.0316;\ g_2 = 1.1474$$

Having obtained the low-pass parameters, the band pass design parameters such as coupling coefficient *'κ'* is calculated by using the formula in [15]:

$$\kappa_{i,i+1} = \frac{fractional\ bandwidth}{\sqrt{g_i\ g_{i+1}}} \quad (3)$$

Thus, the required coupling coefficients among three resonators are $\kappa_{12} = \kappa_{23} = 0.0092$. The external quality factors at the input and output of the resonators are also calculated from a formula from [15]:

$$Q_{ex_n} = \frac{g_n\ g_{n+1}}{fractional\ bandwidth} \quad (4)$$

The required external quality factor for the input and output resonator is found to be 103. After calculating the required coupling coefficients and Q_{ex} for the input and output sections, the physical filter dimensions are obtained using the curves in Fig.5 and Fig.7. The height of the pins and the distance between the pins surface and the upper metal plate are kept unchanged. Also, it is to be noted that the width of the grooves formed by the row of metal pins is always kept at 15.8 mm. After that, full wave simulator Ansoft HFSS is used to simulate the whole structure. The filter with all critical dimensions is shown in Fig.8.

B. 5th order filter design:

Low-pass prototype element values for the 5th order chebyshev filter are:

$g_0 = g_6 = 1$; $g_1 = g_5 = 1.1468$; $g_2 = g_4 = 1.3712$; $g_3 = 1.9750$.

Based on these values, the required coupling coefficients for the resonators and the external quality factor for the I/O resonators are calculated using equation 3 and 4, respectively. The required coupling coefficients are:

$$\kappa_{12} = \kappa_{45} = 0.0079; \kappa_{23} = \kappa_{34} = 0.00607 .$$

Also, the external Q for the I/O resonator is 114.68. In next step, the physical dimensions of the filter are obtained by using Fig.5 and Fig.7. The filter dimensions are shown in Fig. 9.

V. MANUFACTURED PROTOTYPE AND MEASURED RESULTS

Both the 3rd order and the 5th order filters are manufactured with open end walls. The metal used for manufacturing the filters is aluminum as shown in Fig.10. For both the filter prototypes, two end walls are kept open in order to demonstrate the manufacturing flexibility of the above mentioned groove gap waveguide resonator and to show that the Q-factor will be still high even with open end walls. At first both filters are manufactured in two parts: one bottom metal plate with the textured pin surface and the smooth top metal plate. Later on, the two parts are screwed together.

As, mentioned in previous section, only two rows of pins are used to create the groove gap cavity resonators of the proposed filters. The leaked energy becomes almost negligible after just two rows of pins, so the presence of the sidewalls and electrical connections between the two metal blocks become insignificant. Initially, in the prototype filters only the two end walls of the structures are kept open. However, the two sidewalls can also be taken away without any effect on performance. The top metal plate can be placed at the required position only with the help of the spacers. To prove this concept, the 5th order filter prototype was modified later and the remaining two sidewalls were also removed leaving the top metal lid to be placed on four spacers. The pictures of the manufactured parts of both filters are shown in Fig.10, including one without sidewalls that was verified to have the about the same performance as the one with two walls.

At this point, the filters are measured with a vector network analyzer calibrated up to the SMA ports. No further calibration has been done to calibrate out the losses in SMA connectors. Also, no surface treatment or no fine tuning has been done on the manufactured prototype to improve the filter response. The simulated and measured filter responses for the 3rd order and 5th order filters are shown in Fig.11. On the other hand, Fig.12 shows the frequency responses for both the filters over a wider frequency range of 10-21GHz.

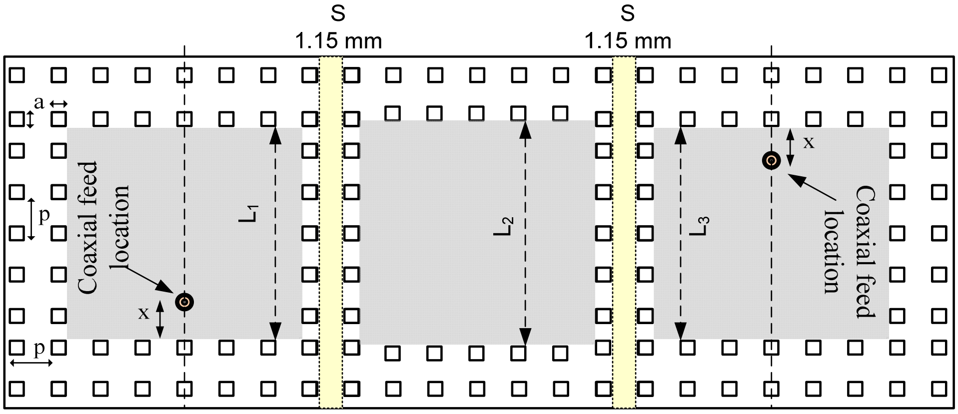

Fig.8. Top view of the 3rd order filter (top metal plate not shown). $L_1 = L_3 = 15.3$ mm; $L_2 = 15.8$ mm; x = 3.8 mm, a = 1mm and p = 3 mm

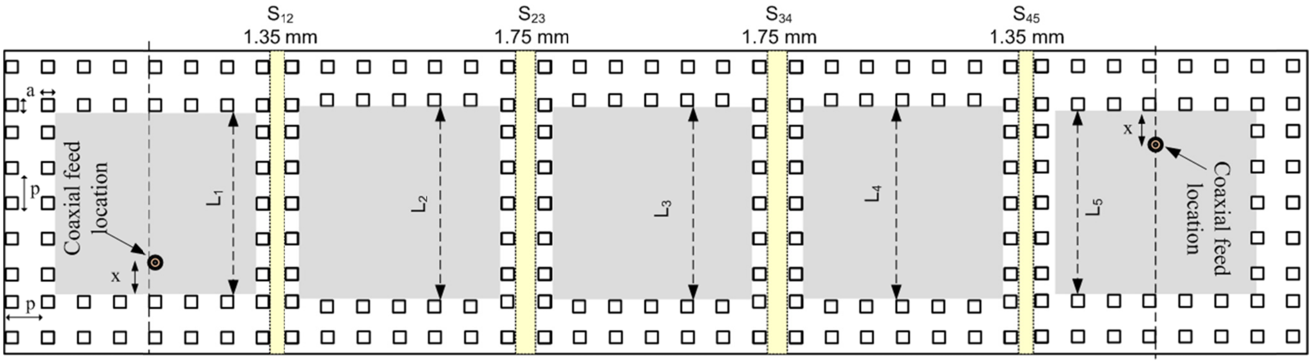

Fig. 9. Top view of 5[th] order filter (top metal plate not shown). $L_1 = L_5$ = 15.35 mm; $L_2 = L_4$ =15.75 mm; L_3 = 15.77mm; x = 3.8 mm, a = 1mm and p = 3 mm

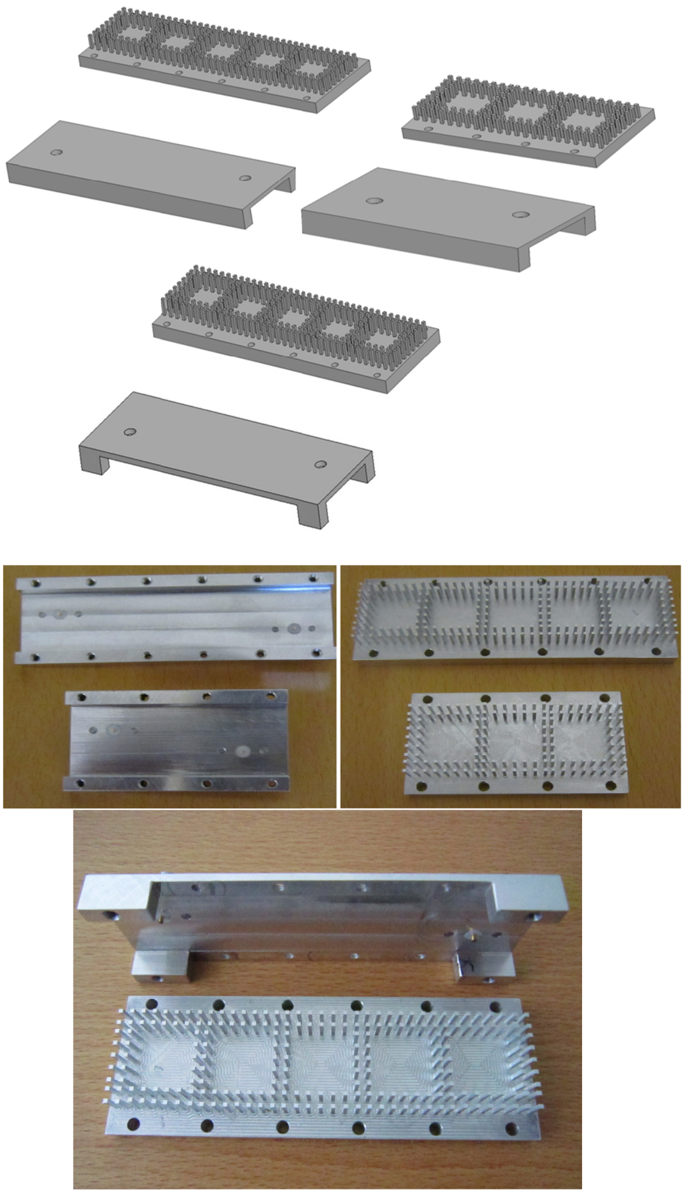

Fig.10. Manufactured 3[rd] order and 5[th] order filter prototype

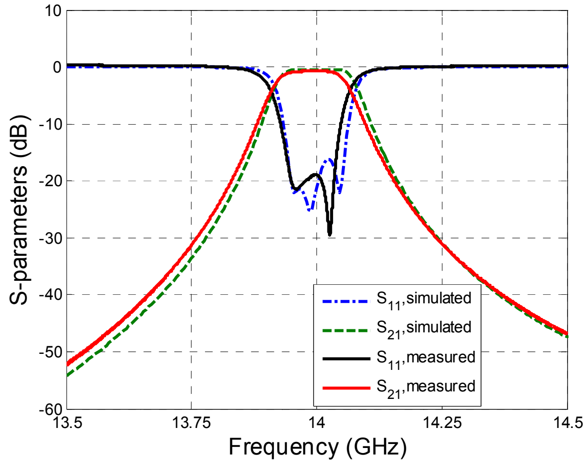

Fig.11 (a). Simulated and measured response for 3[rd] order filter

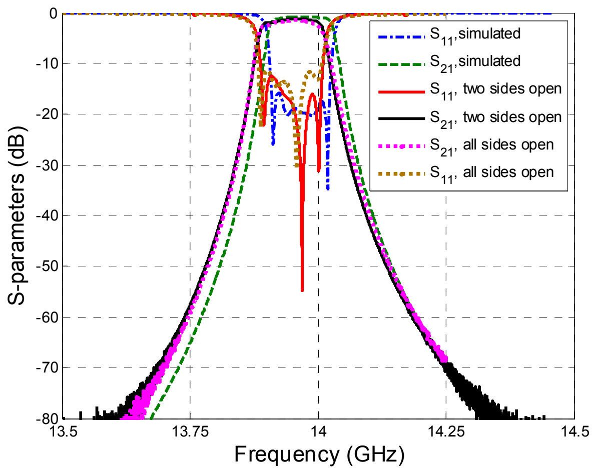

Fig.11 (b). Simulated and measured response for 5[th] order filter

As seen from Fig 11(a), the simulated filter response for the 3[rd] order agrees well with the measurement results. In this case, the measured mid-band insertion loss is about 0.31dB more than the simulated cases. This extra loss is mainly attributed to the losses in SMA connectors and the degradation of the conductivity of aluminum due to surface roughness. SMA connectors from Rosenberger were used in the filters. The insertion loss data available in the connector data sheet shows a possible loss of 0.149 dB per SMA at 14GHz. We have two SMA in our prototype, so the losses due to SMA in our prototype filter can be up to 0.3 dB more already. For the5th order filter, the measured mid band insertion loss is 0.42 dB more than the simulated values.

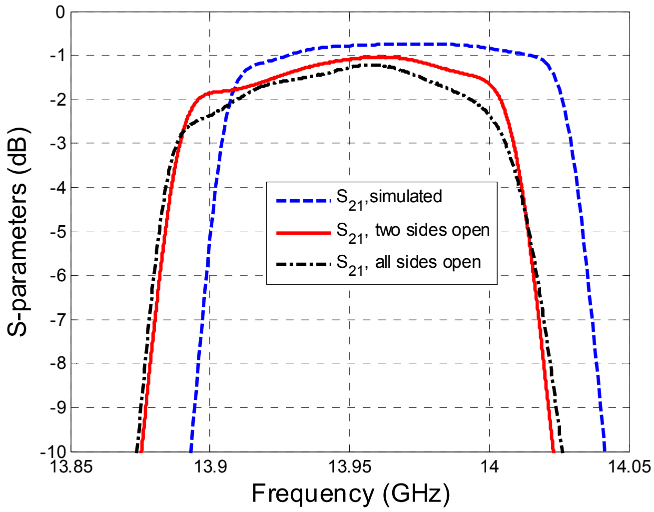

Fig.11 (c). Simulated and measured S_{21} response for 5rd order filter with two sides open and all side open.

The main reasons of increased insertion loss are the same as stated for the 3rd order filter. But the effect of degrading conductivity of Aluminum due to surface roughness also contributes towards increasing the insertion loss in 5th order filter. The measured insertion loss at band edges is 0.84dB and 1.76 dB for the 3rd and 5th order filter respectively which is expected due to the rounding of the passband edges [20]-[21] .

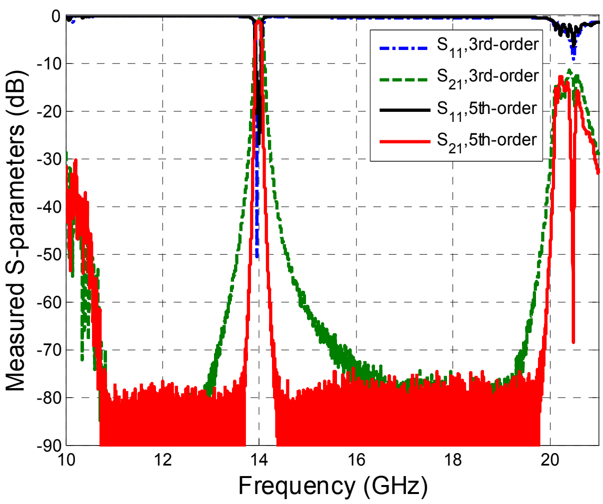

Fig.12. Spurious parallel plate modes around 10 GHz and 20 GHz

In cases of both 3rd and 5th order filters, there are small shift of the frequency band of about 0.08% and 0.16 %, the larger shift for the 5th order filter. Measured center frequency for this filter is 13.947 GHz in comparison of simulated 13.97 GHz value. This frequency shift is attributed mainly to the mechanical tolerance issue which is in the order of 15-20 micrometer in this case. This tolerance is 0.1% of the resonator width, which is of the same relative order as the frequency shift. The out of band rejection of both filters is also found to be as predicted. The parallel plate cut off bandwidth for this pin surface (AMC surface) is 11-20 GHz. So, these structures are expected to allow parallel-plate mode propagation before and after this stop-band, and Fig.12 shows that filter rejection vanishes outside this band.

VI. Conclusion

Newly introduced groove gap waveguide technology is used to design 3rd order and 5th order bandpass filters with a fractional bandwidth of 1%. Good electrical performance is achieved for the manufactured filter structures even without using any side walls. These filters utilize high Q groove gap waveguide resonators which eases the tight mechanical requirements applicable to standard waveguide cavity filters. Therefore, these proposed structures are more suitable for mass production. The filter design was based on using a full wave frequency domain field solver to generate design curves for the coupling coefficients κ's and the external Q-factors Q_{ex}. The measured shape of the filter skirts, s-parameters values and out of band rejection of both the 3rd and 5th order filters show quite good agreement with those of the simulated values, except for a small relative frequency shift that is in the same order of magnitude as relative tolerances of each resonator. The Q value of the resonators and the insertion loss of the filters can be improved even more by doing surface treatment and silver-plating of the prototypes which usually is the case for most commercial filters.

Acknowledgment

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