Design of Ku-band Dielectric Resonator Filter for Satellite Applications

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Abstract— Presented is the design of Ku-Band dielectric resonator filter for satellite applications. The uniqueness of this design is based on the simple feeding arrangement that utilizes a microstrip transmission line section but segmented piecewisely into four stubs, where each stub is offset from the preceding with a distance equal to the diameter of the DRs. Each DR couples electromagnetic energy from the input port via the first to another, and so until, until the final output port. The response of the design is assessed by comparing the simulated and experimental results. The simulated S-parameter results were compared with those obtained experimentally. The simulated and experimental reflection magnitude $|S_{11}|$ is within -6 dB across the band, whereas the transmission magnitude $|S_{12}|$ is -1.66 dB up to 17.5 GHz.

1. INTRODUCTION

Advances in technologies and continued growth of wireless devices has continue to challenge the community to design and produce new alternative but smaller and more multifunctional filters that are suitable and relevant to the current technology. It is evident that the traditional filter such as waveguide filter is no longer attractive, efficient, nor adequate for today's technology that demands small, compact, easy integration and cheap solution to almost every electronic device. Although until recently, alternative filter solutions employing microstrip technology where such filter design consist of only a few rows of parallel-coupled transmission lines has been ground breaking and remarkable. However, such filter becomes unrealizable as the frequency of the filter increases. The gap between the two neighbouring microstrip lines becomes narrower resulting to a very tight coupling. The strait spacing between the rows of parallel-coupled transmission lines complicates the fabrication processes of the filter as the printing machine only supports etch limitation up to a minimum trace and gap of about 0.3 mm.

To overcome this serious challenge therefore, it is contingent to investigate the possibility of an alternative novel-fed dielectric resonator filter for Ku-band applications. Dielectric resonators (DRs) offer more flexibility in filter design. The proposed design is such that three identical cylindrical DRs are excited by four microstrip lines of certain dimensions. The dimensions are multiples of quarter wavelengths of the centre frequency of which the filter operates, and are calculated to coincide with maximum radiation of the transmission lines based on the standing wave ratio formation of the voltage allocation on shorted microstrip transmission line. Besides, these stubs and the DRs were configured in such a way that the DRs were positioned in a manner that their various resonances overlap constructively [1]. In effect, wideband impedance bandwidth was achieved. Instead of the conventional direct side proximity-coupled microstrip feed line common among series feed array, the proposed utilizes the same principle, but rather with broken or piecewise segmented transmission stubs such that each stub is of a dimension of odd multiple of quarter wavelength. The widths are determined by their characteristic impedance equivalency. Not only was the transmission line broken into stubs, the stubs were also offset from each other by the diameter of the cylindrical DRs. By this arrangement, it is therefore evident that each DR is placed in between two transmission lines, and as such, the first DR coupled electromagnetic energy from the input port to the next transmission stub, and so on until the output port. Essentially, the proposed feed will mitigate the effect of unnecessary reflection, and thus deliver a better match, good reflection coefficient, improved bandwidth and performance as it is expected that a good impedance match will naturally improve the filter efficiency. In effect, the proposed will efficiently couple the signal energy from one microstrip to the other while providing a good performance and low insertion loss.

2. THE DESIGN

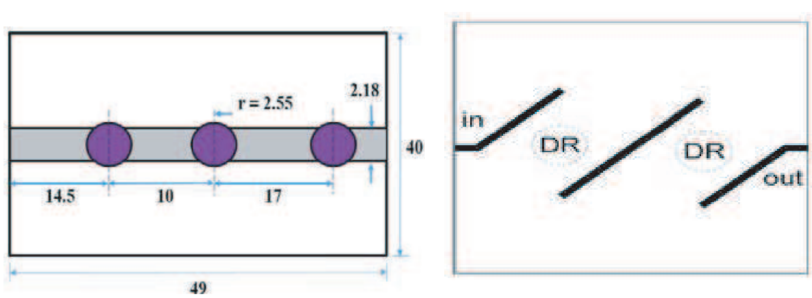
For a filter circuit to pass a certain frequency and rejects or attenuates all the other unwanted component of the signal, two frequencies f_1 and f_2 must be defined as the lower and upper cut-off frequencies respectively. The differential Δf between these two frequencies where ($\Delta f = f_2 - f_1$)

is referred to as the bandwidth. The magnitude of this differential determine whether the filter is a narrowband ($\Delta f < 500$ MHz), and wideband otherwise. One of the serious challenges militating against the DRs in spite of its many attractiveness is the narrow bandwidth limitations. Consequently, a new vista of challenges on bandwidth enhancement became serious issue of concern. Attempt to solve this limitation deepens, and several novel alternative solutions have been reported but with different performance penalties. According to Rezaei et al. [2], there are three broad techniques that can be used to enhance the bandwidth of the DR by either lowering the inherent Q-factor of the resonator or using external matching networks or combine multiple dielectric resonators. Matching networks, such as matching stubs can be used to increase the bandwidth of a DR by transforming its input impedance to better match than that of the coupling circuit. In this study, we have decided to employ the second approach, using a novel feed. This is because using additional DRs can also result in generating extra resonant frequency which can contribute to the wideband DR. In multiple DRs, dielectric resonator can be arranged either in stacked configuration or parasitic co-planar configuration [3]. The idea is that, the merged frequency band can achieve wideband provided that

$$f_1 + \frac{\Delta f_1}{2} \geq f_2 - \frac{\Delta f_2}{2} \tag{1}$$

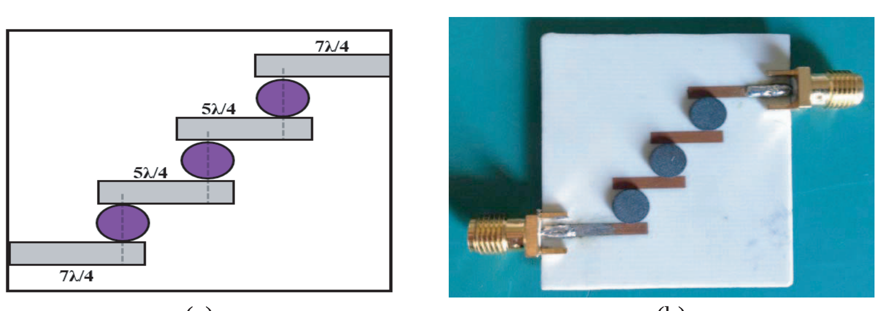
where Δf is the 3-dB bandwidth of S_{11} curve, f_1 and f_2 is the resonant frequency from the two segments with $f_1 < f_2$ [2–4]. Whereas, if the resonant frequencies are spaced further apart, then a dual band operation instead of wide band will result. Two half rectangular DRs, due to their different size with the same permittivity, resonated in two different adjacent frequency merged together to produce wideband DR [1, 3].

In [5], Ain et al. uses three DRs in an arrangement depicted in Figure 1(b). Therefore, the filter performance was consequent on the geometry and the locations of the resonators on the feed line, where all the DRs have the same permittivity and dimensions. Each dielectric resonator resonates at the same mode but at different frequency, thus combines together the DRs respective frequencies to form a wide-bandwidth provided that Equation (1) is satisfied. Authors in [6] employed this technique with certain but significant modifications as shown in Figure 1(b). In his method, the effect of reflections due to coupling mismatch with attendant poor voltage standing wave ratio (VSWR) common to the conventional direct-coupled microstrip transmission line as employed by



(a) (b)

Figure 1: Existing similar works.



(a) (b)

Figure 2: The proposed design. (a) Schematic. (b) Prototype.

Ain et al. was eliminated. Our proposed design is similar to that reported in [6] but differ in terms of the number of DRs and the geometry of the microstrip line. Instead, the microstrip feed stubs are configured as shown in Figure 2. The first (input port) and last (output port) stubs were set at $7\lambda/4$, whereas the two intermediate stubs were set at $5\lambda/4$ where λ is the guided wavelength. The width of each stub is calculated to be 1.9 mm in order to approximate to the characteristic impedance of the microstrip line. All the three DRs are identical, and are made of $CaCu_3Ti_4O_{12}$ (CCTO), each with a radius of 2.75 mm, a dielectric permittivity (ε_r) of 36, and a thickness (h) of 1 mm.

In this design, the signal is coupled to the DRs using a few procedure of parallel coupled transmission stub excitation as earlier said. Each resonator couples electromagnetic energy over $n\lambda/4$ stubs to the neighbouring resonator, thereby forming a staggered line. The DRs are then placed between each line, such that the coupling gap between each DR and the transmission line is 0.2 mm. The value of a quarter-wavelength is calculated to be 2.509 mm at the centre frequency, which is twice smaller than the diameter of each DR of 5.5 mm. Consequently, the length of the transmission line was adjusted to odd multiple integers of the guided wavelength in order to catch the antinodes radiation of the voltage allocation on a short-ended microstrip transmission line.

The proposed design was modeled using finite integration technique (FIT) full-wave CST commercial solver, and the resulting design is fabricated on the Duroid RO4003C microwave laminate board of permittivity (ε_s) of 3.38, with a substrate thickness (h_s) of 0.813 mm, $\tan(\delta)$ of 0.0027, and a metal thickness (t) of 0.035 mm. Subsequently, the fabricated design was measured, and the measured and simulated results were plotted to determine the degree of agreement.

3. RESULTS AND DISCUSSIONS

The TM_{01} resonant mode was excited, and the simulated resonance frequency occurred at 17.1 GHz which is within the Ku-Band of (12 to 18 GHz) as shown in Figure 3. The physical dimensions of the dielectric cylinder were optimized to satisfy the excitation mode resonance. The simulated $|S_{21}|$, $|S_{11}|$, and the bandwidth are −1.66 dB, −15 dB, and 366 MHz respectively as shown in the figure. The measured S-parameter results are shown in Figure 4. The measured $|S_{21}|$, $|S_{11}|$, and the bandwidth are −23.12 dB, −17.66 dB, and 550 MHz respectively.

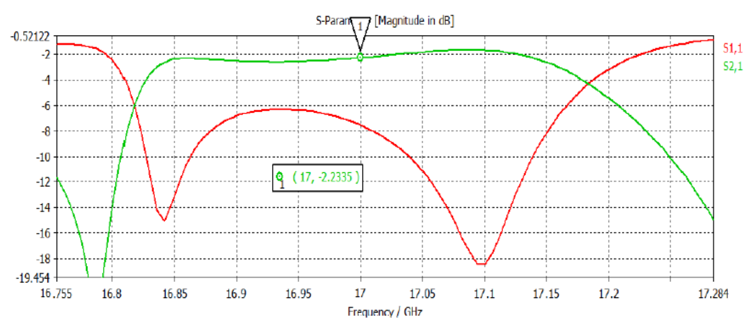
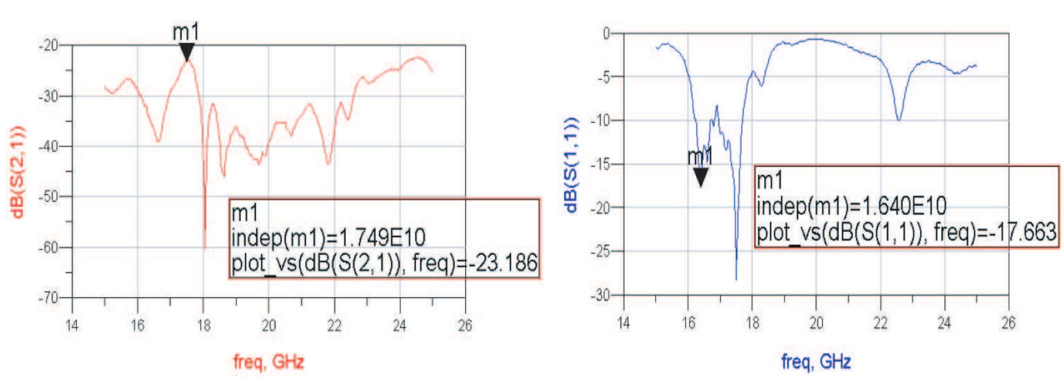


Figure 3: The simulated S-parameters of the proposed design.



(a) (b)

Figure 4: The measured S-parameters of the proposed design. (a) The $|S_{21}|$. (b) The $|S_{11}|$.

Table 1: Simulated and measured results comparison.

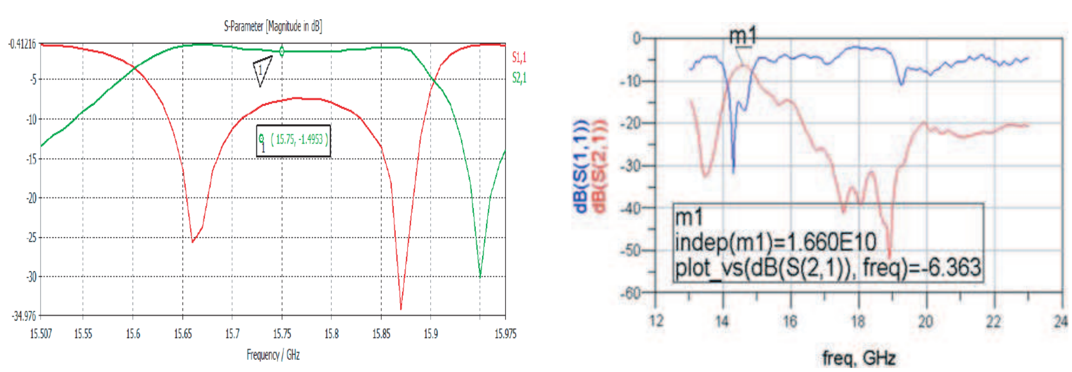
Parameters	Simulation	Measurement
$\lvert S_{21}\rvert$ dB	-1.66	-23.12
$\lvert S_{11}\rvert$ dB	-15.02	-17.66
Bandwidth (MHz)	366	550
f_0 (GHz)	17	17.5

The result obtained from testing the fabricated filter greatly differs from the simulated result. This is seen especially in the measured insertion loss value. The filter shows high insertion loss of -23.186 dB as compared to the simulation of -1.663 dB. The simulated result shows a good, almost 0 dB insertion value but the loss during measurements was significant. The high insertion loss value shows that a large unnecessary portion of the signal power is lost at the input port, resulting to a meagre transmitted signal energy through to the output port. In this case, the filter is therefore not efficient for real life use. Besides, the measured filter response shows a slight shift in the frequency. The operating frequency of the simulated result shows that the filter has a centre frequency of 17 GHz while the measured result shows resonance at 17.49 GHz. This suggests that the operating frequency of the filter shifted for a rough amount of 490 MHz. This situation is a bit difficult to totally avoid as it is pretty difficult to appropriate place all the DRs at exact coordinate as simulated. Alternatively therefore, the electromagnetic waves are not totally confined in the device. For high frequency RF devices, even a millimeter change will greatly affect the response of the device.

Further optimization was done where it was discovered that the measured loss at this point was approximately -3.249 dB. To compensate for the transmission line losses, the measured transmission line loss was subtracted from the measured insertion loss of the filter. Subsequently, issue of inefficient coupling due to relatively large coupling spacing between the DRs and feed line, which however could lead to tangible energy loss was addressed. Eventually, a better and efficient results were obtained as shown in Figure 5 and Table 2.

The insertion loss is better than the early design where the $|S_{21}|$ shows a good flat line that approached 0 dB. The 2 pole filter also exhibits a good return loss of over -15 dB. However the filter operates at a frequency of 15.75 GHz which is different from the previous un-optimized result. The measured result showed improvement in terms of $|S_{21}|$. Comparing this with the previous un-optimized filter, the insertion loss of the final design improves significantly to -6.363 dB. However, the operating frequency deviates from the simulation result. This dissimilarity is due to the inability to accurately position the DRs at the actual coordinate during, thus leading to a change in frequency and bandwidth. The simulated result and the measured result is compared in Table 2.

Moderately, the use of the three DRs is proven to produce a wide bandwidth bandpass filter.



(a) (b)

Figure 5: The optimized S-parameters of the proposed design. (a) Simulated. (b) Measured.

Table 2: Optimized simulated and measured results comparison.

Parameters	Simulation	Measurement
$\lvert S_{21}\rvert_{dB}$	-0.60	-6.36
$\lvert S_{11}\rvert_{dB}$	-25.68	-17.0
Bandwidth (MHz)	299	800
f_0 (GHz)	15.75	14.59

The result shows that each dielectric resonator resonates with different frequency but in the same mode that the combination of all frequency will produce a wider response that could increase the overall bandwidth of the filter.

4. CONCLUSIONS

In this paper, the design of Ku-Band bandpass filter for satellite applications has been investigated. The parametric effect of the feed with respect to the coupling spacing, as well as the efficiency of the coupled energy bearing in mind the decomposed nature of the proposed feed. It was evident that the combination of all the DRs responses lead to a final bandwidth of 800 MHz, which is larger than each DR individual bandwidth. This therefore verify this theory that if there are DRs labelled as DR_1, DR_2 and DR_3, with their discrete resonant frequencies of f_1, f_2 and f_3, and bandwidth of BW_1, BW_2 and BW_3 respectively, then the combination of their composite resonances is a wideband provided a certain inter-element spacing is satisfied.

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