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Technical Report on
Design of Switched Filter Bank for
550-900 MHz Band of GMRT

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Abstract

The purpose of this project is to improve the bandwidth of the existing 610 MHz Front End System. Presently, system bandwidth is 40 MHz, limited by the LNA and Band Pass Filter. The existing Coaxial Feed is being replaced by a wide band Cone Dipole Feed (CDF) with a frequency coverage from 550-900 MHz. The front-end electronics has to be modified with low Loss wide band Quadrature Hybrid, Wide Band Low Noise Amplifier and Broad Band Filter to process the RF signal received by cone dipole feed. Therefore a broad band filter is designed for 550-900 MHz bandwidth. Also, a switched filter bank is designed to provide the users with an option to select a bandwidth of 100 MHz at different centre frequencies over the band. The prototype filters are fabricated and the measurement results are found to be satisfactory. This report covers the design of the Broadband Band-pass Filter 550-900 MHz and Switchable Filter Bank comprising of the sub-band band-pass filters each of 100 MHz bandwidth. The report also covers the basics of filters, design of microstrip filters in general and detailed theoretical design concepts for Hairpin structure that has been used in designing of these Bandpass filters.

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1. Introduction:

Giant Meter wave Radio Telescope (GMRT) has been designed to operate at six different frequency bands centred at 50 MHz, 150 MHz, 235 MHz, 327 MHz, 610 MHz, and L-Band which extends from 1000 MHz to 1450 MHz. The L-band is further splits into four sub-bands centred at 1060 MHz, 1170 MHz, 1280 MHz and 1390 MHz, each having a bandwidth of 120 MHz. The 150 MHz, 235 MHz and 327 MHz bands have a bandwidth of around 40 MHz whereas 610 MHz band has a bandwidth of around 60 MHz.

This report attempts to describe a design of a full-band band pass filter (550-900 MHz) and a switchable sub-band filter bank for upgradation of 610 MHz. The upgradation of 610 MHz band to 550-900 MHz band will help in achieving a broadband bandwidth of 350 MHz and with sub-band filter bank this bandwidth can further be sub-divided into 100 MHz band. The switchable sub-band filter bank consists of four sub-band band pass filter each of 100 MHz. This upgradation of 610 MHz band to broadband bandwidth of 350 MHz will help in achieving a higher dynamic range with increased sensitivity of GMRT receiver chain.

Bandpass Filters designed for 550-900 MHz band are microstrip filter design. Filters are realized using lumped or distributed circuit elements. However with the advent of advanced materials and new fabrication techniques, microstrip filters have become very attractive for microwave applications because of their small size, low cost and good performance. There are various topologies to implement microstrip bandpass filters such as end-coupled, parallel coupled, hairpin, interdigital and combline filters.

This project will present the design of a hairpin microstrip bandpass filter. The basic design specifications that will be used for this bandpass filters are viz. centre frequency and bandwidth while Agilent Advance Design System ADS software is used for simulation. The passband for full-band bandpass filter is 550-900 MHz and for each sub-band bandpass filter of Switchable Filter Bank are 550-650 MHz, 635-735 MHz, 720-820 MHz, and 800-900 MHz. The filters are design to have a minimum attenuation of -40 dB at 10% from band edge frequency and passband ripple of 0.1 dB. The minimum attenuation of -40 dB at 10% from band edges frequency is chosen to have very good rejection for out of band frequency. The filters are designed using ADS design software and implemented on Rogers 1060 LM substrate with dielectric constant of 10.2, loss tangent of 0.0023 and substrate height of 1.27mm.

2. Basics of Filters:

Filters may be classified in a number of ways. An example of one such classification is reflective versus dissipative. In a reflective filter, signal rejection is achieved by reflection the incident power, while in a dissipative filters are used in most applications. The most conventional description of a filter is by its frequency characteristic such as lowpass (a), highpass (b), bandpass (c) and bandstop (d). Typically frequency responses for these difference types of filters are shown in fig. 1. In additional, an ideal filter displays zero insertion loss, constant group delay over the desire passband and infinite rejection elsewhere. However, in practical filters deviate from these characteristics and the parameters in the introduction above are a good measured of performance.

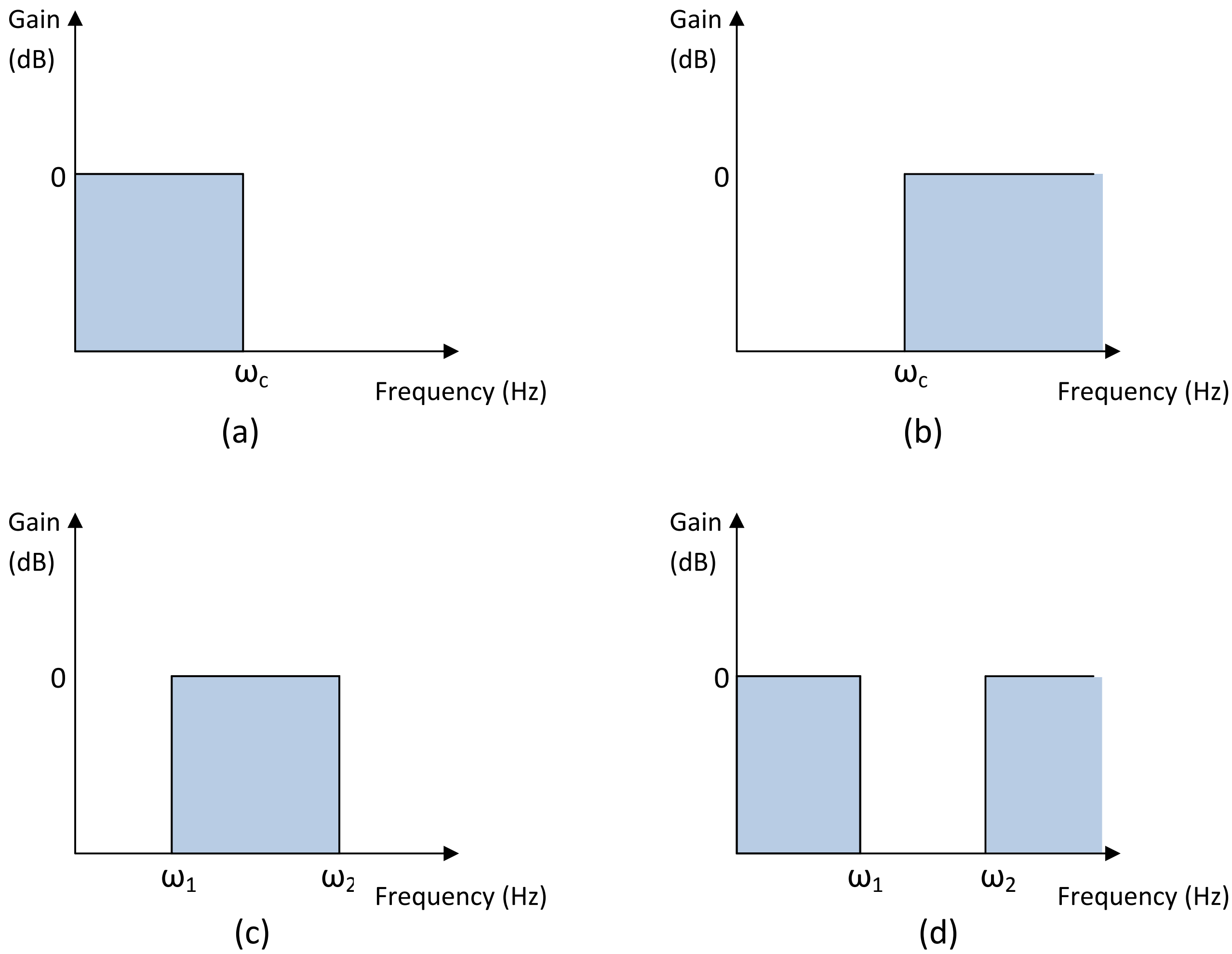


Fig. 1 Types of Filters

3. Types of Frequency Response:

There are basically three types of filters response:

1. Butterworth Filter.
2. Chebyshev Filter.
3. Elliptical Filter.

3.1 Butterworth Filter:

The Butterworth filter is a type of signal processing filter designed to have as flat a frequency response as possible in the passband. It is also referred to as a maximally flat magnitude filter.

The frequency response of the Butterworth filter is maximally flat (i.e. has no ripples) in the passband and rolls off towards zero in the stopband. When viewed on a logarithmic Bode plot the response slopes off linearly towards negative infinity. A first-order filter's response rolls off at −6 dB per octave (−20 dB per decade) (all first-order lowpass filters have the same normalized frequency response). A second-order filter decreases at −12 dB per octave, a third-order at −18 dB and so on. Butterworth filters have a monotonically changing magnitude function with ω, unlike other filter types that have non-monotonic ripple in the passband and/or the stopband.

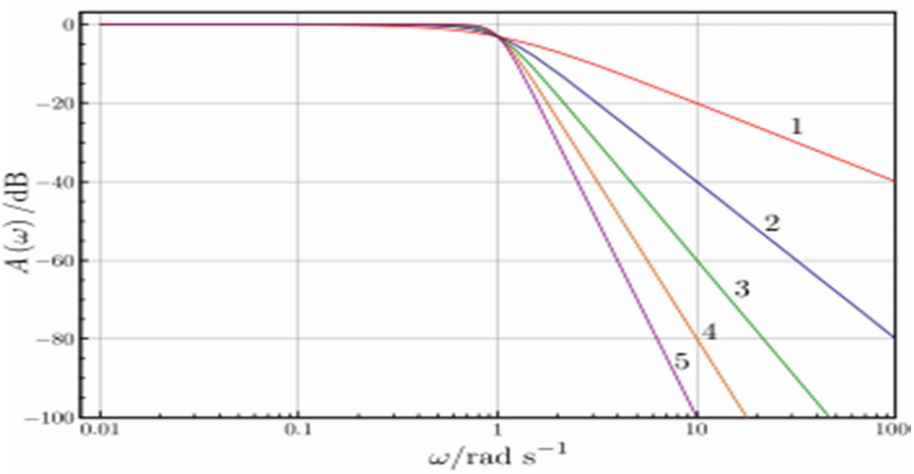


Fig. 2 Plot of the gain of Butterworth low-pass filters of orders 1 through 5, with cutoff frequency ω_o=1.

The gain response as a function of angular frequency ω of the nth-order Butterworth low pass filter is

$$G_n^2(\omega) = |H_n(j\omega)|^2 = \frac{G_O^2}{1 + (\frac{\omega}{\omega_C})^{2n}}$$

where

n => Order of filter,

ω_C => Cutoff frequency (approximately the -3dB frequency),

G_O => DC gain.

3.2 Chebyshev Filter:

Chebyshev filters are analogue or digital filters having a steeper roll-off and more passband ripple (type I) or stopband ripple (type II) than Butterworth filters. Chebyshev filters have the property that they minimize the error between the idealized and the actual filter characteristic over the range of the filter, but with ripples in the passband. Because of the passband ripple inherent in Chebyshev filters, the ones that have a smoother response in the passband but a more irregular response in the stopband are preferred for some applications.

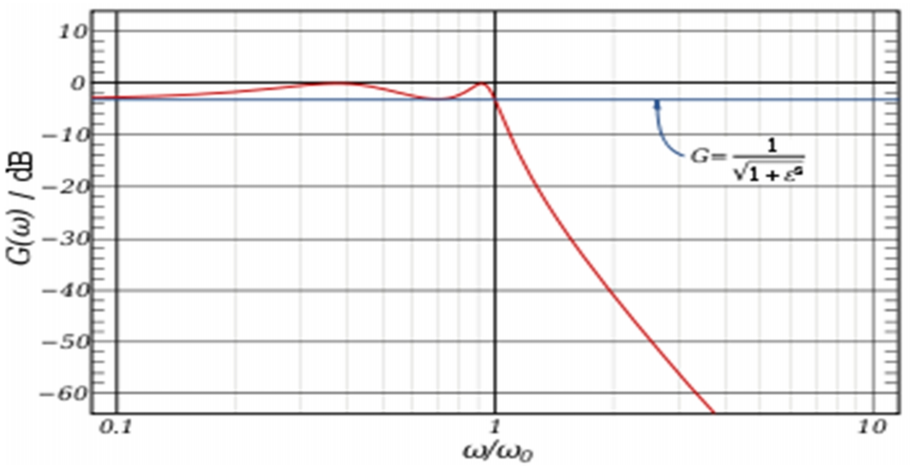


Fig. 3 The frequency response of a fourth-order type I Chebyshev low-pass filter with Ɛ=1.

The gain response as a function of angular frequency ω of the n^{th}-order Chebyshev low pass filter is

$$G_n(\omega) = |H_n(j\omega)| = \frac{1}{\sqrt{1 + \varepsilon^2 T_n^2(\frac{\omega}{\omega_O})}}$$

Where
ε => Ripple factor,
ω_O => Cutoff frequency,
T_n => Chebyshev polynomial of the n^{th} order.

3.3 Elliptical Filter:

An elliptic filter (also known as a Cauer filter, named after Wilhelm Cauer) is a signal processing filter with equalized ripple (equiripple) behaviour in both the passband and the stopband. The amount of ripple in each band is independently adjustable, and no other filter of equal order can have a faster transition in gain between the passband and the stopband, for the given values of ripple (whether the ripple is equalized or not). Alternatively, one may give up the ability to independently adjust the passband and stopband ripple, and instead design a filter which is maximally insensitive to component variations.

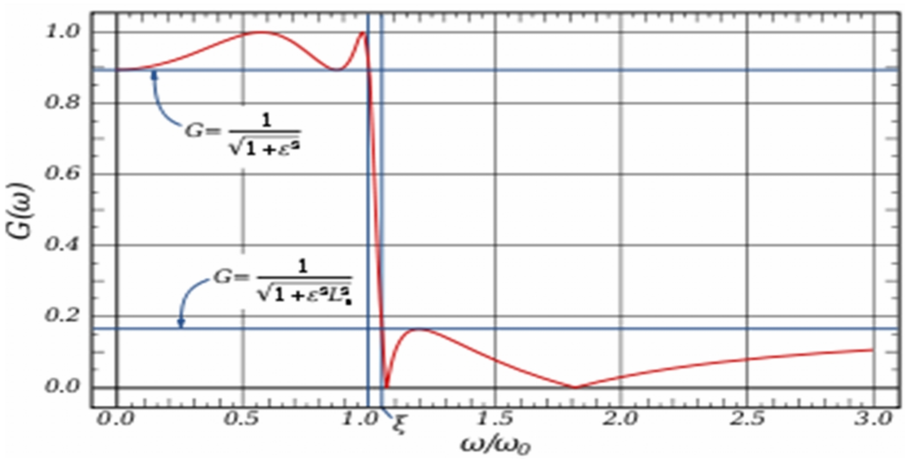


Fig. 4 The frequency response of a fourth-order elliptic low-pass filter with ε=0.5 and ξ =1.05.

As the ripple in the stopband approaches zero, the filter becomes a type I Chebyshev filter. As the ripple in the passband approaches zero, the filter becomes a type II Chebyshev filter and finally, as both ripple values approach zero, the filter becomes a Butterworth filter.

The gain of a lowpass elliptic filter as a function of angular frequency ω is given by:

$$G_n^2(\omega) = |H_n(j\omega)|^2 = \frac{1}{1 + \varepsilon^2 R_n^2\left(\xi, \frac{\omega}{\omega_0}\right)}$$

Where

R_n is the *n*th-order elliptic rational function (sometimes known as a Chebyshev rational function),

ω_o => cutoff frequency,

ε => ripple factor,

ξ => selectivity factor.

4. Microstrip Structure:

The general structure of a microstrip is illustrated in Fig. 4. A conducting strip (microstrip line) with a width W and a thickness t is on the top of a dielectric substrate that has a relative dielectric constant ε_r and a thickness h, and the bottom of the substrate is a ground (conducting) plane.

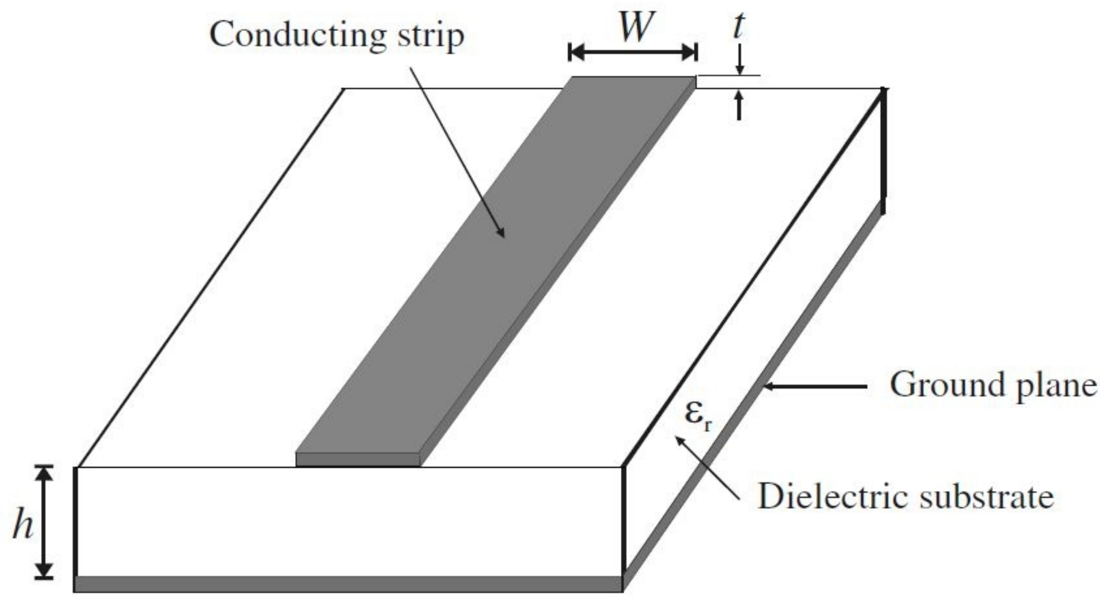


Fig. 5 General Microstrip Structure.

4.1 Waves in Microstrip:

The fields in the microstrip extend within two media air above and dielectric below so that the structure is inhomogeneous. Due to this inhomogeneous nature, the microstrip does not support a pure TEM wave. This is because that a pure TEM wave has only transverse components, and its propagation velocity depends only on the material properties, namely the permittivity and the permeability. However, with the presence of the two guided wave media (the dielectric substrate and the air), the waves in a microstrip line will have no vanished longitudinal components of electric and magnetic fields, and their propagation velocities will depend not only on the material properties, but also on the physical dimensions of the microstrip.

4.2 Quasi TEM Approximation:

When the longitudinal components of the fields for the dominant mode of a microstrip line remain very much smaller than the transverse components, they may be neglected. In this case, the dominant mode then behaves like a TEM mode, and the TEM transmission line theory is applicable for the microstrip line as well. This is called the quasi-TEM approximation and it is valid over most of the operating frequency ranges of microstrip.

4.3 Microstrip Losses:

The loss components of a single microstrip line include conductor loss, dielectric loss and radiation loss, while the magnetic loss plays a role only for magnetic substrates such as ferrites. The propagation constant on a lossy transmission line is complex; namely, $\gamma = \alpha + j\beta$, where the real part α in nepers per unit length is the attenuation constant, which is the sum of the attenuation constants arising from each effect. In practice, one may prefer to express α in decibels (dB) per unit length, which can be related by

$$\alpha \text{ (dB/unit length)} = (20 \log 10\ e)\ \alpha \text{ (nepers/unit length)}$$

$$\approx 8.686\ \alpha \text{ (nepers/unit length)}$$

A simple expression for the estimation of the attenuation produced by the conductor loss is given by

$$\alpha_c = \frac{8.6 \quad R_S}{Z_C\, W} \text{ dB/unit length}$$

in which Zc is the characteristic impedance of the microstrip of the width W, and Rs represents the surface resistance in ohms per square for the strip conductor and ground plane. For a conductor

$$Rs = \sqrt{\frac{\omega\, \mu_\circ}{2\, \sigma}}$$

where σ is the conductivity, $\mu_\circ$ is the permeability of free space, and ω is the angular frequency. The surface resistance of superconductors is expressed differently. Strictly speaking, the simple expression of α_c is only valid for large strip widths because it assumes that the current distribution across the microstrip is uniform, and therefore it would overestimate the conductor loss for narrower microstrip lines. Nevertheless, it may be found to be accurate enough in many practical situations, due to extraneous sources of loss, such as conductor surface roughness.

The attenuation due to the dielectric loss in microstrip can be determined by

$$\alpha_d = 8.686\,\pi \left(\frac{\varepsilon_r - 1}{\varepsilon_r - 1}\right) \frac{\varepsilon_r}{\varepsilon_r} \frac{\tan\delta}{\lambda_g} \quad \text{dB/unit length}$$

where tan δ denotes the loss tangent of the dielectric substrate. Since the microstrip is a semi open structure, any radiation is either free to propagate away or to induce currents on the metallic enclosure, causing the radiation loss or the so-called housing loss.

4.4 Microstrip Components:

Microstrip components, which are often encountered in microstrip filter designs, may include lumped inductors and capacitors, quasilumped elements (i.e., short line sections and stubs), and resonators. In most cases, the resonators are the distributed elements such as quarter-wavelength and half-wavelength line resonators. The choice of individual components may depend mainly on the types of filters, the fabrication techniques, the acceptable losses or *Q* factors, the power handling, and the operating frequency.

5. Hairpin Design:

Hairpin-line bandpass filters are compact structures. They may conceptually be obtained by folding the resonators of parallel-coupled, half-wavelength resonator filters, into a “U” shape. This type of “U” shape resonator is the so-called hairpin resonator. However, to fold the resonators, it is necessary to take into account the reduction of the coupled-line lengths, which reduces the coupling between resonators. Also, if the two arms of each hairpin resonator are closely spaced, they function as a pair of coupled line themselves, which can have an effect on the coupling as well. To design this type of filter more accurately, a design approach employing full-wave EM simulation will be required.

Out of various bandpass microstrip filters, Hairpin filter is one of the most preferred one. The following structure shows a typical hairpin Structure.

(a) (b)

Fig. 6: (a) tapped line input 5-pole Hairpin Filter. (b) coupled line input 5-pole Hairpin Filter.

5.1 Hairpin Resonator:

Fig. 6 shows a single Hairpin Resonator. α is called the slide angle. If the slide angle is small it might lead to coupling between the arms of individual resonator. The voltage at the end of hairpin arms is antiphase, and thus causes the arm to arm capacitance to have seemingly disproportionate effect. The added capacitance lowers the resonant frequency requiring a shortening of the hairpin to compensate. To avoid this, slide angle is kept as large as possible. But by increasing the slide angle the coupling length between two resonators reduces, so as to attain the required coupling, the coupling spacing needs to be reduced which posses a practical limitation. For practical design purpose slide angle is kept twice the strip width to avoid inter-element coupling.

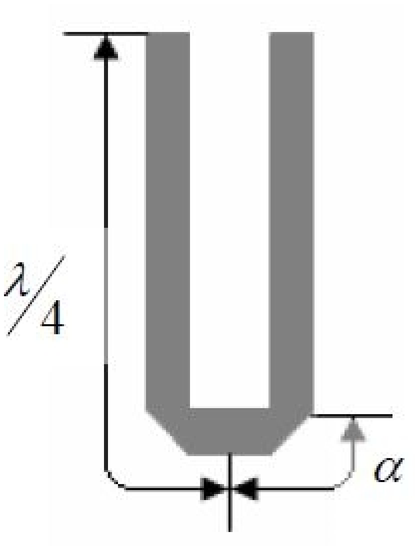


Fig. 7 Hairpin Resonator.

5.2 Tapped Line Input:

Conventional filters employ coupled line input. Tapped line input has a space saving advantage over coupled line input. Further while designing sometime the coupling dimensions required for the input and output coupled line is very small and practically not achievable which hinders the reliability of the design. Thus tapped line input is preferred over coupled line input.

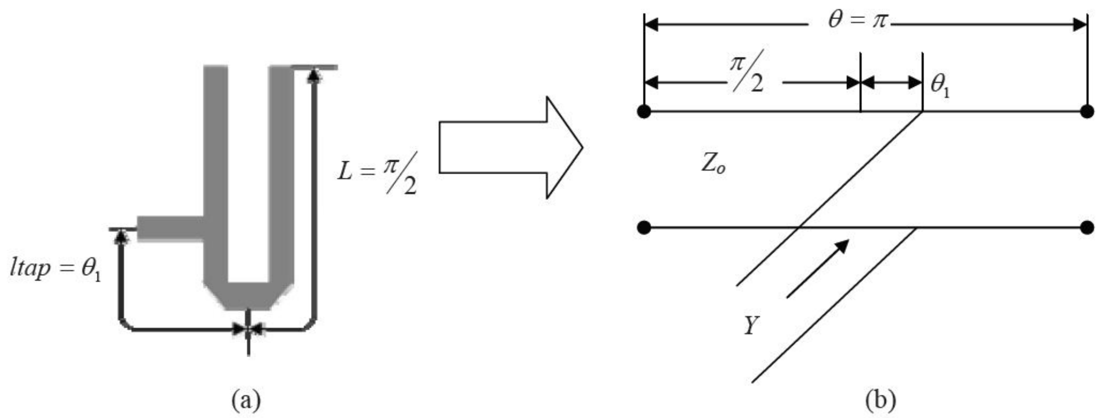


Fig. 8 (a) Tapped Hairpin Resonator Schematic. (b) Equivalent Circuit of a Tapped Hairpin Resonator.

5.3 Design Parameter For Hairpin Filter:

For Designing a Hairpin filter, Full Wave EM simulation is used. For the design purpose the low pass prototype (Butterworth, Chebyshev, Bessel) is selected according to the design requirement.

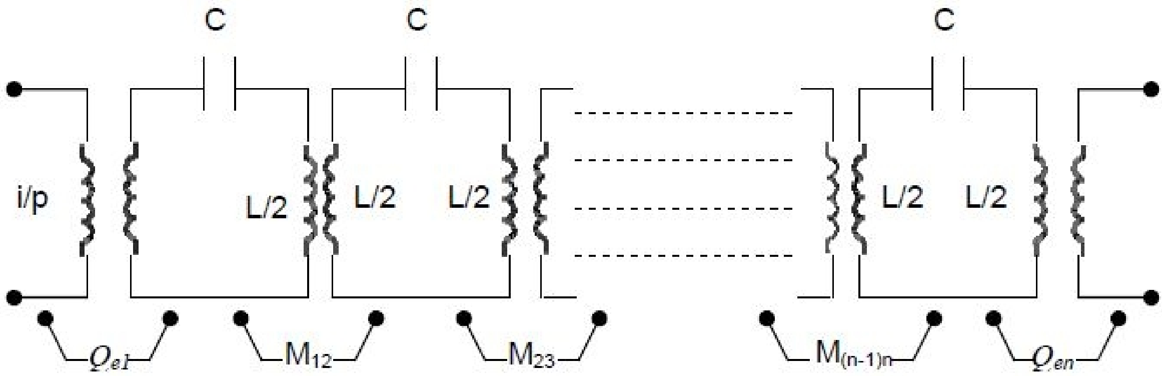


Fig. 9 Equivalent circuit of the n-pole Hairpin Bandpass Filter.

As seen from the equivalent circuit of n pole Hairpin filter, each resonator can be modelled as a combination of inductor and capacitor. The mutual coupling coefficient between two resonators is $M_{i,i+1}$. Q_{el} and Q_{en} are the Quality Factor at the input and output.

Coupling coefficient and Quality Factor can be calculated as

$$Q_{el} = \frac{g_0\, g_1}{\text{Fł}}$$

$$Q_{en} = \frac{g_n\, g_{n+1}}{\text{Fł}}$$

$$M_{i,i+1} = \frac{\text{Fł}}{\sqrt{g_i\, g_{i+1}}} \qquad \textit{for} \quad i = 1 \textit{ to } n\text{-}1$$

where FBW is the fractional bandwidth and $g_{0,1.....n+1}$ are the normalized lowpass element of the desired low pass filter approximation.

The quality factor can be substituted and the *ltap* length can be calculated as

$$ltap = \frac{2L}{\pi}\sin^{-1}\left(\sqrt{\frac{\pi}{2}\frac{Z_\circ/R}{Q_e}}\right)$$

$$2L = \frac{\lambda}{2} = \frac{\lambda}{2\sqrt{\varepsilon_r}} = \frac{C}{2f_\circ\sqrt{\varepsilon_r}}$$

6. 550-900 MHz Filter Bank for GMRT:

As a part of the ongoing upgrade to wideband the GMRT observatory, the front end receiver system is being modified to include wide band, high dynamic range LNAs, wide band filters and octave band polarizers with low insertion loss, amongst other improvements.

The report basically emphasis on designing of wideband filters for the 550-900 MHz observing band of GMRT. The modified Front-End system is shown in fig below.

<u>Front End Box filter location and specification for 550 – 900 MHz</u>

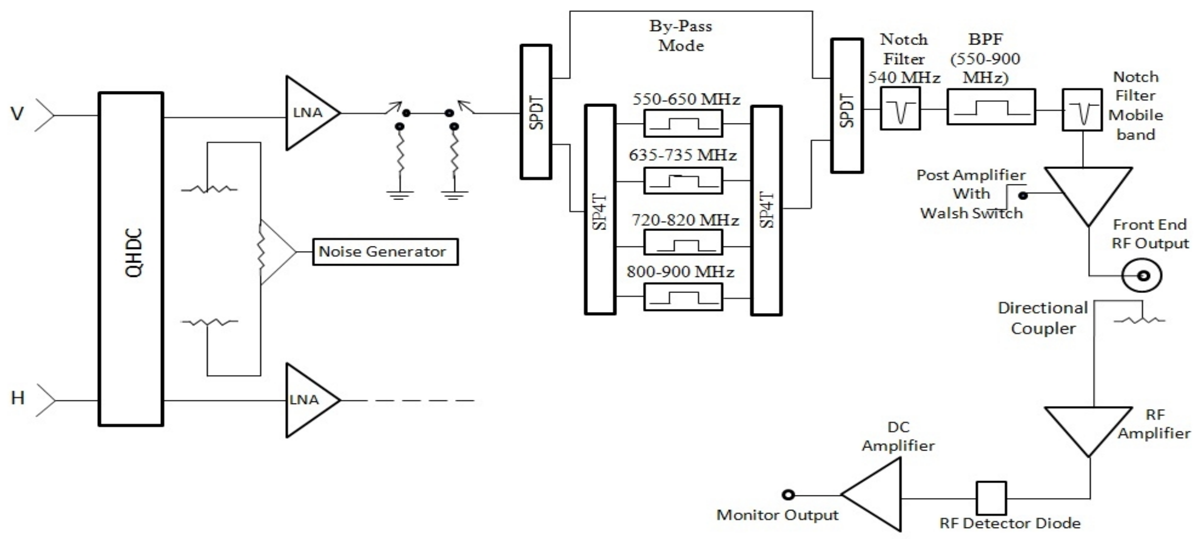


Fig. 10 Upgraded Front-End System 550-900 MHz block diagram.

The block diagram shows a bypass path and switchable sub-band filters in the other path. This type of arrangement gives the freedom to observe either with full bandwidth of 350 MHz by selecting the bypass mode or can observe with 100 MHz bandwidth by selecting any of the sub-band filters. These filters can be selected one at a time by using the 2:1 and 4:1 switch combination. The 2:1 switch switches between the bypass mode and the sub-band filters and the 4:1 switch is used to select any one of the four sub-band filter. The main band pass filter 550-900 MHz along with the notch filters always remains in the RF path even for bypass mode as well as for the sub-band filters.

7. Design Specification:

The design specification given for the full band and sub-band filters are as follows:

6.1. 550-900 MHz Band Pass Filter:

Filter Specification	Values
Centre Frequency	725 MHz
Insertion Loss	4 dB
Lower Cut-off Frequency	550 MHz
Upper Cut-off Frequency	900 MHz
Bandwidth	350 MHz
Passband Ripple	0.1 dB
Min. Attenuation	-40 dB at 10% from band-edge Frequency

6.2. 550-650 MHz Sub-Band Band Pass Filter:

Filter Specification	Values
Centre Frequency	600 MHz
Insertion Loss	4 dB
Lower Cut-off Frequency	550 MHz
Upper Cut-off Frequency	560 MHz
Bandwidth	100 MHz
Passband Ripple	0.1 dB
Min. Attenuation	-40 dB at 10% from band-edge Frequency

6.3. 635-735 MHz Sub-Band Band Pass Filter:

Filter Specification	Values
Centre Frequency	685 MHz
Insertion Loss	4 dB
Lower Cut-off Frequency	635 MHz
Upper Cut-off Frequency	735 MHz
Bandwidth	100 MHz
Passband Ripple	0.1 dB
Min. Attenuation	-40 dB at 10% from band-edge Frequency

6.4. 720-820 MHz Sub-Band Band Pass Filter:

Filter Specification	Values
Centre Frequency	770 MHz
Insertion Loss	4 dB
Lower Cut-off Frequency	720 MHz
Upper Cut-off Frequency	820 MHz
Bandwidth	100 MHz
Passband Ripple	0.1 dB
Min. Attenuation	-40 dB at 10% from band-edge Frequency

6.5. 800-900 MHz Sub-Band Band Pass Filter:

Filter Specification	Values
Centre Frequency	850 MHz
Insertion Loss	4 dB
Lower Cut-off Frequency	800 MHz
Upper Cut-off Frequency	900 MHz
Bandwidth	100 MHz
Passband Ripple	0.1 dB
Min. Attenuation	-40 dB at 10% from band-edge Frequency

7. Design Procedure:

The procedure for designing the Bandpass Hairpin filter basically comprises of 6 steps which are as follows:

1. **Filter Specification:**
   The design specification for the filter has to be properly defined.

2. **Determination of filter order:**
   The next step is to find the order of the filter which depends on the design specification for the filter.

3. **Determination of low pass filter prototype element:**
   After defining the order the next step is to find the element values for the normalised low pass filter.

4. **Low pass to Bandpass transformation:**
   The forth step is to transform the element values of the normalised low pass filter into desired bandpass filter using normalised low pass to bandpass transformation.

5. **Determination of width, spacing and length for the hairpin resonators:**
   The next step is to find the width and length of the transmission line for the given frequency and property of substrate. And spacing or coupling between the hairpin resonators by using the element values.

6. **Implementation and simulation of designed bandpass hairpin filter:**
   The final step is to implement and simulate the theoretically designed bandpass hairpin filter in the simulation software and later tune the filter to obtain the desired results.

8. Schematic Circuit and PCB Layout:

The filters designed are Microstrip based Hairpin Filter Design. The filters are designed on Rogers 6010 LM substrate having a dielectric constant of 10.2. The substrate chosen has a very low loss tangent of about 0.0023 and has a high dielectric constant which helps in reducing the filter size and also helps in reducing the microstrip losses. Depending on the specification given the filter designed are 7^{th} order filters. The 5^{th} order would also have given a good frequency response but in order to achieve a sharper roll-off of about 40 dB from band edges frequency the 7^{th} order filters were designed. The circuit and PCB layout along with the schematic and momentum simulation of full-band band pass filter and sub-band switchable filter bank were done on Agilent Advance Design System (ADS). These filters are finely tuned and optimized in ADS to meet the given specification exactly. Being a microstrip design the schematic circuit looks exactly the same as the fabricated PCB. The chassis for these filters are made with the help of Mechanical Dept. and Workshop. The fabricated layout of full-band BPF and sub-band switched filter bank is shown in fig. 11 & 12.

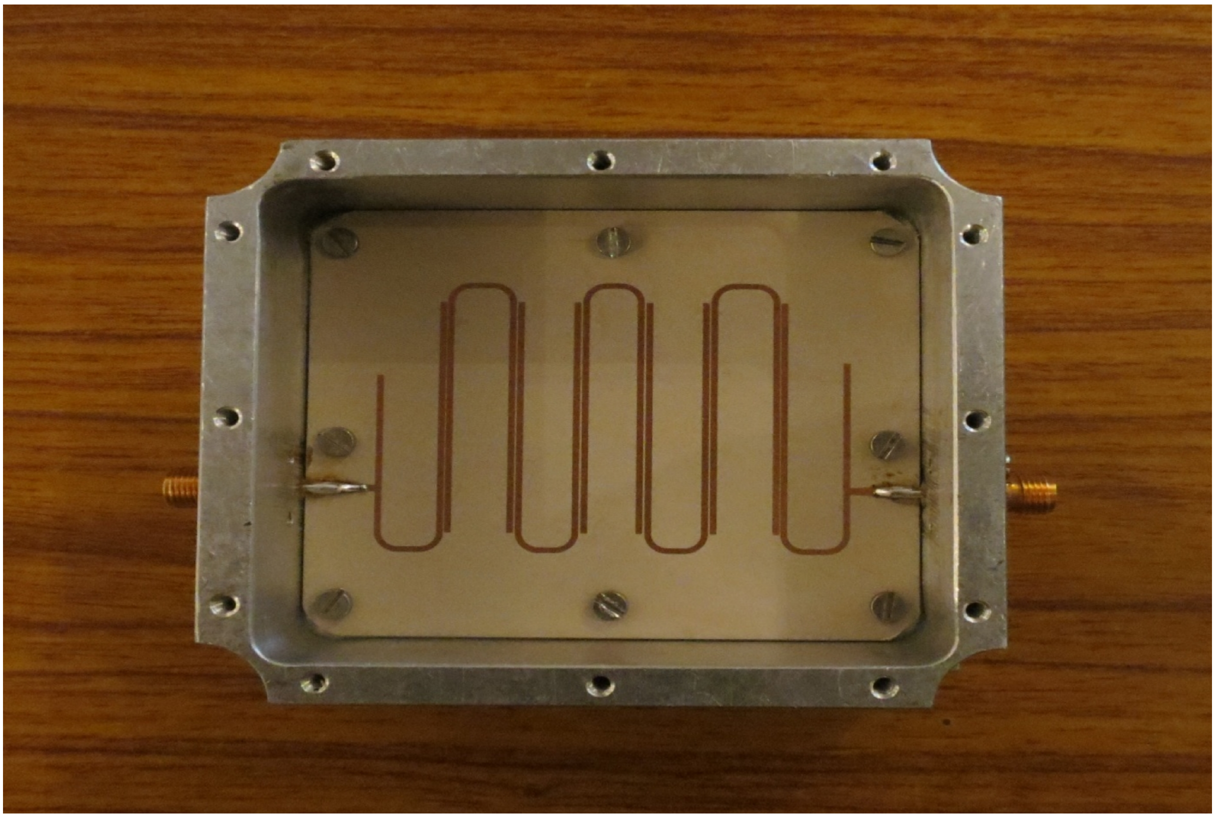


Fig. 11 550-900 MHz Full-Band Band Pass Filter

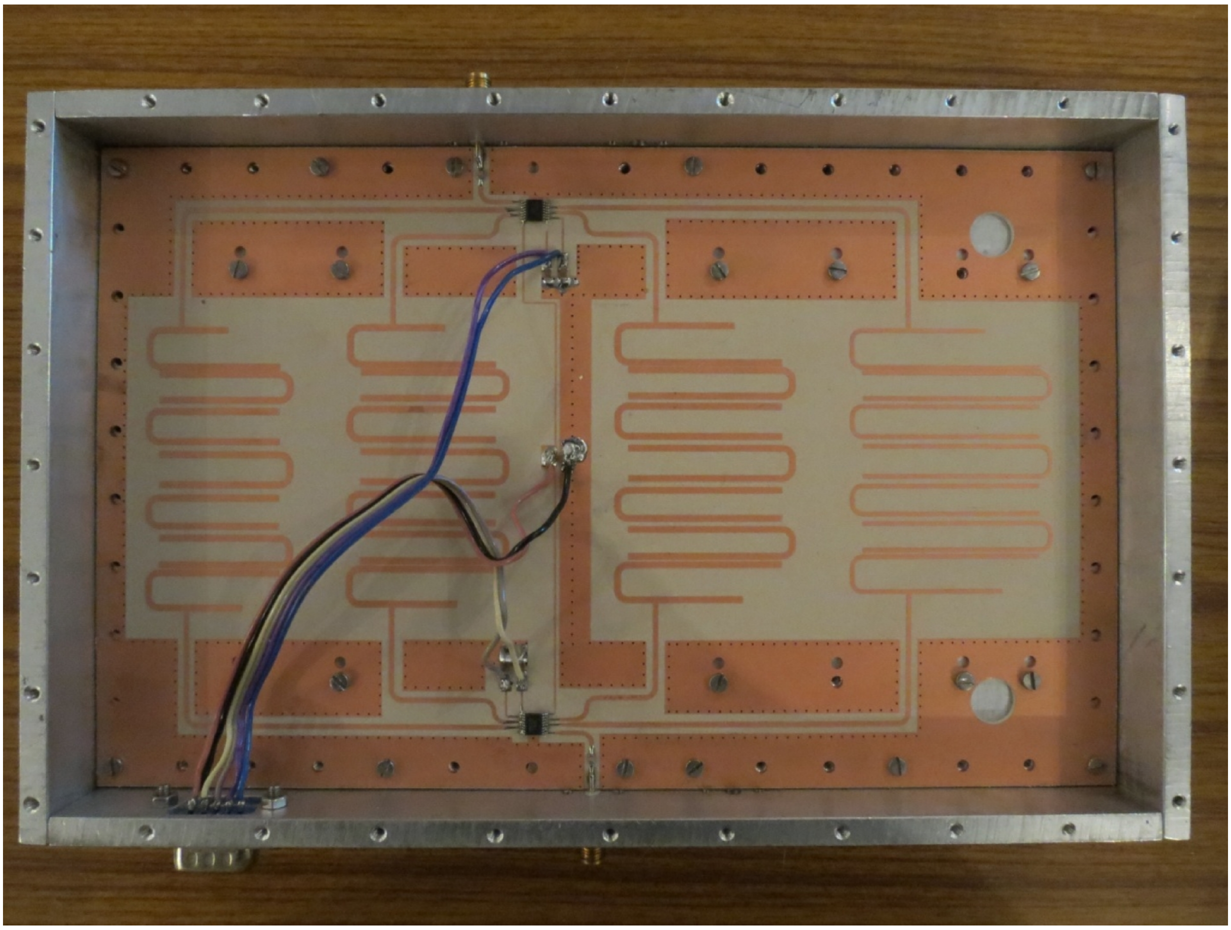


Fig. 12 Switchable Filter Bank with 4 Sub-Band Band Pass Filter

9. Evolution of the Final product:

Struggle & Learning

Initially the prototype of all the individual sub-band band-pass filters were designed and fabricated for the given specifications to have them as a reference and see the effect on their performance after combining all of them on a single board in switchable configuration. The first prototypes of all these filters were found to have their overall response shifted right on the frequency axis by 10-15 MHz. However all other measured parameters like insertion loss, bandwidth, input and output return loss etc. were matching the desired specifications. The frequency shift was later corrected by increasing the length of the hairpin resonators by 1 mm to 1.5 mm.

Later all these sub-band filters were combined on a single board as switched filter bank using Hittite 4:1 RF switches HMC241QS16. The combined structure is simulated along with the switches in ADS and the results showed increase in insertion loss by 2 dB due to the additional loss of the RF switches. Then the PCB for switched filter bank (Filter-Bank Ver.1) was fabricated. All the measured parameters were found matching with the specified values. However a little shift in frequency response was noticed. This shift was caused by increase in the length of tapped lines for connecting the individual filters to RF switch on either side of the filter bank. Also the bandwidths of all the sub-band filters were little more than the specified bandwidth of 100 MHz which was corrected by tuning of the coupling between the resonators.

With the above corrections a PCB for Filter-Bank Ver.2 was fabricated. In this version the grounding pads were added on PCB around the filter circuits and along the tapped lines to improve the band shape. This improved grounding reduces the coupling and the capacitive effect between the two closely spaced tapped lines. A coupling capacitor of 220 microfarad is also added in series with input and output RF path in order to improve the coupling effects. The measured results for this version of circuit were found to match very closely to the specified values for all the parameters. Hence design was concluded.

10. Simulated and Measured Result:

The simulated results and the fabricated or measured results are shown in fig. below. A shift in frequency can be seen in simulated and the measured result. This is due to the di-electric constant of the substrate and thickness of the copper which is not even over the entire area of the substrate. Therefore the filters are shifted by a certain amount in the simulated results so that the measured result matches the given specifications. The S-parameter measurements were done on the filter units using the network analyser in Front-end lab. The insertion loss for these filters are found to 1.5 dB whereas in sub-band filters the addition 2 dB loss has been added due to the 4:1 RF switches which are used at the input and the output of the filter bank. The return loss found is well below 10 dB. And they also offer a very good rejection at 540 MHz and at 850 MHz & 900 MHz.

10.1. 550-900 MHz Full Band BPF:

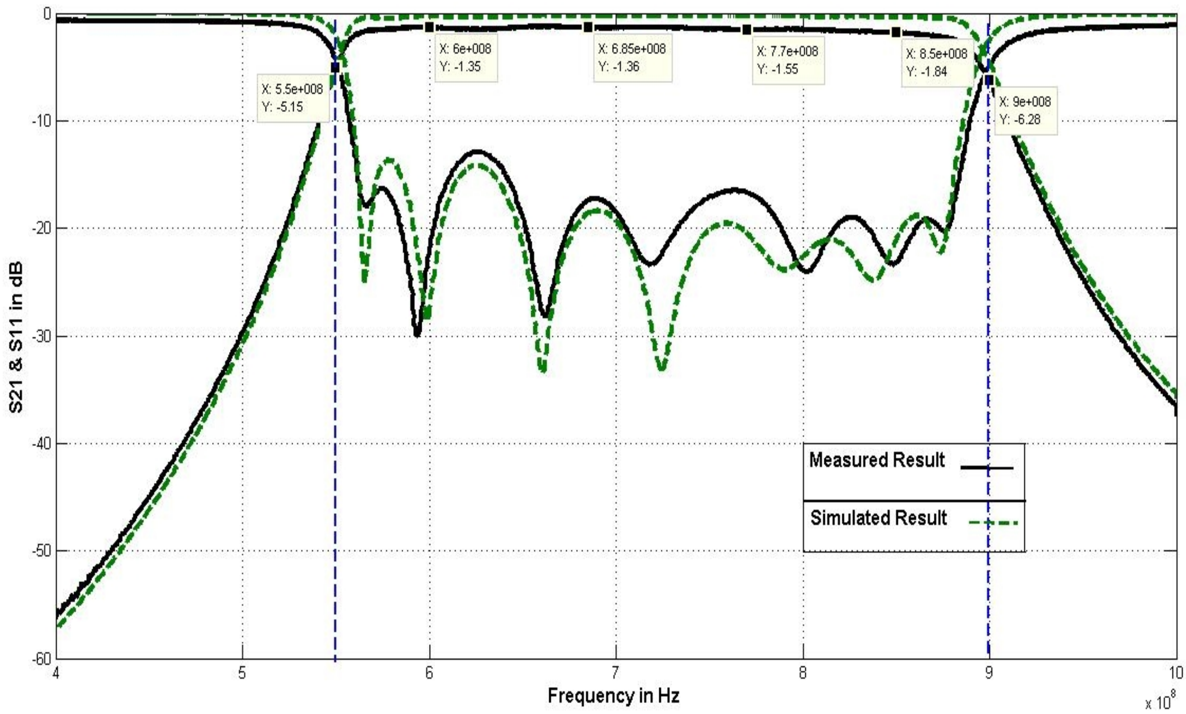


Fig. 13 Simulated Vs Measured Response

10.2. 550-650 MHz Sub-Band BPF* :

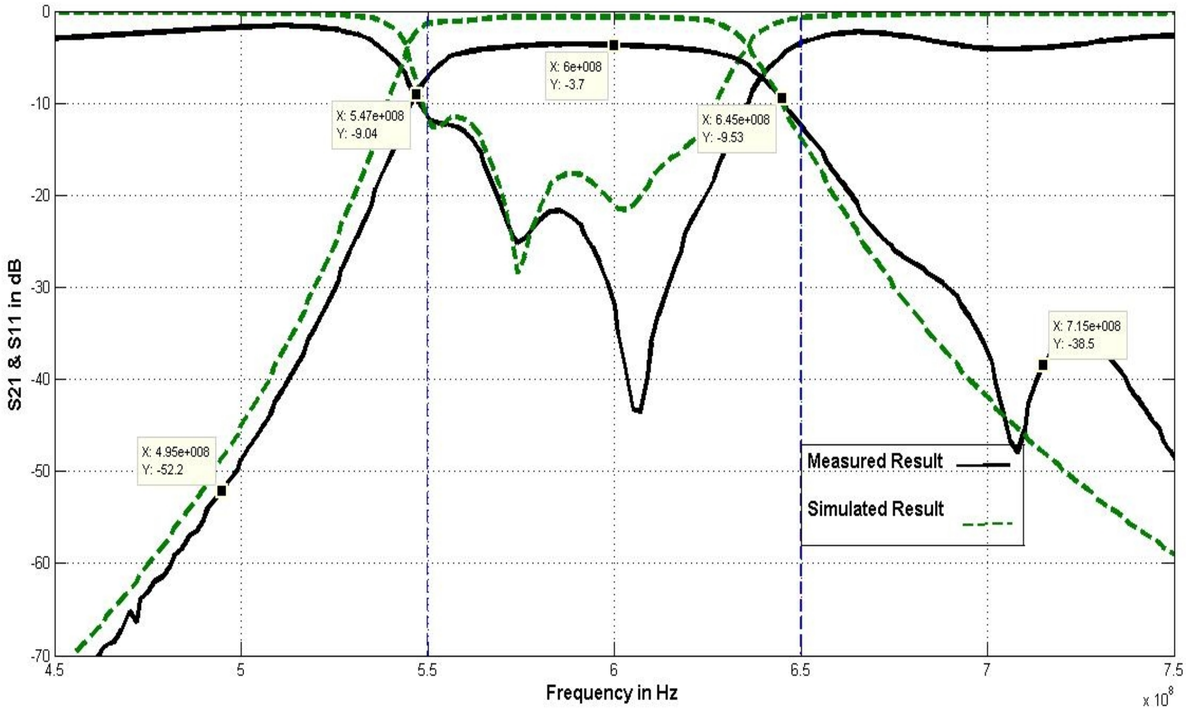


Fig. 14 Simulated Vs Measured Response

10.3. 635-735 MHz Sub-Band BPF* :

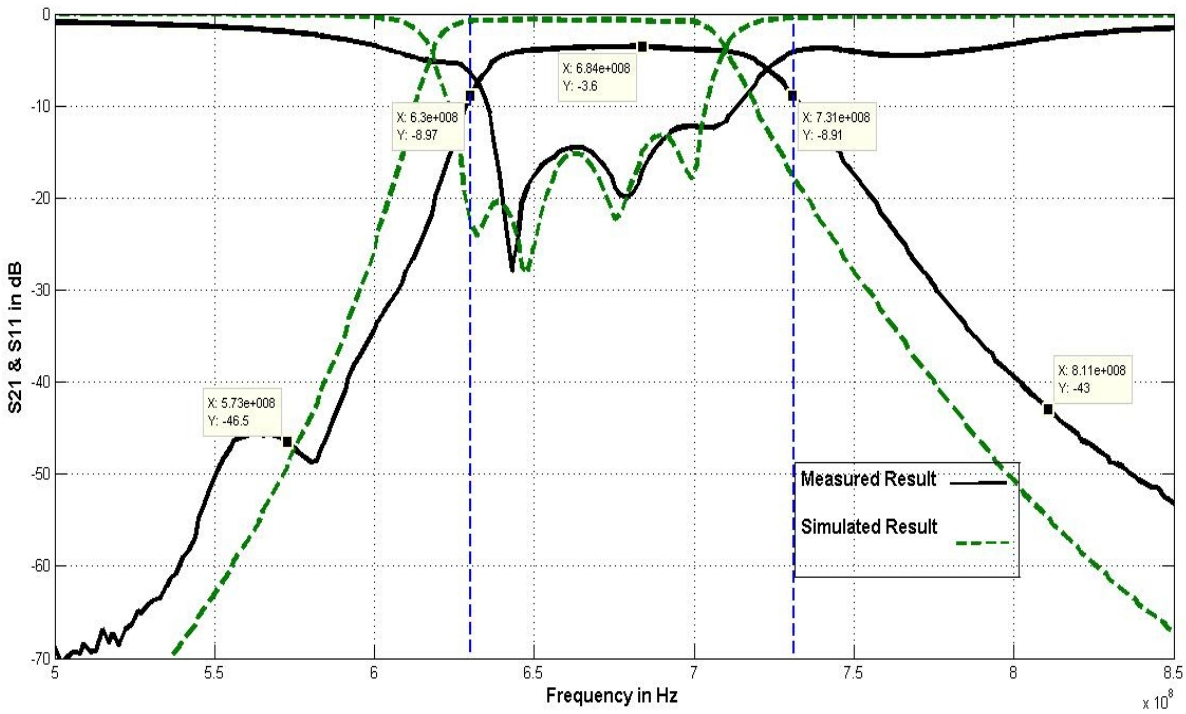


Fig. 15 Simulated Vs Measured Response

10.4. 720-820 MHz Sub-Band BPF* :

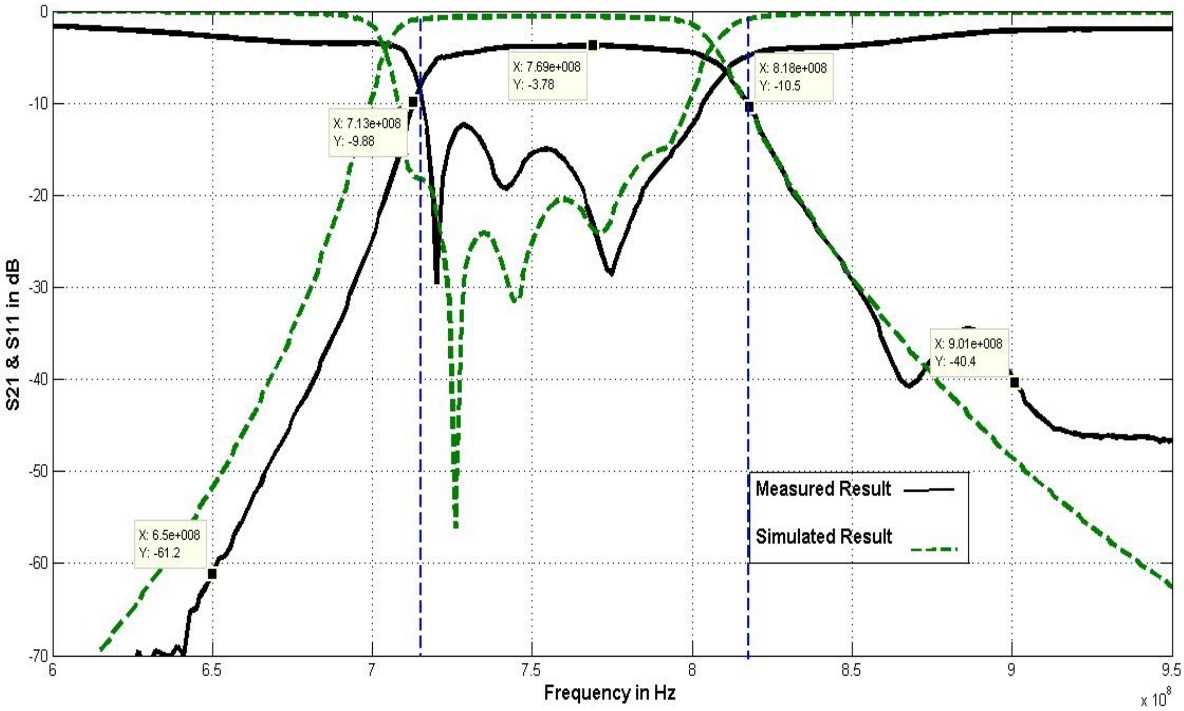


Fig. 16 Simulated Vs Measured Response

10.5. 800-900 MHz Sub-Band BPF* :

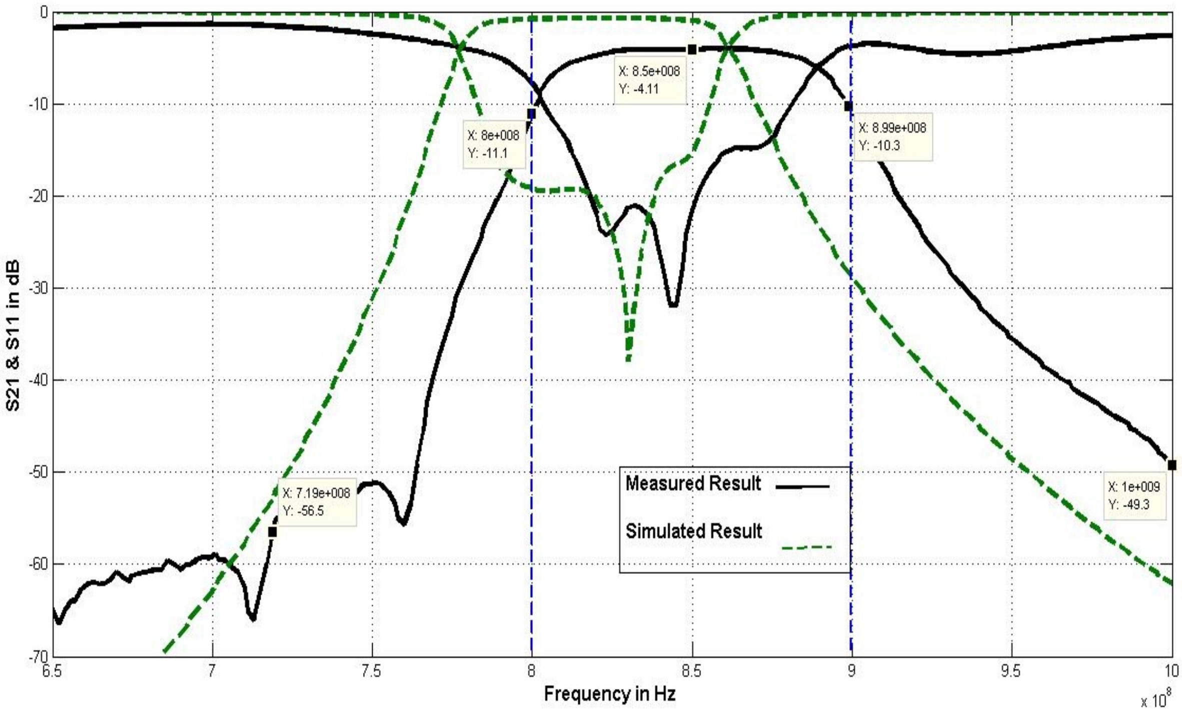


Fig. 17 Simulated Vs Measured Response

*** Note:** The difference in the insertion loss is due to the additional loss added by the RF switches.

10.6. Combined Response of Full Band And Sub bands BPF Filters:

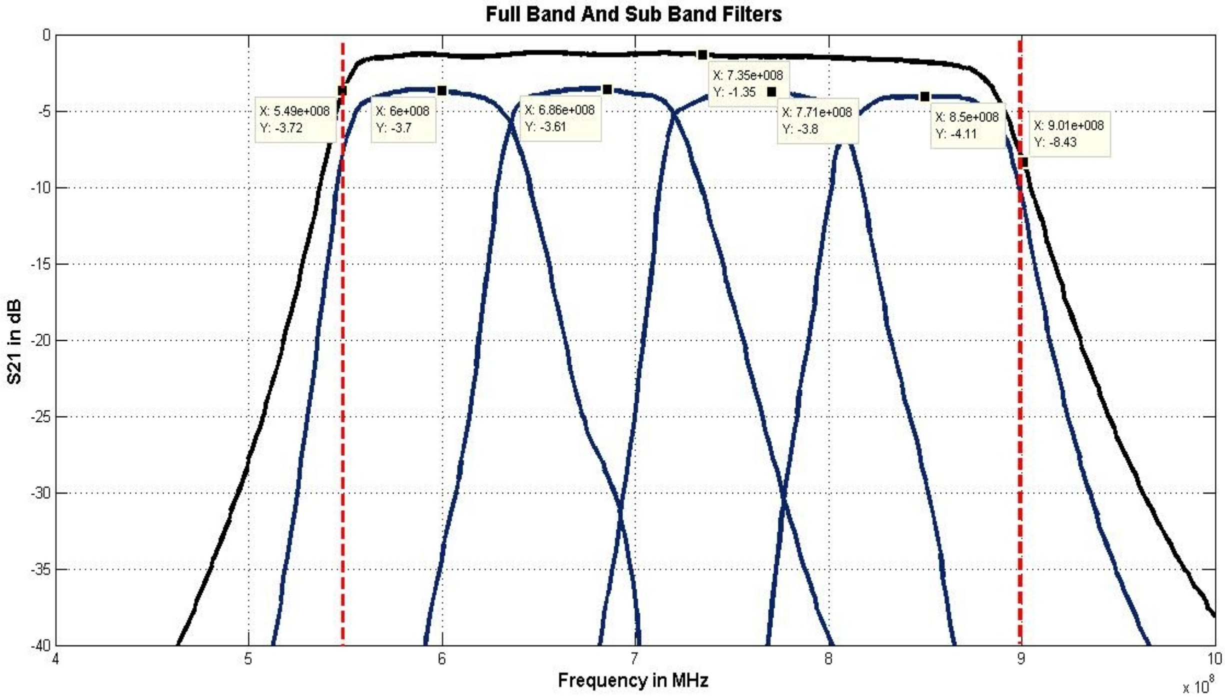


Fig. 18 Combine plot of Full Band BPF with all Sub Band BPF

10.7. Comparison Summary of Simulated & Measured Results:

Filters		6 dB cutoff Points (MHz)	Bandwidth (MHz)	Insertion Loss (dB)
550-900 MHz Full Band BPF	Simulated	548 & 905	357	0.5
	Measured	547 & 903	356	1.2
550-650 MHz Sub Band BPF	Simulated	542 & 640	108	0.6
	Measured	545 & 648	103	3.5
635-735 MHz Sub Band BPF	Simulated	615 & 715	100	0.6
	Measured	630 & 735	105	3.5
720-820 MHz Sub Band BPF	Simulated	708 & 810	102	0.6
	Measured	715 & 818	103	3.7
800-900 MHz Sub Band BPF	Simulated	775 & 865	90	0.7
	Measured	800 & 900	100	4

11. Conclusion:

The filters designed are 7th order Microstrip Hairpin Design filter. The Full-Band BPF filter is found to have an insertion loss (S21) of around 1.5 dB and return loss (S11) well below 10 dB over the desired frequency range of 550-900 MHz band. The Sub-band Filters have a insertion loss of around 3.5 dB to 4 dB (2 dB to 2.5 dB loss is due to the switch used for switching) and having return loss less than 10 dB. All the filters offer a very good rejection of -40 dB at 10% from band edges frequency. Each filter also offers a very good rejection at known radio frequency interference (540 MHz, 800 MHz & 900 MHz). These filters designed meet the given specifications very well.

12. Reference:

1. T. C. Edwards & M. B. Steer "Foundation of Interconnect and Microstrip design".
2. Jia-Sheng Hong & M. J. Lancaster "Filters for RF/Microwave Application".
3. David M. Pozar "Microwave Engineering".
4. Sandeep C. Chaudhari, "Modified 150MHz Front-End System Incorporating Filters For RFI Mitigation".
5. Aarti Sandikar, Manisha Parate & V.B. Bhalerao, "Upgraded Broadband 300-500 MHz Front-End System.
6. Vinod Toshniwal, "RFI Rejection Filter At 150 MHz", STP project report.
7. Bhalerao V. B., 2010,Wide Bandpass filter for 327 MHz Front-End GMRT Reciever.
8. Wikipedia.

Appendix

ROGERS CORPORATION

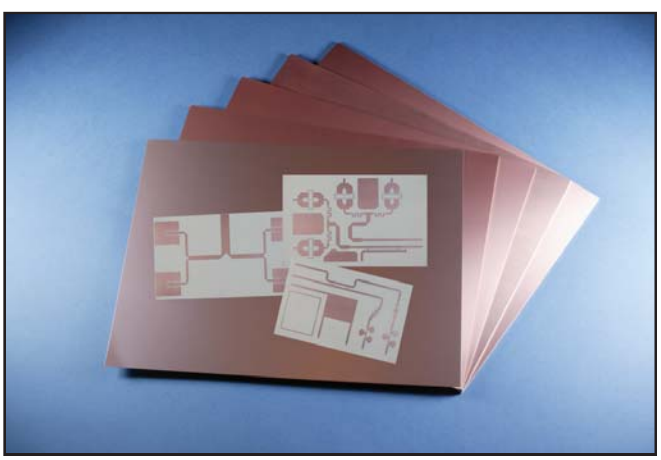
Advanced Circuit Materials

Advanced Circuit Materials Division
100 S. Roosevelt Avenue
Chandler, AZ 85226
Tel: 480-961-1382, Fax: 480-961-4533
www.rogerscorp.com

Data Sheet
1.6000

RT/duroid® 6006/6010LM High Frequency Laminates

Features	
•	High dielectric constant for circuit size reduction.
•	Low loss. Ideal for operating at X-band or below.
•	Low Z-axis expansion for RT/duroid 6010LM. Provides reliable plated through holes in multilayer boards.
•	Low moisture absorption for RT/duroid 6010LM. Reduces effects of moisture on electrical loss.
•	Tight ε_r and thickness control for repeatable circuit performance.
Some Typical Applications	
•	Space Saving Circuitry
•	Patch Antennas
•	Satellite Communications Systems
•	Power Amplifiers
•	Aircraft Collision Avoidance Systems
•	Ground Radar Warning Systems



RT/duroid® 6006/6010LM microwave laminates are ceramic-PTFE composites designed for electronic and microwave circuit applications requiring a high dielectric constant. RT/duroid 6006 laminate is available with a dielectric constant value of 6.15 and RT/duroid 6010LM laminate has a dielectric constant of 10.2.

RT/duroid 6006/6010LM microwave laminates feature ease of fabrication and stability in use. They have tight dielectric constant and thickness control, low moisture absorption, and good thermal mechanical stability.

RT/duroid 6006/6010LM laminates are supplied clad both sides with ¼ oz. to 2 oz./ft^2 (8.5 to 70 µm) electrodeposited copper foil. Cladding with rolled copper foil is also available. Thick aluminum, brass, or copper plate on one side may be specified.

Standard tolerance dielectric thicknesses of 0.010”, 0.025”, 0.050”, 0.075”, and 0.100” (0.254, 0.635, 1.270, 1.905, 2.54 mm) are available. When ordering RT/duroid 6006 and RT/duroid 6010LM laminates, it is important to specify dielectric thickness, electrodeposited or rolled, and weight of copper foil required.

Property	Typical Value [2]		Direction	Units [1]	Condition	Test Method
	RT/duroid 6006	RT/duroid 6010.2LM				
[3]Dielectric Constant ε_r *Process*	6.15± 0.15	10.2 ± 0.25	Z		10 GHz 23°C	IPC-TM-650 2.5.5.5 Clamped stripline
[4]Dielectric Constant ε_r *Design*	6.45	10.9	Z		8 GHz - 40 GHz	Differential Phase Length Method
Dissipation Factor, tan δ	0.0027	0.0023	Z		10 GHz/A	IPC-TM-650 2.5.5.5
Thermal Coefficient of ε_r	-410	-425	Z	ppm/°C	-50 to 170°C	IPC-TM-650 2.5.5.5
Surface Resistivity	$7X10^7$	$5X10^6$		Mohm	A	IPC 2.5.17.1
Volume Resistivity	$2X10^7$	$5X10^5$		Mohm•cm	A	IPC 2.5.17.1
Youngs' Modulus						ASTM D638 (0.1/min. strain rate)
under tension	627 (91) 517 (75)	931 (135) 559 (81)	X Y	MPa (kpsi)	A	
ultimate stress	20 (2.8) 17 (2.5)	17 (2.4) 13 (1.9)	X Y	MPa (kpsi)	A	
ultimate strain	12 to 13 4 to 6	9 to 15 7 to 14	X Y	%	A	
Youngs' Modulus						ASTM D695 (0.05/min. strain rate)
under compression	1069 (155)	2144 (311)	Z	MPa (kpsi)	A	
ultimate stress	54 (7.9)	47 (6.9)	Z	MPa (kpsi)	A	
ultimate strain	33	25	Z	%		
Flexural Modulus	2634 (382) 1951 (283)	4364 (633) 3751 (544)	X	MPa (kpsi)	A	ASTM D790
ultimate stress	38 (5.5)	36 (5.2) 32 (4.4)	X Y	MPa (kpsi)	A	
Deformation under load	0.33 2.10	0.26 1.37	Z Z	%	24 hr/ 50°C/7MPa 24 hr/150°C/7MPa	ASTM D621
Moisture Absorption	0.05	0.01		%	D48/50°C, 0.050" (1.27mm) thick	IPC-TM-650, 2.6.2.1
Density	2.7	3.1				ASTM D792
Thermal Conductivity	0.49	0.86		W/m/°K	80°C	ASTM C518
Thermal Expansion	47 34, 117	24 24,47	X Y,Z	ppm/°C	0 to 100°C	ASTM 3386 (5K/min)
Td	500	500		°C TGA		ASTM D3850
Specific Heat	0.97 (0.231)	1.00 (0.239)		J/g/K (BTU/lb/°F)		Calculated
Copper Peel	14.3 (2.5)	12.3 (2.1)		pli (N/mm)	after solder float	IPC-TM-650 2.4.8
Flammability Rating	V-0	V-0				UL94
Lead-Free Process Compatible	Yes	Yes				

[1] SI unit given first with other frequently used units in parentheses.
[2] References: APR4022.33 DJS 4019.27-32, Internal TR 2610. Tests were at 23°C unless otherwise noted.
[3] Dielectric constant is based on .025 dielectric thickness, one ounce electrodeposited copper on two sides.
[4] The design Dk is an average number from several different tested lots of material and on the most common thickness/s. If more detailed information is required, please contact Rogers Corporation. Refer to Rogers' technical paper "Dielectric Properties of High Frequency Materials" available at http://www.rogerscorp.com/acm.

Typical values are a representation of an average value for the population of the property. For specification values contact Rogers Corporation.

STANDARD THICKNESS:	STANDARD PANEL SIZE:	STANDARD COPPER CLADDING:
0.005" (0.127mm) 0.010" (0.254mm) 0.025" (0.635mm) 0.050" (1.27mm) 0.075" (1.90mm) 0.100" (2.50mm)	10" X 10" (254 X 254mm) 10" X 20" (254 X 508mm) 20" X 20" (508 X 508mm) *18" X 12" (457 X 305 mm) *18" X 24" (457 X 610 mm) (*note: the above 2 panel sizes are not available in the 0.005" (0.127mm) and 0.010" (0.254mm) thicknesses)	¼ oz. (8.5 µm) electrodeposited copper foil. ½ oz. (18 µm), 1 oz. (35µm), 2 oz. (70µm) electrodeposited and rolled copper foil. Heavy metal claddings are available. Contact Rogers' Customer Service.

The information in this data sheet is intended to assist you in designing with Rogers' circuit material laminates. It is not intended to and does not create any warranties express or implied, including any warranty of merchantability or fitness for a particular purpose or that the results shown on this data sheet will be achieved by a user for a particular purpose. The user should determine the suitability of Rogers' circuit material laminates for each application.

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Revised 03/2011 0938-0111-.5CC **Publication: #92-105**

MICROWAVE CORPORATION v04.0404

HMC241QS16 / 241QS16E

GaAs MMIC SP4T NON-REFLECTIVE SWITCH, DC - 3.5 GHz

RoHS EARTH FRIENDLY

Typical Applications

The HMC241QS16 & HMC241QS16E are ideal for:

- Base Stations & Portable Wireless
- CATV / DBS
- Wireless Local Loop
- Test Equipment

Features

RoHS Compliant Product

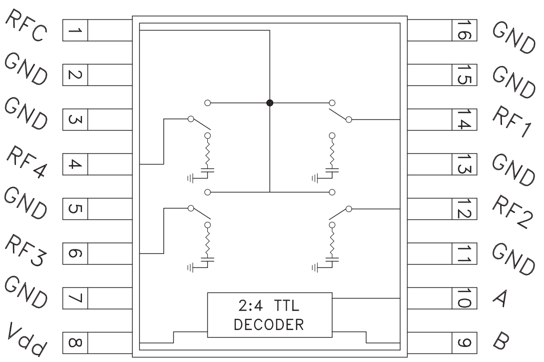
Low Insertion Loss (2 GHz): 0.5 dB

Single Positive Supply: Vdd = +5V

Integrated 2:4 TTL Decoder

16 Lead QSOP Package

Functional Diagram



General Description

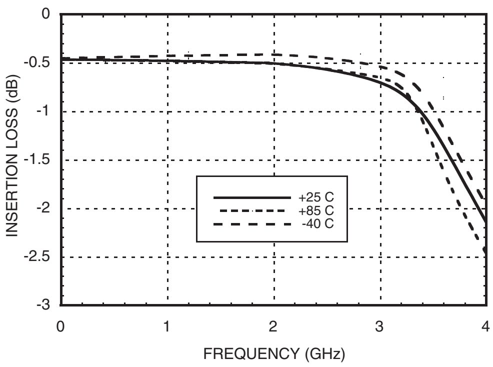
The HMC241QS16 & HMC241QS16E are general purpose low-cost non-reflective SP4T switches in 16-lead QSOP packages. Covering DC - 3.5 GHz, this switch offers high isolation and has a low insertion loss of 0.5 dB at 2 GHz. The switch offers a single positive bias and true TTL/CMOS compatibility. A 2:4 decoder is integrated on the switch requiring only 2 control lines and a positive bias to select each path, replacing 8 control lines normally required by GaAs SP4T switches.

Electrical Specifications, T_A = +25° C, For TTL Control and Vdd = +5V in a 50 Ohm System

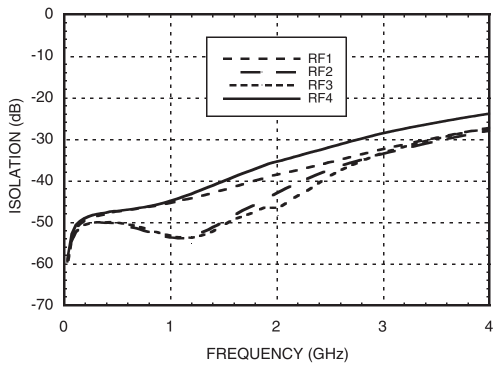
Parameter	Frequency	Min.	Typ.	Max.	Units
Insertion Loss	DC - 1.0 GHz DC - 2.0 GHz DC - 2.5 GHz DC - 3.5 GHz		0.5 0.5 0.6 1.0	0.8 0.8 0.9 1.5	dB dB dB dB
Isolation	DC - 1.0 GHz DC - 2.0 GHz DC - 2.5 GHz DC - 3.5 GHz	40 32 28 23	45 36 32 26		dB dB dB dB
Return Loss "On State"	DC - 2.5 GHz DC - 3.5 GHz	17 9	21 12		dB dB
Return Loss RF1-4 "Off State"	0.3 - 3.5 GHz 0.5 - 2.5 GHz	8 12	12 16		dB dB
Input Power for 1dB Compression	0.3 - 3.5 GHz	22	25		dBm
Input Third Order Intercept (Two-Tone Input Power = +7 dBm Each Tone)	0.3 - 3.5 GHz	40	44		dBm
Switching Characteristics tRISE, tFALL (10/90% RF) tON, tOFF (50% CTL to 10/90% RF)	0.3 - 3.5 GHz		 40 150		 ns ns

GaAs MMIC SP4T NON-REFLECTIVE SWITCH, DC - 3.5 GHz

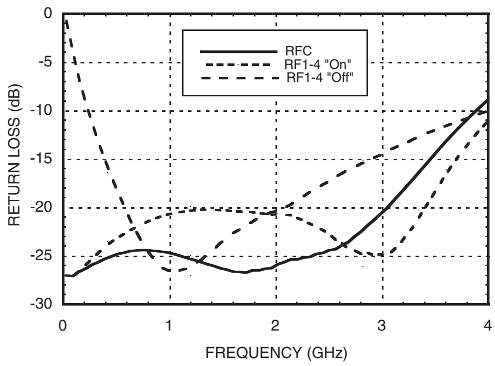
Insertion Loss



Isolation



Return Loss



Bias Voltage & Current

Vdd Range = +5.0 Vdc ± 10%		
Vdd (Vdc)	Idd (Typ.) (mA)	Idd (Max.) (mA)
+5.0	4.0	7.0

Truth Table

Control Input		Signal Path State
A	B	RFCOM to:
LOW	LOW	RF1
HIGH	LOW	RF2
LOW	HIGH	RF3
HIGH	HIGH	RF4

TTL/CMOS Control Voltages

State	Bias Condition
Low	0 to +0.8 Vdc @ 5uA Typ.
High	+2.0 to +5.0 Vdc @ 70 uA Typ.

NOTE:
DC Blocking capacitors are required at ports RFC and RF1, 2, 3, 4.

ALL DIMENSIONS ARE IN mm.

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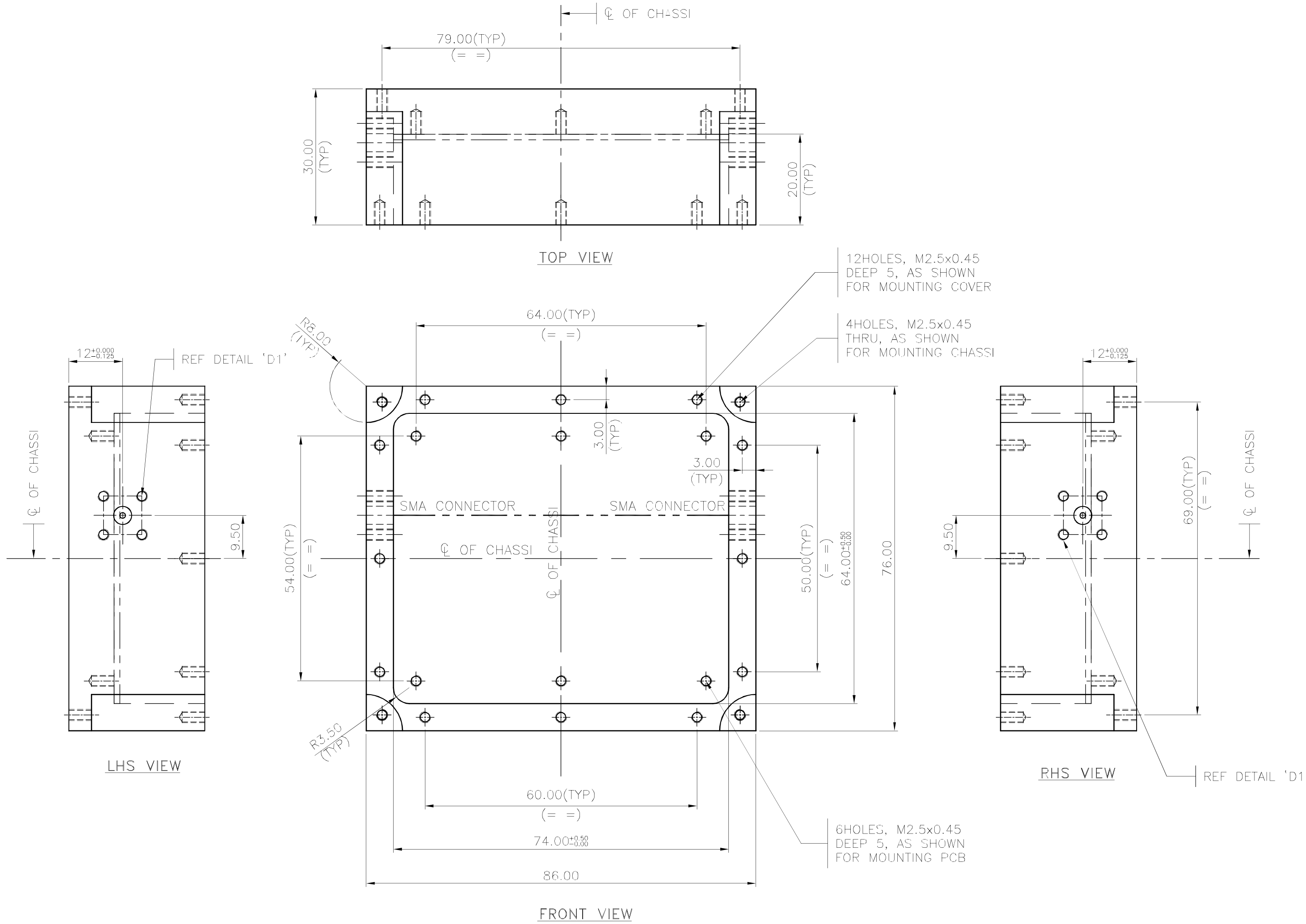
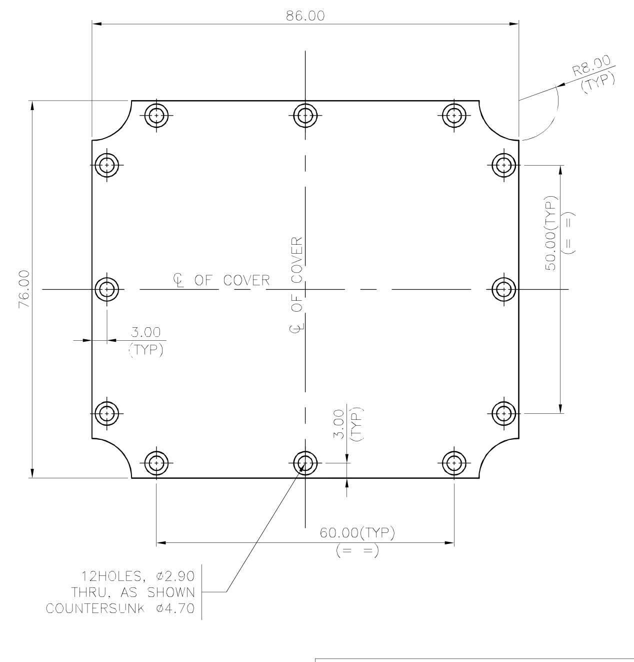
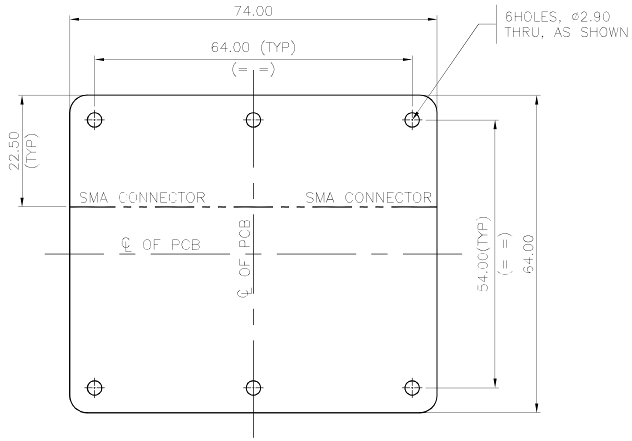




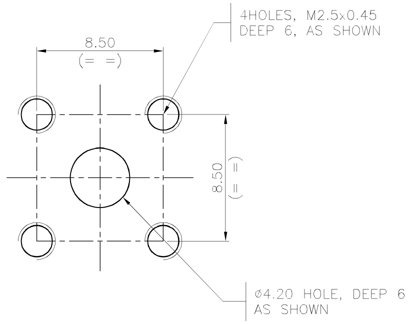
TITLE: CHASSI COVER

SCALE:- 1:1



TITLE: PCB SKETCH

SCALE:- 1:1



TITLE: DETAIL 'D1' (SMA CONNECTOR)

SCALE:- 3:1

NOTES

1) ALL DIMENSIONS ARE IN mm
2) MATERIAL : ALUMINIUM
3) CHASSI WEIGHT : 0.27KG (APPROX)
4) CHASSI COVER : 2mmTHK
5) PCB : 74x64x1.25mmTHK
6) SMA CONNECTOR : 2nos.
7) REMOVE BURRS & SHARP CORNERS.

R.NO	DATE	MODIFICATION	CHKD

REQ NO:- FRO549	DEPARTMENT:- FRONT END
REQ DATE:- 21.06.13	NAME OF REQ:- IMRAN KHAN
DATE OF COMP:- 05.07.13	APPD BY:- S.SURESHKUMAR

GIANT METERWAVE RADIO TELESCOPE
TATA INSTITUTE OF FUNDAMENTAL RESEARCH
KHODAD, P.B.NO:-6, NARAYANGAON, TALUKA-JUNNAR, PUNE:-410504

- DRAWING NOT TO BE SCALED.
- FILLET SHARP EDGES.
- ALL DIMENSIONS ARE IN mm UNLESS OTHERWISE STATED.
- TAPPED HOLES CHAMFER TO FIRST THREAD x 45.°

TITLE:-
FILTER CHASSI
(550-900MHz) 74x64mm
MATERIAL : ALUMINIUM

UNITS:- mm		SHEET SIZE:- A2	
	DATE	NAME	SIGN
DRN	21.06.13	RKL	
CHD			
APPD			

SCALE 1:0.75

SHT.NO:- 1 OF 1

DRG NO:- GMRT-07-DD-027-00

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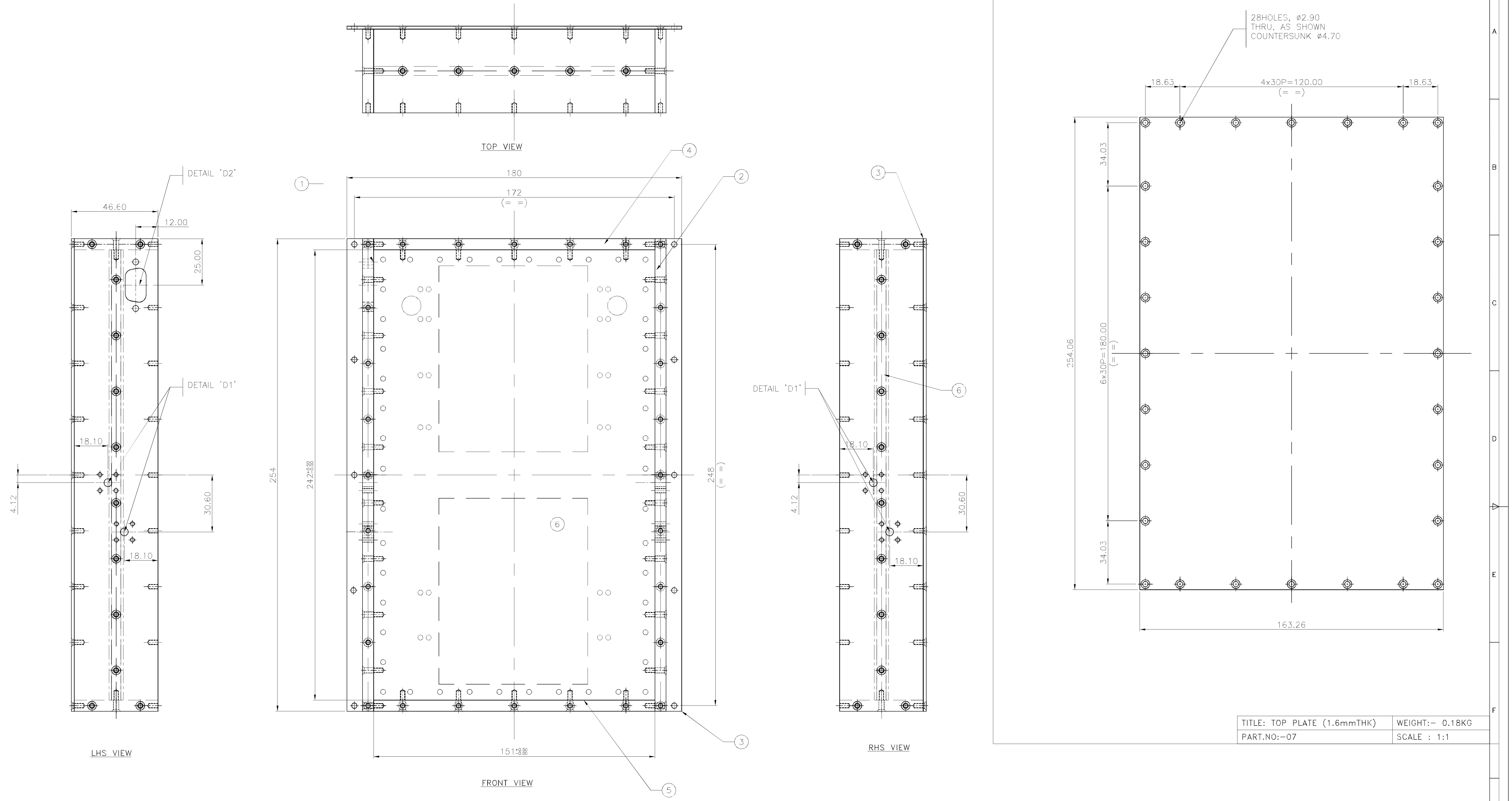
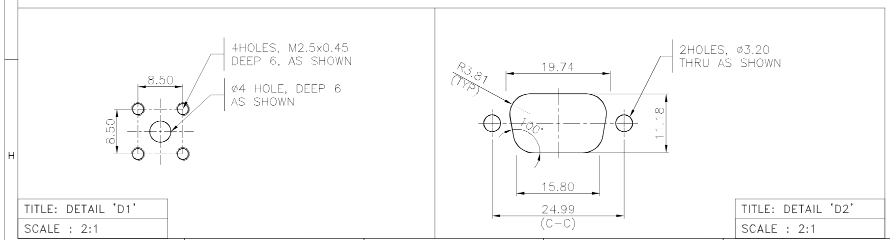




NOTES

1) ALL DIMENSIONS ARE IN mm
2) MATERIAL : ALUMINIUM
4) CHASSI WEIGHT : 1.25KG (APPROX)
5) CHASSI COVER : 1.6mmTHK AL PLATE
6) SMA CONNECTOR : 4nos
7) PCB MOUNTING HOLES AS PER SAMPLE PCB
8) PCB MOUNTING HOLES : M2.5x0.45
9) D-TYPE CONNECTOP (9-PIN) : 1no.

R.NO	DATE	MODIFICATION	CHKD
R1	15.07.14	PARTITION & BASE PLATE CHANGED	RKL

REQ NO:-	DEPARTMENT:- FRONT END
REQ DATE:-	NAME OF REQ:- IMRAN KHAN
DATE OF COMP:- 24.04.14	APPD BY:- S.SURESHKUMAR

NCRA · TIFR

GIANT METERWAVE RADIO TELESCOPE
TATA INSTITUTE OF FUNDAMENTAL RESEARCH
KHODAD, P.B.NO:-6, NARAYANGAON, TALUKA-JUNNAR, PUNE:-410504

* DRAWING NOT TO BE SCALED.
* FILLET SHARP EDGES.
* ALL DIMENSIONS ARE IN mm UNLESS OTHERWISE STATED.
* TAPPED HOLES CHAMFER TO FIRST THREAD x 45°

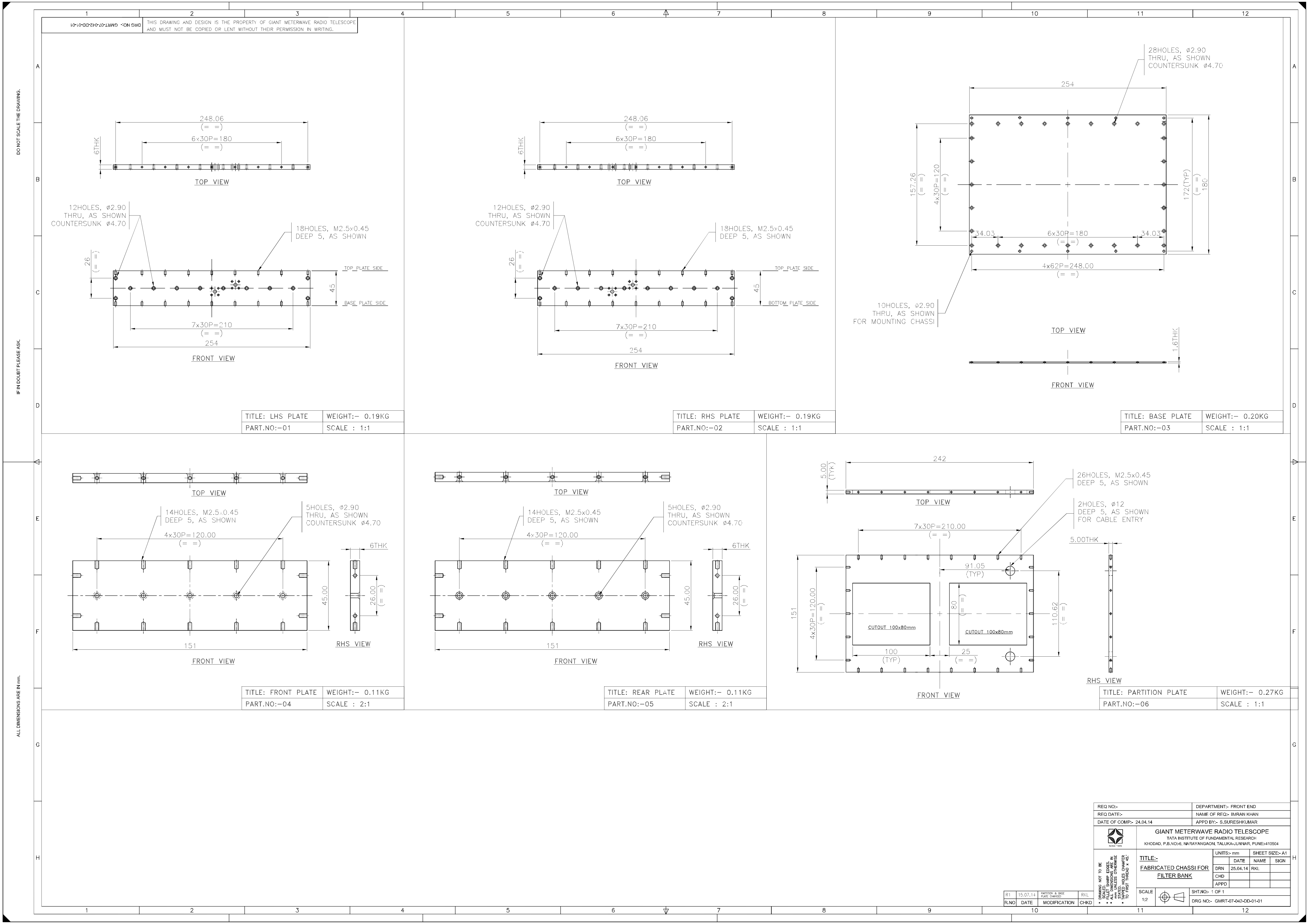
TITLE:-
FABRICATED CHASSI FOR FILTER BANK

UNITS:- mm		SHEET SIZE:- A1	
	DATE	NAME	SIGN
DRN	25.04.14	RKL	
CHD			
APPD			

SCALE 1:2

SHT.NO:- 1 OF 1

DRG NO:- GMRT-07-042-GA-01-01



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248.06
6x30P=180
6THK
TOP VIEW
12HOLES, ø2.90
THRU, AS SHOWN
COUNTERSUNK ø4.70
18HOLES, M2.5x0.45
DEEP 5, AS SHOWN
26
TOP PLATE SIDE
BASE PLATE SIDE
BOTTOM PLATE SIDE
45
7x30P=210
254
FRONT VIEW
TITLE: LHS PLATE
WEIGHT:- 0.19KG
PART.NO:-01
SCALE : 1:1
TITLE: RHS PLATE
WEIGHT:- 0.19KG
PART.NO:-02
SCALE : 1:1
28HOLES, ø2.90
THRU, AS SHOWN
COUNTERSUNK ø4.70
157.26
4x30P=120
172(TYP)
180
34.03
4x62P=248.00
10HOLES, ø2.90
THRU, AS SHOWN
FOR MOUNTING CHASSI
1.6THK
TITLE: BASE PLATE
WEIGHT:- 0.20KG
PART.NO:-03
SCALE : 1:1
14HOLES, M2.5x0.45
DEEP 5, AS SHOWN
5HOLES, ø2.90
THRU, AS SHOWN
COUNTERSUNK ø4.70
4x30P=120.00
45.00
26.00
151
RHS VIEW
TITLE: FRONT PLATE
WEIGHT:- 0.11KG
PART.NO:-04
SCALE : 2:1
TITLE: REAR PLATE
WEIGHT:- 0.11KG
PART.NO:-05
SCALE : 2:1
242
5.00
(TYK)
26HOLES, M2.5x0.45
DEEP 5, AS SHOWN
2HOLES, ø12
DEEP 5, AS SHOWN
FOR CABLE ENTRY
7x30P=210.00
5.00THK
91.05
(TYP)
80
110.62
CUTOUT 100x80mm
100
(TYP)
25
TITLE: PARTITION PLATE
WEIGHT:- 0.27KG
PART.NO:-06
SCALE : 1:1
REQ NO:-
REQ DATE:-
DATE OF COMP:- 24.04.14
DEPARTMENT:- FRONT END
NAME OF REQ:- IMRAN KHAN
APPD BY:- S.SURESHKUMAR
GIANT METERWAVE RADIO TELESCOPE
TATA INSTITUTE OF FUNDAMENTAL RESEARCH
KHODAD, P.B.NO:-6, NARAYANGAON, TALUKA-JUNNAR, PUNE-410504
UNITS:- mm
SHEET SIZE:- A1
DATE
NAME
SIGN
DRN
25.04.14
RKL
CHD
APPD
TITLE:-
FABRICATED CHASSI FOR
FILTER BANK
• DRAWING NOT TO BE SCALED.
• FILLET SHARP EDGES.
• ALL DIMENSIONS ARE IN mm UNLESS OTHERWISE STATED.
• TAPPED HOLES CHAMFER TO FIRST THREAD x 45.
SCALE
1:2
SHT.NO:- 1 OF 1
DRG NO:- GMRT-07-042-DD-01-01
R1
15.07.14
PARTITION & BASE PLATE CHANGED
RKL
R.NO
DATE
MODIFICATION
CHKD