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Giant Meter-wave Radio Telescope
National Center for Radio Astrophysics
Tata Institute of Fundamental Research

Wideband 327 MHz Front-End for GMRT

By

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Contents

Introduction

Filters

Band Pass Filter Design

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Abhay and Ankit

Introduction

Overview of GMRT Receiver System

Giant Meter-wave Radio Telescope (GMRT) currently operates at five observing bands centered at 150 MHz, 235 MHz, 327 MHz, 610 MHz and an L-band extending from 1000 to 1450 MHz. The L-band is split into four sub bands centered at 1060 MHz, 1170 MHz, 1280 MHz and 1390 MHz, each with a bandwidth of 120 MHz. The L-band receiver also has a bypass mode in which the entire RF band can be brought down to the Antenna Base Receiver (ABR). The 150 MHz, 235 MHz, 327 MHz bands of GMRT have 40 MHz bandwidth and 610 MHz band has about 60 MHz of bandwidth.

Lower frequency bands from 150 to 610 MHz have dual circular polarization which are named as CH1 and CH2 for right hand circular polarization and left hand circular polarization respectively. The higher frequency L-band has dual linear polarization (Vertical and Horizontal polarization) named CH1 and CH2 respectively.

At the lower frequencies the polarizer placed before the Low Noise Amplifier (LNA) converts the received linear polarization to circular. At L-band, in order to keep the system temperature low, this element is not inserted into the signal path, and the linear polarized signals are fed directly to the LNA. To calibrate the gain of the receiver chain, it is possible to inject an additional noise signal (of known strength) into the input of the LNA. It is possible to inject noise at any one of four levels. These are called Low cal, Medium cal, High cal and Extra high cal and are of monotonically increasing strength.

To minimize crosstalk between different signals a phase switching facility using separate Walsh functions for each signal path is available at the RF section of the receiver.

At the Common Box the signals go through one additional stage of amplification. The common box has a broad band amplifier which covers the entire frequency range of the GMRT (10 – 1800 MHz). The band selector in the common box can be configured to take signals from any one of the six RF frontends. The common box (and the entire receiver system) has the flexibility to be configured for receiving either both polarizations at a single frequency band or a single polarization at each of two different frequency bands. It is also possible to swap the polarization channels whenever required.

For observing strong radio sources like Sun, solar attenuators of 14 dB, 30 dB or 44 dB are available in the common box. In addition there is a power monitor whose output can be continuously monitored to verify the health of the subsystems upstream of the common box.

The first synthesized local oscillator converts the RF band to an IF band centered at 70 MHz. The synthesized local oscillator has a frequency range of 100 MHz to 1795

MHz . The frequency range 100 MHz to 600 MHz is covered by synthesizer-1 and 605 MHz to 1795 MHz is covered by synthesizer-2. The local oscillator frequency from 100 MHz to 354 MHz can be set with a step size of 1 MHz and the frequency range from 355 MHz to 1795 MHz can be set with a step size of 5 MHz. At the IF stage, bandwidth of 5.5 MHz, 16 MHz or a full available RF bandwidth can be selected. The IF at 70 MHz is then translated to a second IF at 130 MHz and 175 MHz for CH1 and CH2 respectively. The maximum bandwidth available at this stage is 32 MHz for each polarization channel . This frequency translation is done so that they can be transported to Central Electronics Building (CEB) over a single fiber-optic cable. An Automatic Level Control (ALC) facility is provided at the output stage of IF which can be bypassed whenever required.

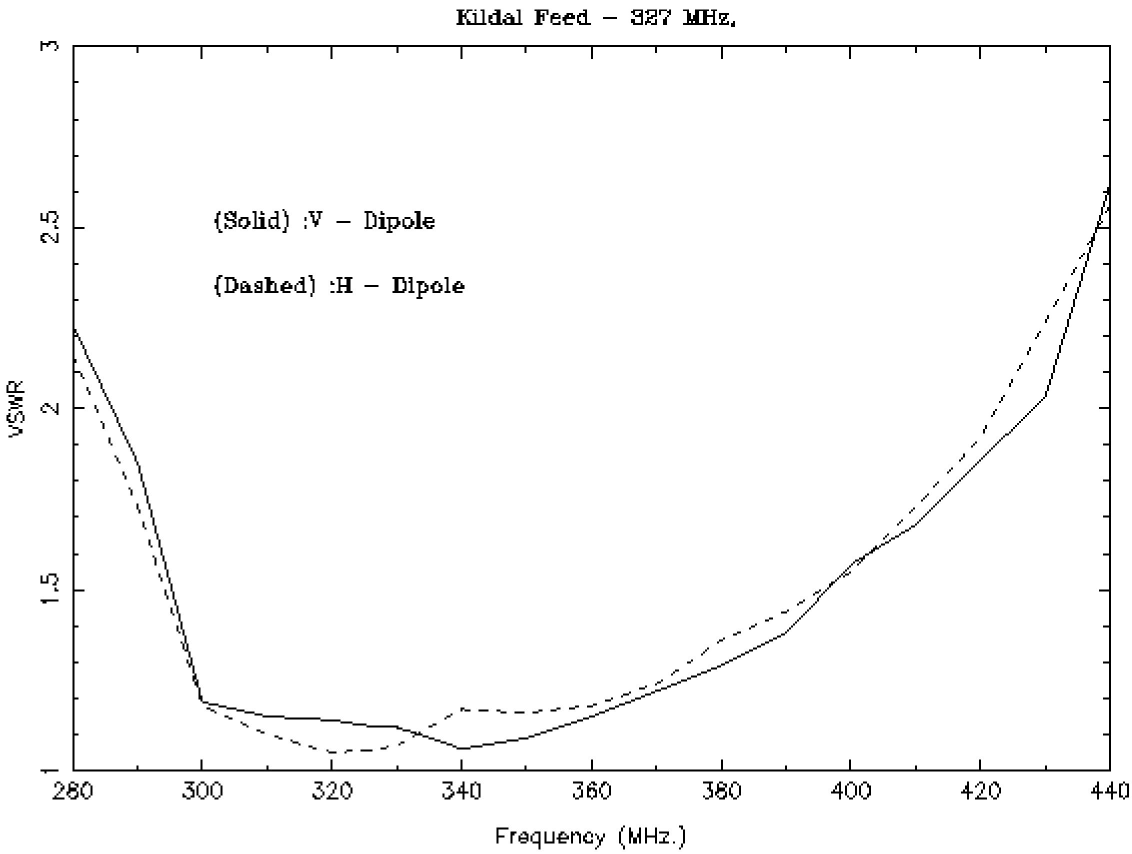
The IF signal at 130 MHz and 175 MHz along with telemetry and LO round trip phase carriers directly modulate a laser diode operating at 1300 nm wavelength which is coupled to a single mode fiber-optic link between the receiving antennas and the CEB.

At the CEB these signals are recovered with a PIN photo diode detector and suitably amplified. The 130 MHz and 175 MHz signals are then separated out and then sent for baseband conversion. There is a monitor port available at the fiber-optic receiver front panel at CEB, where all the received signals can be monitored.

The baseband converter section converts 130 MHz and 175 MHz IF signals to 70 MHz using 3rd LO (200 MHz & 105 MHz respectively.)The 70 MHz signals are then converted to baseband consisting of upper and lower sidebands for each sidebands for each polarization channel using a tunable LO which can be set from 50 MHz to 90 MHz in steps of 100 Hz. The BB system bandwidths can be set to any one of the bandwidths out of 62.5 KHz, 128 KHz, 256 KHz, 512 KHz, 1 MHz, 2 MHz, 4 MHz, 8 MHz and 16 MHz as per the user requirements . An ALC is incorporated at the output of Baseband converter that can be bypassed whenever required.

Scope of Work

The purpose of our project is to improve the bandpass filter in the 327 MHz Front-end Box to have a larger bandwidth and hence enable the front-end system to receive a wider band of frequencies. The existing 327 MHz Kildal feed has a broad frequency coverage. The plot of VSWR v/s Frequency for the 327 MHz Kildal feed is as shown below:



Clearly we see that the feed has useful band of frequencies ranging from about 290 MHz to 410 MHZ since the VSWR over this range is less than 1.5, which is desirable. However the present bandpass filter has a bandwidth of only 40 MHz around the 327 MHz frequency. So to fully exploit this available bandwidth from the feed, the bandwidth of the bandpass filter is proposed to be increased to 100 MHz, from 300 MHz to 400 MHz. The report concentrates on the designing of this improved bandpass filter to have good response characteristics such as increased bandwidth, sharp roll-off, flat group delay, proper image rejection as required by the system and to integrate the same with the present front-end system. The report also contains detailed description of various types of filters, their comparisons and the reason for the choice of the filter chosen.

Existing Front-end System

The existing system of front-end receiver gets its input from an arrangement known as Kildal Feed located at the prime focus of the GMRT antenna. It consists of orthogonal pairs of two crossed dipoles, each having a beam forming ring above it, in a plane parallel to the reflector supported by the dielectric rods.

The present front-end system consists of the following sub-systems:

Polarizer

The polarizer present in the front-end box transforms linear vertical and horizontal polarized signals received from the orthogonal dipole feed into right and left circularly polarized signals. The two circularly polarized channels are known as CH1 and CH2.

The polarizer design is based on Quadrature Hybrid with low loss since the polarizers are placed before the LNA. The insertion loss of the existing polarizer is 0.18 dB.

Low Noise Amplifier (LNA)

The low noise amplifiers are placed immediately after the low loss polarizers. Usually LNA is placed in the front-end portion of the system and its noise temperature determines the system noise properties. This is because by Friis' formula, if the gain of the first device is very high and noise temperature is very less, then the noise temperature of the entire receiver system is governed by its noise temperature.

The present 327 MHz LNA design is based on Low noise GaAs FET technology using Agilent Technologies ATF 10135 and ATF 10136 Low noise GaAs FETs. It has a gain of 36 dB, a bandwidth of 50 MHz, a noise figure of 0.41 dB and a noise temperature of 30 °K. .

To calibrate the gain of the receiver chain, it is possible to inject an additional noise signal(of known strength) into the input of LNA. It is possible to inject noise at any one of four levels:

- **Low cal - 10 %**
- **Medium cal - 40 %**
- **High cal - 100 %**
- **Extra high cal - 400 %**

This is done using the microstrip Directional Coupler .

Band Pass Filter

The band pass filter after the low noise amplifier is used to select a wide band of frequencies around the 327 MHz specified frequency. It is also useful for rejecting unwanted interference signal and for image rejection. The present band pass filter has a bandwidth of only 40 MHz around central frequency. Moreover the filter has an asymmetric and gradual roll-off, hence there is a scope of improvement in this regard.

Post Amplifier and Phase Switch

The bandpass filter is followed by a post amplifier to have the required output power at the end of front end box. The amplifiers being used presently are the Mini-circuits MMIC amplifiers MAV-11 with a gain of 12.5 dB and noise figure of 3.6 dB.

Thereafter the signal is modulated with Walsh function using phase switching to minimize the effect of crosstalk between different signals. Mini-circuits double balanced mixer SBL-1MH is used for phase-switching.

RF on/off facility is provided for connecting and disconnecting a channel by means of RF switch SW-239 having an insertion loss of about 0.50 dB.

Band Selector

All frequency bands converge at the band selector which then selects any one of the six bands and passes it on for further processing in the Common Box. The band selector design is based on GaAs FET MMIC RF switches. It can be configured either to receive both polarizations at
at a single frequency band or a single polarization at each of the two frequency bands, if any such need arises.

Solar Attenuator

This attenuator is used while observing Sun or any other strong radio sources. If the radiations are too high then they are attenuated to prevent the following circuitry from getting saturated. Solar attenuator has a set of two attenuators i.e. 14 dB and 30 dB. A maximum of 44 dB and a minimum of 0 dB can be attained using these two. Also this facilitates the band selector and the channel swap to terminate on 14 dB and 30 dB variable attenuators respectively.

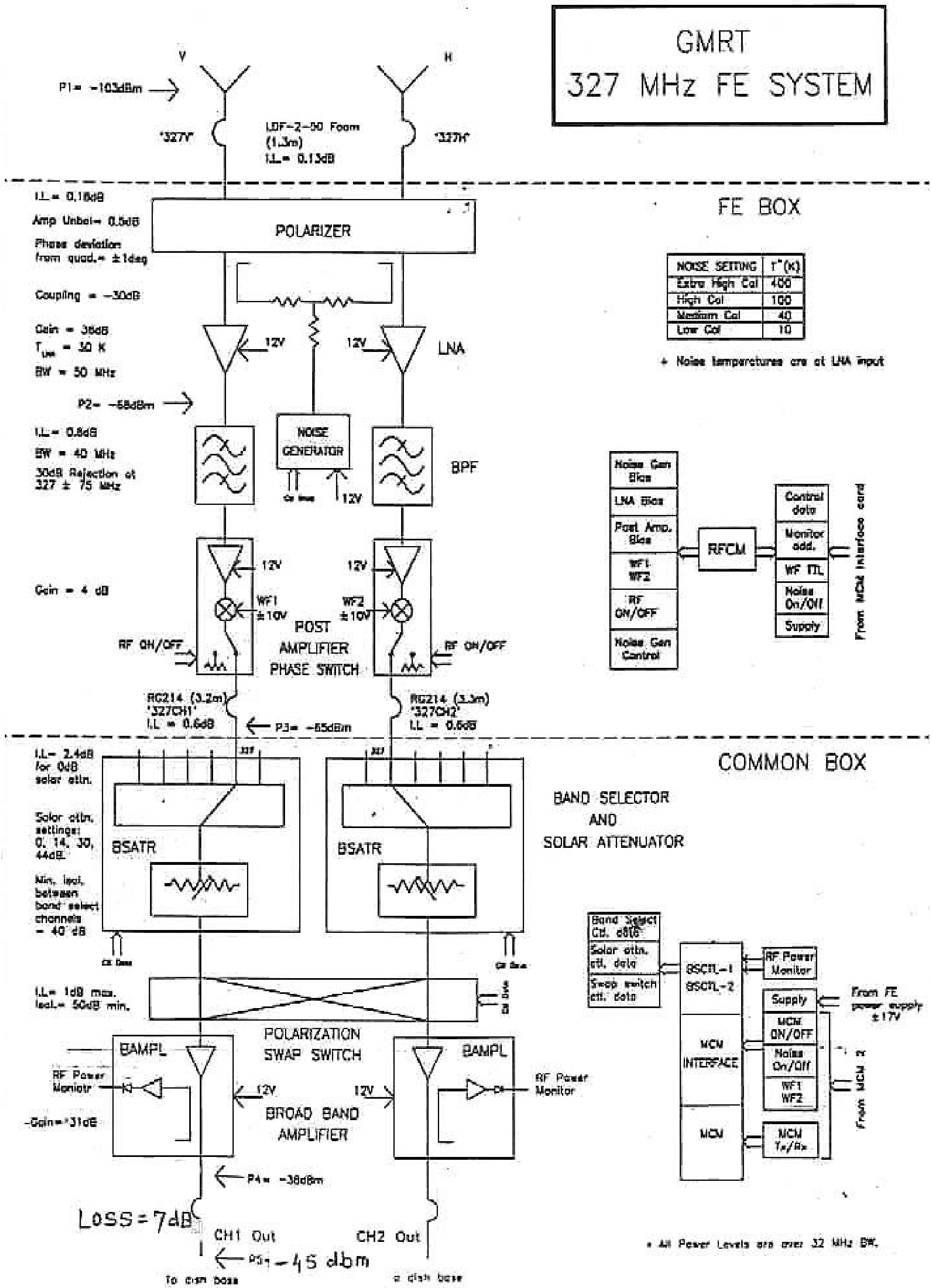
Channel Swap Switch

The channel swap switch is designed using a set of GaAs FET MMIC RF switches. Using this facility, the polarization channels can be swapped whenever required. It also helps in fault detection.

Broadband Amplifier

It is a conglomerate of 3 devices in cascade. They are Amplifiers INA-10386(Agilent), MAR-3(Mini-Circuits), and Directional Coupler PDC 10-5(Mini-Circuits). It has a gain of approximately 31 dB.

After this follow a pair of low-loss foam cables which transport the signals from the front end box to the Antenna Base Receiver.


GMRT
327 MHz FE SYSTEM
V
H
P1= -103dBm
'327V'
'327H'
LDF-2-50 Foam
(1.3m)
I.L= 0.13dB
FE BOX
I.L= 0.18dB
Amp Unbal= 0.5dB
Phase deviation from quad.= ±1deg
POLARIZER
Coupling = -30dB
Gain = 38dB
BW = 50 MHz
12V
LNA
P2= -68dBm
I.L= 0.8dB
BW = 40 MHz
30dB Rejection at 327 ± 75 MHz
NOISE GENERATOR
BPF
NOISE SETTING
Extra High Cal
High Cal
Medium Cal
Low Cal
400
100
40
10
+ Noise temperatures are at LNA input
Noise Gen Bias
LNA Bias
Post Amp. Bias
WF1 WF2
RF ON/OFF
Noise Gen Control
RFCM
Control data
Monitor add.
WF TTL
Noise On/Off
Supply
From MCM Interface card
Gain = 4 dB
WF1 ±10V
WF2 ±10V
POST AMPLIFIER PHASE SWITCH
RF ON/OFF
RG214 (3.2m) '327CH1' I.L = 0.6dB
RG214 (3.3m) '327CH2' I.L = 0.6dB
P3= -65dBm
COMMON BOX
I.L= 2.4dB for 0dB solar attn.
Solar attn. settings: 0, 14, 30, 44dB
Min. isol. between band select channels = 40 dB
BSATR
BSATR
BAND SELECTOR AND SOLAR ATTENUATOR
Band Select Ctl. data
Solar attn. ctl. data
Swap switch ctl. data
BSCTL-1
BSCTL-2
RF Power Monitor
Supply
From FE power supply ±17V
MCM ON/OFF
Noise On/Off
WF1 WF2
MCM INTERFACE
MCM
MCM Tx/Rx
From MCM 2
I.L= 1dB max.
Isol.= 50dB min.
POLARIZATION SWAP SWITCH
BAMPL
BAMPL
RF Power Monitr
RF Power Monitor
Gain= 31dB
12V
BROAD BAND AMPLIFIER
P4= -38dBm
LOSS = 7dB
CH1 Out
CH2 Out
P5= -45 dbm
To dish base
* All Power Levels are over 32 MHz BW.

FILTERS

Generally signals contain both wanted and unwanted informations.Filters separate signals based on their frequency differences by offering little opposition to certain frequencies while blocking others. Filters vary in attenuation characteristics (high-pass,low-pass, band-pass and notch), power-handling capabilities and in their passband and stopband frequencies.

The names given to various filters are based on their uses. The range of frequencies that a filter does not attenuate significantly is known as the filter's **passband**. The range of frequencies that a filter attenuates is known as its **stopband**.

Pole-Zero Concept :

The frequency response of a filter can be expressed as ratio of two polynomials describing the relationship between input and output of a system. Mathematically it can be stated as

$$\mathbf{H(s) = O(s) / I(s)}$$

where s = jw (j =√ -1) .

The roots of the numerator are called zeros and that of the denominator as poles .Their representation in s-plane gives pole-zero plot.

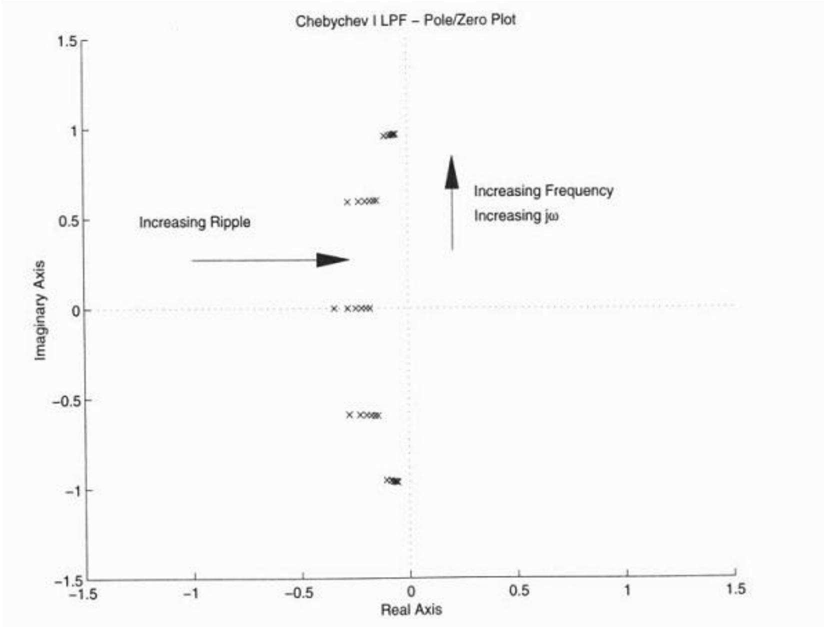
There are certain restrictions on poles and zeros for the filter to be realizable:

- The poles must have negative real coordinate.
- Complex poles or zeros must occur in conjugate pairs.

Dominant Poles and Zeros :

Whenever the distance between the imaginary axis (jw) in the s-plane and any pole/zero is small in relation with the other distances on the pole/zero plot, the transfer function is almost entirely dominated by the closest one. The areas of high attenuation in the transfer function are generally the result of dominant zeros.

This is clearly reflected by the graph shown :



Some Important Terms:

- **Frequency-Scaling Factor (FSF)**

It is the the ratio of a reference frequency of the desired response to the corresponding reference frequency of the given filter. Usually 3-dB points are selected as reference frequencies of low-pass and high-pass filters and the center frequency is chosen as the reference for band pass filters.

$$\textbf{FSF} = \frac{\textbf{desired reference frequency}}{\textbf{existing reference frequency}}$$

Since FSF is a dimensionless quantity, so both numerator and denominator must have same units, generally chosen to be 'rad/sec'.

The filter's response can be shifted to a different frequency range by dividing the reactive elements by the frequency scaling factor. Frequency scaling a filter multiplies all points on the frequency axis by the factor FSF. It also scales the poles and zeros by the same factor.

- **Impedance-Scaling Factor(Z)**

Sometime in the circuit, the component values that we get are not very practical.Such a situation can be resolved by impedance scaling. Any linear active or passive network maintains its transfer function if all resistor and inductor values are multiplied by an impedance scaling factor **Z** and all capacitors are divided by the same factor **Z**.

$$Z = \frac{\textbf{desired input impedance}}{\textbf{existing input impedance}}$$

This occurs because the Zs cancel in the transfer function of the system.

- **Steepness Factor(A)**

This factor gives the extent of steepness of the filter at the edge of the pass band i.e. how fast the roll off occurs.

For a low-pass filter, $\mathbf{A = f_s / f_c}$

For a high-pass filter, $\mathbf{A = f_c / f_s}$

where, 'fs' is the frequency having the minimum required stop-band attenuation
'fc' is the limiting frequency or the cutoff of the passband, usually the 3-dB point

- **Phase Delay**

$$\mathbf{P(\omega) = -\Theta(\omega)/\omega}$$

where, $\mathbf{\Theta(\omega)}$ is the complex angle of the frequency response $H(e^{j\omega})$ or the phase response of the filter
and $\boldsymbol{\omega}$ is the frequency of interest.

The phase delay gives the time delay, in seconds, experienced by each sinusoidal component of the signal input to the system.It is all known as **carrier delay**.

In working with phase delay, care must be taken to ensure all appropriate multiples of 2π have been included in $\Theta(\omega)$.By discarding multiples of 2π, as is done in the definition of complex angle, the phase delay is modified by multiples of the sinusoidal period. Since LTI filter analysis is based on sinusoids without beginning or end, one cannot in principle distinguish between "true" phase delay and a phase delay with discarded sinusoidal periods.

- **Group Delay**

The group delay of a Linear Time Invariant (LTI) filter H(z), with phase response Θ(ω) is defined by

$$\mathbf{D(\omega) = -d\ \Theta(\omega)\ /d\omega}$$

For any phase function, the group delay D(ω) may be interpreted as the time delay of the amplitude envelope of a sinusoid at frequency ω. The bandwidth of the amplitude envelope in this interpretation must be restricted to a frequency interval over which the phase response is approximately linear. At this point, it is evident that the group delay at the carrier frequency gives the slope of the linear phase of the translated spectrum.

If the group delay is not constant over the bandwidth of the modulated signal, waveform distortion will occur. In fact it is a common practice to use group-delay variation as a criterion to evaluate phase non-linearity and subsequent waveform distortion.

For linear phase responses, the group delay and the phase delay are identical, and each may be interpreted as time delay.

- **Voltage Standing Wave Ratio(VSWR)**

When in a circuit, a transmission line is not matched to its load, some of the energy is absorbed by the load and some is reflected back down the line towards the source. The interference of incident and reflected waves create standing waves on the transmission line.
VSWR is basically the ratio of the maximum to minimum amplitude of these standing waves.

$$\mathbf{VSWR = Maximum\ Amplitude\ /\ Minimum\ Amplitude}$$
$$\mathbf{= (V_i + V_r) / (V_i - V_r)}$$

where $\mathbf{V_i}$ is the incident voltage
and $\mathbf{V_r}$ is the reflected voltage

- **Reflection Coefficient(ρ)**

It is defined as the ratio of the reflected voltage to incident voltage at the load impedance in a circuit.

$$\boldsymbol{\rho} = \mathbf{Reflected\ Voltage\ /\ Incident\ Voltage}$$

$$\mathbf{= (Z_L - Z_0)/(Z_L + Z_0)}$$

where, $\mathbf{Z_L}$ is the load impedance
and $\mathbf{Z_0}$ is the characteristic impedance of the transmission line

It is a measure of how well the system is matched. The value of reflection coefficient varies from -1 to +1 depending on the magnitude of reflection.

- $\rho = 0$ indicates a perfect match with no reflection.
- $\rho = -1$ indicates a short circuited load.
- $\rho = +1$ indicates an open circuit.

Reflection Coefficient and VSWR are related to each other as follows:

$$|\rho| = (\mathbf{VSWR} - 1) / (\mathbf{VSWR} + 1)$$

$$\mathbf{VSWR} = (1 + |\rho|) / (1 - |\rho|)$$

- **Quality Factor (Q)**

The Q factor or quality factor is a measure of the "quality" of a resonant system. Resonant systems respond to frequencies close to their natural frequency much more strongly than they respond to other frequencies. For an electrically resonant system, the Q factor represents the effect of electrical resistance.

The Q factor is defined as the resonant frequency (or center frequency) f_c divided by the bandwidth Δf :

$$\mathbf{Q = f_c / \Delta f}$$

where, the bandwidth $\Delta f = f_2 - f_1$
$\mathbf{f_2}$ is the upper cutoff frequency
$\mathbf{f_1}$ is the lower cutoff frequency

The reactive components of the circuit also have a Q value which indicates their quality or figure of merit since losses can be defined in their terms. Using finite Q-valued passive components in the circuit can lead to more rounded response at pass band edge, diminished ripples in the passband, increased insertion loss and reduced relative attenuation between passband and stop-band.

Active v/s Passive Filters

- **Definitions**

Active Filters: In electronics, an active component is one that can be used to provide gain in an electronic circuit .An active filter is a type of analog electronic filter, distinguished by the use of one or more active component i.e. one which provides some form of power amplification. Typically this will be a vacuum tube, transistor or operational amplifier.

Passive Filters: A passive component is an electronic component that does not require a source of energy to perform its intended function. Examples of passive components include resistors, capacitors, and inductors. A passive filter is an electronic filter made entirely from passive components.

- **Frequency Considerations**

Active Filters: Active filters are more advantageous at sub-audio frequencies because they can be designed at higher impedance levels so that the capacitor magnitudes are reduced whereas LC filters at this frequency require higher values and hence much more space.

Passive Filters: Passive filters are better choices at higher frequencies since most commercial-grade operational amplifiers have insufficient open loop gain for the average active filter requirement at this frequency.

- **Ease of Adjustment**

Active Filters: Most active filters are not easily tunable. They may contain RC sections where two or more resistors in each section have to be varied in order to control resonances but these are very cumbersome.

Passive Filters: In LC filters, tuning is much easier. Capacitor below a few hundred pico-farads can be made variable. Inductors also can be easily adjusted since most coil structures provide a mean for tuning, such as an adjustment slug.

Types of filters

Depending on the relationship between the desired and undesired frequencies, receiver front ends may require low-pass, high-pass, band-pass or notch filters at the antenna input .Theoretically filters provide ideal response but in real practice they undergo certain limitations .So the filters can be broadly divided into ideal and real filters as follows :

- ## Ideal Filters

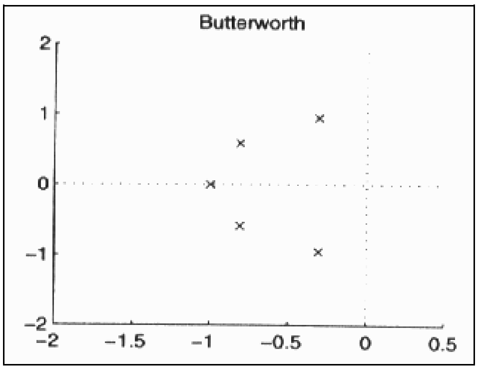
An ideal filter transmits frequencies in its pass band without attenuation and phase shift while offering infinite attenuation to the signal components in the stop-band .These filters have a sharp transition . Ideal filters are impossible to realize and the filters we make use are all real filters .

- ## Real Filters

These are non ideal filters which attenuate the unwanted signal to some finite limit .More importantly it considers how much attenuation is needed and the unwanted signals can be tolerated. Thus its design takes into account the requirements and perform suitable compromises to get the best possible output.

Some **real filters** are :

- ### Butterworth Filter



These are 'all-pole' filters. The poles when plotted in s-plane lie on a circle centered about the origin .The frequency response of the Butterworth filter is maximally flat (has no ripples) in the passband, and rolls off towards zero in the stopband.

This circular arrangement prevents any pole to be dominant and thus the magnitude response of the filter is maximally flat or monotonic. The attenuation increases steadily as the operating frequency increases beyond the 3-db cutoff frequency .

The attenuation for the low pass filter is given by

$$\mathbf{A_{db} = 10 \log [1+ (f_x/f_c)^{2n}]}$$

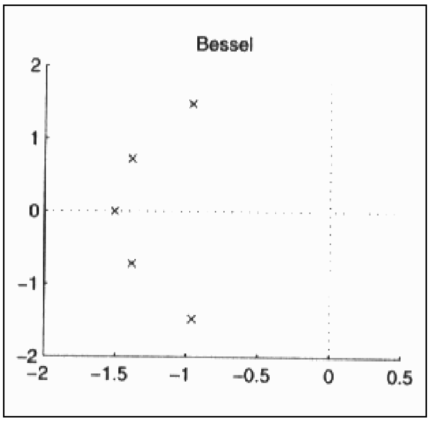
where, f_c is the cutoff frequency
f_x is the frequency of interest
n is the order of the filter.

The poles of the normalized filter lying on a unit circle are given by

$$P_k = -\sin[(2k-1)\Pi / 2n] + j\cos[(2k-1)\Pi / 2n]$$

where k = 1,2,3,......,n.

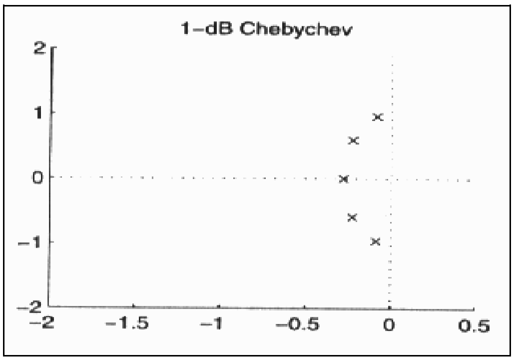
- ## Bessel Filters



Bessel filters are also 'all-pole' filters. The poles lie on an ellipse with imaginary axis as its major axis. Here the frequency response is much less selective than other filters but the phase and group delays are comparatively much better than others.

The group delay response remains flat throughout the filter's passband then drops off slowly beyond the cutoff frequency. These filters show very little ringing or overshooting and their rise time is nearly optimal .These filters are ideal for handling pulses.

- **Chebychev Filters**



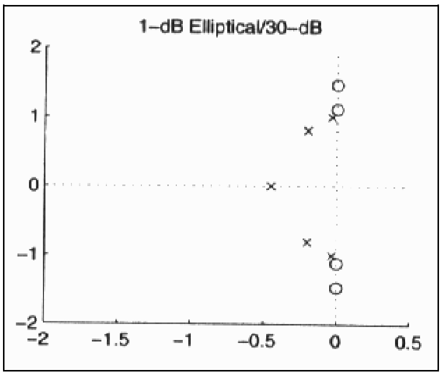
The magnitude response of these filters exhibits ripples in the passband but a steeper initial roll- off than the butterworth filters .It also contains only poles which lie on an ellipse . As the frequency increases and passes by each pole, the closest pole near the imaginary axis in s-plane appears dominant which causes ripples in the magnitude response .

Chebychev filters could also be classified as Chebychev I and Chebychev II. In the previous case the passband exhibits ripple and the stopband attenuation increases monotonically whereas in the second case the passband is flat and the stopband shows amplitude ripples.

The slope in the phase response is steepest in the neighborhood of the 3-db cutoff frequency thus the group delay being it's derivative shows a peak at the cutoff frequency. Moreover the group delay peak, in this case, is higher than the one in the butterworth.

The ripples are directly proportional to the SWR and Reflection coefficient. The cutoff frequency is specified at an attenuation equal to passband ripple. The Chebyshev filter offers more attenuation than the butterworth and the difference increases if more ripple is allowed in the passband.

- **Elliptic Filter**



Elliptic filters contain both poles and zeros in their transfer function.The dominant poles considerably increases the magnitude response whereas zeros are responsible for large attenuations. Thus proper placement of zeros reduce the transition region so that extremely sharp roll-off can be obtained. These filters are equirippple in both passband and stopband.

The dominant zeros also cause rapid phase change as well as significant magnitude change. The sudden jump in the phase response causes jumps in the group delay response at the same frequency.

The zeros provide sharp transition into the stopband but the improved performance is obtained at the expense of return lobes in the passband. The response in passband is similar to that of chebychev filters except that the attenuation at 1 rad/sec is equal to the passband ripple instead of 3 dB.

Comparison of filters

Butterworth, Chebychev and Bessel filters all have the same final roll-off. But the elliptic filters have comparatively lesser ultimate roll-off. It provides most attenuation immediately after the passband.

In Buterworth and Bessel filters none of the poles become dominant and the magnitude response of these filters are smooth in the passband. The poles in the case of Chebychev and Elliptic filters are closer to the imaginary axis in the s-plane and thus acts as dominant pole.

Generally the decreasing order of flatness in the group delay occur in the order of Bessel, Butterworth, Chebychev and finally Elliptic filters. Except the Bessel filters group delay becomes uneven near the cutoff frequency. The group delay usually peaks at the cutoff frequency.

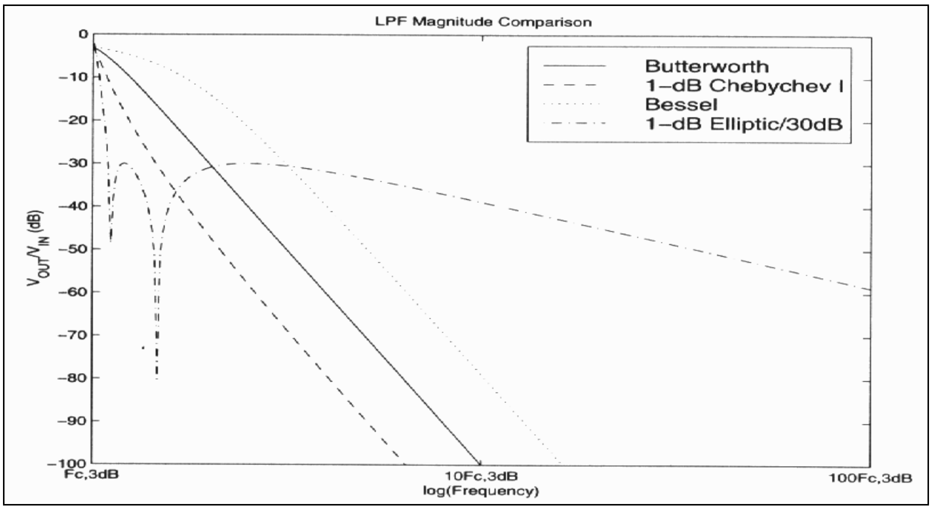


Fig. Comparison of Magnitude responses of filters

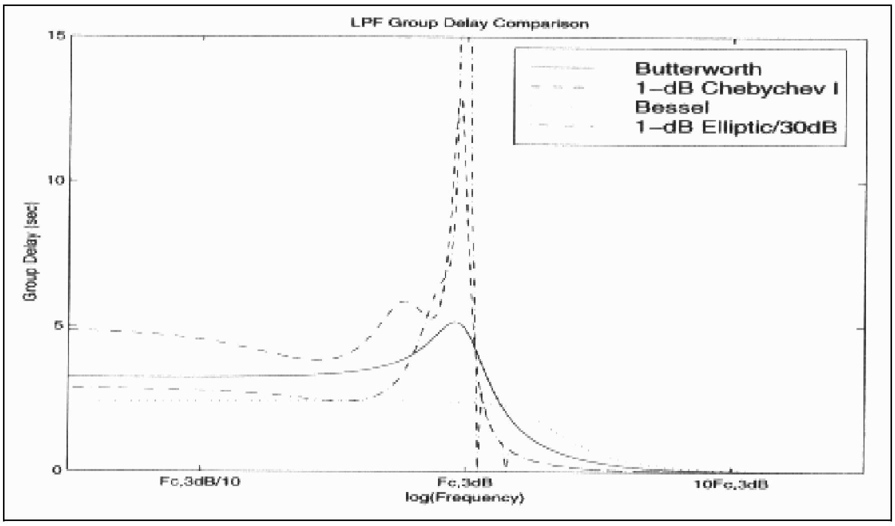


Fig. Comparison of Group Delay responses of filters

Band Pass Filter Design

Choice of Filter

In order to choose the correct filter its essential to point out the **prime requirements** and make **suitable compromises**. A major attribute of a filter is its frequency characteristic. In choosing a filter, both the desired and unwanted frequencies must be considered.

- **Elliptic filter**

If the cut off rate specification is stringent the classical Butterworth and Chebychev filters result in higher order filters .This adds complexity to the filter and the resulting design is more difficult to tune. The sensitivity of the filter to its component also increases. So for obtaining high attenuation immediately after the passband, accounting for high selectivity, elliptic filter is used in the design of the bandpass filter.

LC filter

They are better choice at higher frequencies since most commercial-grade operational amplifiers have insufficient open loop gain for the average active filter requirement at this frequency. Capacitors below a few hundred pico-farads can be made variable. So in LC filters, tuning is much easier.

- **LPF And HPF Combination**

In the LPF and HPF combination , zeros for both the filters can be separately selected .Thus it allows flexibility in designing the roll-off characteristics separately according to the requirement . Also adjusting the transition band of the filters (on lower as well as upper side of frequencies) individually, allows better symmetry around the center frequency.

Image Frequency Considerations

The GMRT receiver system is of the so-called superheterodyne type .To have a common signal processing path for signals of all the frequency bands, a combination of mixer and first local oscillator provide a frequency conversion or heterodyning function, whereby the incoming signal is converted to an IF band centered at 70 MHz.

This frequency conversion is achieved without disturbing the relation of the sidebands to the carrier.

$$f_{IF} = | f_{RF} - f_{LO} |$$

where, f_{LO} is the frequency of local oscillator
f_{RF} is the central frequency of incoming signal
f_{IF} is the intermediate frequency

Image Frequency is defined as an undesired input frequency that is capable of producing the same intermediate frequency (f_{IF}) that the desired input frequency produces. The mixer will develop an intermediate frequency output when the input signal frequency is greater or less than the local oscillator frequency by an amount equal to intermediate frequency. That is, there are two input frequencies, namely, $| f_{LO} \pm f_{IF} |$ which will result in f_{IF} at the mixer output.

Since the signal path thereafter depends only on the intermediate frequency, both signals get transmitted even though, we would not like to allow image signal to pass. So it is important to include circuitry for the purpose of image rejection before frequency down conversion in order to favor the desired signal and discriminate against the unwanted signal. Hence this is achieved in the Front-end section of the receiver itself using the band pass filter to have the desired response.

For the case of 327 MHz feed, the proposed Band Pass Filter is proposed to have a bandwidth of 100 MHz centered around 350 MHz. For the purpose of IF conversion there are two possible LO frequencies of 280 MHz (350-70) and 420 MHz (350+70).

- **Case 1** : LO frequency 280 MHz

Corresponding to this frequency, there will be an image bands centered around 210 MHz ranging from 160 MHz to 260 MHz .

- **Case 2** : LO frequency 490 MHz

For this frequency, there will be an image bands centered around 490 MHz ranging from 420 MHz to 540 MHz.

While designing the filter care should be taken to have proper attenuations over these image bands so that it does not interfere with the band of frequencies under consideration.

Low Pass Filter Design

- **Low-pass Filter Design procedure**

  - List out the specifications like cutoff frequency, stop-band frequency, minimum stop-band attenuation, input and output impedances etc.
  - Calculate the steepness required, from the specifications, and estimate the order of the filter which best suits the requirements.
  - Design a normalized LPF filter having cutoff frequency of 1 rad/sec, and input and output impedance of 1 ohm.
  - Next frequency scale the circuit to the desired cut-off frequency and impedance scale the above designed filter components by using following formulas for components scaling

  For capacitor

  $$C' = C / (FSF * Z)$$

  For inductor

  $$L' = L * Z / FSF$$

  where, FSF is Frequency Scaling Factor
  and Z is the impedance scaling factor

- **Low-pass Filter Specifications**

  - Minimum inductor elliptic
  - Cauer-chebyshev Normal.
  - 3 dB cutoff at a frequency of 400 MHz.
  - Input & Output impedance is 50 ohms.
  - Stop band frequency is 420 MHz.
  - Minimum stop band attenuation = 30 dB.

- **Order of the Filter**

Using cutoff frequency and stop band frequency firstly the steepness factor (A_s) is calculated as follows

$$\begin{aligned} A_s &= f_s / f_c \\ &= 420 / 400 \\ &= 1.05 \end{aligned}$$

Corresponding to worst case VSWR = 1.5, ρ= 20 % was chosen for the Amin of 30 dB and steepness of As <= 1.05

Actually, **VSWR = (ρ+ 1)/ (ρ-1)**

= 1.2 / 0.8

= 1.5 ....The worst case VSWR.

And passband ripple RdB is related to reflection coefficient ρ as

$$\mathbf{R_{dB} = -10 \log (1-\rho^2)}$$

So, $R_{dB} = -10 \log (1-(0.2)^2) = 0.1773$ dB

Thus ρ= 20% corresponds to A(ρ) = 13.9 dB

So total attenuation is $A_{min} + A_\rho = 30 + 13.9 = 43.9$ dB

Using As = 1.05 and total attenuation of A_s + A(ρ), a suitable order filter is chosen from the standard curve for estimating the order of the elliptic function filter. A filter of order n = 7 meets the required specifications.

Now the Return Loss is given as

$$\mathbf{RL = 20 \log (1/\rho)}$$
$$= 20 \log (1/0.2)$$
$$= -13.979 \text{ dB}$$

Therefore the theoretical attenuation in this case is 32 dB

- ## Expected Filter Response Properties

  - VSWR = 1.5
  - Return Loss = -13.979
  - Stopband Attenuation = 32 dB
  - Passband Ripple = 0.1773 dB

- ## Normalized LPF Design

This consists of designing the values of components of the LPF for the cutoff frequency of 1 rad/sec from the standard tables.

The angle θ determines the steepness of the filter and is given as

$$\theta = \sin^{-1}(1/A_s)$$
$$= \sin^{-1}(1/1.05)$$
$$= 72.24^{o}$$

Now using the table for elliptic filter of order 7, we choose θ = 73 º which satisfies the steepness criteria of As <=1.05

This normalized LPF can be expressed as

C n ρ θ = (C 7 20 73º),

where, C – Cauer filter
n – order
ρ – reflection coefficient
θ – Steepness angle

Now for these specifications of LPF, the normalized values for the filter elements are chosen from the tables and are as shown in the circuit below:-

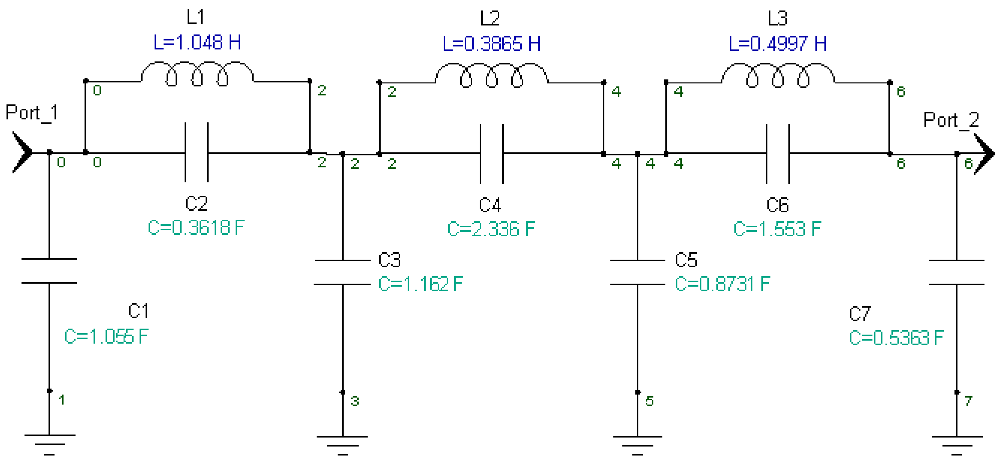


Fig. Normalized LPF Circuit Diagram

- ## Frequency And Impedance Scaling

The designed normalized filter has a cutoff frequency of 1 rad/sec and an input impedance of 1 Ohm. In order to scale the circuit for desired frequency response, frequency and impedance scaling is done by replacing capacitors and inductors with scaled values as follows.

Here,

$$FSF = 2 * \pi * 400 * 10^6 = 2.513 * 10^9$$

and impedance scaling factor is Z = 50.

All capacitors C are to be replaced by capacitors C', where

$$C' = C / (FSF * Z)$$

So the actual circuit capacitor values are

$C_1' = 83.95$ pF

$C_2' = 2.879$ pF

$C_3' = 9.247$ pF

$C_4' = 18.589$ pF

$C_5' = 6.948$ pF

$C_6' = 12.358$ pF

$C_7' = 4.268$ pF

Again all the inductors L are to be replaced by inductors L', where

$$L' = L * 50 / FSF$$

So the actual circuit inductor values are

$L_2' = 20.849$ nH

$L_4' = 7.689$ nH

$L_6' = 9.941$ nH

The parallel resonant sections containing L_2 C_2 , L_4 C_4 and L_6 C_6 produce the zeros. Now using the table values we find theoretical zeros to be at 649.64 MHz, 420.97 MHz and 454.087 MHz for these three branches respectively.

With the above impedance and frequency scaling the actual LPF is designed for the given specifications.
The circuit diagram is as shown below:

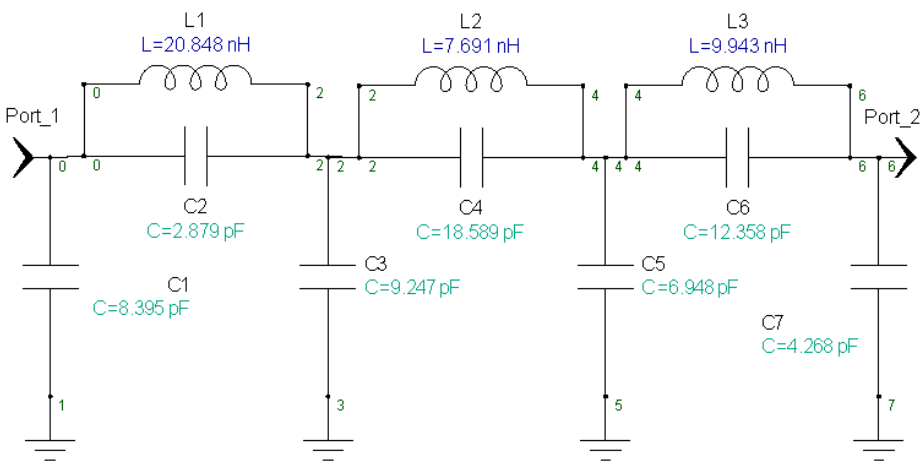


Fig. Frequency Scaled LPF Circuit Diagram

These finite transmission zeros are added to increase the steepness of the filter response . This is the theoretical filter design as per the specifications. The final filter circuit response is as shown on the next page

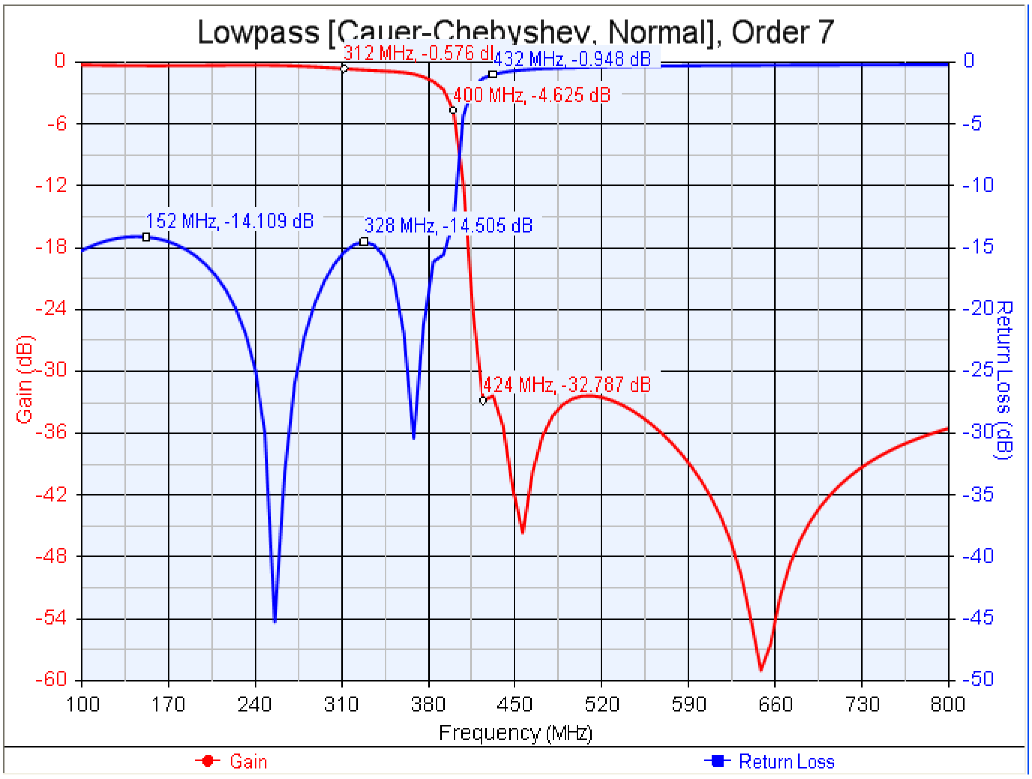


Fig. Simulated Filter Response of LPF

High Pass Filter Design

- **HPF Filter Design Procedure**

  - List out the specifications like cutoff frequency, stop-band frequency, minimum stop-band attenuation, input and output impedances etc.
  - Calculate the steepness required, from the specifications, and estimate the order of the filter which best suits the requirements.
  - Design a normalized LPF filter having cutoff frequency of 1 rad/sec, and input and output impedance of 1 ohm.
  - For LPF to HPF conversion, replace each capacitor by an inductor and each inductor by a capacitor in the circuit and invert the values found out for the normalized LPF.

$$C_{hp} = 1 / L_{lp}$$

$$L_{hp} = 1 / C_{lp}$$

  - Next frequency scale the circuit o the desired cut-off frequency and impedance scale the above designed filter components by using following formulas for components scaling

For capacitor

$$C' = C / (FSF * Z)$$

For inductor

$$L' = L * Z / FSF$$

where, FSF is Frequency Scaling Factor
and Z is the impedance scaling factor

- ## High-pass Filter Specifications

  - Minimum inductor elliptic.
  - Cauer-chebychev Normal.
  - 3 dB cutoff at a frequency of 300 MHz.
  - Input & Output impedance is 50 Ohms.
  - Stop band frequency is 280 MHz.
  - Minimum stop-band attenuation is 30 dB.

- ## Estimation of Order

From cutoff frequency and stop-band frequency specifications given, we can calculate the steepness factor (A_s) for HPF

$$A_s = f_c / f_s$$
$$= 300 / 280$$
$$= 1.071$$

With A_{min} = 30 dB and A_s <= 1.071, we choose reflection coefficient ρ = 20% which corresponds to VSWR = 1.5 , the worst case pass-band VSWR.

Actually, $\boldsymbol{\rho = (VSWR - 1) / (VSWR + 1)}$

So, $VSWR = (\rho + 1)/ (\rho - 1)$
$= 1.2 / 0.8 = 1.5$ ....The worst case VSWR.

And passband ripple R_{dB} is related to reflection coefficient, ρ, as

$$R_{dB} = -10 \log (1 - \rho^2)$$

So $R_{dB} = -10 \log (1 - (0.2)^2) = 0.1773$ dB

Thus ρ = 20% corresponds to $A(\rho)$ = 13.9 dB

So total attenuation is As + $A(\rho)$ = 30 +13 = 43.9 dB

Using As = 1.071 and total attenuation of $A_s + A(\rho)$, a suitable order filter is chosen from the standard curve for estimating the order of the elliptic function filter. A filter of order n = 7 meets the required specifications.

Return Loss is given as

$$RL = 20 \log (1 / \rho)$$
$$= 20 \log (1 / 0.2)$$
$$= -13.979 \text{ dB}$$

Therefore theoretical attenuation in this case is 37.2 dB.

- **Expected Filter Response Properties** :
    - VSWR = 1.5
    - Return Loss = -13.979 dB
    - Stop Band Attenuation = -30 dB
    - Passband ripple = 0.1773 dB

- **Normalized LPF for HPF Design**

This consists of designing the values of components of the LPF for the cutoff frequency of 1 rad/sec from the standard tables.

The angle θ determines the steepness of the filter and is given as

$$\theta = \sin^{-1}(1/A_s)$$
$$= \sin^{-1}(1/1.071)$$
$$= 69.02^{\circ}$$

Now using the table for elliptic filter of order 7, we choose θ = 69 ° which satisfies the steepness criteria of As <=1.071,

This normalized LPF can be expressed as

$C\ n\ \rho\ \theta = (C\ 7\ 20\ 69^{\circ})$

where, C – Cauer filter
n – order of the filter
ρ – reflection coefficient
θ – Steepness angle

Now for these specifications of LPF, the normalized values for the filter elements are chosen from the tables and are as shown in the circuit on the next page

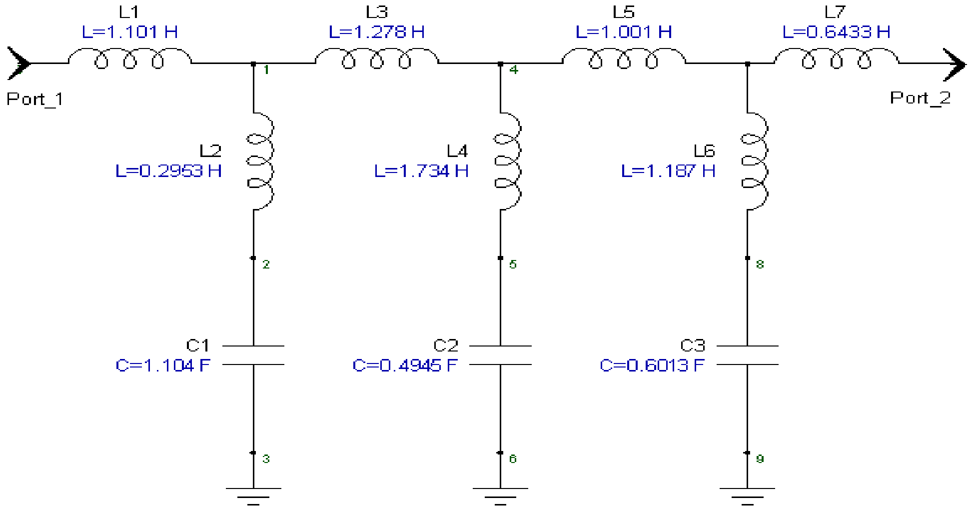


Fig. Normalized LPF For HPF Design

- ## Normalized HPF Design

The normalized HPF can be designed from the normalized LPF by just replacing capacitors by inductors and vice versa. The values for inductors and capacitors are found by taking reciprocal of the corresponding capacitor and inductor in the normalized LPF circuit as described earlier.

The normalized HPF filter circuit is as shown below:

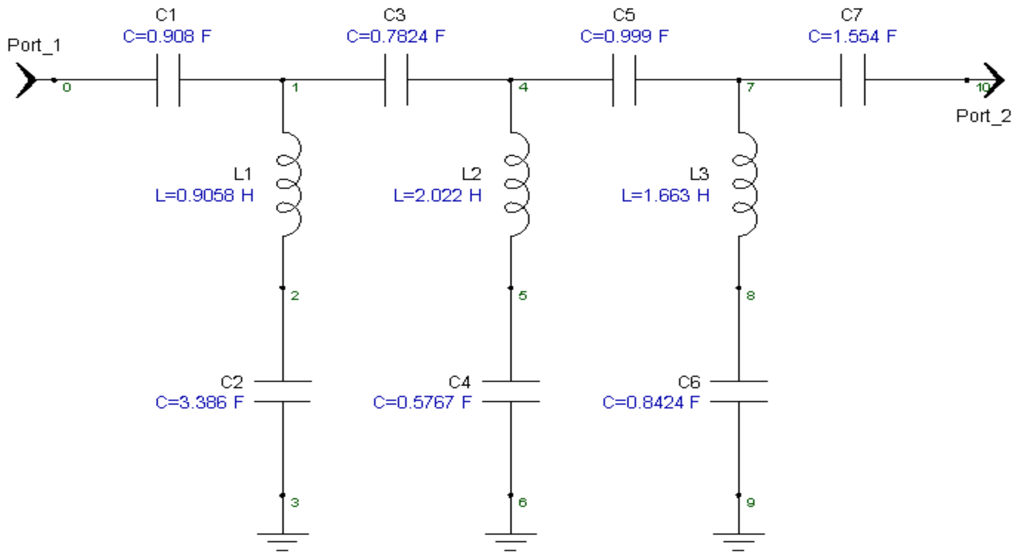


Fig. Normalized LPF To HPF Converted Circuit

- **Frequency And Impedance Scaling**

The designed normalized filter has a cutoff frequency of 1 rad/sec and an input impedance of 1 Ohm. In order to scale the circuit for desired frequency response, frequency and Impedance scaling is done by replacing capacitors and inductors with scaled values as follows.

The Frequency Scaling Factor (FSF) for the circuit is

$$FSF = 2 * \pi * 300 \text{ MHz}$$
$$= 1.885 * 10^9 \text{ Hz}$$

and Impedance Scaling Factor is Z = 50.

So all capacitors C are to be replaced by capacitors C', where

$$C' = C / (FSF * Z)$$

So the actual circuit capacitor values are

$C_1' = 9.64$ pF

$C_2' = 35.92$ pF

$C_3' = 8.3$ pF

$C_4' = 6.12$ pF

$C_5' = 10.6$ pF

$C_6' = 8.94$ pF

$C_7' = 16.49$ pF

Again all the inductors L are to be replaced by inductors L', where

$$L' = L * 50 / FSF$$

So the actual circuit inductor values are

$L_1' = 24.03$ nH

$L_2' = 53.63$ nH

$L_3' = 44.11$ nH

The series resonant sections containing L_2 C_2 , L_4 C_4 and L_6 C_6 produce the zeros. Now using the table values we find theoretical zeros to be at 171.28 MHz , 277.77 MHz and 253.41 MHz for the three branches respectively.

With the above impedance and frequency scaling the actual HPF is designed for the given specifications and the circuit diagram is as shown below. Also resonant frequency of each parallel branch series resonant section that produces zero at definite frequency is also shown below:

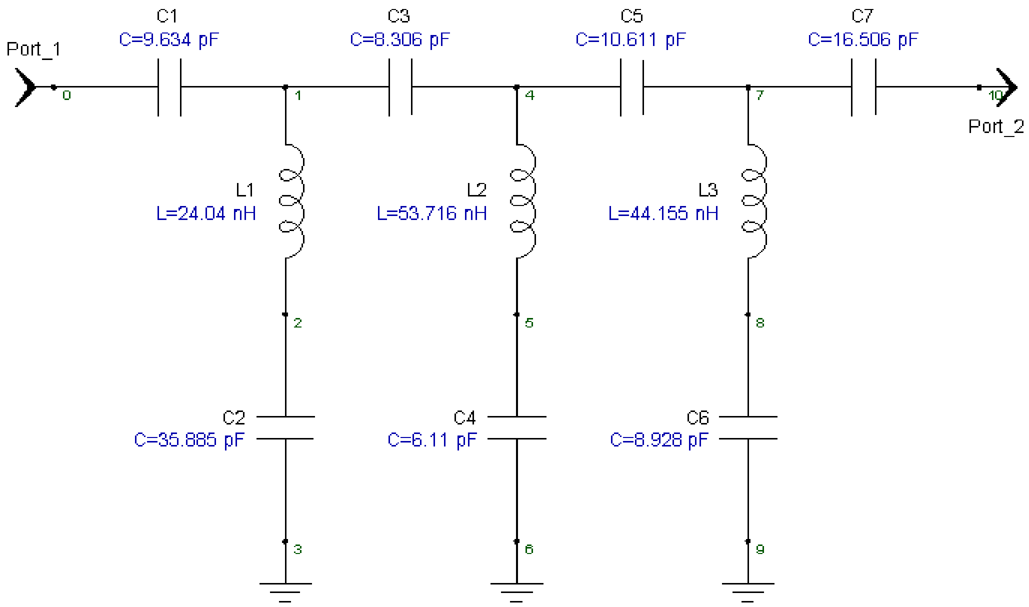


Fig. Frequency Scaled HPF Circuit

These finite transmission zeros are added to increase the steepness of the filter response. The overall response of the HPF is as shown below:

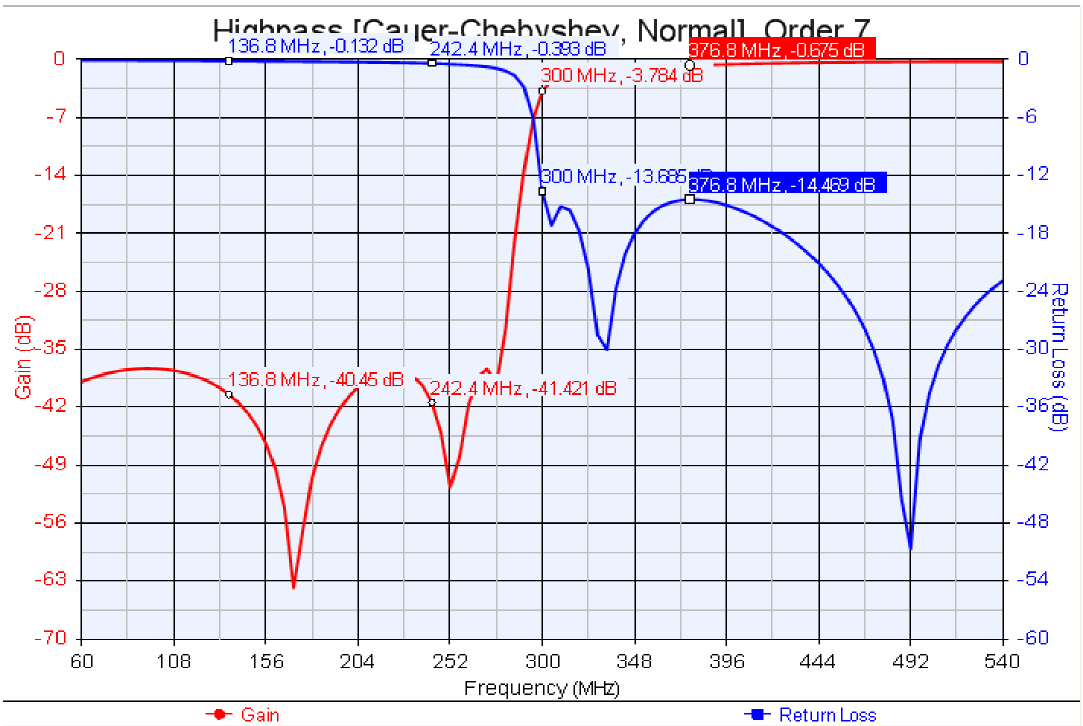


Fig. Simulated Filter Response of HPF

Now the overall band-pass filter is obtained by combining the LPF in series with the HPF. However on doing so the response obtained does not meet the required specifications. Therefore the components were tuned using the simulation software **Eagleware.** The final filter circuit obtained after tuning is as shown below:

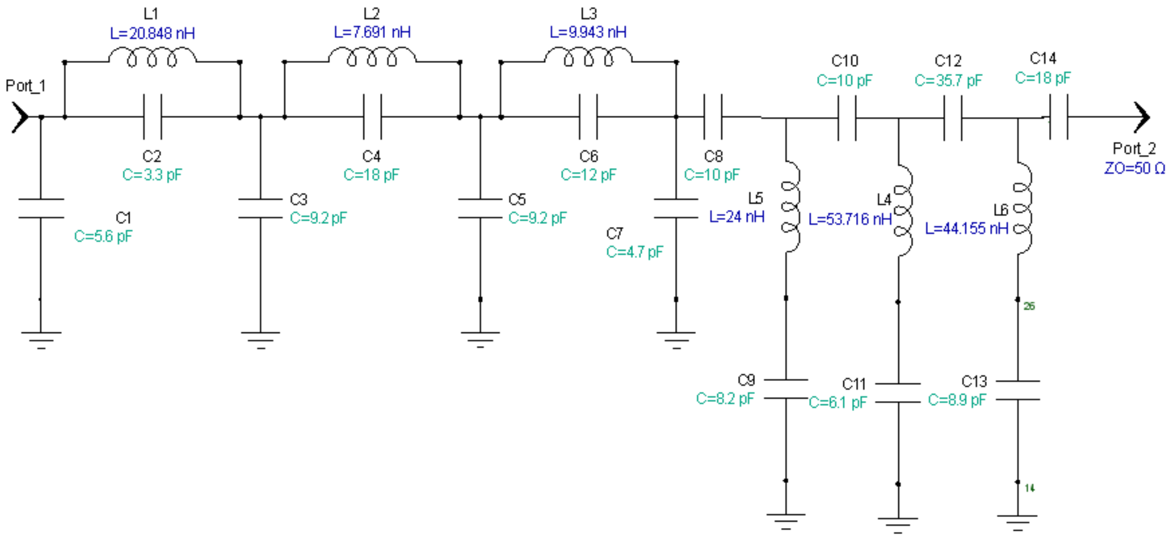


Fig. Final BPF circuit

The overall bandpass filter response is as shown below:

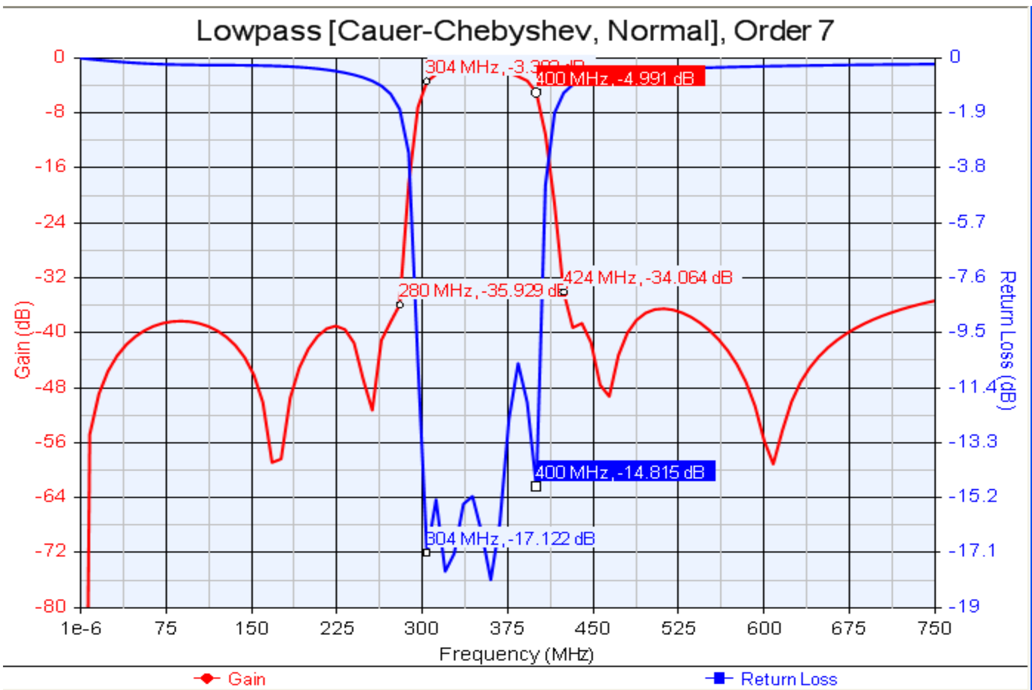


Fig. BPF response after tuning on Eagleware

Practical Bandpass Filter Tuning

Since the bandpass filter consists of series connection of an LPF and an HPF, both these filters can be individually tuned to have desired filter response. The tuning of the individual filters can be accomplished as follows

- **Lowpass filter tuning**

Objectives:

3 dB cutoff frequency = 400 MHz
Stopband frequency is 420 MHz
Insertion loss > -1 dB from in the pass band (from DC to 400 MHz)
Return loss < -10 dB in the passband (from DC to 400 MHz)

Procedure:

In the low pass filter major tuning part is the series arm circuits, which provide a zero/null at particular frequencies and thus provide a required degree of steepness.

- The first series arm parallel resonant circuit (L_1,C_2) provides null at 650 MHz. So by adjusting this arm capacitor C_2 improves rejection on higher frequency side. Adjust this capacitor so as to get minimum attenuation at the side lobes.

- The second series arm parallel resonant circuit (L_2,C_4) provides null at 421 MHz. So adjust this arm capacitor C_4 to get cutoff at 400 MHz and give sharp steepness to the high pass filter in its transition band.

- Third series arm parallel resonant circuit (L_3,C_6) provides null at a finite frequency of 454 MHz. So adjust it to get desired attenuation at 454 MHz, hence restricting the attenuation of the return lobes.

- Capacitor C_1 and C_7 are input and output capacitors respectively. These are used to match impedances of input and output circuit to whom they are connected. So adjust them so that insertion loss and the return loss is optimum.

- Capacitor C_3 and C_5 are impedance matching capacitors in between two series arm parallel resonant circuits to optimize insertion and return loss.

- ## Highpass Filter Tuning

Objectives:

- 3 dB cutoff frequency = 300 MHz.
- Stopband frequency is 280 MHz.
- Insertion loss > -1dB in the pass band (from 300 MHz to infinity).
- Return loss < -10 dB in the pass band (from 300 MHz to infinity).

Procedure:

In the high pass filter major tuning part is the the shunt arm series resonant circuits,which provide a null/zero at a finite frequencies and thus provide a required degree of steepness.

- The first shunt arm series resonant circuit(L_1,C_2) provides null/zero at finite frequency of 171 MHz . This improves rejection at lower frequency side. So the capacitor C_2 can be tuned to adjust the response at lower frequency.

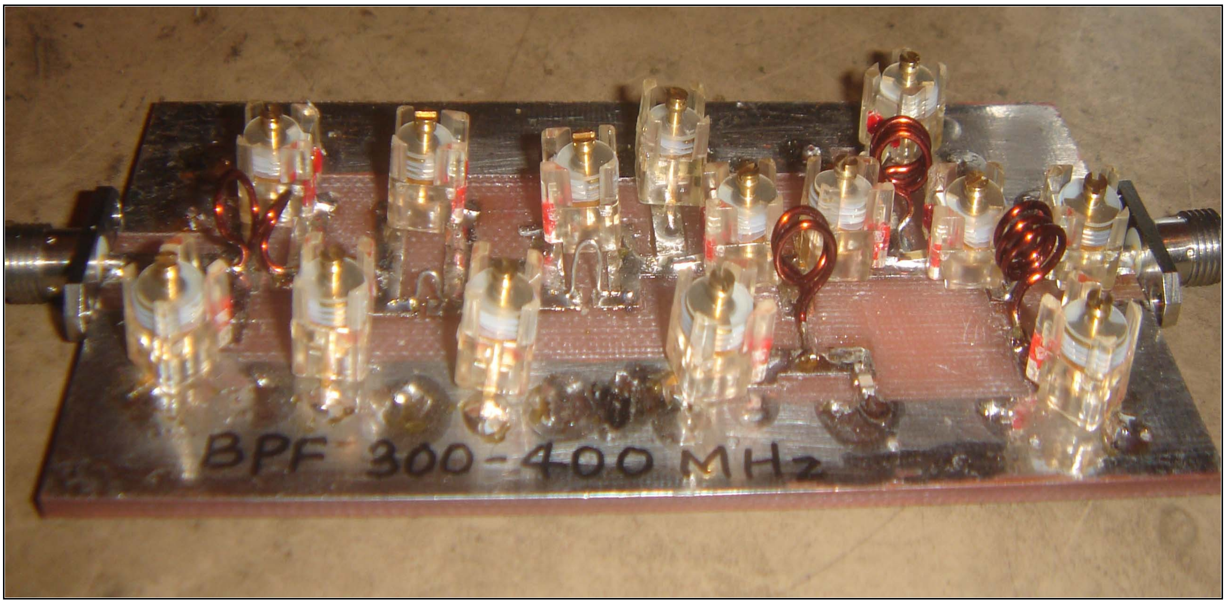
- The second shunt arm series resonant circuit(L_2,C_4) provides null/zero at a finite frequency of 278 MHz . So adjust this arm capacitor C_4 to get cutoff at 300 MHz and give sharp steepness to the high pass filter in its transition band.

- Third shunt arm series resonant circuit(L_3,C_6) provides null/zero at 253 MHz to improve rejection on lower frequency side. Since this frequency lies very close to stopband frequency, capacitor C_6 is tuned to have high attenuation at the side lobes of the HPF.

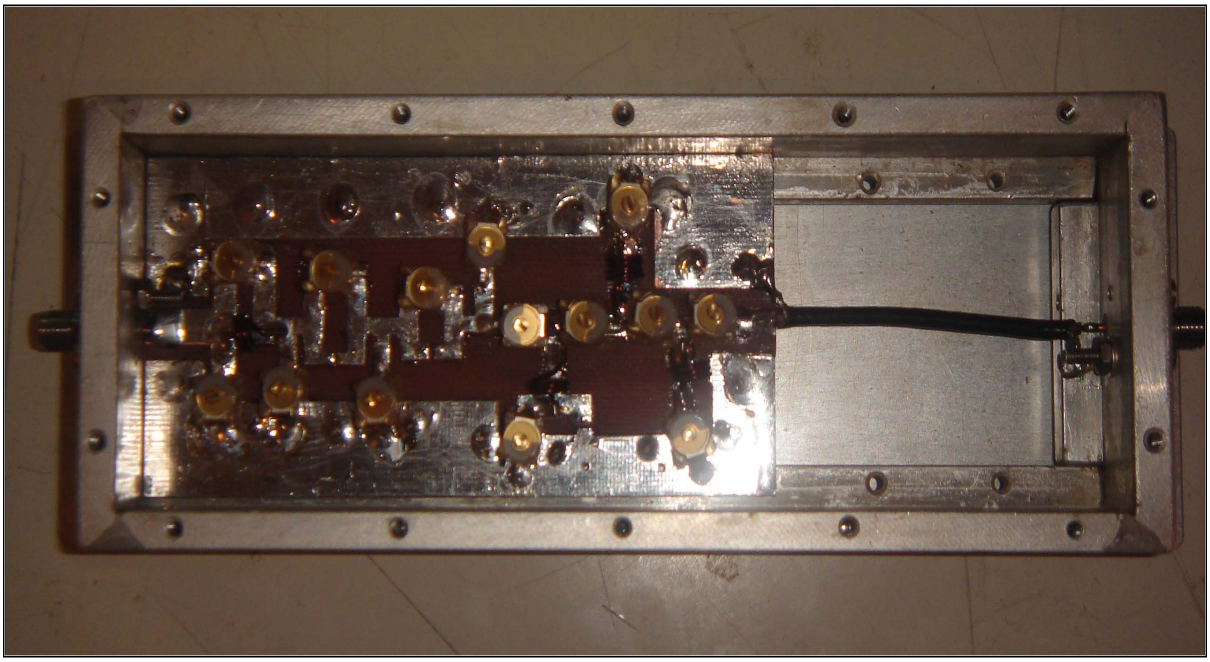
- Capacitor C_1 and C_7 are input and output capacitors respectively. These are used for input output impedance matching, thus reducing insertion loss and return loss.

- Capacitor C_3 and C_5 are impedance matching capacitors in between two shunt arm series resonant circuit to optimize the return loss and insertion loss .

Prototype of the Final Circuit



Final Bandpass Filter Chassis



Results after Practical Realization

The designing of the improved bandpass filter has been completed successfully. The filter has a bandwidth of 100 MHz with 3 dB cutoff at 300 MHz and 400 MHz frequencies. It has sharp roll-off characteristics on both the low pass and the high pass side, providing enough attenuation for the out of band signals and strong interference frequencies.

The filter has minimum stopband attenuation of 30 dB , with nearly the same attenuation at 280 MHz and 420 MHz providing the necessary image rejection.

The overall response of the bandpass filter that was realized is shown in Appendix A.

The phase response plot for the practical designed filter that was realized is shown in Appendix A.

The group delay for the filter is nearly constant over the entire passband, with slope near the midband accounting for the corresponding distortion in the phase response plot. The group delay shows surges at the cut-off frequencies of the BPF indicating steep slope in the phase transfer function of the filter in the neighborhood of these frequencies. The group delay plot of the practical designed filter that was realized is shown in Appendix A.

Conclusion and Future Scope

In this project report, we have tried to highlight the overview of the GMRT receiver system and the existing 327 MHz Front-End system. We have successfully designed and implemented a pair of filters having a bandwidth of 100 MHz from 300 MHz to 400 MHz. The report also deals with the detailed aspects of design of filters.

Currently a Low Noise Amplifier with a band coverage of 200 – 500 MHz is being developed in the Front-End lab. If these filters are integrated with the new LNA, we can get a very good frequency response for the front-end from 300 MHz to 400 MHz.

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- Sandeep C. Chaudhari**, *Modified 150MHz Front-End System Incorporating Filters for RFI Mitigation*,** NCRA Graduate Trainee Project Report, 2004-05, Guided by Mr Anil Raut and Mr. A. Praveen Kumar.

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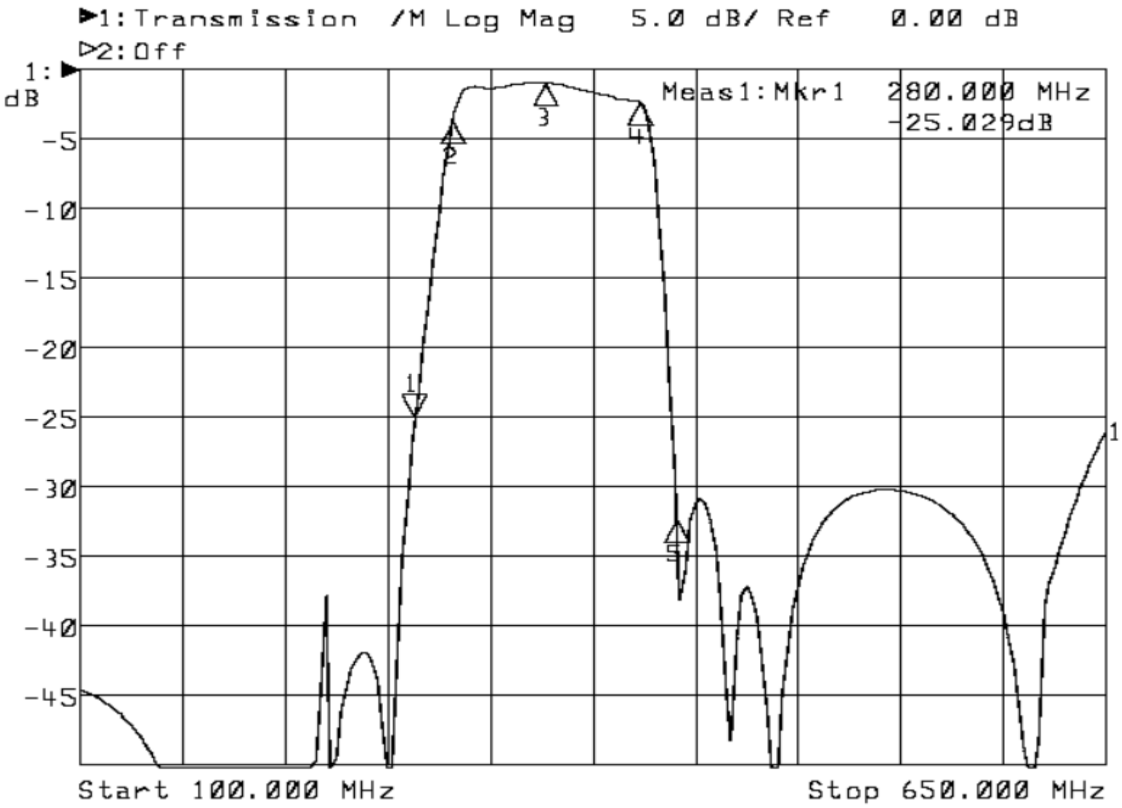
- A. Praveen Kumar and M. Srinivas, *Signal Flow Analysis of the GMRT Receiver System,* GMRT Internal Technical Report, October 1996.

- A. Praveen Kumar and Anil Raut, ***Improvement of GMRT Receiver for better Dynamic Range***, GMRT Internal Technical Report, November 2003.

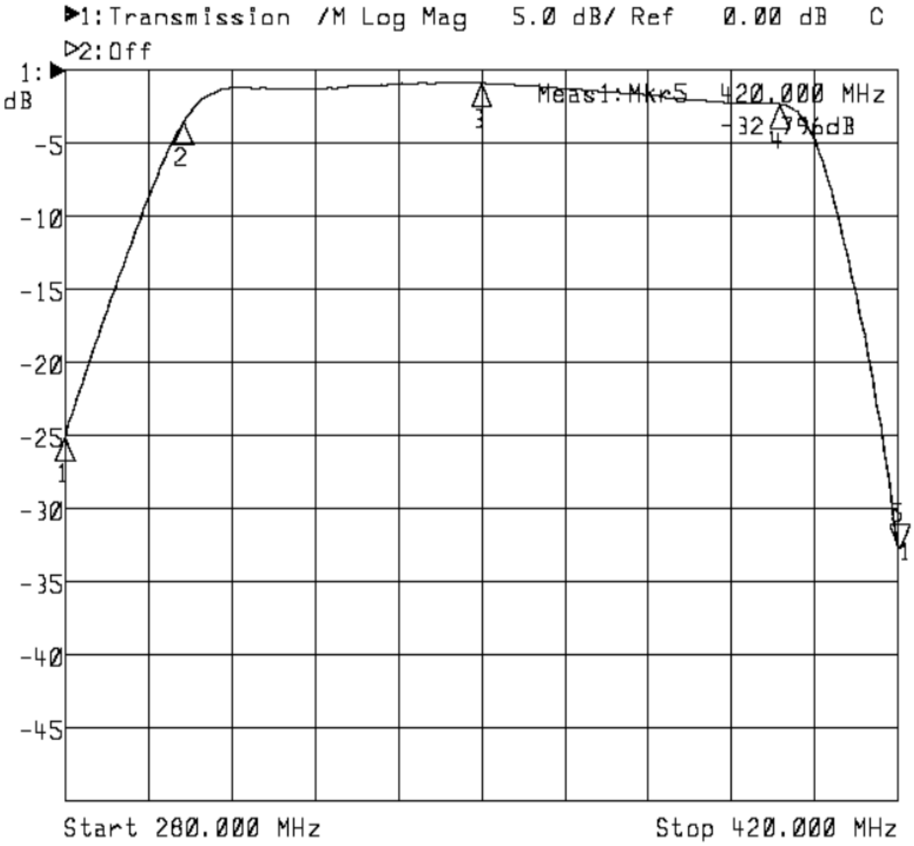
- Eagle-ware GENSYS software manual for M-filter design.

Appendix



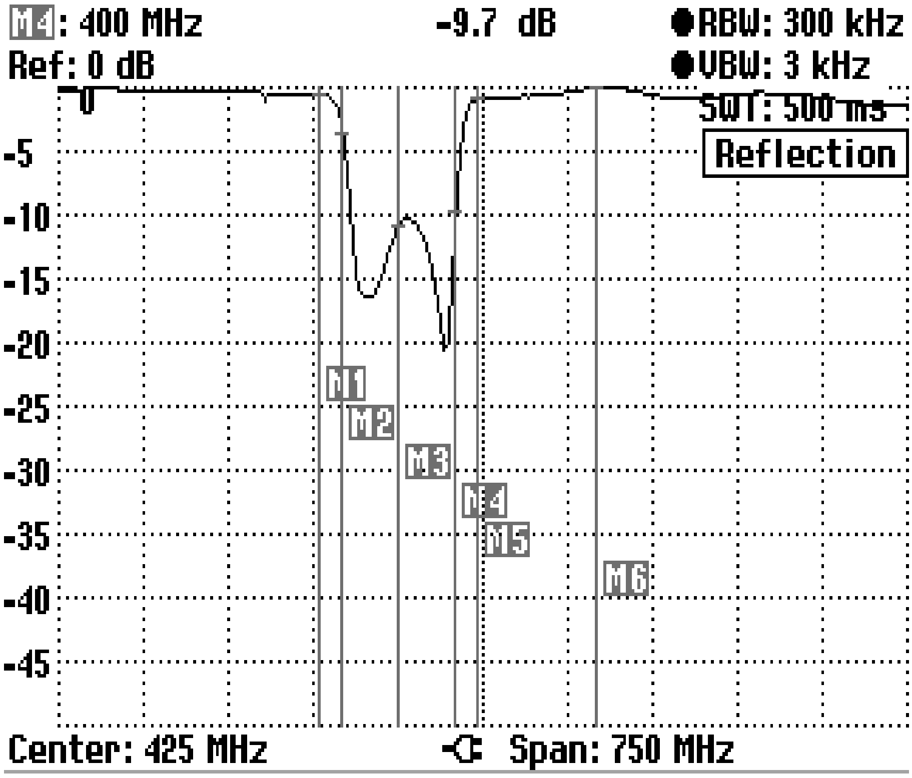
Magnitude Response

1:Mkr (MHz)	dB
1> 280.0000	-25.029
2: 300.0000	-3.693
3: 350.0000	-0.971
4: 400.0000	-2.318
5: 420.0000	-32.337



Passband Magnitude Response

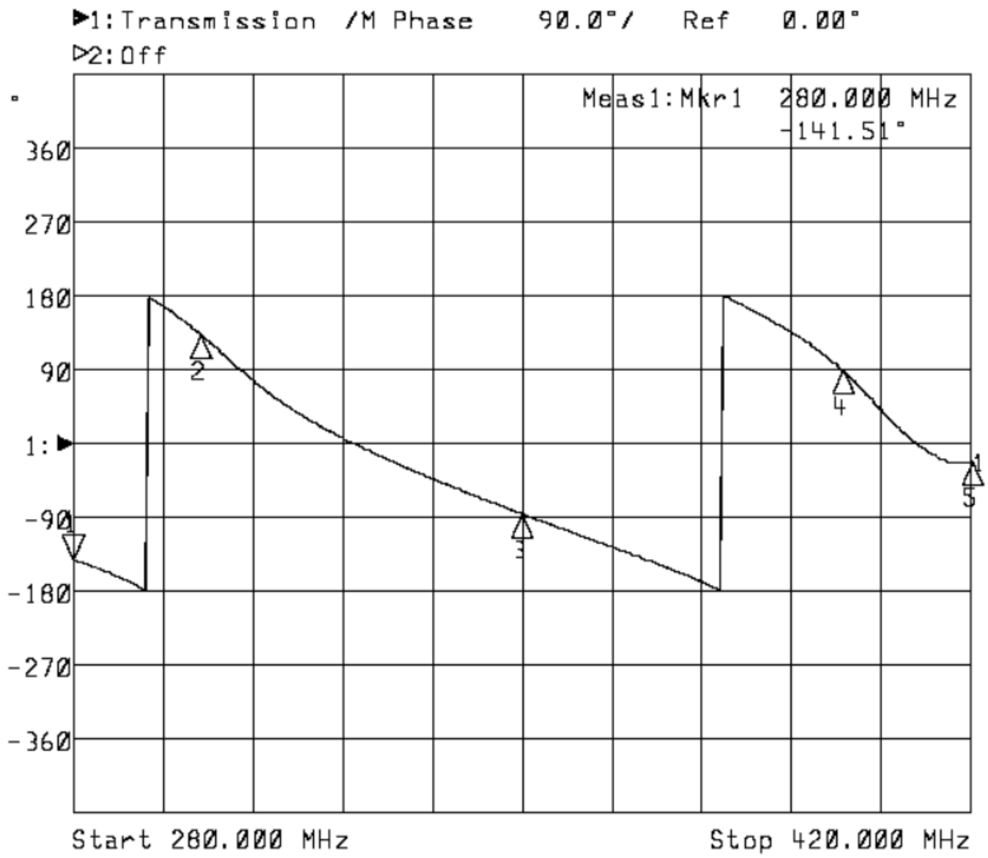
1:Mkr (MHz)		dB
1:	280.0000	-25.080
2:	300.0000	-3.558
3:	349.9533	-0.910
4:	400.0000	-2.320
5>	420.0000	-32.796



Return loss

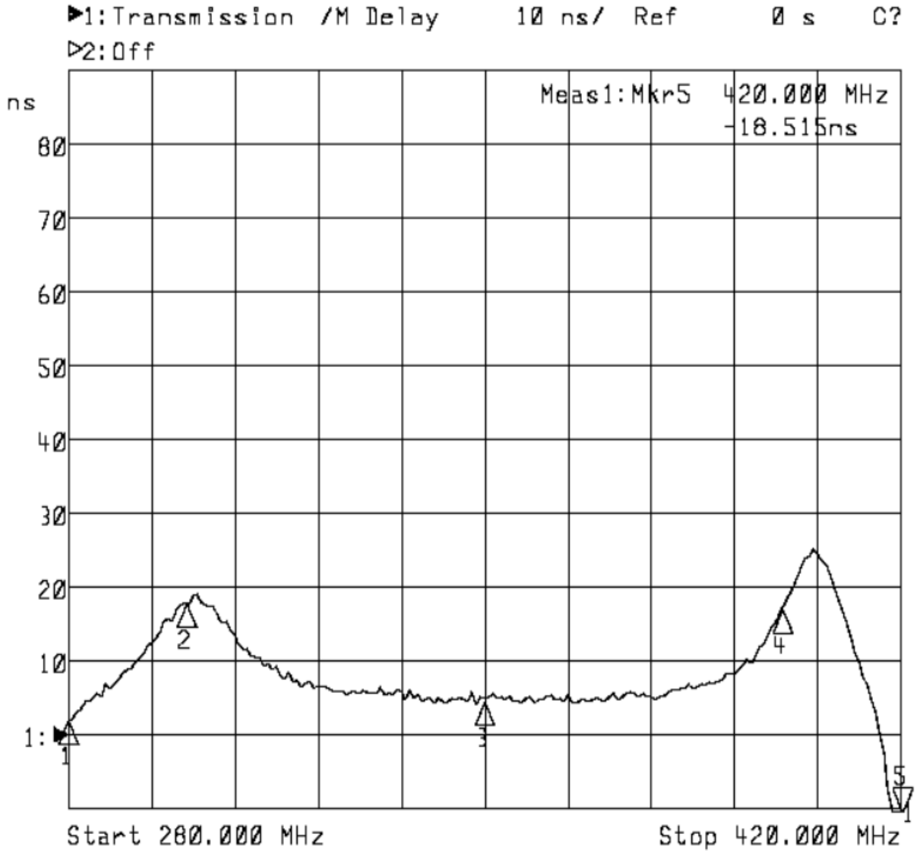
Markers

M1	**280 MHz**
M2	**300 MHz**
M3	**350 MHz**
M4	**400 MHz**
M5	**420 MHz**



Phase Response

1:Mkr (MHz)	Deg
1> 280.0000	-141.51
2: 300.0000	132.27
3: 350.0000	-85.317
4: 400.0000	89.388
5: 420.0000	-21.506



Group Delay

1:Mkr (MHz)	s
1: 280.0000	1.9219n
2: 300.0000	17.747n
3: 350.0000	4.4577n
4: 400.0000	17.075n
5> 420.0000	-18.515n