DESIGN AND ANALYSIS OF A BASEBAND RECEIVER
for
The Giant Metrewave Radio Telescope

M.Tech Dissertation

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Submitted in partial fulfillment of the requirements for the award of the degree of
Master of Technology in Electrical Engineering of the
Indian Institute of Technology, Bombay.

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ABSTRACT

A Baseband Receiver is developed for the Giant Metrewave Radio Telescope (GMRT). This report describes in detail the system design aspects and the circuit implementation. An introduction to radio telescopes and various types of receivers is given. The receiver system for GMRT is described and the requirements of the Baseband system highlighted. The system specifications are arrived at so as to improve the facilities available in the telescope receiver and to facilitate a wide variety of observations. The design of various units is done using advanced high frequency components and the circuits are wired and tested for their performance. Sophisticated RF equipments are used in testing the parameters of the Baseband Receiver. A suitable mounting scheme for the circuits is planned so as to reduce any RF interference to the observations. Alongwith the measured performance of units the response is calculated and compared with the actual values obtained.

DISSERTATION APPROVAL SHEET

The dissertation titled *"Design and Analysis of a Baseband Receiver for the Giant Metrewave Radio Telescope"* submitted by Ajith Kumar B. (Roll No. 97307416) is approved for the degree of Master of Technology.

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SKrishnan		

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Ajith Kumar B
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Chapter 1

INTRODUCTION

1.1 RADIO ASTRONOMY AND TELESCOPES

Radio Astronomy deals with the science of observing the outer space at radio frequencies. This relatively young branch of science has its beginnings in the experiments by Karl G Jansky in 1931. Optical astronomy and radio astronomy utilize two transparent bands of earth's atmosphere and ionosphere which are referred to as the optical and radio windows respectively. The radio sky contains thousands of radio emitting objects some of which emit only in the radio window. Observations at frequencies outside these windows is possible only with deep space probes. Radio astronomy offers several advantages over optical astronomy in the detection of celestial objects such as observations during both day and night, a large observing frequency coverage (10^4:1 as compared to 2:1 in case of optical) and its ability to penetrate our dust covered galaxy. Utilizing these advantages radio astronomers have discovered very interesting astrophysical objects like quasars, pulsars and complex organic molecules in space. Radio astronomy has also provided unprecedented angular resolution (10^{-4} arcsec) by the use of ground and space based interferometry. Hence it is a very powerful technique available for studying the distant parts of the universe.

The main function of a radio telescope is to detect and measure the radio emission of very low average power levels (of the order of 10^{-17} to 10^{-25} Watts/Hz) from radio sources but the total system noise power levels are much higher than this. However, high sensitivity can be achieved in the detection by reducing the output fluctuations from the receiver by measuring them over a sufficiently long integration time and over a large enough bandwidth. This makes the fluctuations decrease by the square root of the integration time (in seconds) and bandwidth (in Hertz), over the average power. Such measurements do require high stability in the telescope receiver. Special techniques are required to reduce the gain variations in the receiver systems. When observations are done on highly varying solar bursts or giant pulses from pulsars etc. it is also necessary that the receiver system of the telescope has adequate dynamic range (defined as the ratio of the largest to the smallest detectable signal).

A hypothetical radio telescope is shown in Fig 1.1. However the systems used varies between telescopes depending on the specifications. The antenna with feeds collect the radio power from distant celestial sources. After pre-amplification at the feed level, the signal is fed to the receiver system. Here the signal is detected, integrated and recorded, either on an analog chart recorder or stored in magnetic tapes for further off-line processing by high speed computers.

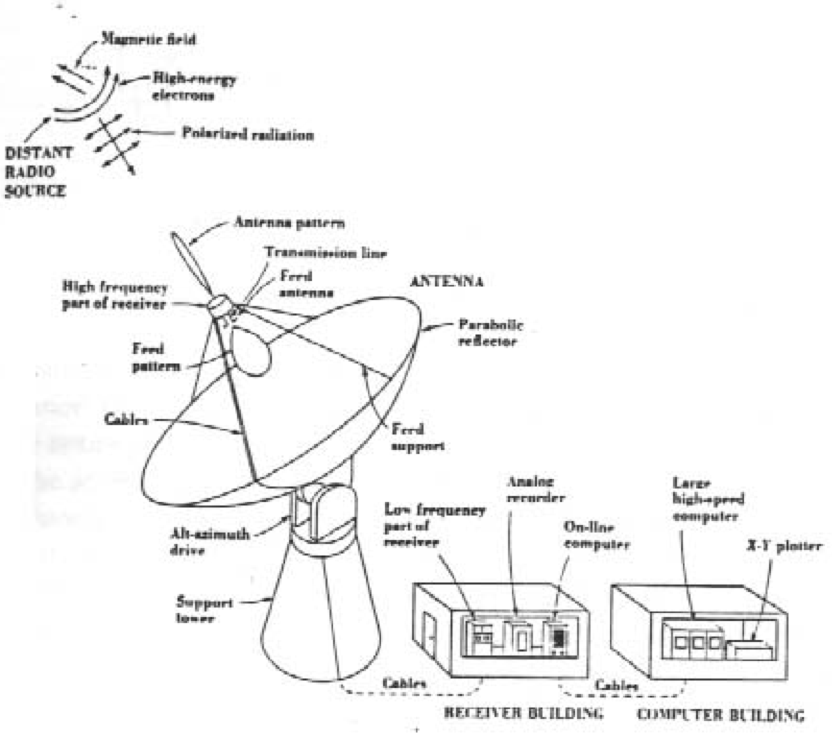

Fig 1.1 - Modern Radio Telescope

1.2 RADIO TELESCOPE RECEIVERS

A Radio Telescope Receiver (RTR) is similar in construction to an ordinary radio receiver. Any receiver produces an unavoidable noise signal which is indistinguishable from that collected by the antenna. It is in minimizing this to a very significant extent that the RTR is a great improvement over an ordinary radio receiver. The complexity of an RTR can be anywhere between a simple direct detection receiver to a multi-channel superheterodyne receiver. Radio telescope receivers may be classified into several categories. A detailed explanation of these different categories is given in [5]. A detailed explanation of Synthesis Radio Telescopes is given in [6].

A total power receiver is one which measures the total noise power from the antenna and the receiving system, unlike some receivers which measure the difference in power from the antenna and some reference. Fig 1.2 shows a simple total power receiver. One of the major drawbacks of a total power receiver is the sensitivity reduction due to receiver instability. In order to stabilize the gain of the receiver, Ryle and Vonberg and later Dicke used very frequent calibrations. The input to the receiver was switched at the rate of several tens of hertz between the antenna and a resistance kept at a constant temperature, T_0 so that effectively the receiver noise is cancelled or greatly reduced.

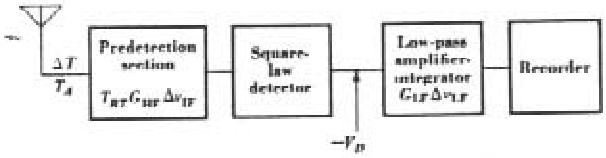

Fig 1.2 - Total Power Receiver

A radio interferometer receiver consists of two or more antennas separated from each other by a distance of several wavelengths. Each antenna is equipped with its own preamplifiers and entire pre-detection sections with identical electrical characteristics. (see Fig 1.3). When the antennas are pointing to the same source, equal signal voltages will be induced with a phase difference depending on the source direction. The IF signals from the receiver outputs are added and fed to a square law detector. The use of two antennas and two receivers increases the maximum sensitivity in case of an interferometer receiver to twice as that of a total power system with same system temperature. However sensitivity is still affected by gain instabilities similar to a total power receiver. Such sensitivity variations may be stabilized either by synchronous switching in both receivers or by correlation method.

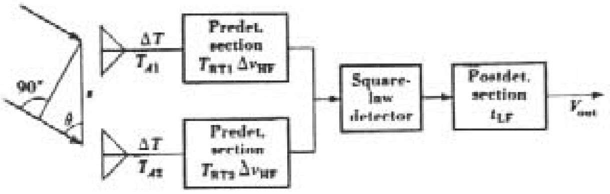

Fig 1.3 - Interferometer Receiver

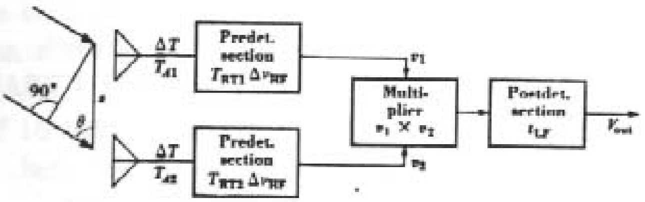

Fig 1.4 - Interferometer with Correlation Receiver

In a correlation receiver the IF output voltages are multiplied as shown in Fig 1.4. Uncorrelated noise voltages when multiplied gives an average value of zero. Only correlated noise voltages give a dc output and hence voltage gain instabilities will not affect the sensitivity of the correlation receiver. It will only change the calibration of the receiver. Random phase variations in the amplifiers in the predetection section are undesirable and hence scintillations in ionosphere can reduce the sensitivity. A correlation receiver can be designed to have a much lower noise temperature as there is no switch (hence no extra losses) between the antenna and the receiver.

1.3 GIANT METREWAVE RADIO TELESCOPE

The Giant Metrewave Radio Telescope (GMRT) is set up by the National Centre for Radio Astrophysics (NCRA), Pune of the Tata Institute of Fundamental Research (TIFR) as a national facility, at village Khodad near Narayangaon in Pune District. The GMRT is the most powerful facility in the world for astronomical research in the metre and decimetre wavelengths (30 to 1420 MHz).

There are many outstanding and challenging astrophysical problems which can be studied only at metrewavelengths. India offers a suitable environment for radio astronomy observations at metrewavelengths due to low levels of man-made RF interference compared to developed countries, wider coverage of both northern and southern skies from a suitable location in South India, labour intensive nature of the metrewave radio telescopes, comparatively cheaper labour available in India, and the experience gained by TIFR in the construction of low cost antennas for the Ooty Radio Telescope (ORT).

The GMRT is designed to investigate a wide variety of celestial phenomena, reaching from our solar system to the very edge of the observable universe [1]. It still retains the versatility to allow its use for a wider range of studies. The major scientific objectives of the GMRT include high resolution radio imaging of the sun and other galactic and extra galactic sources, search for massive neutral Hydrogen clouds, expected to have existed some 15 billion years ago according to the Big Bang model, search for millisecond and binary pulsars.

The GMRT consists of 30 fully steerable parabolic dish antennas each with a diameter of 45 metres arranged in the shape of "Y". 12 of these antennas are kept in a central array of about 1Km X 1Km. The remaining antennas are arranged in the arms of the "Y" with 6 antennas in each arm. Each of these arms extend over 15 kms. Fig 1.5 shows a sketch of the location of the 30 antennas. The mechanical design of the dishes uses a novel concept called SMART (Stretched Mesh Attached to Rope Trusses). The back up structure consists of 16 radial parabolic frames made of tubular steel which are connected to a 12 mtr central hub. The permissible rms surface errors in the reflecting surface are 8, 10, 12 mm in the central, middle and outer areas of the dish. One of the dishes of the GMRT is shown in Fig 1.6.

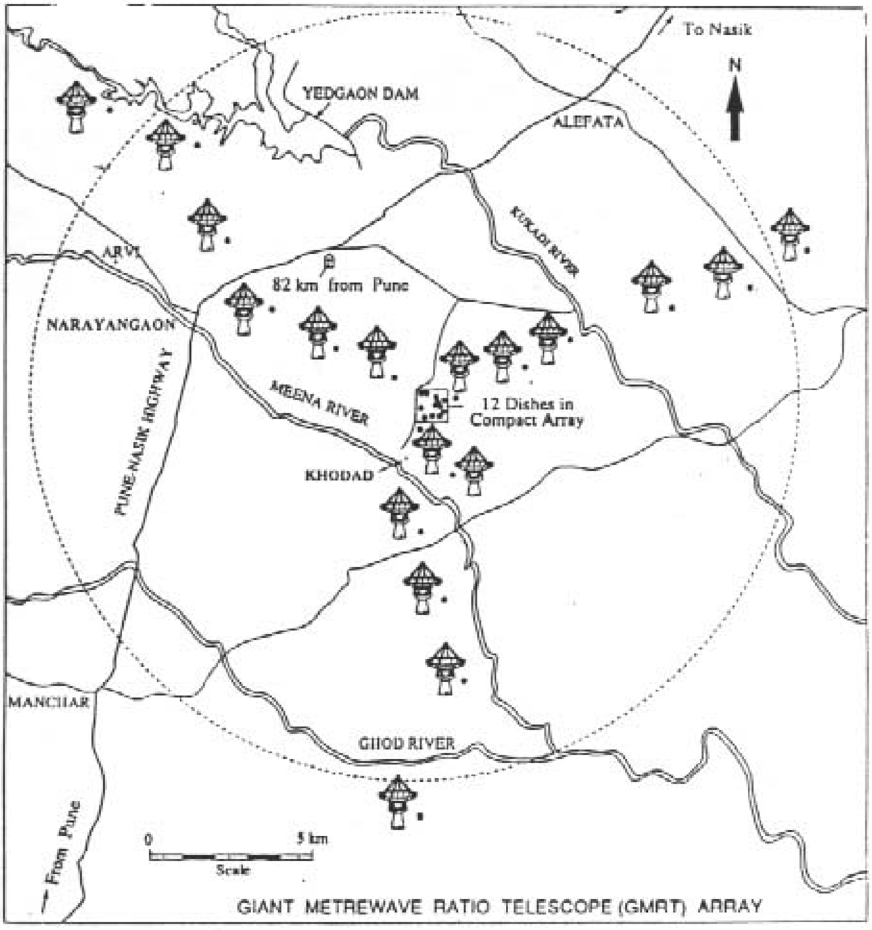

Fig 1.5 - Y Shaped GMRT Array Configuration

Fig 1.6 - GMRT Dish Antenna

The Feed and the Receiver electronics are designed to provide observing capability at 150, 233, 327, 610, and 1420 MHz bands. The specifications for GMRT represent a major leap over the existing radio telescopes in terms of its sensitivity and high resolution imaging at metrewavelengths. The GMRT will operate as an Earth Rotation Synthesis Radio Telescope and provide a resolution of 2 arc second at 1420 MHz band. It will provide a sub-milli Jansky sensitivity (1 Jansky = 10^{-26} W/m^2/Hz). The total effective area of GMRT is 35,000 sq. mtrs (approx.) upto 600 MHz and 27,000 sq. mtrs (approx.) upto 1420 MHz.

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Chapter 2

GMRT RECEIVER SYSTEM

The Giant Metrewave Radio Telescope (GMRT) uses the principle of correlator interferometer to achieve a high sensitivity. The GMRT uses highly sophisticated receiver electronics that is capable of making simultaneous observations at two frequencies or at two polarisations. It uses superheterodyne receivers with a very low noise system to collect the signals at each of the antennas, bring them over to a central station without loosing the phase information, and then combine them to provide a high signal to noise ratio. The signals received at the feeds are processed in low noise systems at the dish focus and then brought down to the dishbase where it is converted to an intermediate frequency (IF) signal using coherent local oscillator (LO) systems. The IF signals are then brought over to the Central Electronics building (CEB) where it is further down converted to baseband (BB) frequencies and then digitized before correlation. A simplified block schematic of the GMRT receiver system is shown in Fig 2.1.

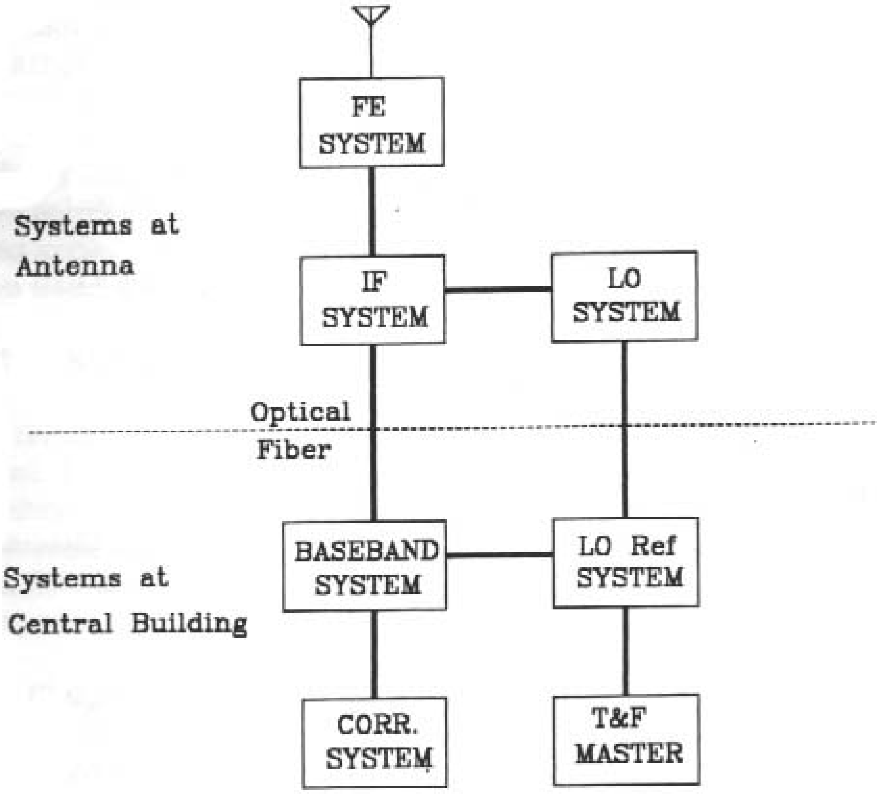


Fig 2.1 - GMRT Receiver System (simplified)

2.1 FEEDS AND FRONT END SYSTEMS

The signals received by the dish antenna are converted to electrical signals using separate feeds located at the dish focus, for each of the observation frequencies. There are different feeds currently operating at 150, 233, 327, 610, and 1420 Mhz, arranged on the four faces of the rectangular structure. This feed structure may be rotated by a feed positioning system to bring the required feed to prime focus. 610 and 233 Mhz feeds are mounted on the same side and hence simultaneous observation at these two frequencies is possible. The feed system provides low loss impedance matching baluns and other associated circuitry.

The signals are then processed through low noise front end systems located at the prime focus just behind the feed system. The Front-end Electronics system converts the two linearly polarized signals (orthogonal to each other) received by the feed system into two circularly polarized signals for further processing. Very low noise amplifiers are used to reduce the overall system temperature. Front end electronics also includes circuits for walsh modulation of the signal from each antenna so as to reduce interference and cross talk. Calibration facility using a noise source with switchable input noise power is also provided. This facility may be used as a secondary calibration system, the primary being the standard celestial sources.

Another facility provided in the front end electronics is the solar attenuator, which is basically a switchable attenuator with large attenuation value. This attenuator is generally switched in, while observing strong sources like sun or sources close to it, to avoid saturation of the electronic systems. At each frequency two channels; one for left circularly polarized and one for right circularly polarized signals are available. A band selector switch at the front end system selects any two of the available channels, it may be two polarisation channels of same frequency or same polarisation channel of two different frequencies. The selected RF signals are then brought over to the dish base using two low-loss shielded RF cables.

2.2 SYSTEMS AT THE DISHBASE

At the dish base the signals are further processed in electronic racks kept in a shielded room. The RF signals are first converted to an intermediate frequency signal using a local oscillator. The IF signal processing includes band shaping using switchable filter banks, switchable attenuators for power level adjustment, an automatic level control circuitry etc. The filter banks may be used to remove any interfering signals and the ALC circuits limit the power fed to the optical transmitter within the acceptable range.

The IF signals from each of the antennas are then brought over to a Central Electronics Building (CEB) through optical fiber cables. As the same fiber is carrying both the IF signals from each antenna, one IF is set to 130 Mhz and the other to 175 Mhz before sending over the fiber. The IF signals are sent as analog signals using a 1400 nm optical signal.

2.3 SYSTEMS AT THE CENTRAL BUILDING

The IF signals are recovered in optical receivers and then further down converted and processed at the CEB using baseband electronics. Each IF channel is brought down to frequencies below 16 Mhz using high image rejection single sideband mixers. An active switchable filter at baseband frequencies provides variable bandwidths for observation. The IF to baseband conversion local oscillator can also be used to do doppler tracking of moving celestial sources. The baseband system is explained in detail in the next chapter.

The signals are then fed to a correlator system which digitizes the signals from each polarization channel and feeds to a 512 channel FFT machine. Details of GMRT Correlator system is given in [12,13]. The resulting spectra are then cross multiplied, adjacent channels averaged after rejecting channels with RF interference, and integrated to produce 238,000 (approx.) outputs to be read by the GMRT computer system. The cross multiplication and accumulation of signals over long time periods improve the overall signal to noise ratio. Normally the integration time is variable from 0.1s to 30s.

2.4 LOCAL OSCILLATOR SYSTEM

Coherency between the local oscillators at remote locations, used in frequency conversions as mentioned above, is critical for proper operation of a synthesis radio telescope. A master time and frequency reference at the CEB is used as the standard frequency source. An LO reference signal generated from this master source is sent to all antennas through the forward optical fiber link, where this signal is used to generate various local oscillator frequencies using phase locked loop techniques. A portion of LO reference signals sent to the antennas is brought back over the down link to the CEB and is used to calculate the phase variation over the fiber in the 100-200 Mhz range. The same master reference is used in the generation of local oscillator signals for the IF to baseband conversion also.

*** *** ***

Chapter 3

BASEBAND SYSTEM - AN OVERVIEW

The major function of the baseband system at the Central Electronics Building (CEB) is to convert the IF signals received from the antennas to frequencies suitable for the sampler units in the correlator [14]. Baseband system also facilitates processing of baseband signals by providing switchable bandwidths for the signal for various types of observations. The baseband system provides level control circuits so that the power fed to the sampler is always within the acceptable range. It also provides buffer circuits for driving the sampler units.

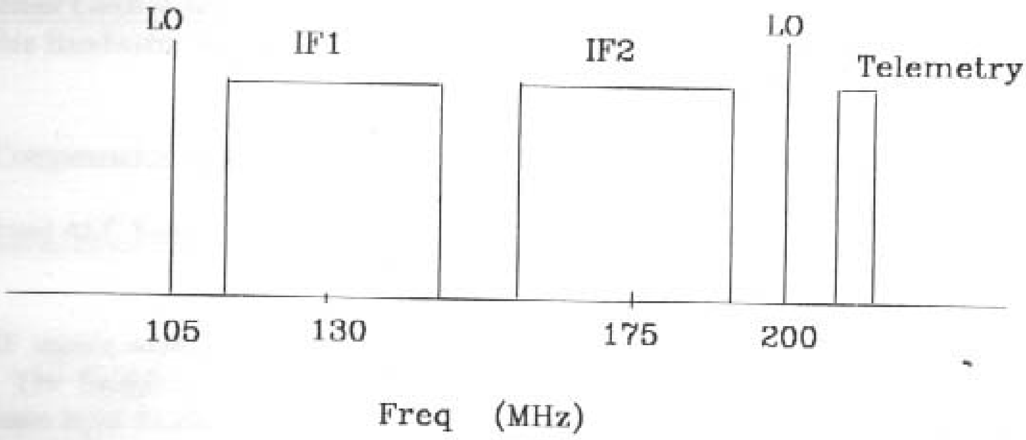


Fig 3.1 - Optical Receiver Output Spectrum

The spectrum of the input signal received through the optical fiber links at the CEB is shown in Fig 3.1. The total integrated power at the output of optical receiver will not exceed 0 dBm. This includes two LO reference signals of -25 dBm and two IF channels of -17 dBm each along with the telemetry signals. The baseband system provides the facility for splitting the incoming spectrum and separate out the LO, IF and telemetry signals before feeding them to the respective units. The 105 and 200 MHz LO signals are used for round trip phase measurements and hence goes to the LO system units. Similarly the telemetry signal is given to the telemetry system for further processing.

3.1 BASEBAND SYSTEM SPECIFICATIONS

The major specifications of the baseband system are as given in Table 3.1. These specifications are of three categories, the first one based on the astronomical requirements for observation, the second one based on the optimum performance of the receiver electronics, and the third category includes other specifications defined to improve the overall performance of the baseband system.

Table 3.1 System Specifications

Input IF Characteristics	
Input IF Signal Centre Frequency	130 Mhz, 175 Mhz
IF Signal Bandwidth	32 Mhz
Input IF power (in each channel)	-17.0 dBm
Baseband Signal Characteristics	
Baseband Output Frequency Range	10 Khz to 16 Mhz
Baseband Signal Power Level	0 dBm ± 0.4 dBm
Baseband Image Rejection	> 25 dB
Passband Ripple	< 0.5 dB
Stopband Attenuation	> 40 dB
ALC Time Constant	1.0 sec
Baseband Control Settings	
Possible Bandwidth Selections	62 Khz, 125 Khz, 250 Khz, 500 Khz, 1.0 Mhz, 2.0 Mhz, 4.0 Mhz, 8.0 Mhz, 16.0 Mhz
BW Compensation Gain Selections	0 dB, 3dB, 6dB, 9dB, 12dB, 15dB, 18dB, 21dB, 24dB
Baseband ALC Switch	On / Off

The IF signals received at CEB are at 130 and 175 Mhz each with a bandwidth of 32 Mhz. The Sampling frequency used in correlator system is 32 Mhz and hence the maximum input frequency to the sampler is 16 Mhz. Hence each IF channel has to be split into two sidebands of maximum 16 Mhz each before sampling. Thus at the output of Baseband System there are four sideband signals corresponding to each antenna. The power in each IF band in the received spectrum is limited to -17.0 dBm. The sampler units power requirement is approx. 0 dbm ± 0.4dB. Hence the Baseband system should provide the necessary gain to the signal.

Some of the specifications are based on the requirements for astronomical observations. These include the specifications of various bandwidths and gains [11], the time constant in the ALC units, facilities for doppler tracking etc. The facility for varying the baseband signal bandwidth from 62 KHz to 16 MHz in 9 different octave steps also helps in removing any unwanted interfering signal. Whenever the bandwidth is varied, the system gain has to be adjusted suitably so as to maintain a constant power to the sampler units. Doppler tracking is facilitated by providing a frequency selectable synthesizer for the IF to baseband frequency conversion. The time constant of the ALC circuits are decided as per the astronomical requirements. A facility for switching the ALC function on/off is also required for pulsar observations. The system should also provide necessary buffers for driving the sampler circuits.

Other specifications important from the system point of view include amplitude ripple of overall system less than 0.4dB peak to peak, Stop band attenuation higher than 40 dB/octave, coupling between sidebands less than -25 dB etc. The system should also have facilities for remote control & monitor through computer, all parameters like the system bandwidth, gain, ALC on/off, baseband LO frequency should be controllable through computer. Also the intermodulation and harmonic products should be of sufficiently low level.

3.2 BLOCK SCHEMATIC

The Fig 3.2 shows the power level and frequency changes at different stages of the system. The overall block schematic of the Baseband system is shown in the Fig 3.3.

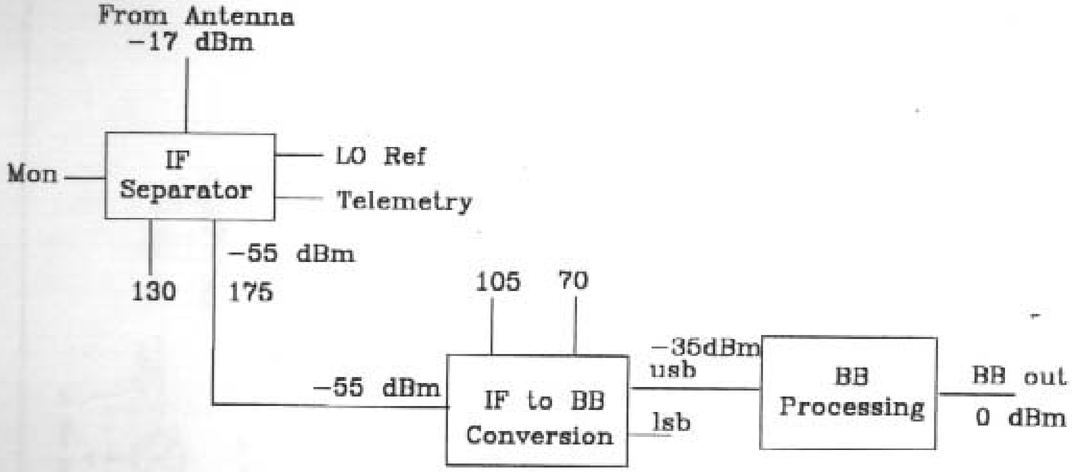


Fig 3.2 - Baseband System Simplified Block Schematic

3.2.1 Intermediate Frequency Circuits

The intermediate frequency circuits at the CEB include all systems for the recovery and processing of the IF signals from the optical fiber output. A torroidal power divider with suitable SAW filters are used at the input of the baseband circuitry to separate out the IF channels send through the optical fiber alongwith LO reference and other telemetry signals. The received signals are then converted to a centre frequency of 70 Mhz using double balanced frequency mixers and low pass filters to cut off the higher frequencies. An MMIC amplifier with a gain of 22 dB (approx.) provides necessary signal power for the IF to baseband conversion. Chapters 4 and 5 gives the design details of the IF circuits and the IF to Baseband conversion circuits respectively.

3.2.2 Baseband Processing Circuits

The baseband processing includes conversion of the IF signals to baseband as well as the processing at baseband frequencies. A high image rejection SSB mixer is used to convert each IF channel to two baseband channels of maximum 16 Mhz bandwidth. The baseband

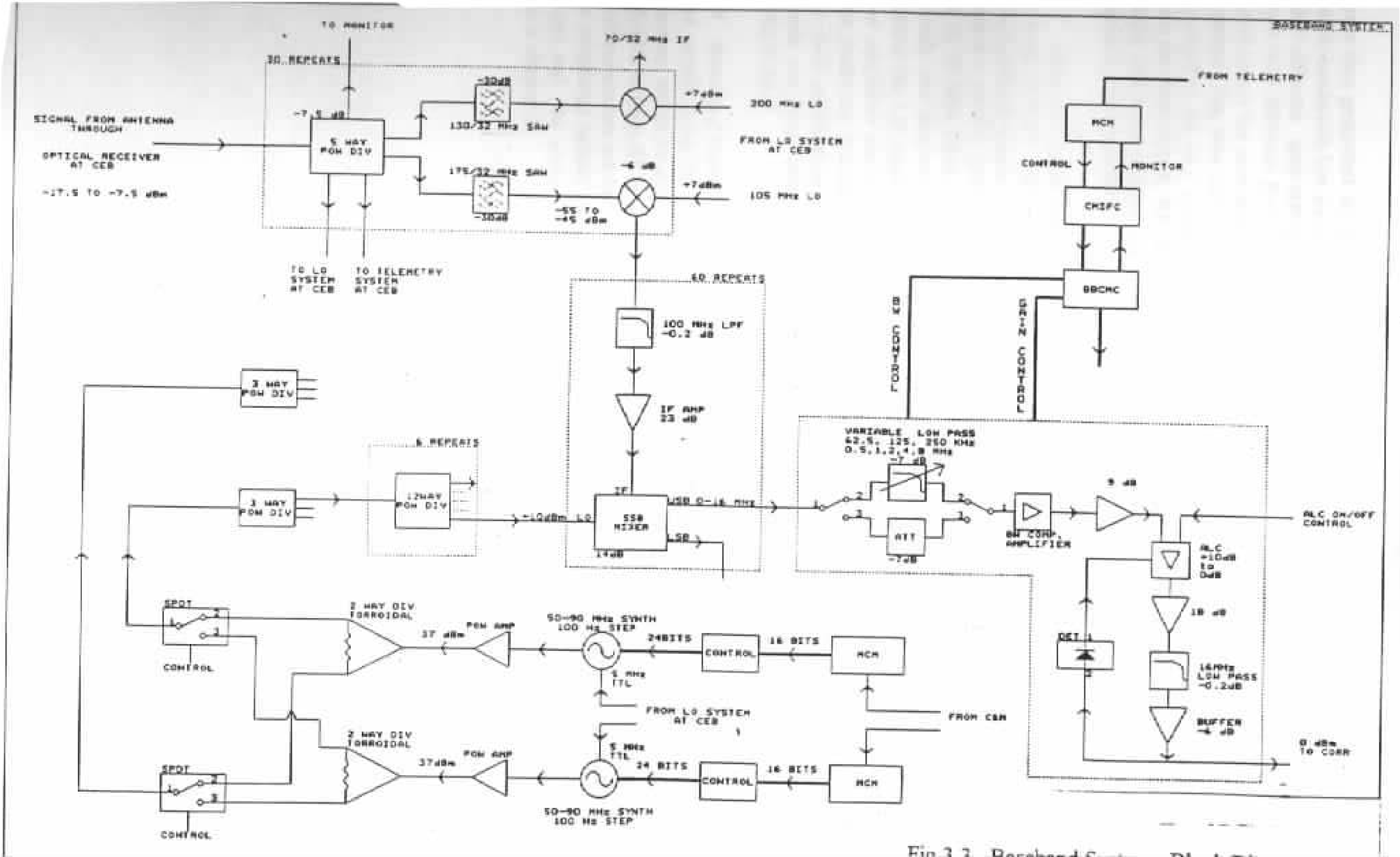


Fig.3.3 Baseband System Block Di

signals thus generated are processed in different circuits for proper bandshaping etc. A variable bandwidth low pass filter with 9 computer selectable bandwidths and an amplifier with 9 different gain settings are used together to provide a constant baseband power output with different frequency bands. The output circuitry includes an automatic level control circuit for providing a constant power to the sampler irrespective of system gain variations. The time constant of this ALC is 1.0 Sec so that faster variations are not attenuated by the ALC circuit. A high speed buffer circuit based on a sampler driver chip forms the final circuit, for driving long cable lengths from the baseband racks to the sampler racks and to drive the sampler input circuitry. Chapter 6 describes the Baseband processing circuits in detail.

3.2.3 Baseband Local Oscillator Circuits

The local oscillator circuits generate the necessary LO signals for frequency conversions in the baseband system. A 5 Mhz reference signal from the time and frequency reference standard is used to generate the local oscillator for the IF to baseband conversion. A 50 Mhz to 90 Mhz Direct Digital Synthesizer (DDS) based circuit is used as the oscillator. Normally this frequency is kept at 70 Mhz. The system has two synthesizers which generates the LO signals for all the antennas. The LO power is amplified through linear power amplifiers and divided and distributed to various SSB circuits. An LO switching circuit facilitates driving all antennas from same LO source or to have one synthesizer drive each polarization channel systems. In the second scheme the two polarization channels can have different frequencies of observation. Chapter 7 describes the LO circuits in detail.

3.2.4 Baseband Control & Monitor Circuits

These circuits form the interface between the computer and the analog circuits whose parameters are to be controlled. The main parameters to be controlled are the gain and bandwidth settings for each sideband of each polarization channel in every antenna. Other control parameters are the LO frequency setting, and ALC on/off setting. The parameters to be monitored include signal power levels at different points in the circuits, dc supply voltages and the control word settings. Control and Monitor circuits are explained in detail in chapter 8.

*** *** ***

Chapter 4

INTERMEDIATE FREQUENCY CIRCUITS

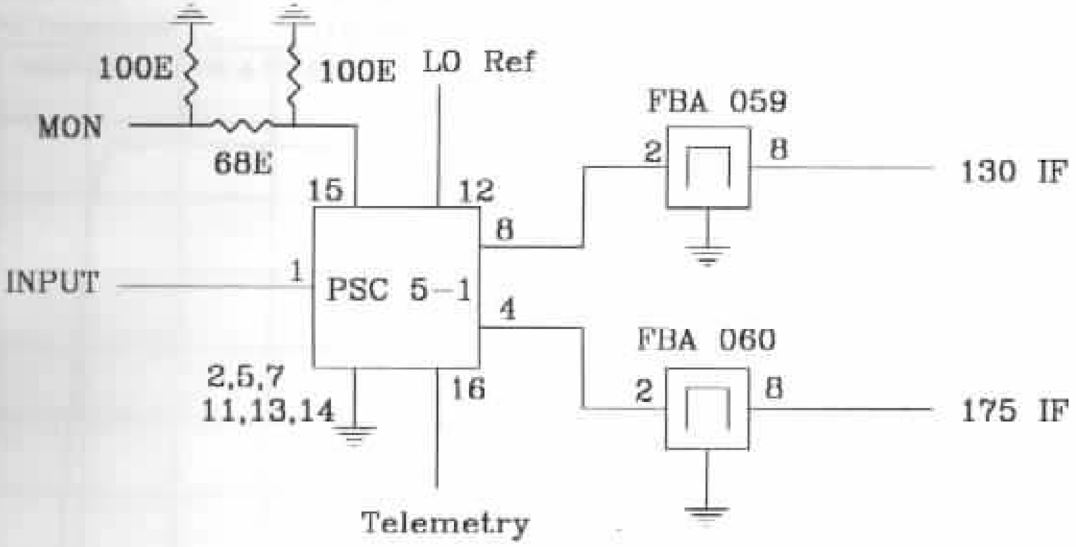
This chapter gives the design details of the IF circuits used in the baseband system. These circuits separate out the IF signals at the output of the optical receiver and then processes them before down conversion to the baseband frequencies. The specifications set for these circuits is mentioned in the previous chapter. The design procedures and finalised circuit diagrams are given here. PCB layouts for these circuits are developed and the circuits wired and tested. The circuits are tested using RF vector network analyzer. Using this equipment the frequency of the input CW signal is varied and the response of the unit may be determined. Any tuning required as in the case of filters is done and the response plotted. The performance of various circuits are given and a comparison is done with the calculated response.

4.1 IF SEPARATOR CIRCUITS

The optical receiver for each antenna at the central electronics building provides the IF signals along with the LO and telemetry return signals. The function of the IF separator circuits is to separate out the IF signals from the LO and telemetry return signals and pass it on for further processing. This circuit should also provide a monitor point for measuring the signal strength as received from the antenna.

This is achieved by first splitting the incoming signal from optical receiver using a low loss torroidal power divider into five outputs. From one output the 130 Mhz IF is filtered out while from the other 175 Mhz IF is filtered out. Two other outputs of the divider are directly fed to the telemetry and LO systems to separate out the return signals. The fifth output port is used as a monitor port for monitoring the power levels of various signals from the antenna. This gives an idea about the health of different systems at the antenna.

Mini Circuits Lab make 5 way torroidal power divider, PSC5-1 is used to split the signal from the optical receiver. This is a low loss divider and provides good return loss characteristics at the input port. This is important for the unit to be impedance matched to the optical receiver. IF signals extend from 114 Mhz to 146 Mhz and from 159 Mhz to 191 Mhz. While separating out these signals using filters care should be taken to minimize the interference from the unwanted channel. This means that the response of the filters should be such that it should pass all signals in the pass band with least attenuation and still provide an attenuation higher than 50 dB at 13 Mhz away. Such a response is possible using Surface Acoustic Wave (SAW) band pass filters. Hence SAW filters from Thompson CSF, FBA-059 (130 Mhz) and FBA-060 (175 Mhz) are used in the IF separator circuit. The monitor port will be brought over to the front panel of the unit for checking the power level or spectrum of the signal and hence should be isolated from any impedance mismatch that can exist on the front panel port. For this purpose a fixed resistive attenuator of 10 dB is wired at this output. This will ensure a return loss of atleast 20 dB for this port. Fig 4.1 shows the circuit diagram of IF separator unit.



PSC 5-1 : PINS 3,6,9,10 ARE NC
SAW Device : Other pins to GND

Fig 4.1 - IF Separator Circuit

The important characteristics of this circuit are the insertion loss to each output and the input reflection coefficient. The input reflection coefficient should be high in order to provide a good matching to the previous circuits. The insertion loss should be as low as possible as this will reduce the signal strength. The insertion loss to the LO/Telemetry port is due to the 5 way divider and the theoretical value may be calculated as follows,

Insertion loss (5 way divider) = 10.log (1/5)
= -6.98 dB
(= -7.5 dB as per PSC 5-1 data sheet)

For the monitor port there will be additional loss due to the resistive attenuator and at the IF ports the SAW filters will introduce extra loss. This circuit is tested for the insertion loss to each output port with a CW signal fed to the input. Fig 4.2 shows the insertion loss plots for the IF ports at 130 Mhz and 175 Mhz and also the LO/Telemetry port and the monitor port. The plots give the loss in dB. The reflection coefficient is also measured at the input port. The results are summarised in Table 4.1. The calculated values of these parameters from the data available in the data sheets is also given for comparison.

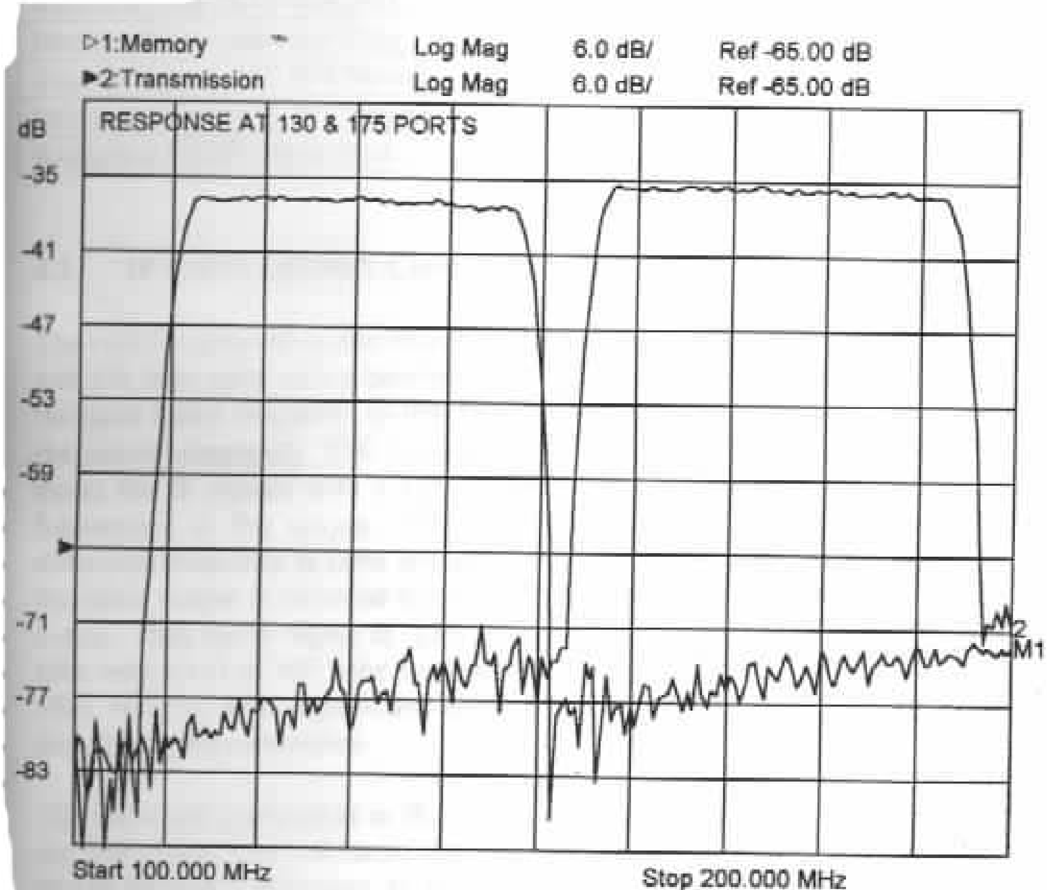
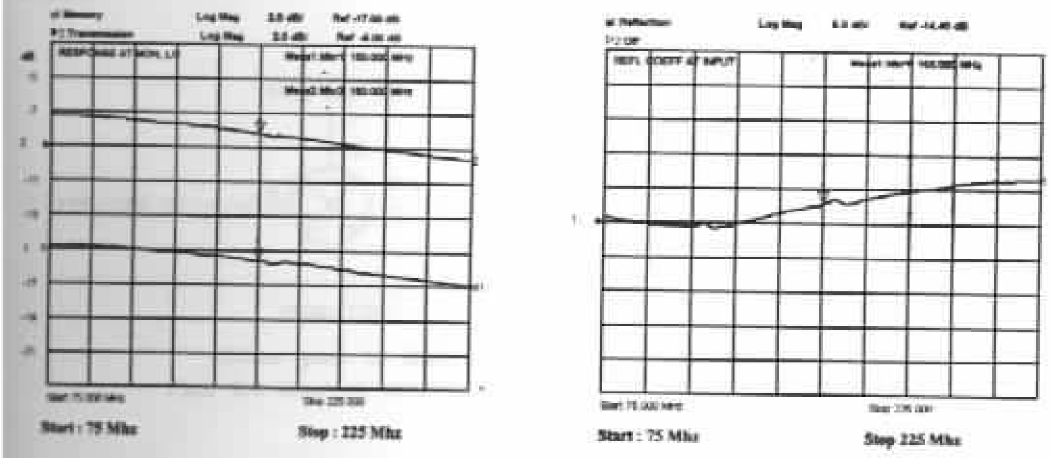


Fig 4.2 (a) - Response at 130,175 IF ports


Start : 75 Mhz Stop : 225 Mhz

Start : 75 Mhz Stop 225 Mhz

ig 4.2 (b) - Response at LO/Monitor Ports (left) and Reflection Coefficient (right)

Table 4.1 Measurement Results

Parameter	Calculated	Actual
Insertion Loss - LO/Tele port	7.5 dB	7.7 dB
Insertion Loss - Monitor Port	17.5 dB	17.8dB
Insertion Loss - 130 IF Channel	35.6 dB	36.8 dB
Insertion Loss - 175 IF Channel	33.8 dB	35.3 dB
Reflection Coeff - Input Port	-	-10 dB

4.2 IF CONVERSION CIRCUITS

The two IF signals from the antenna are received at the optical receiver output at 130 Mhz and 175 Mhz each with a bandwidth of 32 Mhz. These IF signals are to be converted to the same centre frequency so that the rest of the circuits are identical and thereby reducing the circuit complexity. This is achieved by using a standard double balanced mixer which mixes the IF signals with a CW signal thereby producing both the sum and difference frequencies at the output. The CW frequency can be chosen such that the output difference frequency is same irrespective of the input IF frequency. The sum frequency at the mixer output is removed by using a low pass filter of suitable cut off frequency at the output. Thus the IF signal at 130 Mhz is mixed with a LO of 200 Mhz and the IF at 175 Mhz with a LO of 105 Mhz to generate IF frequency of 70 Mhz with a bandwidth of 32 Mhz. 105/200 Mhz signals are available in the LO reference system which can be used directly for this conversion.

The important parameters in this unit are the leakage between the various ports and also the conversion loss. Of these the LO leakage at the output is very important, and this should be kept a minimum, to avoid interfering CW signals at the Baseband output. The conversion loss should be a minimum so as to avoid loss of signal strength. These parameters are considered while selecting the mixer module to be used in the circuit. Mini Circuits Lab make double balanced frequency mixer SCM-1NL is used to down convert the received IF signals from 130/175 Mhz to 70 Mhz. Fig 4.3 shows the circuit diagram in detail.

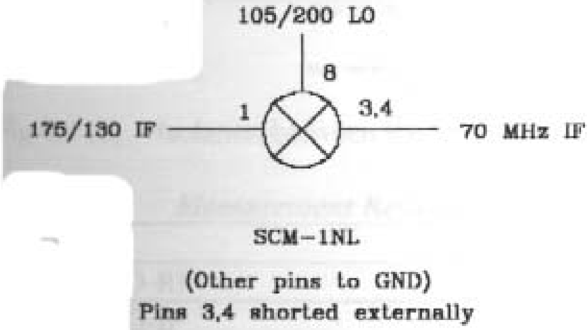


Fig 4.3 - IF Conversion Circuit

These circuits are tested for various parameters like signal leakage in the ports, like the RF to IF leakage, LO to IF leakage, LO to RF leakage. The conversion loss is measured using two CW sources and output IF power measured using a spectrum analyzer. The results are summarised in Table 4.2 and a comparison is done with the calculated values from datasheet. Fig 4.4 gives the isolation response and conversion loss over the frequency range of interest.

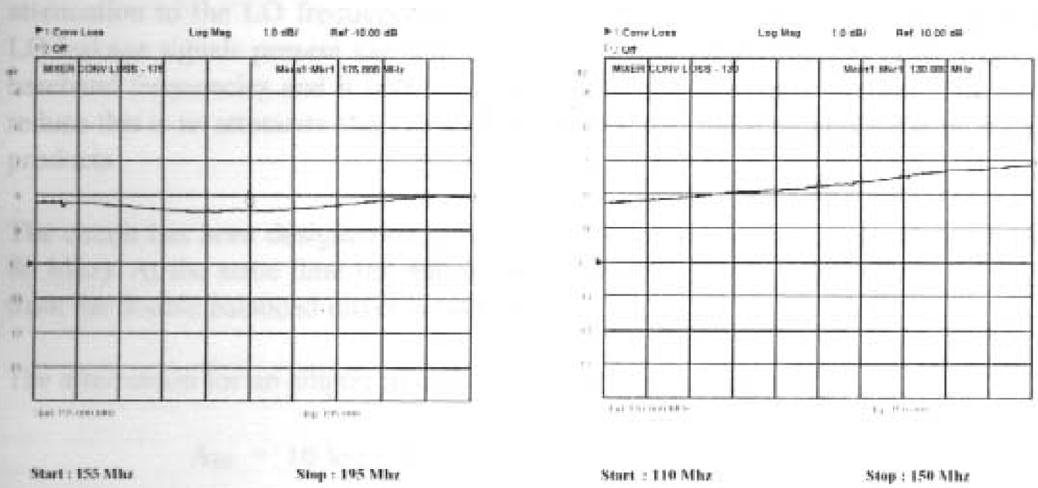


Fig 4.4 (a) - Conversion Loss at 175 IF (left) and 130 IF (right)

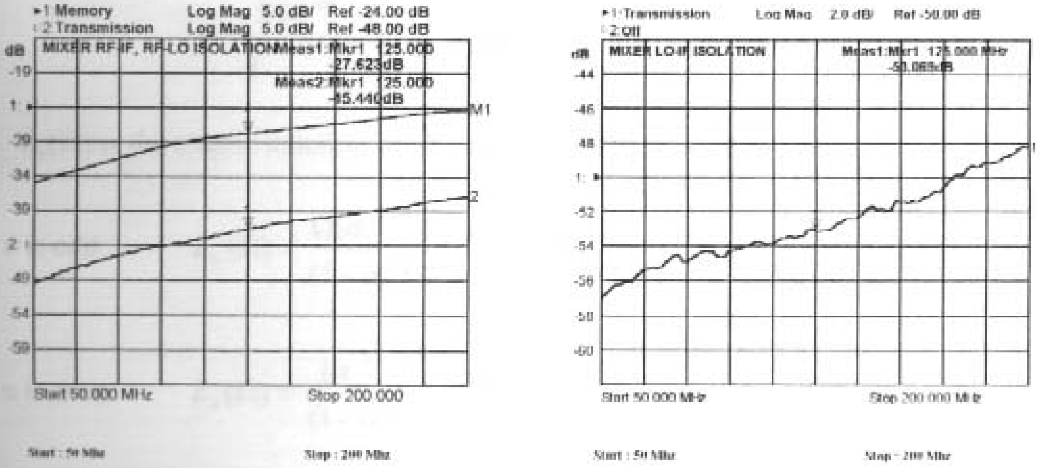


Fig 4.4 (b) - Isolation between RF/IF and RF/LO (left) and LO/IF (right)

Table 4.2 ***Measurement Results***

Parameter	Calculated	Actual
Isolation LO-RF	45 dB	45 dB
Isolation LO-IF	45 dB	53 dB
Isolation RF-IF	-	27 dB
Conversion Loss	5.72 dB	8.5 dB

4.3 LOW PASS FILTER FOR 70 Mhz IF

The main purpose of this circuit is to filter out the IF signal (54 Mhz to 86 Mhz) at the output of the double balanced mixer. This will remove all other signals outside this band which can interfere in the performance of the later systems. Especially this filter will remove the 105/200 Mhz LO signal leakage and the image frequencies at the output of the mixer. Even though the mixer with very low LO leakage has been selected a further attenuation to the LO frequencies is preferable to avoid CW signals at later stages. Any LO leakage signals present can introduce inter-modulation products in later circuits at the baseband frequencies and it will be passed on to the correlator system. The only way to reduce this is to attenuate the LO leakage signals and there by avoid the inter-modulation products.

The circuit has been designed to provide a flat insertion loss over the IF band (54 Mhz to 86 Mhz). At the same time the null is designed to be at 105 Mhz so that any LO leakage from the double balanced mixer is taken care of.

The attenuation for an elliptic filter is given by the expression [7],

$$A_{dB} = 10.\log [\, 1 + \varepsilon^2 Z_n^2(\Omega) \,]$$

where ε is based on the allowable ripple given by the expression,

$$\varepsilon = \sqrt{10^{R/10} - 1}$$ where R is the ripple in dB

$Z_n(\Omega)$ is the elliptic function of nth order and may be written as

n is odd $$Z_n(\Omega) = \frac{\Omega(a_2^2 - \Omega^2)(a_4^2 - \Omega^2)......(a_m^2 - \Omega^2)}{(1 - a_2^2\Omega^2)(1 - a_4^2\Omega^2)......(1 - a_m^2\Omega^2)}$$ where m = (n-1)/2

n is even $$Z_n(\Omega) = \frac{(a_2^2 - \Omega^2)(a_4^2 - \Omega^2)......(a_m^2 - \Omega^2)}{(1 - a_2^2\Omega^2)(1 - a_4^2\Omega^2)......(1 - a_m^2\Omega^2)}$$ where m = n/2

Here a_2 ... a_m can be derived from the elliptic integral and these are the zeroes of $Z_n(\Omega)$ and the poles are $1/a_2$ ... $1/a_m$. Elliptic function filters provide very high attenuation in the stop band and hence suitable in this application.

A 7th order elliptic low pass filter is used for this purpose. The circuit has a 0.25dB ripple designed upto 86 Mhz and 50 dB point at 105 Mhz. The circuit is wired using Philips make trimmer capacitors which helps to fine tune the frequency response of the filter, if required before installation. Detailed circuit diagram is shown in Fig 4.5.

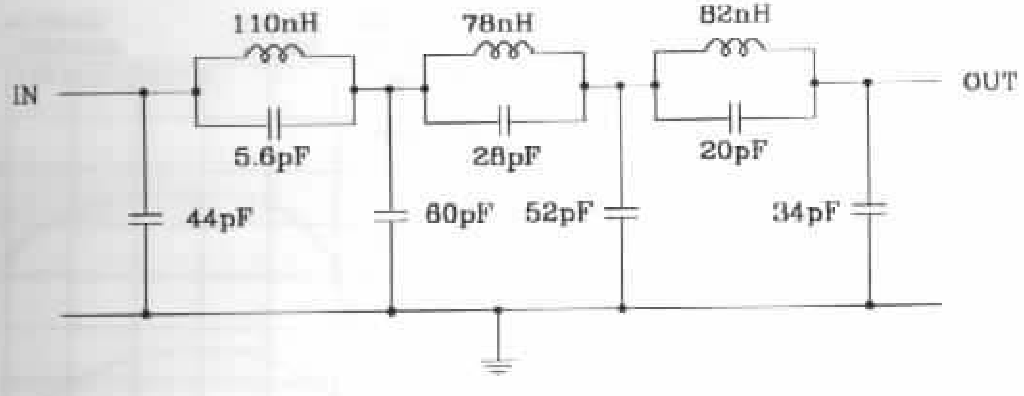
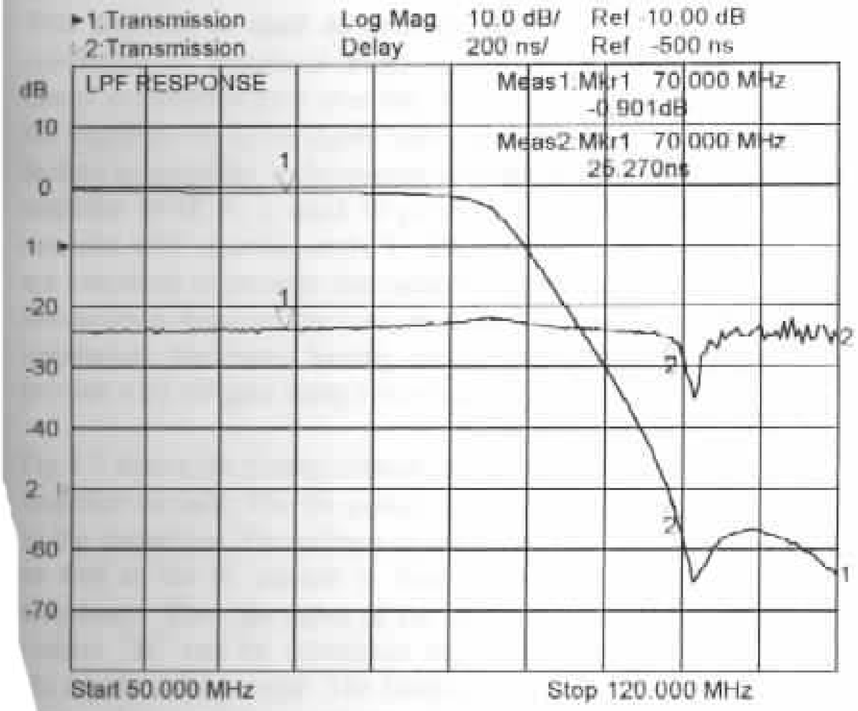


Fig 4.5 - IF Low Pass Filter

The important parameters of this filter are the frequency response and the reflection coefficient. The gain response of this filter is measured and shown in Fig 4.6 (a). It gives both the magnitude and group delay. It is seen that the response falls by 50 dB at 105 Mhz, and the response varies by less the 0.5 dB in the IF band. The input and output reflection coefficients are given as Fig 4.6 (b). It is seen that both input and output reflection coefficients are better than 15 dB in the pass band.



4.6 (a) - IF Low Pass Filter Response

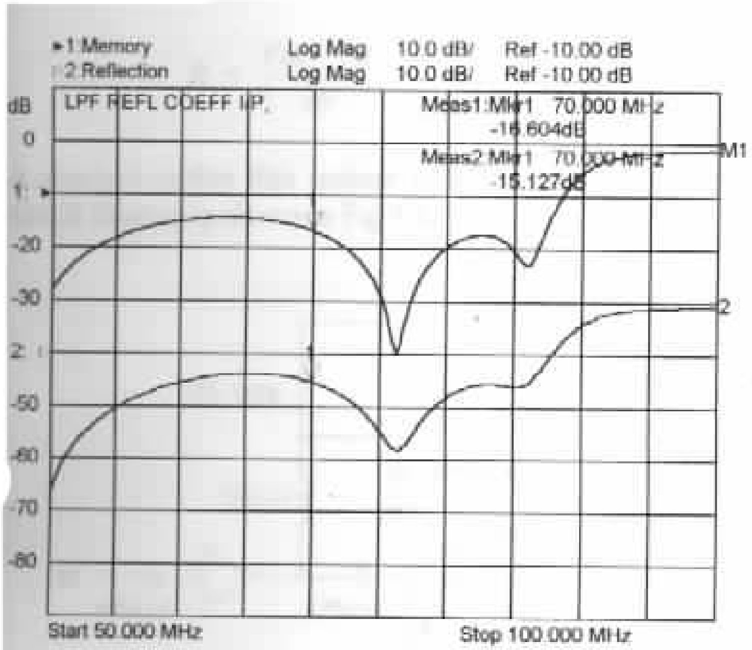


Fig 4.6 (b) - IF Low Pass Filter Reflection Coefficient

4.4 IF AMPLIFIER CIRCUITS

This amplifier is used to provide sufficient gain to the signal so that further down conversion to Baseband frequencies can be done. The important requirements of this circuit are uniform gain over the IF frequency range and a gain of the order of 22 dB. Also the amplifier has to be highly stable and should have a low noise figure. Considering these factors an amplifier is developed around MMIC units from Minicircuits lab. A monolithic amplifier MAR-3, is used to provide a stable amplifier. These are wide band amplifier modules with approximately 13 dB gain under proper biasing conditions. Two such units are cascaded to provide the necessary gain along with a facility to introduce a resistive attenuator in between the two amplifier stages so as to fine tune the overall gain during the installation. Necessary biasing circuits are designed for proper operation of the circuit to provide a 22 dB gain using two stages.

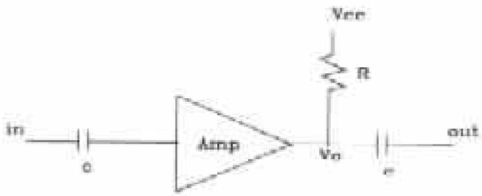
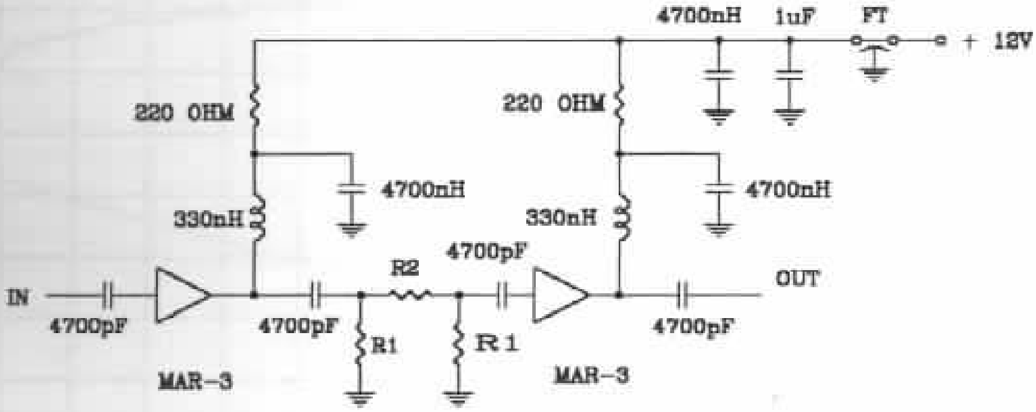


Fig 4.7 - Biasing of MAR Amplifier

Fig 4.7 shows the biasing circuits for the MAR amplifier module. The dc power supply is fed to the output pin. The voltage at the output pin as well as the dc current is fixed as per the datasheets. Then the value of the dc dropping resistor "R" can be calculated depending on the supply voltage used. The biasing resistor is designed to provide a current of 35 mA at a voltage of 5 V at the output port as per the data sheet. 12 Vdc supply is used to power the amplifier unit.

$$R = \frac{Vcc - Vo}{Io} = \frac{12V - 5V}{35mA} = 200\ \Omega$$

A standard carbon film resistor of 220 Ω is used as the dropping resistor. The detailed circuit diagram is shown in Fig 4.8.



Adjust value of R1, R2 to match the gain between units

Fig 4.8 - IF Amplifier Circuit

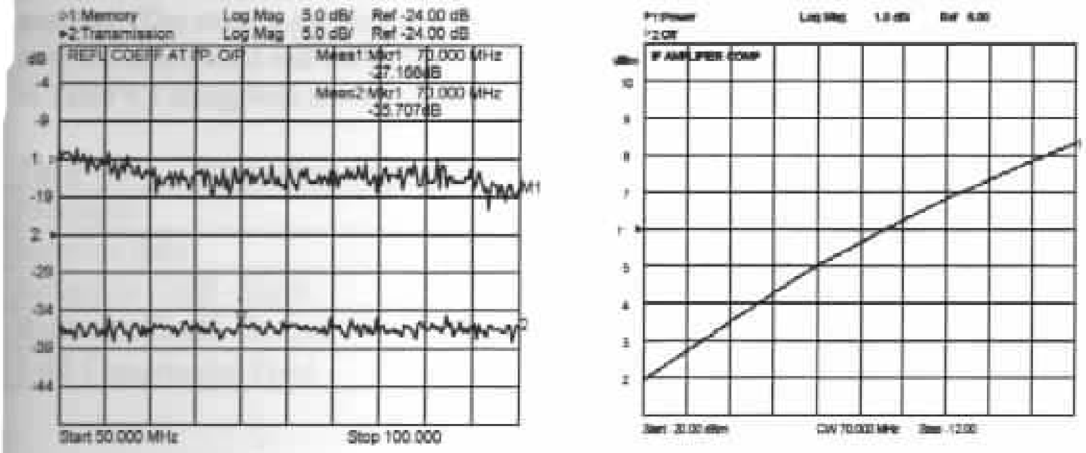


Fig 4.9 (a) - IF Amplifier Reflection Coeff (left) and Compression Point (right)

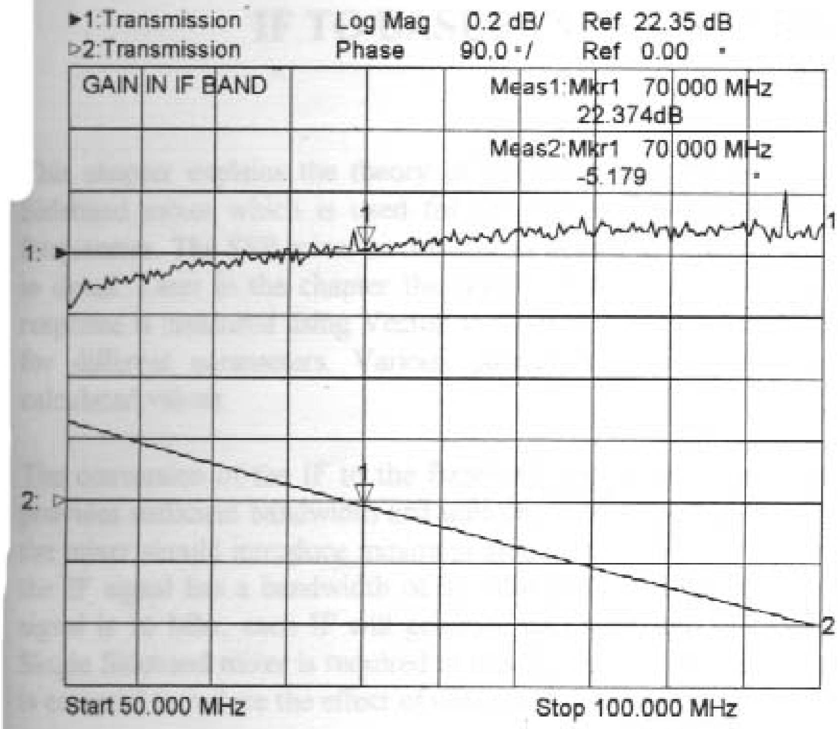


Fig 4.9 (b) - IF Amplifier Gain Response

The main parameters in this case are the gain response and the reflection parameters. The gain and reflection coefficient of the amplifier circuit are measured using a network analyzer. The results are given in Fig 4.9. It may be seen that the gain is fairly constant over the entire band and the reflection coefficient is also acceptable. The summary is given in Table 4.3 along with the calculated values from the datasheet.

Table 4.3 Measurement Results

Parameter	Calculated	Actual
Insertion Gain	25 dB	22.3 dB
Reflection Coeff - Input	-	26.5 dB
Reflection Coeff - Output	-	35 dB
1 dB Compression Point	-15 dBm	-15 dBm

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Chapter 5

IF TO BASEBAND CONVERSION

This chapter explains the theory of operation and the design procedure for the Single Sideband mixer which is used for the conversion of the IF signals to the Baseband frequencies. The SSB mixer circuit and its design parameters and constraints are explained in detail. Later in the chapter the results of the circuits wired and tested is given. The response is measured using Vector Voltmeter, Network Analyzer and Spectrum Analyzer for different parameters. Various parameters are measured and compared with the calculated values.

The conversion of the IF to the Baseband should be done in such a way that the circuit provides sufficient bandwidth and uniformity of response over this bandwidth. In addition the mixer should introduce minimum amount of noise and distortion to the signals. Since the IF signal has a bandwidth of 32 Mhz and the maximum frequency of the Baseband signal is 16 Mhz, each IF will generate two sidebands at baseband frequencies. Hence a Single Sideband mixer is required in this conversion. A high image rejection for this mixer is essential to reduce the effect of unwanted sideband, which can reduce the signal to noise ratio.

5.1 THEORY OF SSB CIRCUITS

The basic principle of the circuit used is shown in the Fig 5.1. The theory of operation of this circuit is given in detail in [8].

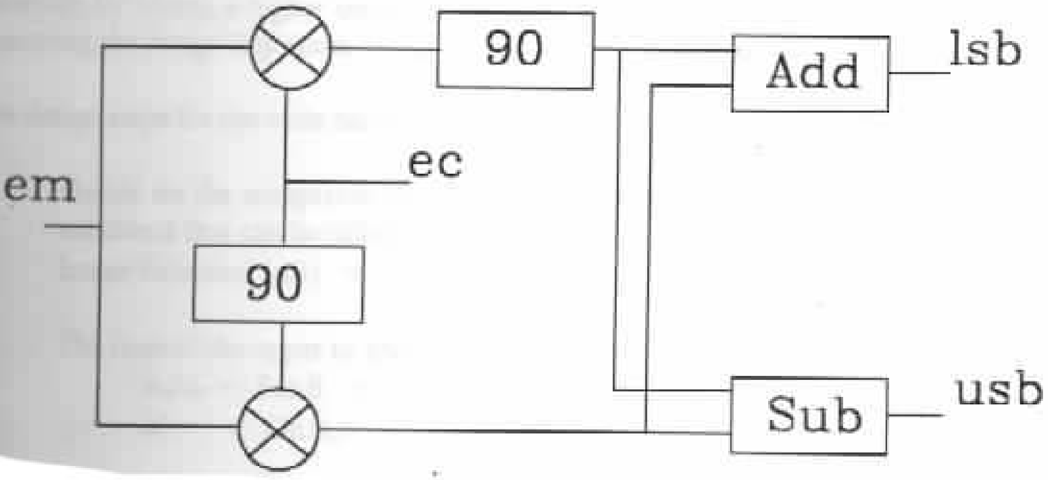

Fig 5.1 - Principle of operation of SSB Mixer

The IF signal is split in phase and fed to two double balanced frequency mixers, the LO signals for the mixers are 90° out of phase. This generates two signals at the Baseband frequencies at the output of the mixers. One of these is further delayed by 90° compared to the other using wide band phase shifter at Baseband frequencies. The resulting signals are passed through adder and substractor circuits to separate out the lower sideband and the upper sideband signals.

IF signal $e_m = E_m.\mathrm{Sin}(\omega_c+\omega_u).t + E_m.\mathrm{Sin}(\omega_c-\omega_l).t$
LO signal $e_c = E_c.\mathrm{Sin}(\omega_c.t)$
Adder output = LSB $= (1/2).E_m.E_c.\mathrm{Cos}(\omega_l.t + 90)$
Substractor output = USB $= (1/2).E_m.E_c.\mathrm{Cos}(\omega_u.t + 90)$

The Phase shifter at Baseband frequencies has to work over several octaves (from 10 Khz to 16 Mhz). The amplitude and phase unbalance in this phase shifter is mainly responsible for leakage of signals from one sideband to the other. This leakage is defined as the Image Rejection of the mixer circuit ie, the amount of unwanted sideband power relative to the required sideband. This can be calculated as

Image Rejection (dB) $= 20.\log\{\cot(\delta/2)\}$
where δ is the maximum phase unbalance between two channels

5.2 WIDE BAND PHASE SHIFTER NETWORK

Here the wide band phase shifter is designed using all pass filter techniques [2,9,10]. In this case a series of all pass networks arranged on both channels provide an overall phase difference of 90° at the output. The amount of phase unbalance (phase variation over and above the 90° difference) is depended on the number of stages of all pass network used. Generally by having a higher order network the phase unbalance can be reduced, thereby improving the Image Rejection.

The design steps for the wide band phase shifter using all pass networks is given below.

1. Decide on the acceptable level of image rejection. Calculate the amount of phase unbalance that can be tolerated from the expression,
   Image Rejection (dB) $= 20.\log\{\cot(\delta/2)\}$

2. The ratio of the upper to lower frequencies of the desired frequency band,
   $\omega_u/\omega_l = \mathrm{Sec}\,\theta = 1/k'$
   $k' = \mathrm{Sin}\,\theta$, $k = \mathrm{Cos}\,\theta$

3. Find K and K′ as the complete elliptic integral of the first kind of modulus k and k′. Complete elliptic Integral is

$$K(k) = \int_0^{\pi/2} \frac{1}{\sqrt{1-k^2 \sin^2(\theta)}} d\theta$$

4. Define the quantity "q" as $q = \exp\{-\pi (K'/K)\}$
Then the phase unbalance $\delta = 4 . q^n$
where n is the order of the network
and δ is the phase unbalance.

The number of stages "n" is determined based on the image rejection required.

5. The location of the poles of the network can now be calculated as

$$\text{Pole Location,} \quad P_i = \sqrt{\frac{\omega_u}{\omega_l}} . \frac{cn(U_i, K)}{sn(U_i, K)}$$

sn and cn are elliptic functions,

$$U_i = \frac{4j+1}{2n} . K; \quad j = 0,1,2,3 \ldots\ldots (n-1)$$

The pole locations are used to design the All Pass network as the denormalised pole location is given by the reciprocal of the RC product. The All Pass structure of first order is as shown in Fig 5.2, a higher order network is obtained by cascading these networks [7].

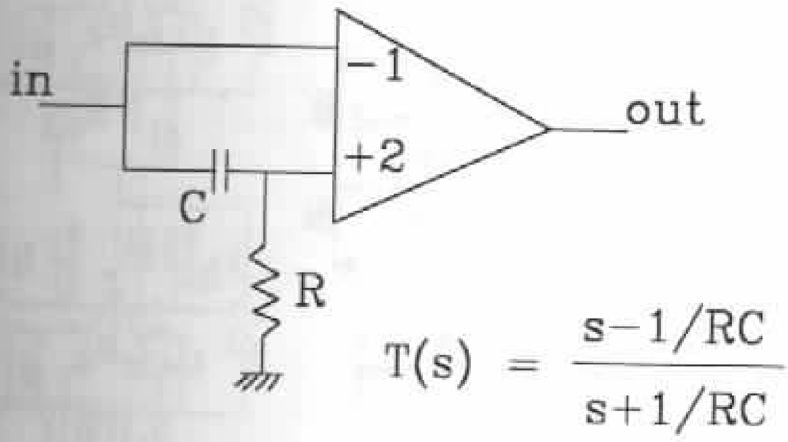

Fig 5.2 - First order All Pass Network

5.3 CIRCUIT IMPLEMENTATION

The overall circuit schematic is shown in Fig 5.3. Mini Circuits make torroidal power dividers PSC 2-1, and PSCQ 2-90 are used in splitting the IF and LO signal for down conversion to the Baseband frequencies. The double balanced mixer used is Mini Circuits

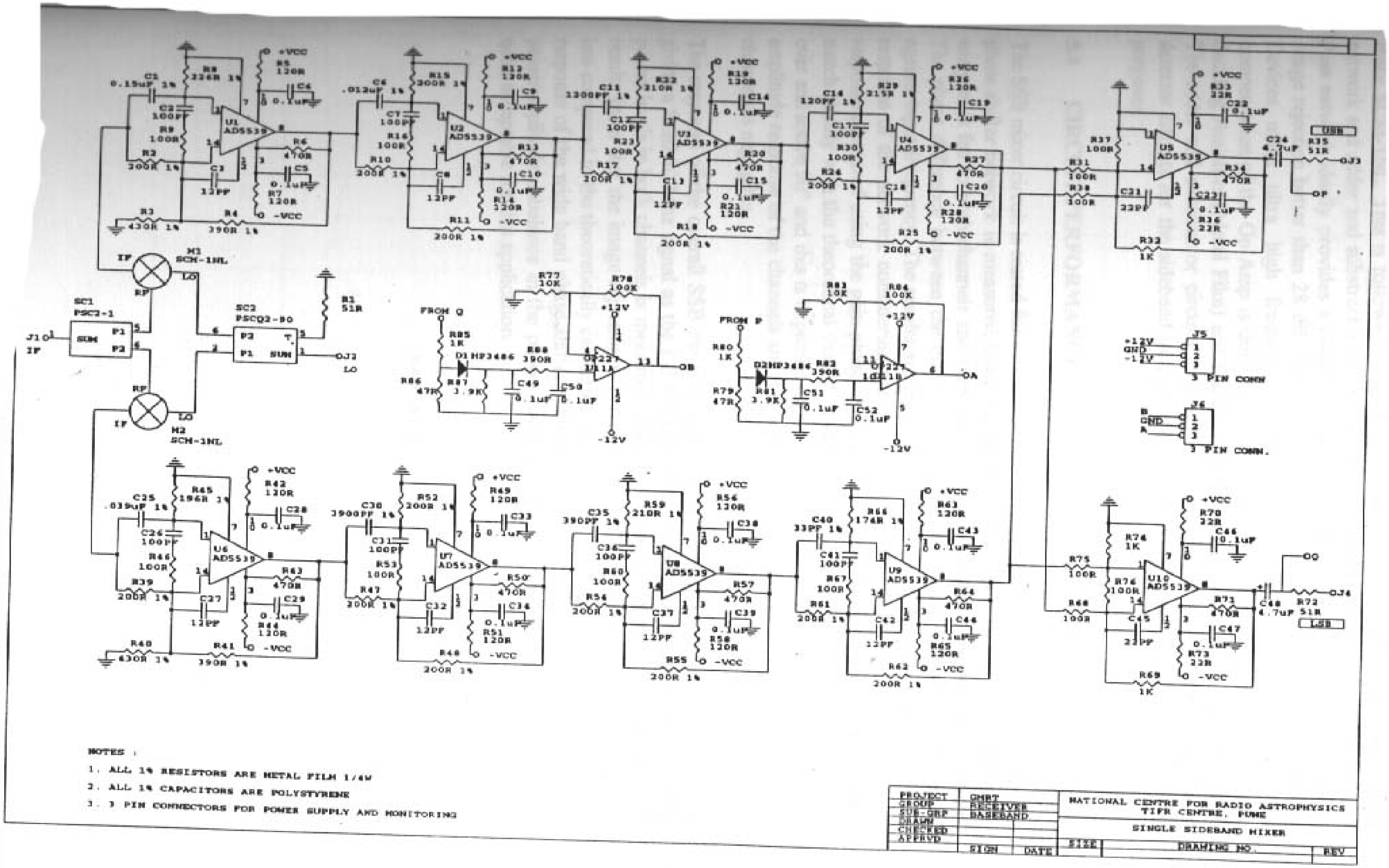

Fig 5.3 - SSB Circuit Schematic

make SCM-1NL. This is followed by an 8 stage wide band 90 degrees phase shifting network and adder and substractor circuits to separate out the sidebands. An 8 stage All Pass network ideally provides a phase unbalance less than ± 3°, which in turn gives an image rejection better than 28 dB. The Active all pass network is based on the Analog Devices make ultra high frequency operational amplifier, AD 5539. Frequency compensation of the Op Amp is done as per the guidelines in the data sheets. The pole deciding Resistors (Metal Film) and Capacitors (Polystyrene) are of 1% tolerance type. The adder and substractor circuits are also based on the same Op Amp. Power level detector circuits for the sideband outputs are also included in the circuit for monitoring purposes.

5.4 CIRCUIT PERFORMANCE

The SSB mixer circuit is tested for various circuit parameters. The performance of the 90° phase shifter network is measured using a gain-phase analyzer. The CW signal from the analyzer is fed to both channels and then the frequency is varied over the desired range. The phase difference between the output of the two channels is measured and plotted against input frequency. The results are given in Fig 5.4. Fig 5.4(a) gives the theoretical response of the network considering the 8 stage network, and Fig 5.4(b) is the actual response measured using the gain phase analyzer. It may be seen that the practical results match closely with the theoretical values. A variation of ± 3° is observed in the response over and above 90° and this is expected for an 8 stage network. The figure also shows the amplitude response of the channels and it is seen that the amplitude unbalance between the channels is negligible.

The response of the overall SSB mixer circuit is shown in Fig 5.5. This is measured by giving a fixed 70 Mhz signal as the LO frequency and varying the IF input. The resulting power levels in both channels is measured using an RF vector voltmeter and plotted. The results show that the image rejection is better than 24 dB in the frequency band. This is less compared to the theoretically calculated value of 28 dB, considering the 90° ± 3 phase response of the wide band phase shifter. This reduction in the image rejection is due to phase/amplitude unbalance in the power dividers used. An image rejection of 24 dB is quite acceptable for this application.

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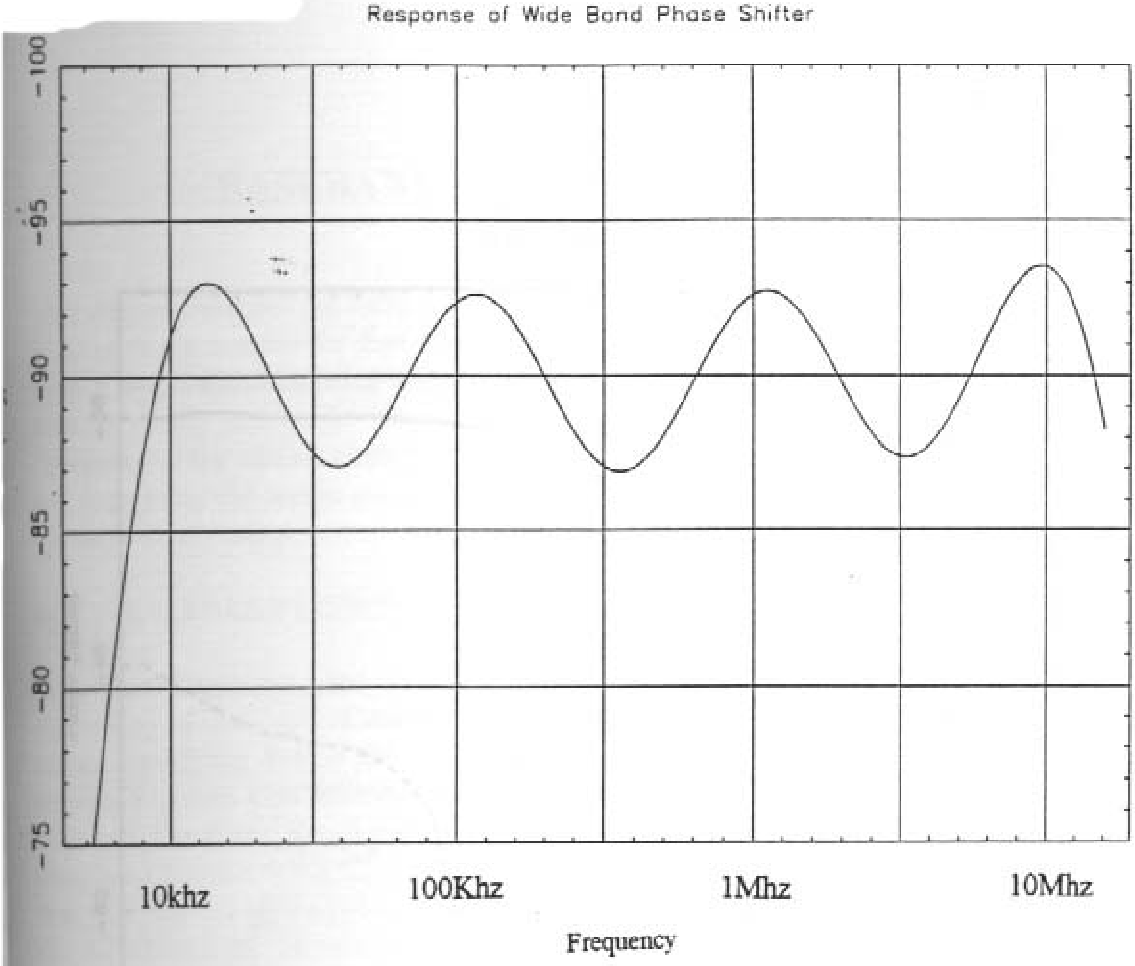
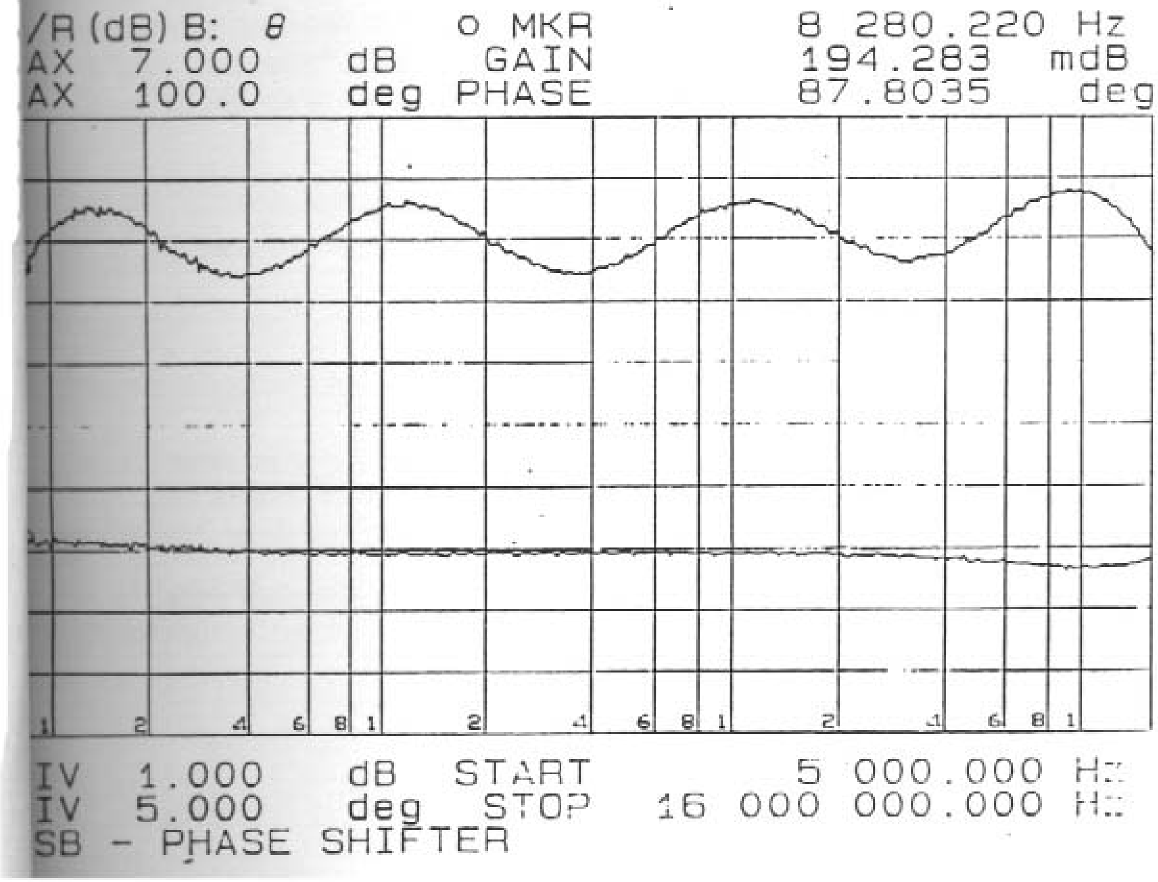


Fig 5.4 (a) - Theoritical Phase Response of All Pass Network



/R (dB) B: θ ○ MKR 8 280.220 Hz
AX 7.000 dB GAIN 194.283 mdB
AX 100.0 deg PHASE 87.8035 deg

IV 1.000 dB START 5 000.000 Hz
IV 5.000 deg STOP 16 000 000.000 Hz
SB - PHASE SHIFTER

Fig 5.4 (b) - Measured Phase Response of All Pass Network

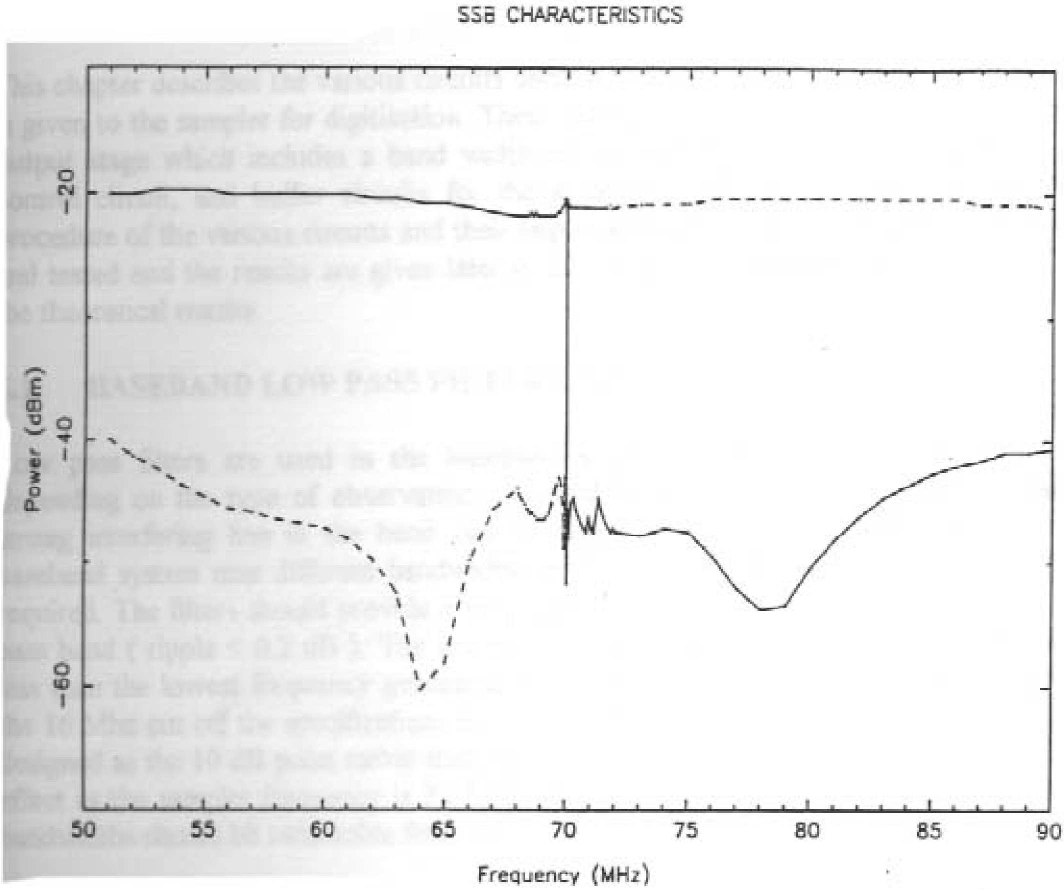


Fig 5.5 - Response of SSB Mixer Circuit

Chapter 6

BASEBAND PROCESSING CIRCUITS

This chapter describes the various circuits used in processing the Baseband signal before it is given to the sampler for digitisation. These include the baseband low pass filter, and an output stage which includes a band width compensation amplifier, an automatic level control circuit, and buffer circuits for the sampler. This chapter describes the design procedure of the various circuits and their implementation. These circuits have been wired and tested and the results are given later in the chapter. A comparison is also done with the theoretical results.

6.1 BASEBAND LOW PASS FILTER CIRCUITS

Low pass filters are used in the baseband system to remove unwanted frequencies depending on the type of observation. A variable cut off filter is required so that any strong interfering line in the band can be removed. As per the specifications of the baseband system nine different bandwidths in octave steps from 62.5 Khz to 16 Mhz is required. The filters should provide a very high attenuation (> 45 dB/octave) and a flat pass band (ripple < 0.2 dB). The lowest frequency that these filters can pass should be less than the lowest frequency generated by the SSB mixer at the sideband outputs. For the 16 Mhz cut off the specifications are more stringent than above, also 16 Mhz is to be designed as the 10 dB point rather than the 3 dB point. This is done to reduce the aliasing effect as the sampler frequency is 32 Mhz. Another requirement for this filter is that the bandwidths should be switchable from the central computer.

As a flat response is required in the passband a Butterworth filter is more suitable for this application. A first order Butterworth filter will give an 6 dB/octave attenuation in the stop band. Thus an 8th order filter is required to achieve the given specifications. The response for a Butterworth low pass filter [7] is

$$A_{dB} = 10.\log\left[1+\left(\frac{\omega_x}{\omega_c}\right)^{2n}\right]$$

where ω_x and ω_c are the test frequency and the cut off frequency respectively and n is the order of the filter.

Cut off frequencies upto 8 Mhz are achieved by using active low pass filters based on RC circuits. The R and C values vary depending on the cut off frequency. For a second order filter R and C values may be calculated from the expression

$$\text{Cut off Frequency} \quad f_c = \frac{1}{2.\pi.R\sqrt{C1.C2}}$$

where R, C1 and C2 are values of resistors and capacitors
is shown in Fig 6.1.

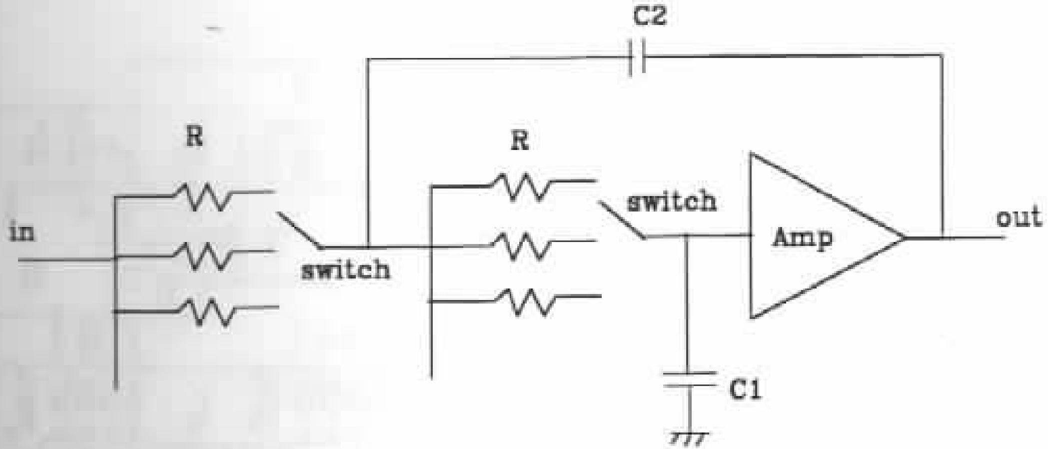


Fig 6.1 - Principle of BW Switching in Filter

An 8th order filter can be designed as a cascade of 4 second order filters and this approach simplifies the design procedure. The values for R, C1 and C2 is calculated for each of the cut off frequencies upto 8 Mhz. In order to have the facility for selecting the bandwidths from central computer, multipole switches are used to select the necessary resistor values while the capacitor values are kept constant for any selected bandwidth. See Fig 6.1. The switches used are FET based switch arrays with TTL level control inputs so that it can be directly controlled from a computer.

AD 5539 is used as the active device since this op amp has a higher gain bandwidth product and higher gain at frequencies upto 20 Mhz. Siliconix make FET switch arrays SD 5002 is used in selecting the resistors while changing the bandwidth. Highly stable metal film resistors and polystyrene capacitors of 1% tolerance are used as frequency selecting devices. While using these components one should take into account the "ON" resistance and capacitance of SD 5002. This value is approx. 30 Ω and 3 pF respectively at the proper biasing voltage.

Detailed circuit diagram of the 8th order filter is given in Fig 6.2.

The 8 stage Butterworth filter is tested using the gain phase analyzer and the results are given in Fig 6.3. The bandwidth is varied by changing the control bits and the corresponding response is plotted for different bandwidth settings. It is seen that the amplitude variation in passband is less than 0.2 dB and the attenuation in the stop band is more than 45 dB/octave. A lower cut-off is seen in the response at 3 Khz due to the series coupling capacitors used in the circuit. However this is much less than the lower end of the baseband frequencies, 10 Khz.

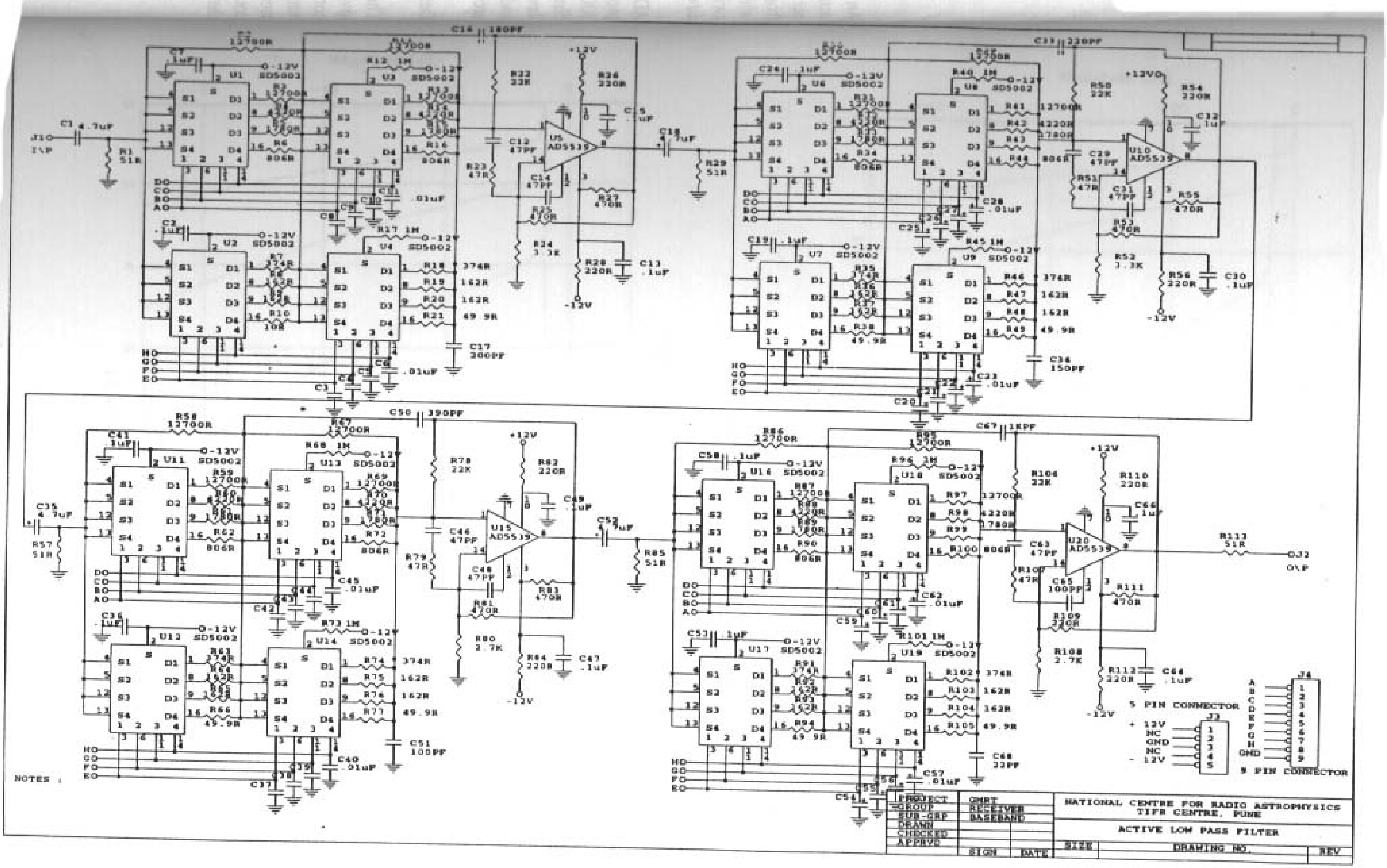


Fig 6.2 - Baseband Low Pass Filter Circuit

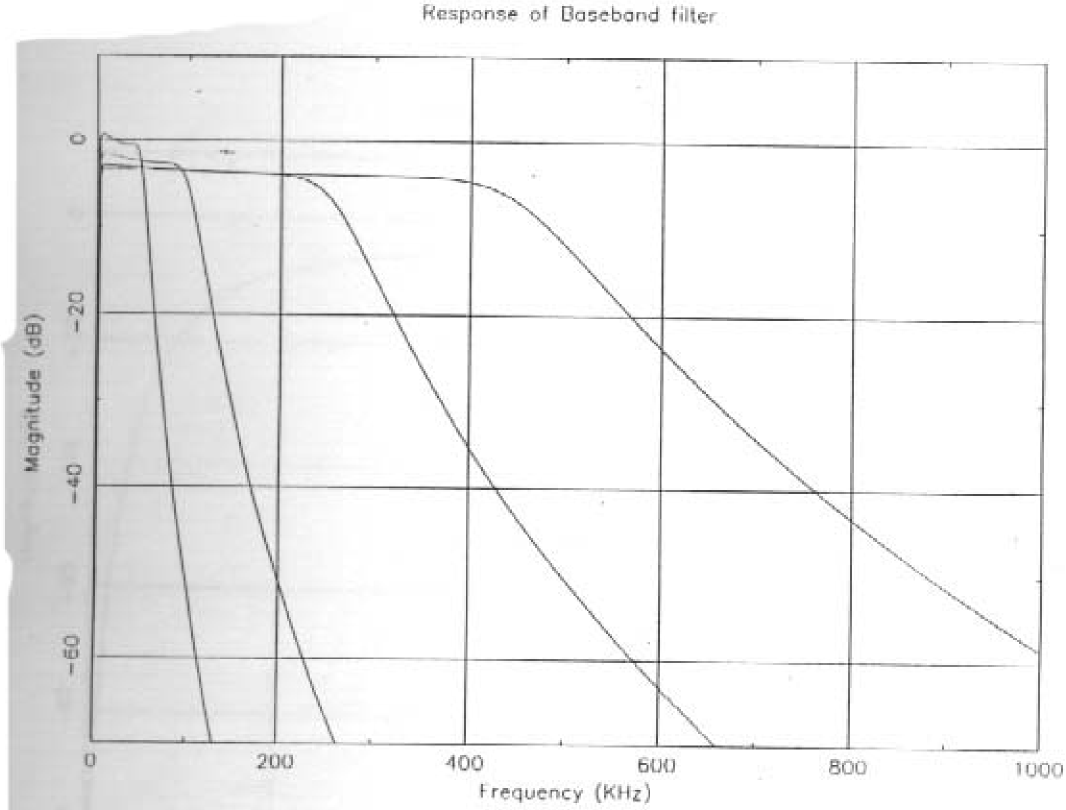
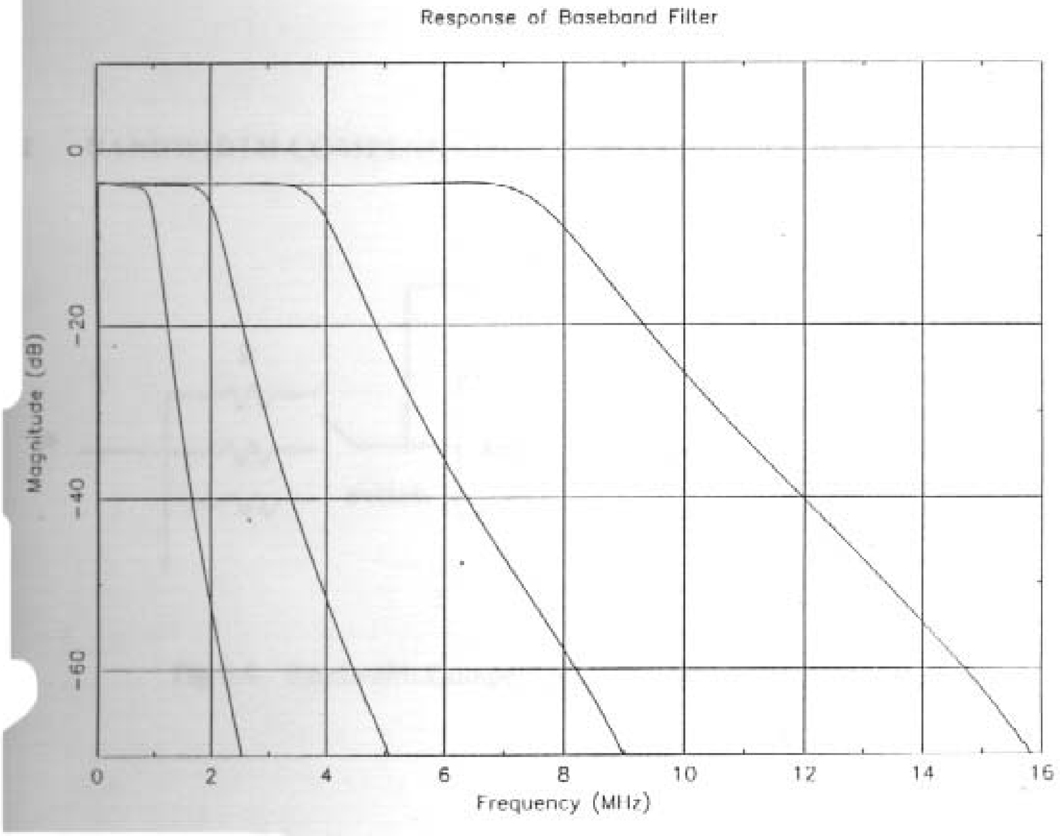


Fig 6.3 (a) - Baseband Filter Response



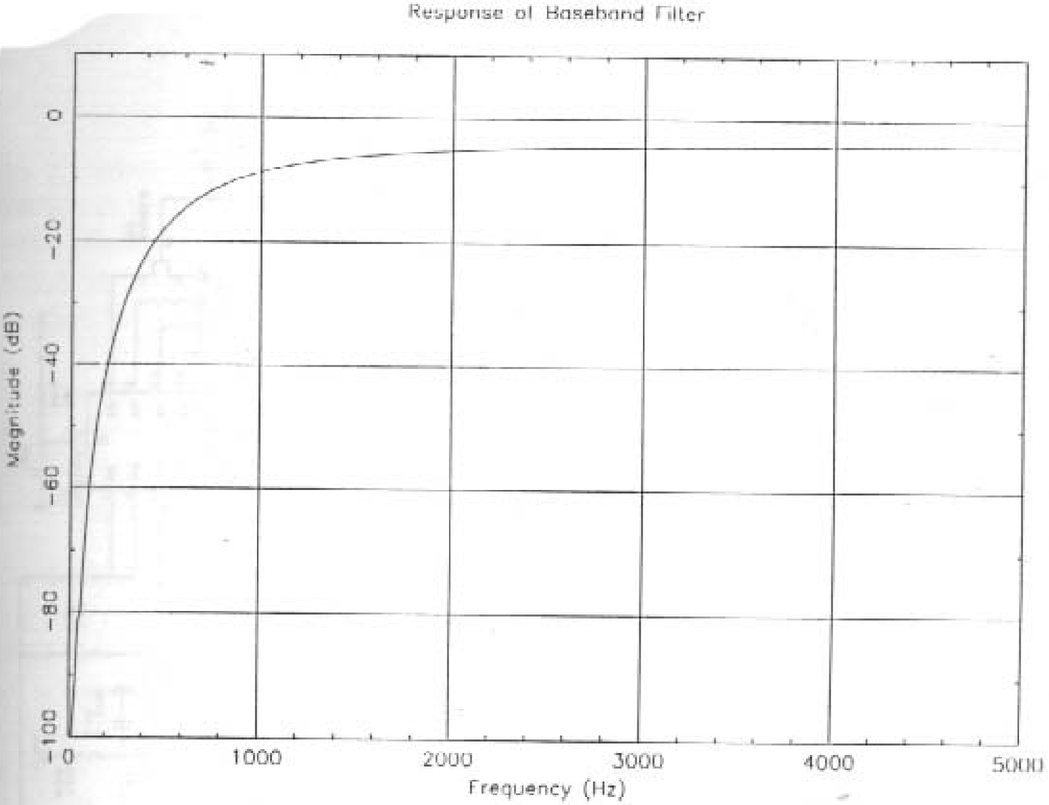


Fig 6.3 (b) - Filter Response at Low Frequencies

6.2 BANDWIDTH COMPENSATION CIRCUIT

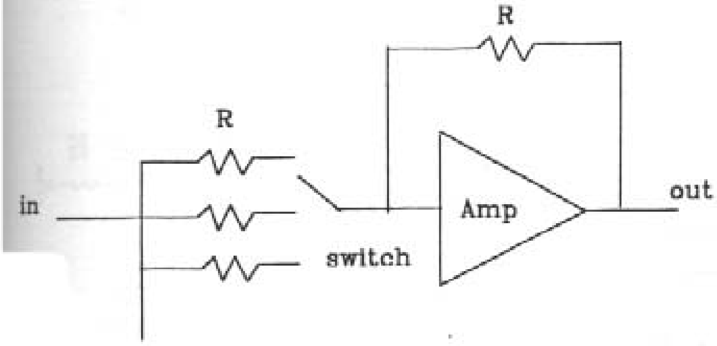


Fig 6.4 - Bandwidth Compensation Amplifier - Principle of operation

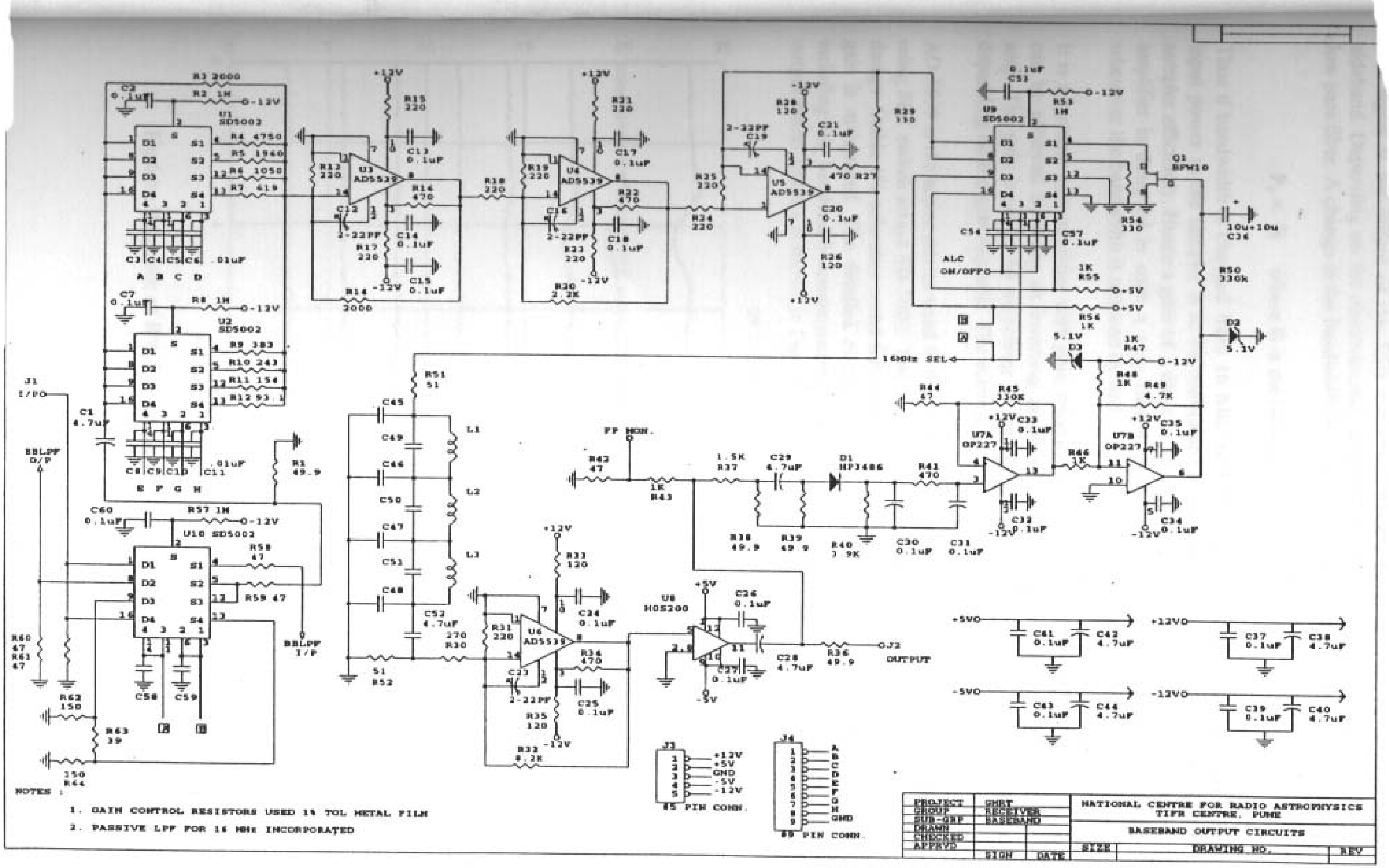


Fig 6.5 - Baseband Output Stage Circuit

The signal at the output of the SSB mixer will have a bandwidth of 16 Mhz in each sideband. Depending on the observational requirements the bandwidth is varied using the low pass filter. A change in the bandwidth will vary the total power output.

$P_o \propto B$ where B is the bandwidth

Thus if bandwidth is changed from 16 Mhz to 8 Mhz, power will reduce by 3 dB. The input power to the sampler is to be maintained a constant in order to use all the bits in sampler efficiently. Hence a gain of 3 dB is to be introduced. The bandwidth compensation amplifier is designed in such a way that it increases the gain of the signal by 3 dB whenever the bandwidth is reduced by half.

It is a variable gain amplifier with gain selection from 0 dB to 24 dB in 3 dB steps. This can be achieved by using an inverting amplifier configuration using high frequency op amp. The gain is varied by switching in the correct resistor value for the input resistor depending on the gain required. The scheme used is shown in Fig 6.4.

AD 5539 is the active device used in the inverting mode, and the resistors are switched in using FET switch arrays SD 5002. The "ON" resistance (approx. 30 Ω) is taken care of in design. Stable 1% tolerance metal film resistors are used as the feedback resistors so that gain is maintained. The detailed circuit diagram of the baseband output stage circuits including the bandwidth compensation amplifier, ALC, 16 Mhz low pass filter and the output buffer circuits is shown in Fig 6.5.

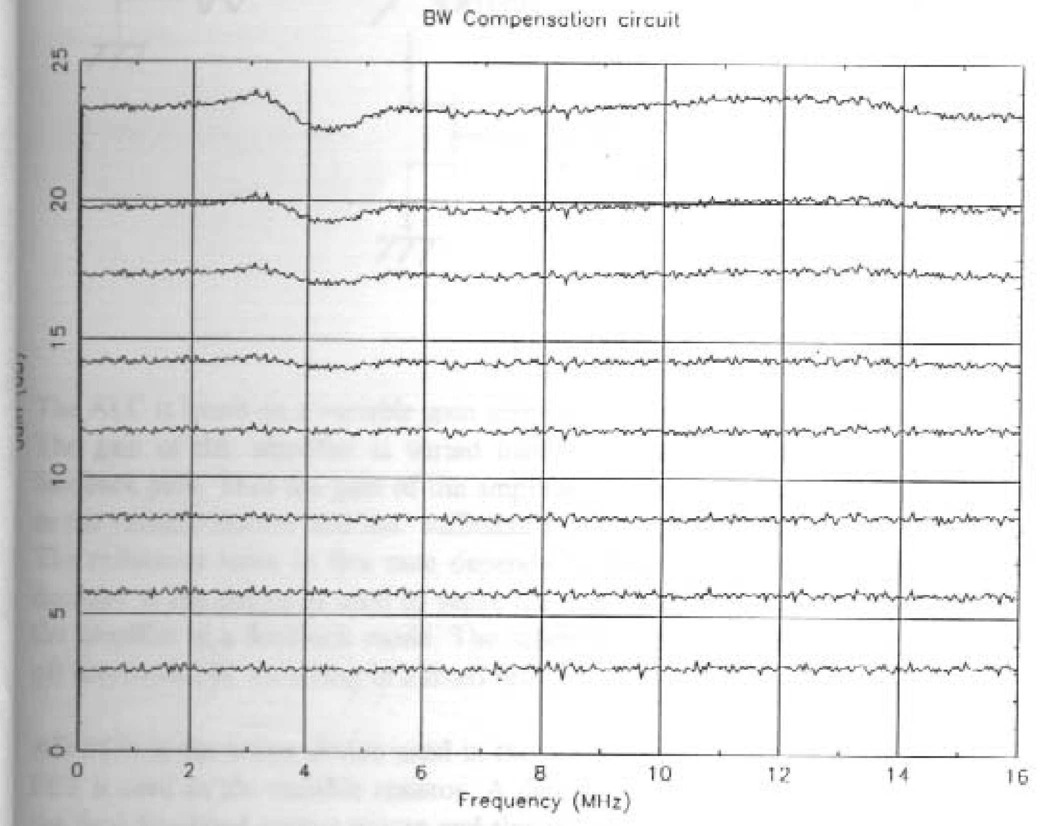


Fig 6.6 - Response of BW Compensation Amplifier

The performance of the bandwidth compensation amplifier measured using a gain-phase analyzer and the response is plotted for different gain settings. The results are shown in Fig 6.6.

6.3 AUTOMATIC LEVEL CONTROL CIRCUIT

An automatic level control circuitry is used at the output of the baseband system so that the power fed to the sampler is maintained a constant irrespective of variations in the gain of systems preceding it. Thus an ALC circuit with about 8 dB range is required. Some observations require the ALC to be switched off and hence this facility is also included in the final circuit developed.

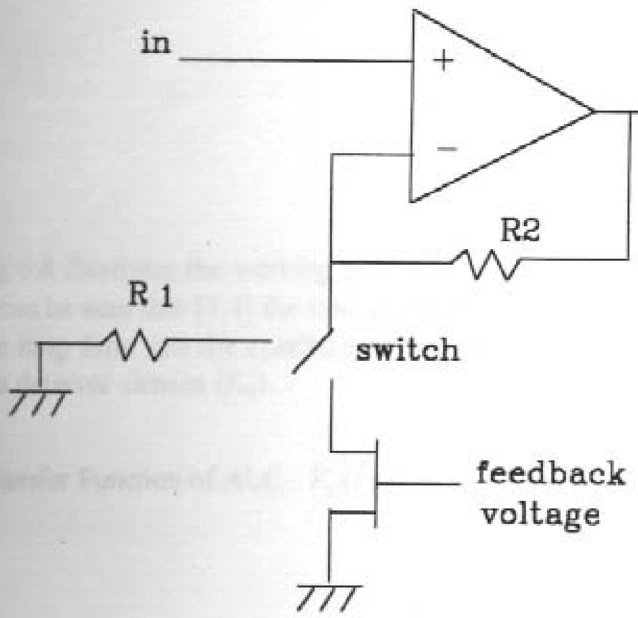


Fig 6.7 - Principle of ALC Circuit

The ALC is based on a variable gain amplifier using an op amp in the non-inverting mode. The gain of this amplifier is varied using a voltage controlled variable resistor in the feedback path. Thus the gain of the amplifier can be varied continuously. An FET is used as the variable resistor and has sufficient range to provide linear gain variation of 10 dB. The resistance value in this case depends on the dc voltage at the gate input. A signal detector at the output is used to sense the power level and is used to control the gain of the amplifier in a feedback mode. The scheme used is shown in Fig 6.7. Facility for ALC off is provided by switching in a fixed resistor in place of the FET.

AD 5539 is the active device used in the non-inverting configuration. N channel Junction FET is used as the variable resistor. A detector using HP 3486 diode is used to monitor the final Baseband output power and this is used to control the FET gate voltage so that the output power is kept a constant. The time constant in the ALC circuit is designed to be 1 sec by selecting proper RC values. The actual circuit of the ALC is shown in Fig 6.5.

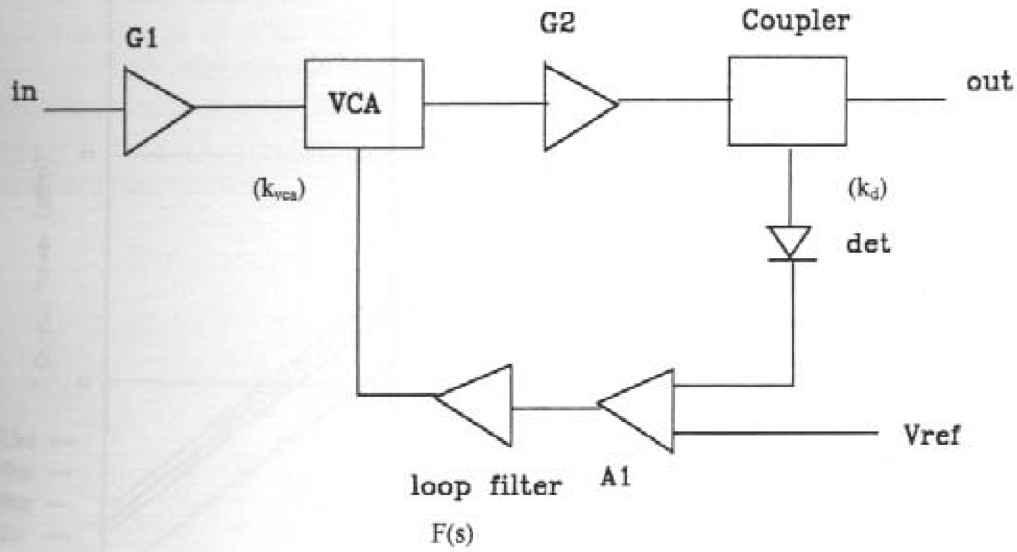


Fig 6.8 - ALC Circuit Analysis

Fig 6.8 illustrates the working of a ALC circuit, the overall transfer function is calculated. It can be seen that [3,4] the time constant of the ALC depends on the time constant (τ) of the loop filter and the coefficients of conversion of the variable gain amplifier (K_{vca}) and the detector circuits (K_d).

Transfer Function of ALC, $V_0(s) = \dfrac{1}{1 + k_{vca} A_1 k_d F(s)} = \dfrac{s}{s + \dfrac{k_{vca} k_d}{RC}}$

where F(s) is the transfer function of the loop filter
and gain A_1 is assumed to be unity

The response to a step input may be written as

$$V_0(t) = e^{\frac{-t}{T}}$$

where $T = \dfrac{RC}{k_d k_{vca}}$ is the ALC time constant

The response of the ALC circuit is tested by varying the input power and recording the output power in a Gain Phase analyzer. The results are shown in Fig 6.9. Rise time and fall time of the ALC circuit is also measured using a step change in input RF power level and ALC feedback voltage is recorded in a storage oscilloscope. See Fig 6.10.

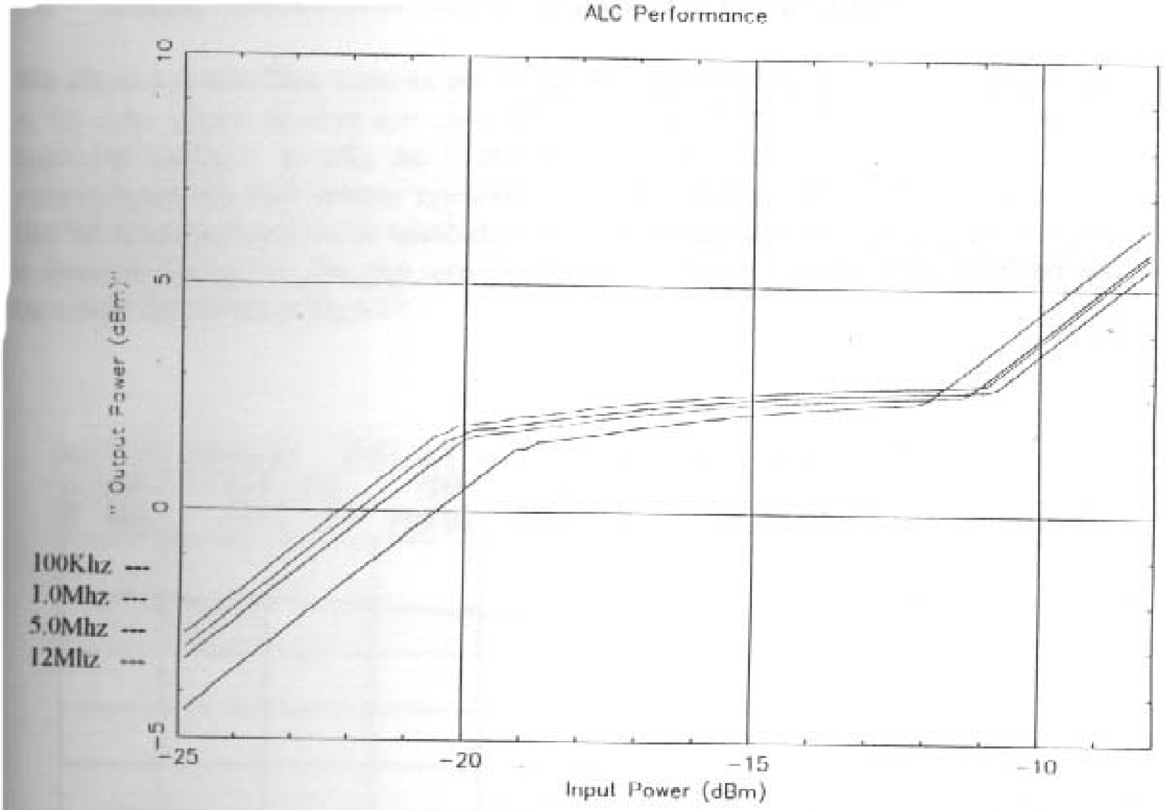


Fig 6.9 - ALC Circuit Response

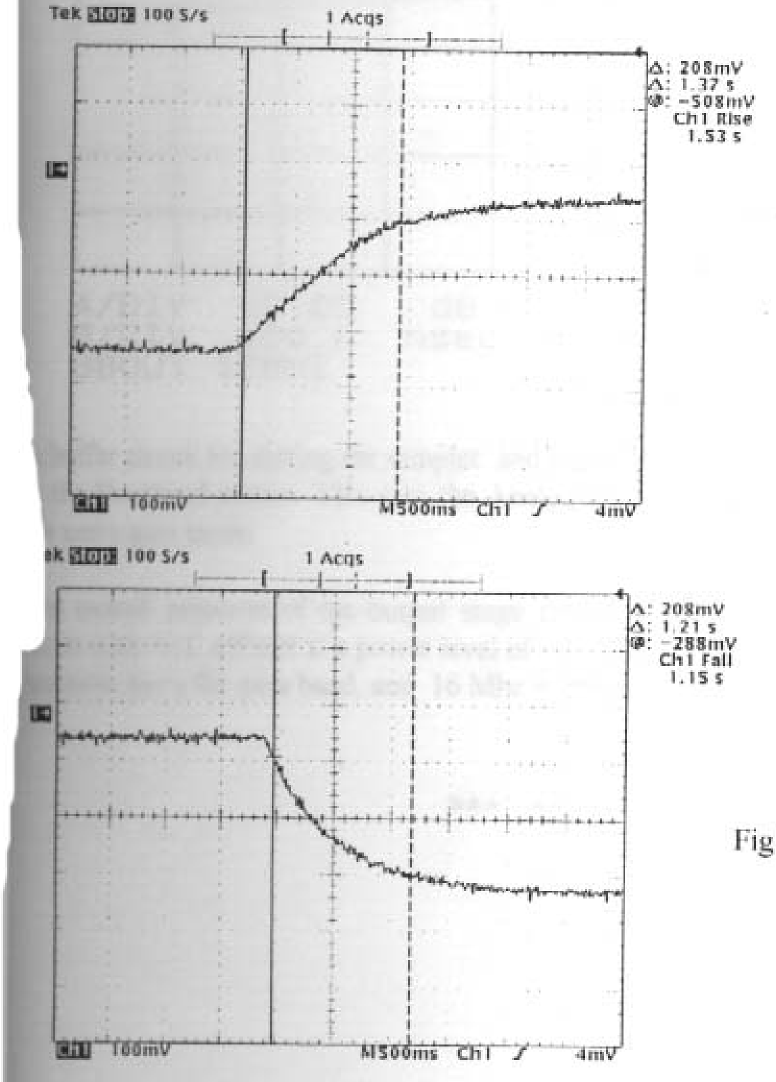


Fig 6.10 - Rise and Fall Time of the ALC

Input Power Switched Between
-27 dBm and -26.5 dBm

6.4 OTHER CIRCUITS IN THE BASEBAND OUTPUT STAGE

The elliptic low pass filter removes any unwanted interfering signals at higher frequencies. A 7th order elliptic function low pass filter with cut off at 16 Mhz is used. When the bandwidth setting is 16 Mhz this is the only filter in the baseband signal path as the variable bandwidth filter remains bypassed. This filter is wired using trimmer capacitors so that the actual response can be tuned at the time of installation. The actual circuit diagram is shown in the Fig 6.5. The gain response is measured using the gain-phase analyzer and the results are plotted in Fig 6.11.

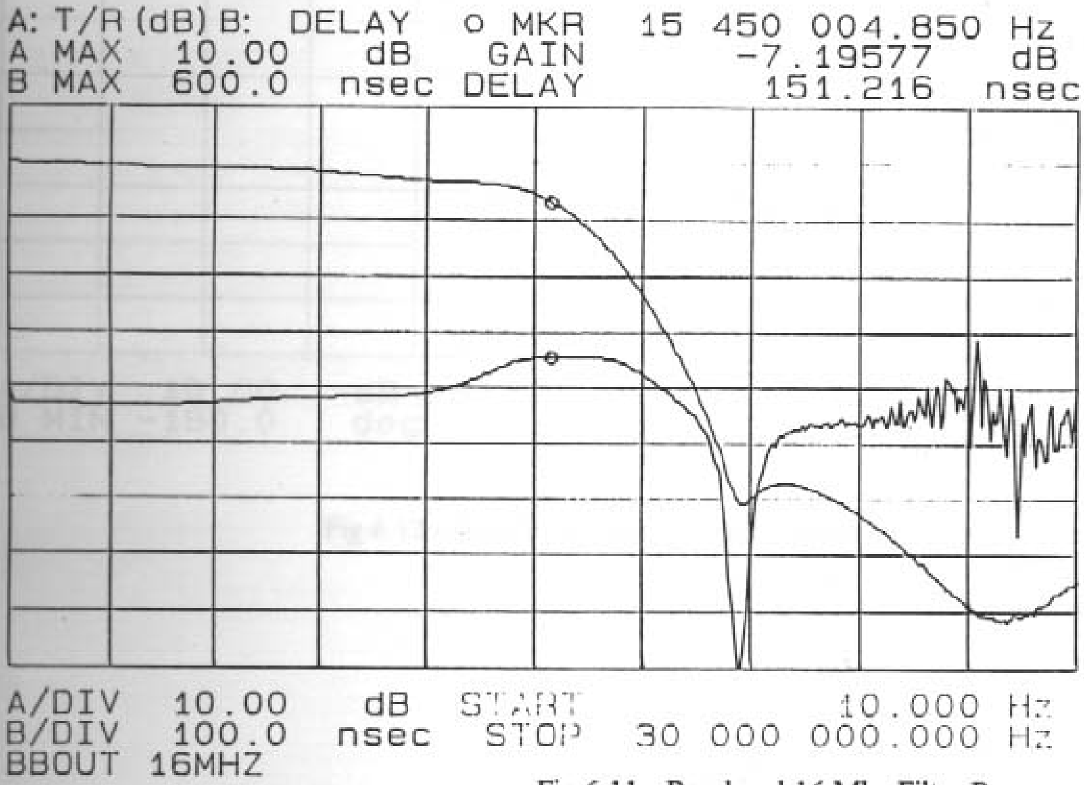


Fig 6.11 - Baseband 16 Mhz Filter Response

A buffer circuit for driving the sampler and long cable lengths is included as the final stage in the Baseband system. This uses the Analog Devices make buffer amplifier HOS 200 in the unity gain mode.

The overall response of the output stage circuit is shown in Fig 6.12. The response is taken with ALC off and at a power level of -40 dBm at the input. It is seen that the overall response has a flat pass band, and 16 Mhz is designed to be the 10 dB point.

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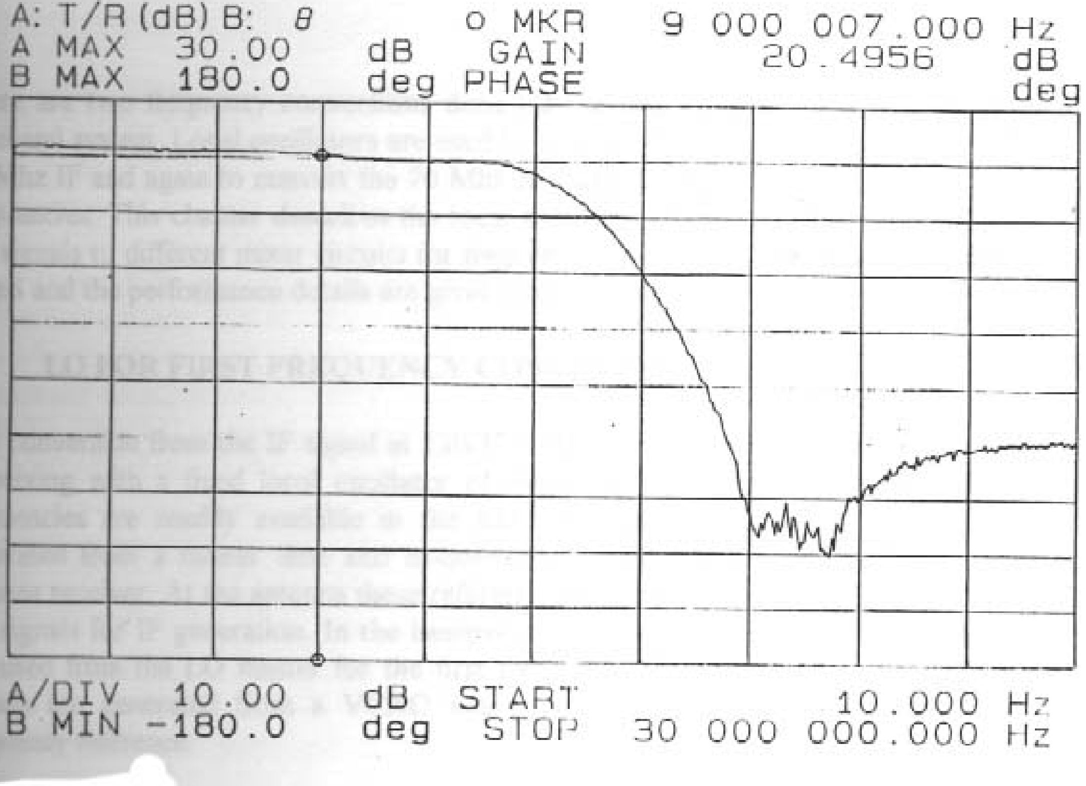

Fig 6.12 - Baseband Output Circuit - Overall Response

Chapter 7

LOCAL OSCILLATOR CIRCUITS

There are two frequency conversions done on the intermediate frequency signal in the baseband system. Local oscillators are used to convert the IF signal from 130/175 Mhz to 70 Mhz IF and again to convert the 70 Mhz IF signal to the sideband signals at baseband frequencies. This chapter describes the local oscillator circuits and the distribution of the LO signals to different mixer circuits for frequency conversion. The circuits are developed tested and the performance details are given later in the chapter.

7.1 LO FOR FIRST FREQUENCY CONVERSION

The conversion from the IF signal at 130/175 Mhz to an IF signal at 70 Mhz is achieved by mixing with a fixed local oscillator of frequency 200/105 Mhz respectively. These frequencies are readily available in the LO reference system where these signals are generated from a master time and frequency reference for onward transmission to the antenna receiver. At the antenna these reference signals are used to generate the necessary LO signals for IF generation. In the baseband system also the same LO reference signals are used from the LO master for the first frequency conversion. The 105/200 Mhz LO signals are generated from a VCXO which is phase locked to the master time and frequency reference.

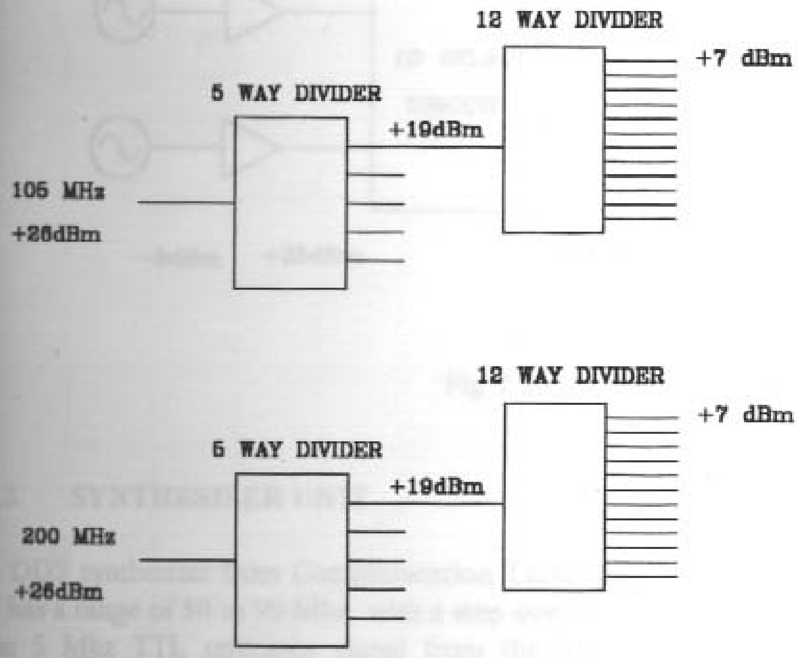

Fig 7.1 - LO 105/200 Mhz Distribution Scheme

Hence only the distribution of these LO signals to various baseband units is to be taken care of in the baseband system. The scheme used for this distribution is shown in Fig 7.1. A standard 5 way torroidal power divider from Minicircuits lab is used to split the signals and then at each baseband rack a 12 way torroidal divider is used to distribute this signal for each system.

7.2 LO FOR IF TO BASEBAND CONVERSION

The IF to Baseband conversion is done using a 70 Mhz local oscillator. This LO is common to all antennas. The main requirement in case of this LO is to have the facility to shift the lower end of baseband output and thereby scan the full bandwidth available. This may be used for avoiding any interfering lines and get a clean baseband signal to the correlator system. This is achieved by using synthesizer circuits which can be tuned from 50 Mhz to 90 Mhz . By varying the LO frequency in this range, the zero frequency of the Baseband output is shifted over the available IF bandwidth. The step size has been fixed at 100 Hz. This LO system has facilities to provide different LO frequencies for the 130 and 175 Mhz IF channels if required. Two synthesizers are used as the local oscillator with a selector switch so that either both polarisation channels may be connected to the same synthesizer or to different synthesizers. The synthesizers may be tuned independently. The block schematic of the unit is shown in Fig 7.2. The distribution scheme used is similar to the one shown in Fig 7.1.

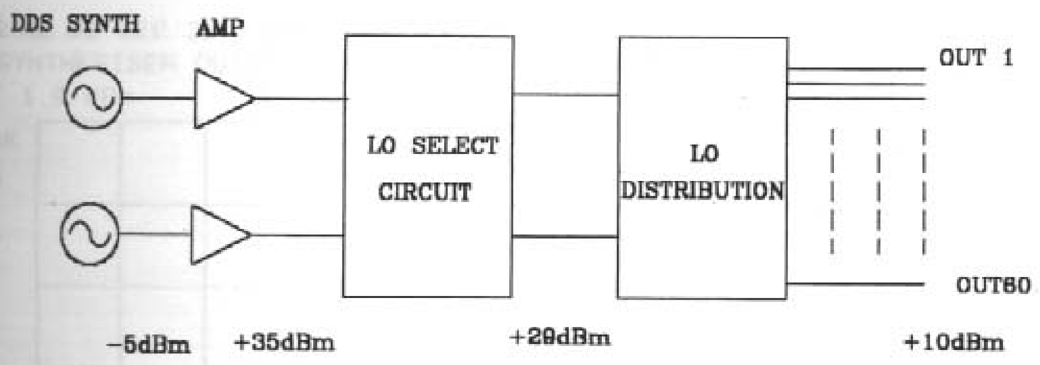

Fig 7.2 - 70 Mhz LO Generation & Distribution

7.3 SYNTHESISER UNIT

A DDS synthesizer from Communication Techniques Inc. is used as the Local Oscillator. It has a range of 50 to 90 Mhz with a step size of 100 Hz. This synthesizer is locked on to the 5 Mhz TTL reference signal from the Master LO unit. The frequency of these synthesizers is set from central computer through control and monitor circuits. The important specifications of the Synthesizer used are given in Table 7.1.

Table 7.1 Specifications

Frequency Range	50 - 90 Mhz, 100 Hz step
Tuning	TTL levels, BCD format
Non-harmonic Spurious	< -60 dBc
Harmonic Spurious	< -30 dBc
Phase Noise	< -80 dBc at 100 Hz
Switching Speed	< 100 mSec

The frequency is selected using a 6 byte control word that specifies the frequency required. The control word is decided as follows,

Frequency required : FE.DCBA Mhz
Control word : (F-5) E D C B A in HEX

For example 51.2345 Mhz can be set using the control word 01 23 45. This data is given to the control input in the BCD format.

Each of these synthesizer outputs are amplified so as to drive all the SSB mixer units. The LO distribution scheme in Fig 7.2 above shows the power levels at various stages of the system. Thus an output of +35 dBm at the synthesizer output is required. This is done using Minicircuits Amplifier module ZHL 5W-1.

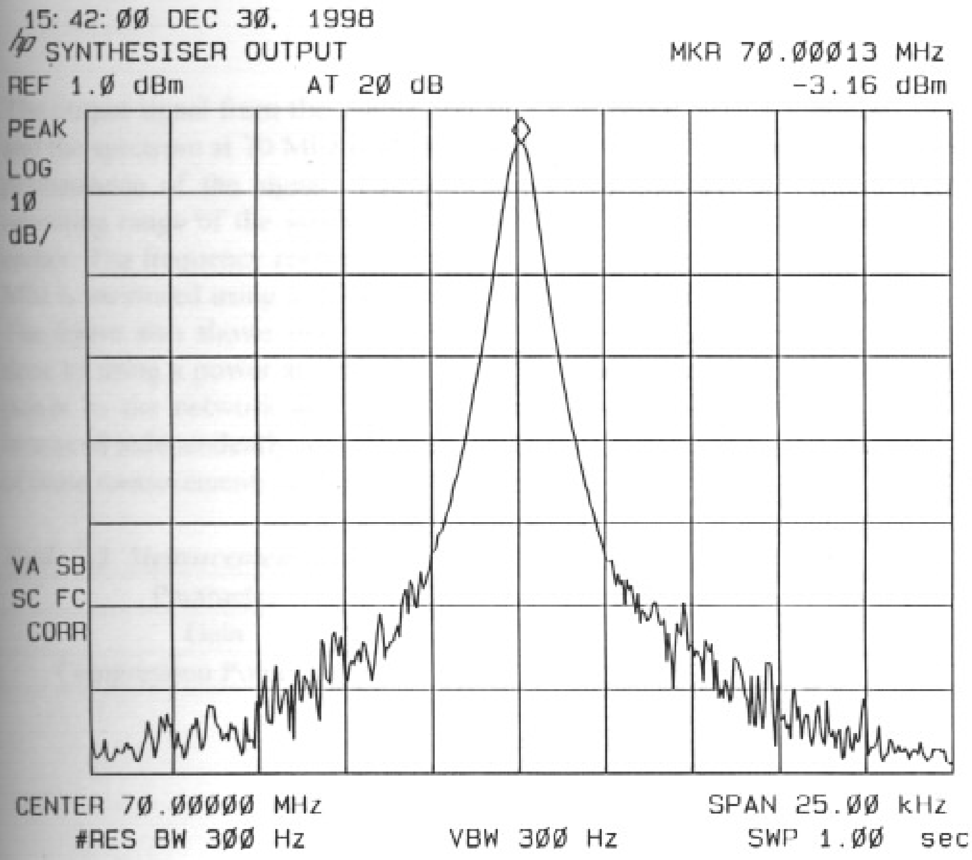

Fig 7.3 (a) - Synthesiser Output Spectrum

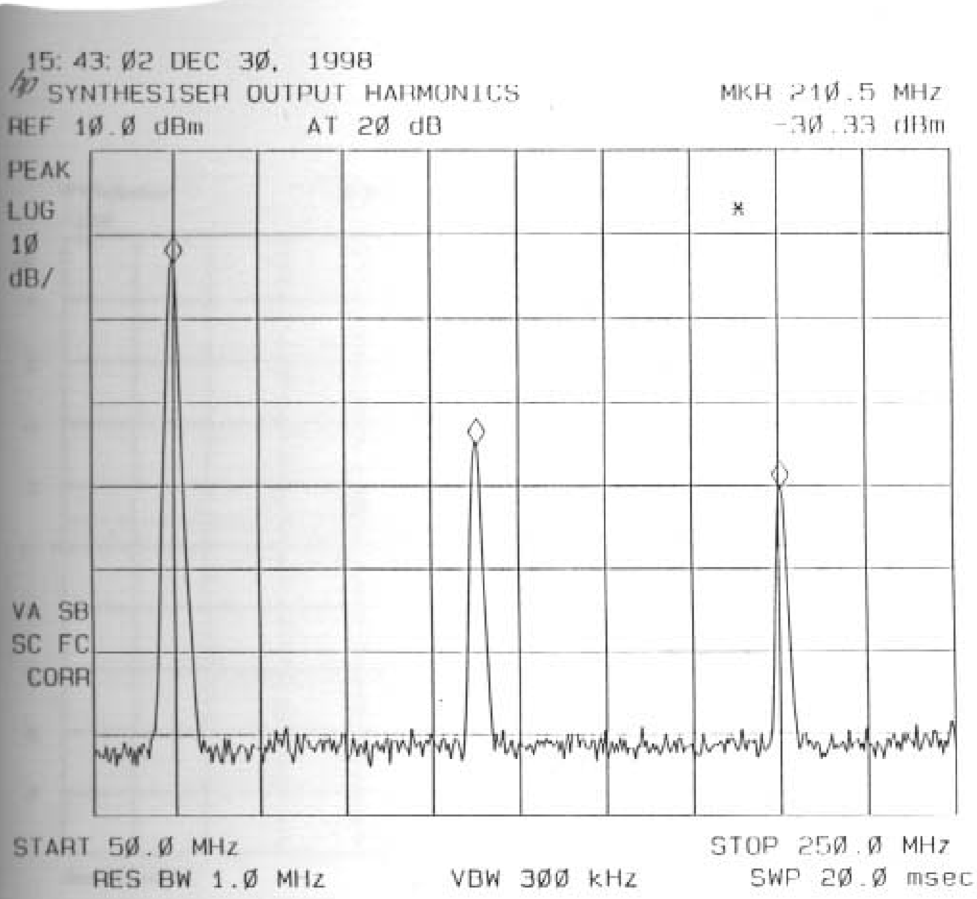

Fig 7.3 (b) - Harmonics at the Synthesiser Output

The output signal from the synthesizer unit is checked with a spectrum analyzer, HP8590 and the spectrum at 70 Mhz is plotted as shown in Fig 7.3. The figure shows the harmonic performance of the signal. The performance is checked at various frequencies in the operating range of the synthesizer and is found to satisfy the requirements as mentioned earlier. The frequency response of the amplifier module ZHL 5W-1 in the range 50 to 90 Mhz is measured using a network analyzer HP8714 and the results are plotted in Fig 7.4. The figure also shows the 1 dB compression point. Compression point measurement is done by using a power attenuator of 40 dB at the output of the amplifier so that the input power to the network analyzer is within the safe limits, the loss of the cable is then measured independently and the performance of the amplifier is calculated. The summary of these measurements is shown in the Table 7.2.

Table 7.2 Measurement Results

Parameter	Calculated	Measured
Gain	40 dB	37.8 dB
Compression Point (input)	-3 dBm	-7 dBm

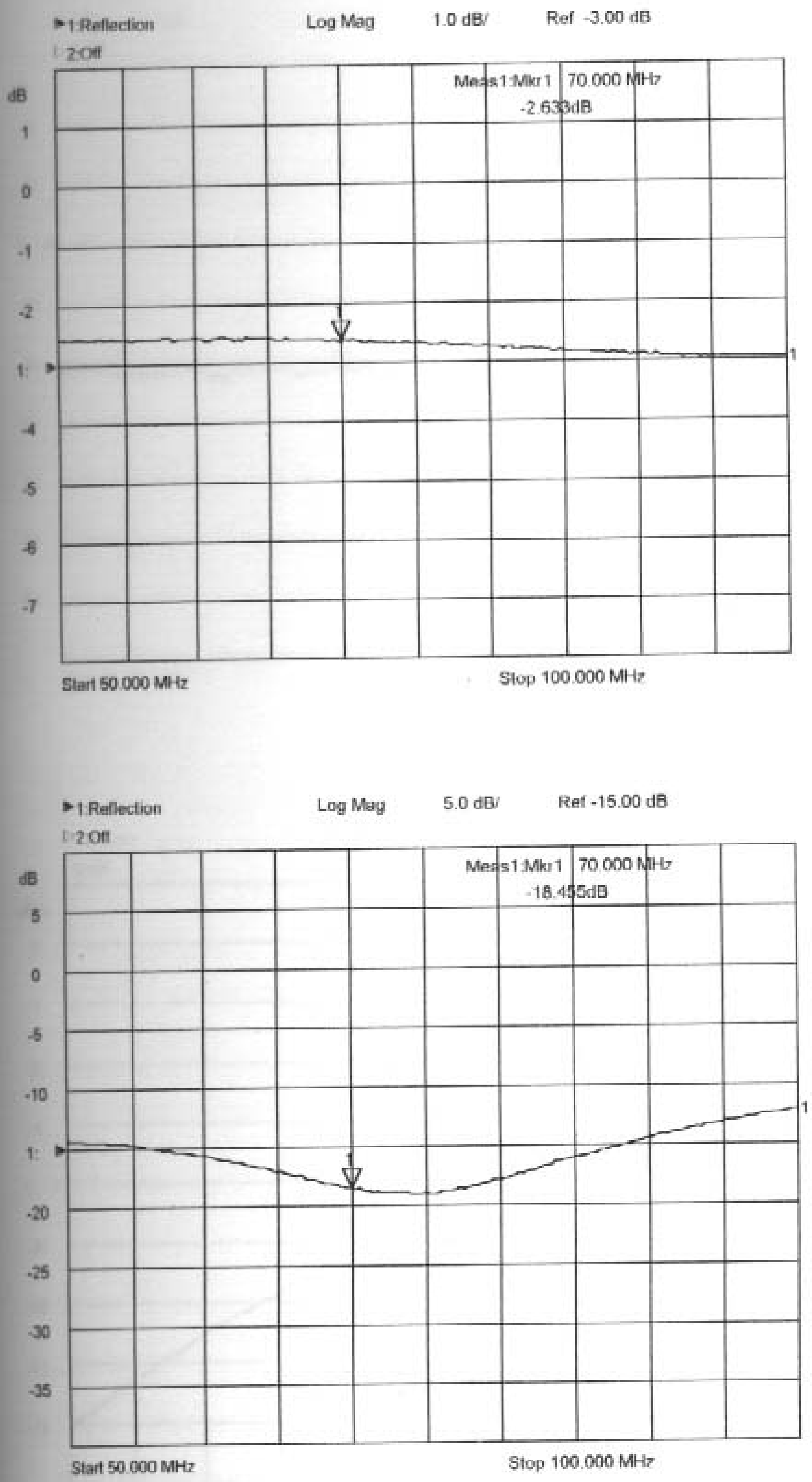

Fig 7.4 (a) - LO Amplifier Reflection Coefficient at input (left) and output (right)

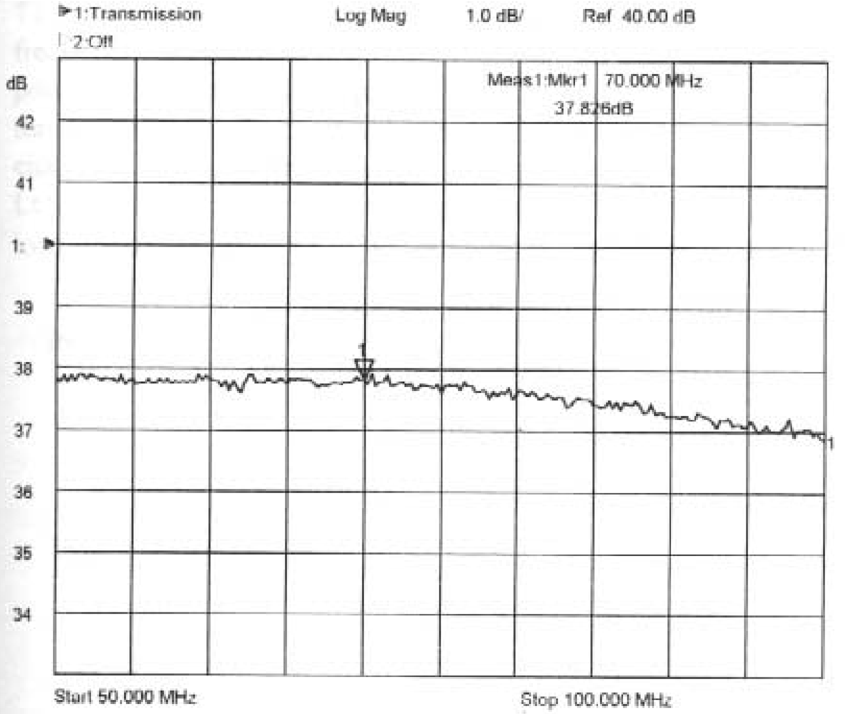

Fig 7.4 (b) - Response of LO Amplifier

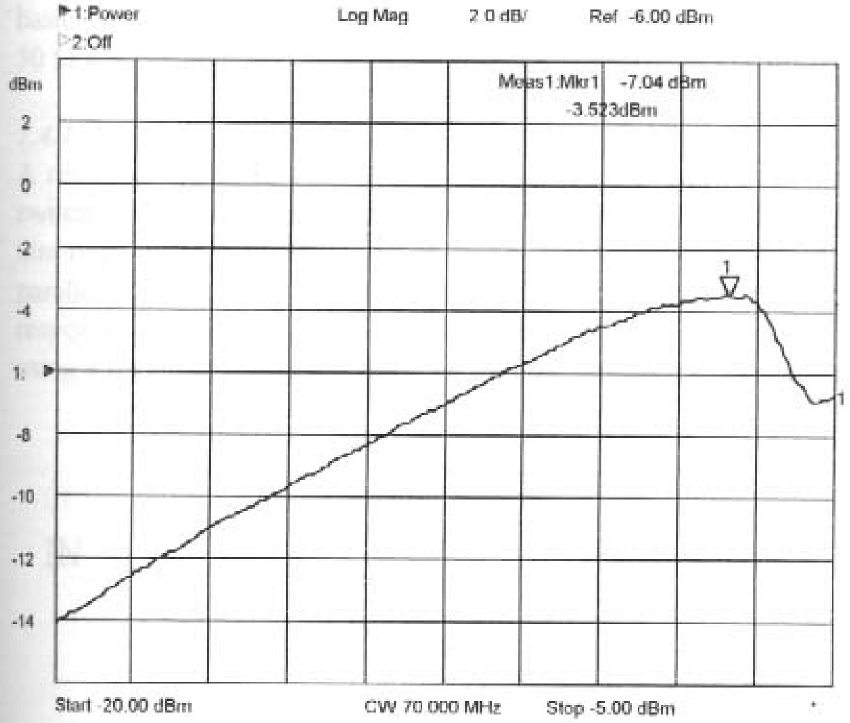

Fig 7.4 (c) - Compression Point, LO Amplifier (with a 40 dB attenuator at output)

7.4 LO SELECTOR UNIT

The main purpose of this unit is to provide either the same LO frequency or different LO frequencies to the two polarisation channels from each antenna. A home made two way power divider is used to split the synthesized signals. The output signals are then passed through the LO select circuit so as to select the LO frequency for the different polarisation channels. An important parameter for these circuits is the power handling capacity as each LO out is approximately 5W output power. The selector circuit should also offer a fixed load impedance of 50 Ω to the amplifier outputs irrespective of the switch position.

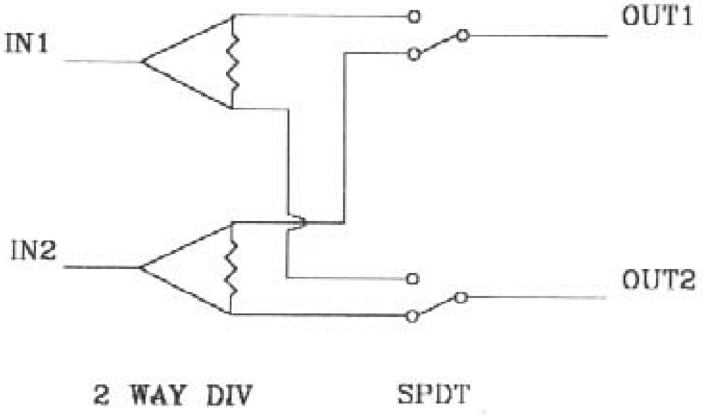

Fig 7.5 - LO Select Schematic Diagram

LO selection is achieved by first splitting both synthesizer outputs and then selecting the necessary synthesizer for each of the polarisation channels using an absorptive switch. The basic schematic is shown in Fig 7.5. Here two reflective switches are wired along with a 50 Ω load of high power handling capacity to form an absorptive switch.

7.4.1 Resistive Power Splitters

A resistive power divider is used in splitting the LO signals before feeding them to the switching device. The power divider is the common T splitter, but high wattage carbon film resistors (5W) are used as shown in Fig 7.6. The resistance of 16.6Ω is obtained as a parallel combination of 22Ω and 68Ω. Such a splitter gives a wideband frequency response and higher power handling capacity. The response of this splitter is measured using a network analyzer (Fig 7.7) and the results are summarised in Table 7.3.

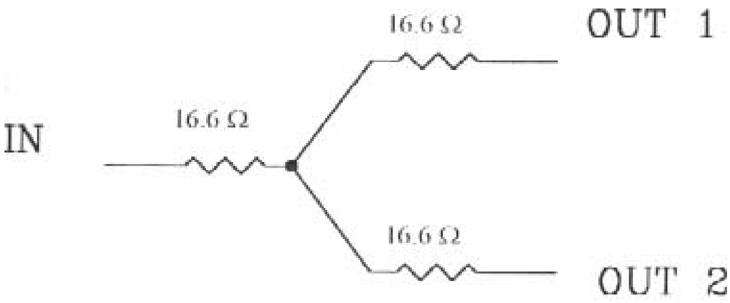

Fig 7.6 - Resistive Power Splitter

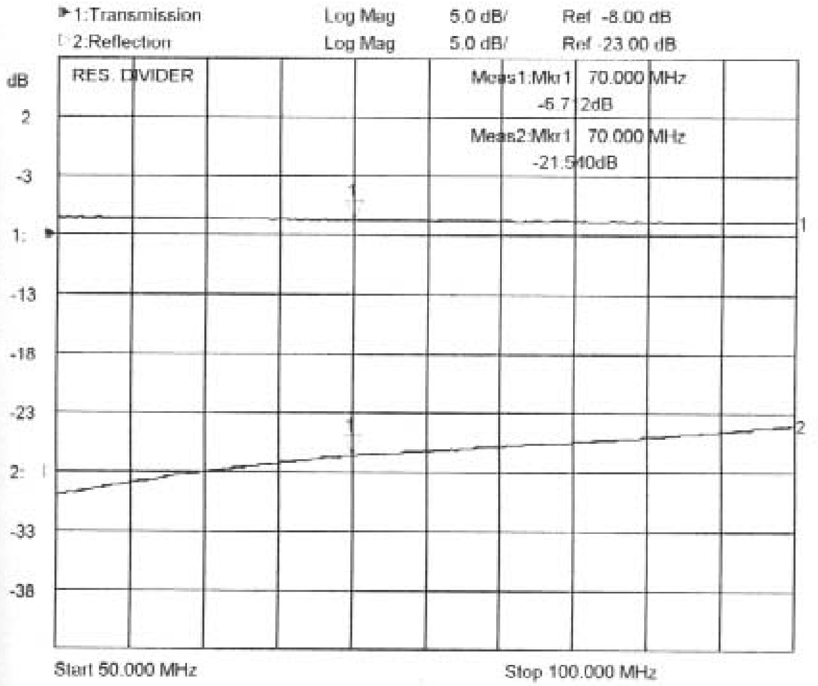

Fig 7.7 - Power Splitter - Response

Table 7.3 Measurement Results

Parameter	Calculated	Measured
Insertion Loss	6 dB	6.7 dB
Reflection Coefficient	-	-21.5 dB

7.4.2 Absorptive Switches

Two reflective SPDT microwave relays from Jennings TOH-54 are wired along with a high power matched load to form an absorptive switch. Necessary driver circuits for the relay is designed based on switching transistor 2N2222. The circuit diagram of the switch is shown in Fig 7.8. The circuit directly works on a TTL control input and hence can be connected to a computer.

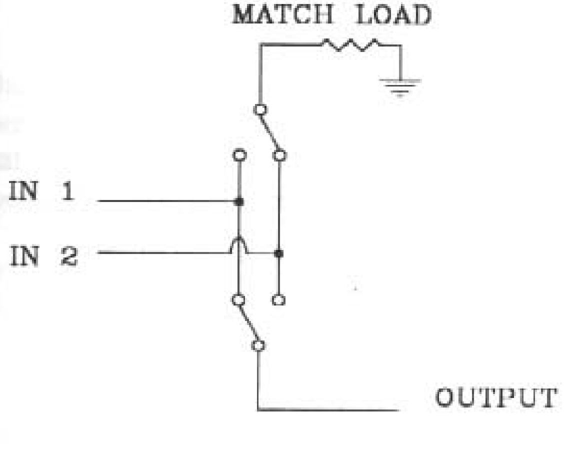

Fig 7.8 (a) - SPDT Absorptive Switches

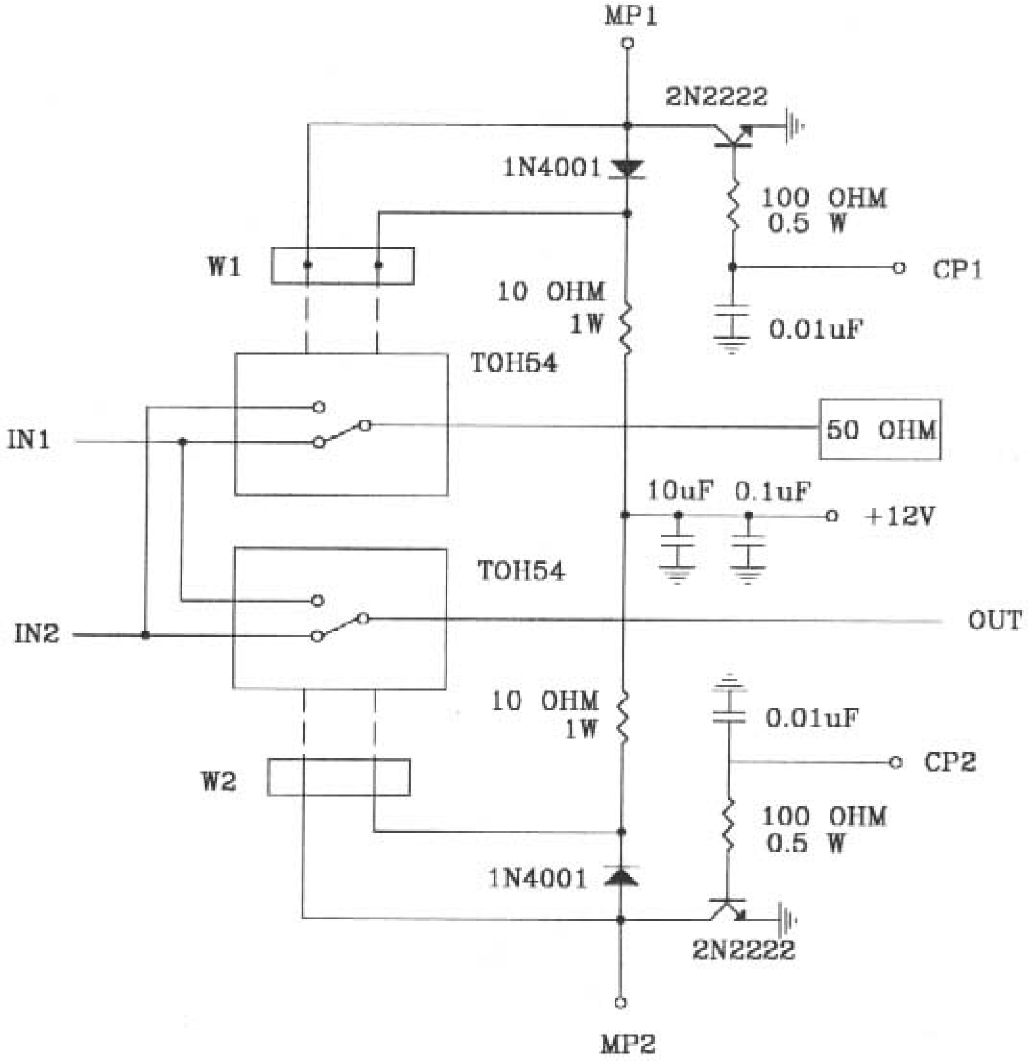

Fig 7.8 (b) - SPDT Switch Circuit Diagram

The insertion loss and the reflection coefficients at the input and output of this circuit are measured using a network analyzer and the results are plotted in Fig 7.9. It can be seen that the circuit provides a very low insertion loss to the signal and the reflection coefficients are acceptable. Fig 7.10 shows the isolation between the ports.

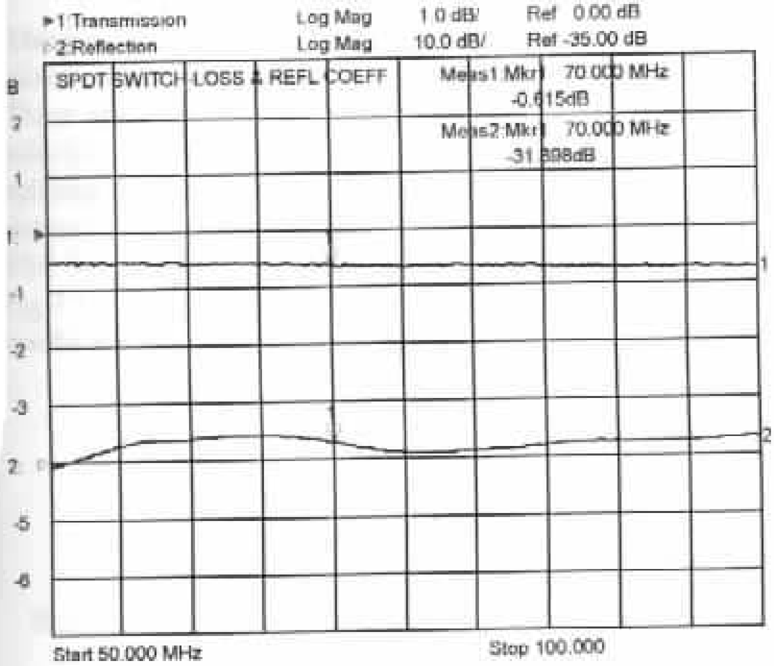

Fig 7.9 - SPDT Switch, Insertion Loss and Reflection Coefficient

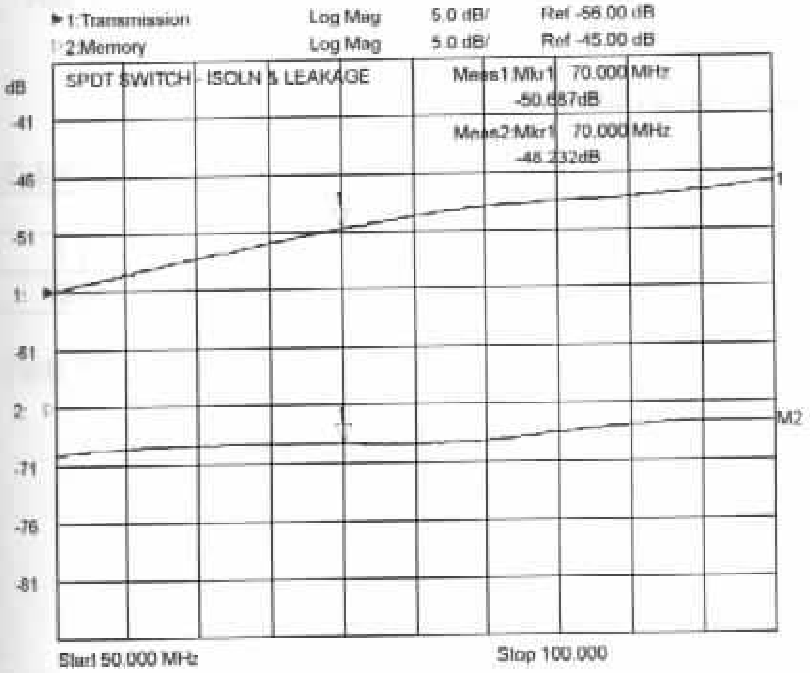

Fig 7.10 - SPDT Switch, Leakage at the unselected input and output

7.5 LO DISTRIBUTION UNIT

These circuits distribute the LO power output from the selector unit to the different SSB mixer circuits so that the required power of +10 dBm is provided to each SSB circuit. There are a total of 30 SSB mixer units in each polarisation channel, this means each selector output is to be split into a minimum of 30 outputs. This is achieved by first splitting using a 5 way divider, one output is used for monitoring the LO power and other outputs are further split using 12 way dividers. Minicircuits make torroidal power dividers PSC 5-1 and PSC 12-1 are used for this purpose. The detailed circuit diagram is shown in Fig 7.11. The insertion loss of these splitters is measured using a network analyzer and the results are plotted as in Fig 7.12.

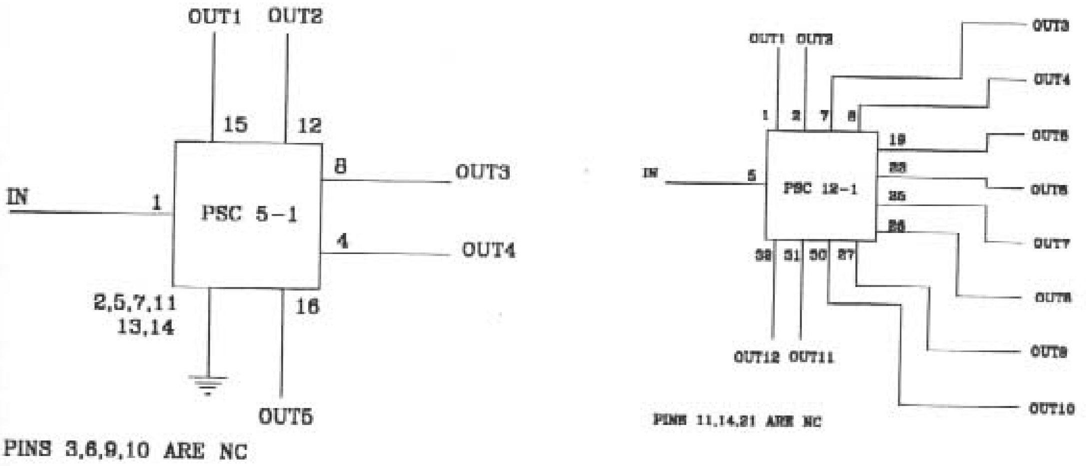

Fig 7.11 - 5 Way (left) and 12 Way Power Splitter Circuits (right)

Table 7.4 Measurement Results

Device	Parameter	Calculated	Measured
5 Way Splitter	Insertion Loss	7.5 dB	7.7 dB
12 Way Splitter	Insertion Loss	11.5 dB	11.8 dB

able 7.4 gives a comparison between the measured and calculated values.

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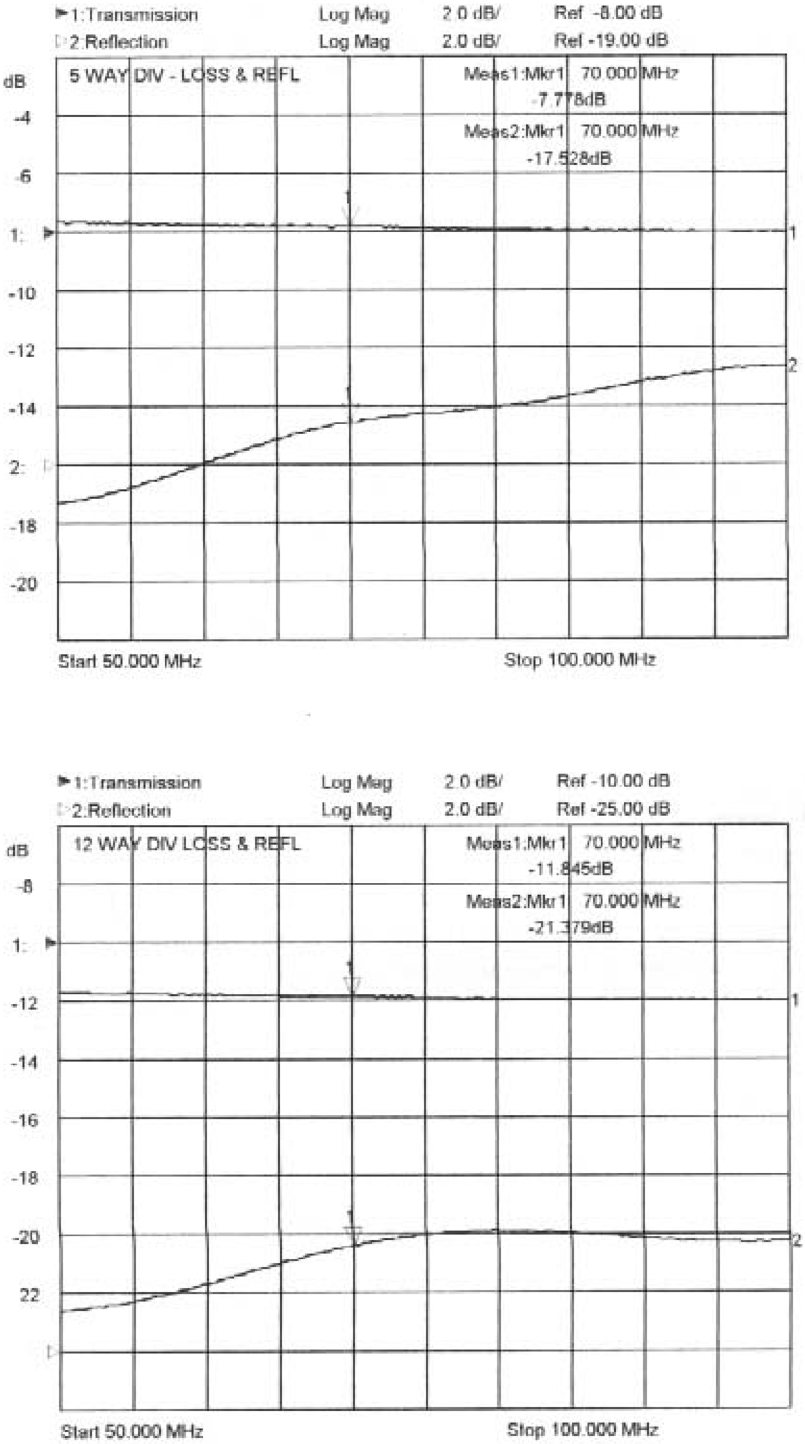


Fig 7.12 - Insertion Loss Plots for PSC 5-1 (above) and PSC 12-1 (below)

Chapter 8

CONTROL & MONITOR CIRCUITS

This chapter describes the C&M circuits used in baseband system. These circuits facilitates the central computer to control all the parameters of the baseband system. The control and monitor operations from the computer are managed by the telemetry systems and baseband C&M provides the necessary interface to the baseband system circuits. The design of the interface cards and other details are given here.

The C & M circuits are used so that the baseband system parameters like gain, bandwidth, LO frequency etc. can be set directly from the central computer. This also facilitates remote monitoring of the set parameters and also other parameters to study the health of the system like the dc voltage and current in each unit, power levels at various stages etc.

All control and monitoring commands to the system are send from the central computer through a telemetry system. The telemetry system provides a Measurement, Control and Monitor (MCM) circuit for sending digital control data to the circuits and for acquiring analog monitor data [15,16,17]. It is basically a micro-controller with digital output channels and analog input channels. The MCM card has a 16 bit digital data output and a 62 channel analog input. The digital data is of TTL levels whereas the analog monitor channels may be configured to any one among the three range settings, ±10V, ±5V, and 0-5V. The digital data is available on a 20 pin FRC type connector and the analog channels are accessible through a 64 pin euro connector. An RS 485 cable connects the MCM card to the telemetry systems for interconnection to the computer.

8.1 C&M FOR THE BASEBAND CIRCUITS

As mentioned in the earlier chapters the baseband system for each antenna will have two IF input channels and four output sidebands to the correlator. Thus there are four active low pass filters and four bandwidth compensation amplifiers for each antenna. The filter and bandwidth compensation amplifier has 8 bit control each. These control points are monitored back so as to ensure that the settings are done correctly before the observation start. In addition to the control points to be monitored, there are many other voltages also to be monitored like the dc supply voltage to the units and detector and ALC feed back voltages. The supply voltage and current in each unit is measured so that any fault can be detected easily. For monitoring the current the supply to the unit is given through a series resistor of 10 Ω and the dc voltage on both ends of the resistor is monitored. The current drawn from the dc supply can be calculated using ohm's law. This measurement is very important from the maintenance point of view as any component failure will be immediately indicated as a change in the current drawn from supply.

The exact number of the control and monitor points to be handled by each MCM depends also on the actual mounting scheme of baseband units as explained in chapter 9.

Table 8.1 Control Bits per Antenna

Controlled Parameter	Control Bits	Units per Antenna	No. of Bits
Low Pass Filter Bandwidth	8	4	32
Bandwidth Compensation Gain	8	4	32

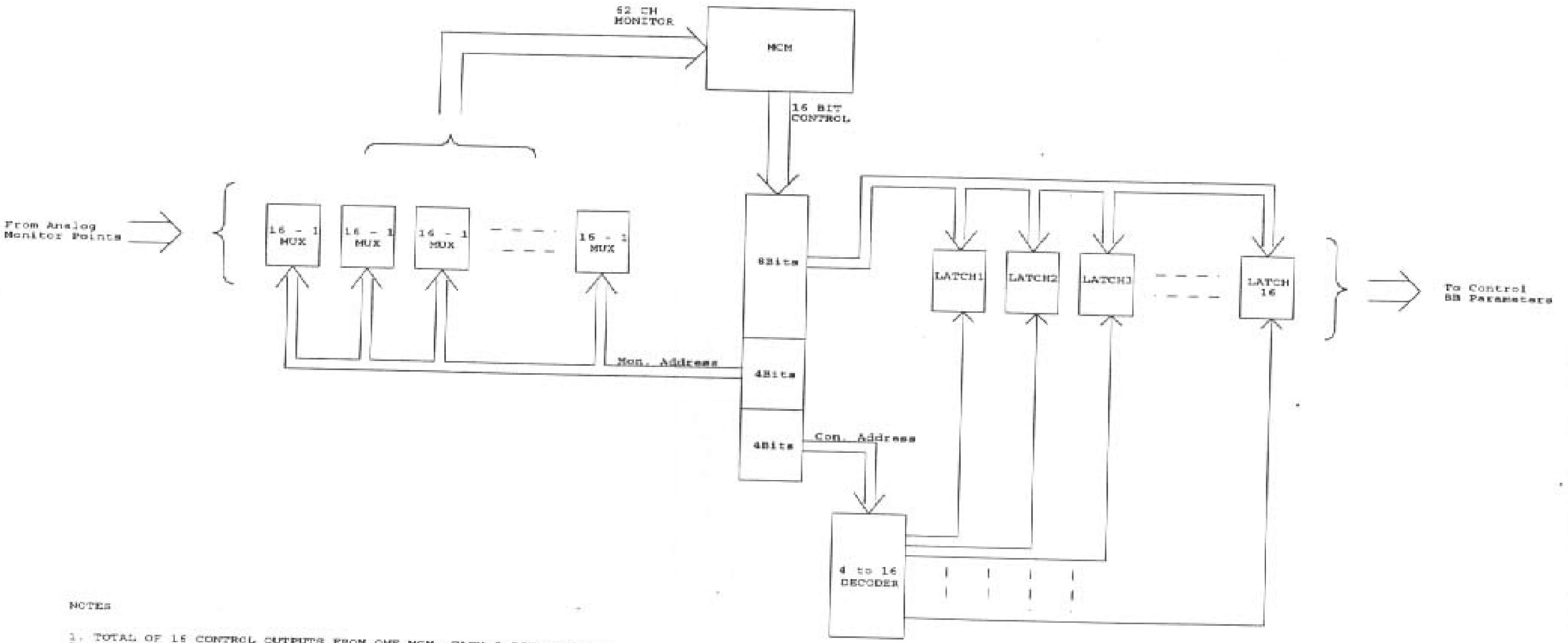
Table 8.2 Monitor Points per Antenna

Monitored Parameter	Monitor Channels	Units per Antenna	No. of Channels
Bandwidth Setting	8	4	32
Gain Setting	8	4	32
DC supply - Mixer unit	4	2	8
DC supply - Filter unit	8	4	32
ALC on/off	1	4	4
Detector - Mixer unit	2	2	4
Detector - Filter unit	1	4	4
16 Mhz Bandwidth setting	1	4	4

Thus each antenna requires 64 control bits and 120 monitor points, it needs a minimum of 2 MCM cards per antenna. In order to reduce the number of MCM cards some sort of multiplexing is used so that a far less number of MCM cards can handle all activities for the baseband system.

The interface circuits are designed so that one MCM card can now handle two antenna baseband system. The 16 bit digital data from MCM is divided into three parts, an 8 bit data field, 4 bit control address field and a 4 bit monitor address field. The control address is varied so that the baseband settings for different sidebands can be done from same MCM. The 4 bit control address can be used to latch the 8 bit data in 16 different latch ICs, thus the same MCM card may be used to set upto 16 different parameters. That is a maximum of 8 sidebands or two antennas as each sideband has both bandwidth and gain controls. Similarly the 4 bit monitor address may be used to switch in any one of 16 analog monitor points. The selected monitor level is again connected to one of the analog channels in the MCM card. Thus these 4 bits can increase the number of monitor points in the MCM by 16 times if required. The scheme used is shown in Fig 8.1.

This is achieved by using two circuits, one is the C&M Interface card at the output of each MCM and the other is the baseband C&M card at each filter circuit. The C&M interface card splits the control data and generates the 8 bit data to be latched in the individual baseband C&M cards and also decodes the 4 bit control address to 16 latch signals to be used by baseband C&M cards for latching the data. The baseband C&M cards multiplexes the analog monitor data and passes it to the interface card for connecting to the MCM. The connections between the MCM, C&M interface card and the baseband C&M card is shown in Fig 8.2.



NOTES

1. TOTAL OF 16 CONTROL OUTPUTS FROM ONE MCM, EACH 8 BIT PARALLEL
2. EACH MONITOR CHANNEL IS FURTHER EXPANDED TO 16 CHANNELS

PROJECT	GMRT	NATIONAL CENTRE FOR RADIO ASTROPHYSICS TIFR CENTRE, PUNE
GROUP	RECEIVER	
SUB-GRP	BASEBAND	
DRAWN		CONTROL & MONITOR SCHEME
CHECKED		
APPRVD		
SIGN	DATE	SIZE | DRAWING NO. | REV

Fig 8.1 - Control and Monitor - Basic Scheme

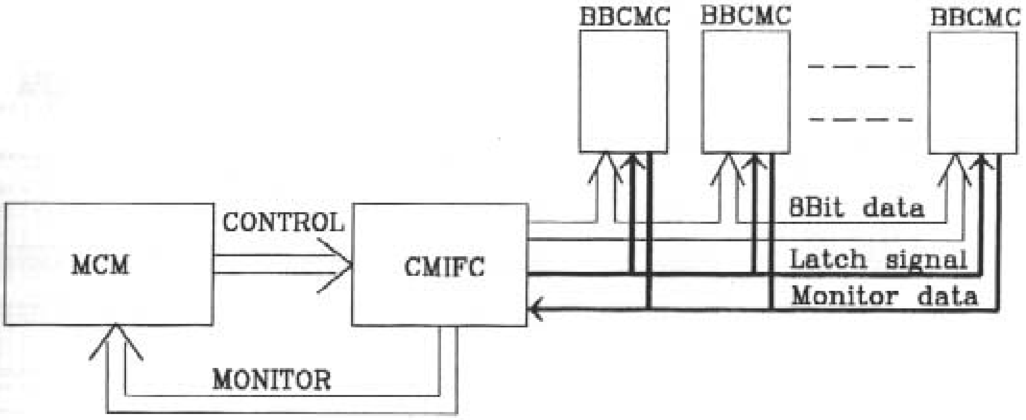


Fig 8.2 - Control & Monitor cards - Interconnections

8.2 C&M INTERFACE CARD

The function of the C&M interface card is to generate an 8 bit buffered data along with 16 latch signals, and 4 monitor addresses from the 16 bit control word in the MCM. This circuit also takes the monitored signals from the baseband C&M card and passes it on to the MCM analog channels with necessary buffering.

The circuit diagram for the C&M interface card is given in Fig 8.3. This circuit uses 74LS241 to provide buffering to the control signals from the MCM. This improves the fan out and allows the data bits and the monitor address bits to be send over long cable lengths. The control address bits are decoded using a 4 to 16 bit decoder 74LS154. These bits are used to enable the latches available in the baseband C&M card so that the 8 bit data is latched permanently. The interface card also provides buffering to the analog data monitored at the baseband C&M card using LM324 op amp circuits. A facility for changing the analog voltage range setting for the monitored channels is available (switchable between ±5 V and 0-5V). These monitored channels are fed to the analog channels of the MCM card.

Each MCM in the baseband system is followed by one C&M interface card and this can handle all control and monitoring for two antennas.

8.3 BASEBAND C&M CARD

The function of this circuit is to latch the data from the C&M Interface card and pass it on to the baseband system with buffering. It also provides a facility to multiplex analog monitor data into two channels to be connected to the MCM.

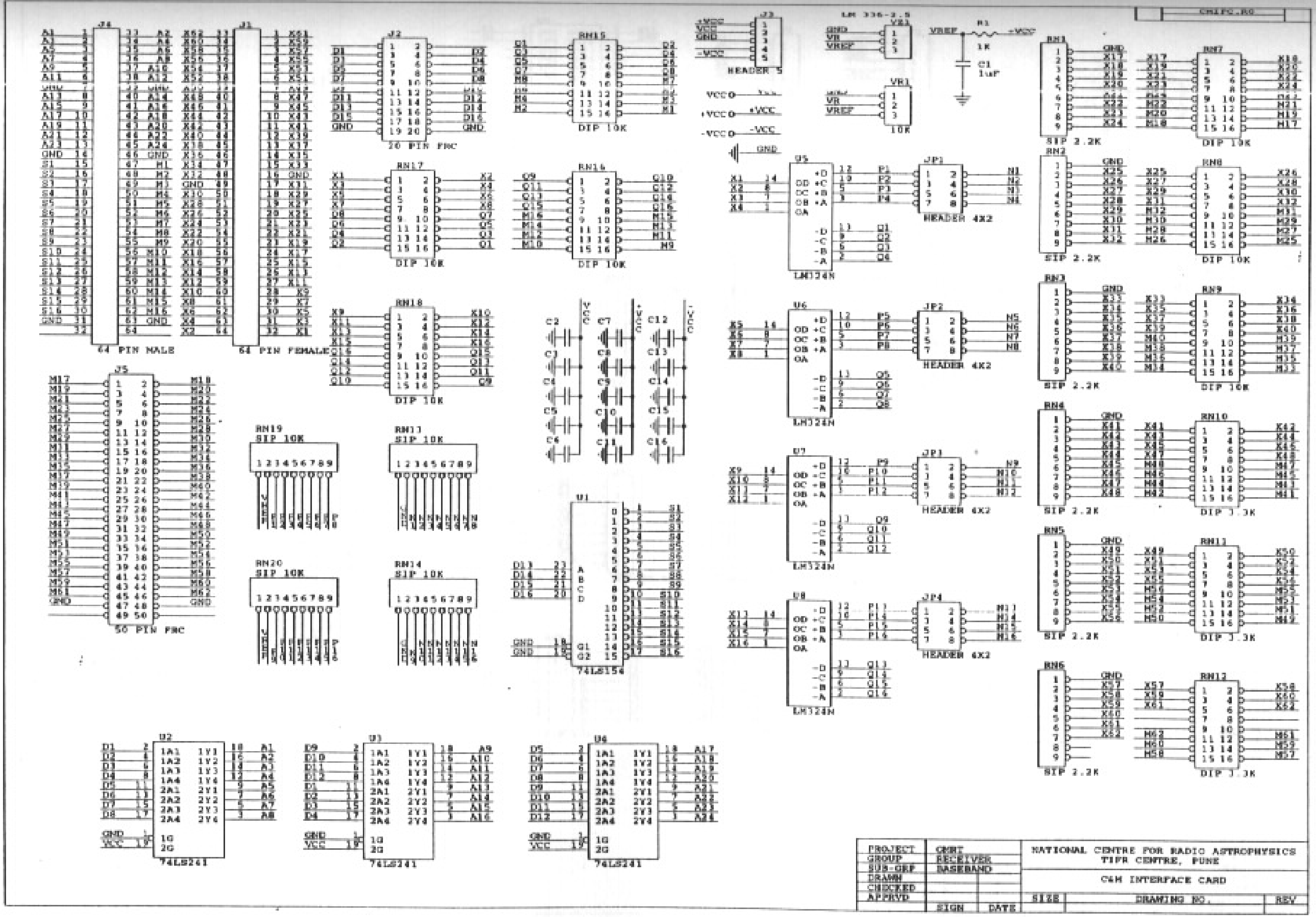


Fig 8.3 - Control & Monitor Interface Card

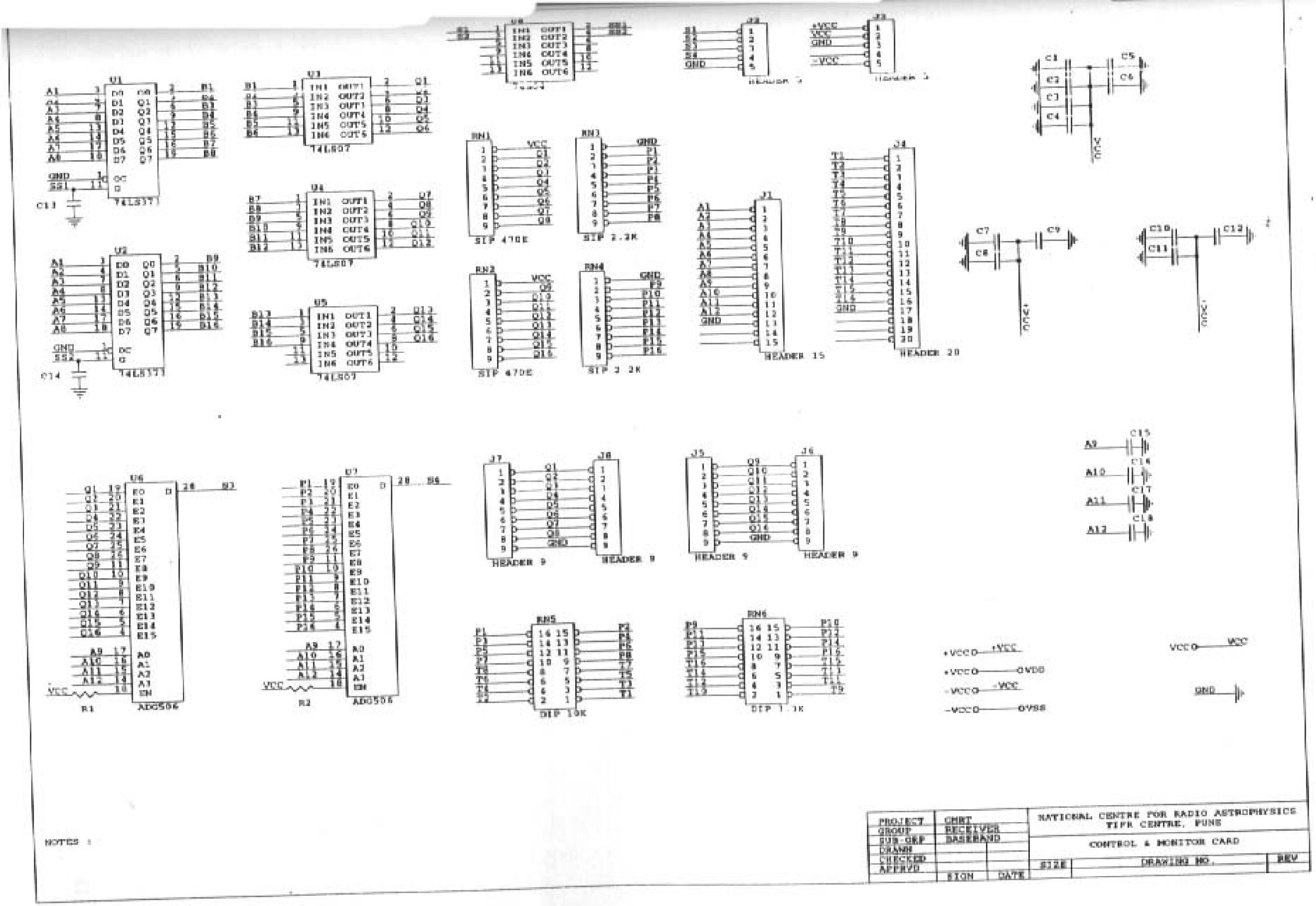


Fig 8.4 - Baseband Control & Monitor Card

The detailed circuit diagram is shown in the Fig 8.4. 74LS373 is used to latch the data and then 74LS07 is used as buffer before sending the data to the baseband system. These TTL level signals can directly control the parameters of the baseband system like gain, bandwidth. 74LS04 is used as buffer and inverter for the latching signals. Analog multiplexer ADG 506 is used for switching in one of the 16 monitor data. Two such Ics are used, one for the monitoring of the set parameters and the other for monitoring the voltage and current in the system. Thus each baseband C&M card uses two analog channels of the MCM.

Each baseband C&M card handles all control and monitoring for one sideband, one polarisation. Thus 4 such cards are required per antenna.

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Chapter 9

MOUNTING SCHEME

This chapter describes the mounting of various units in plug in modules and the installation details of these modules in the standard electronics racks. Detailed drawings and photographs of various units are given later.

The baseband circuits are mounted in standard plug in modules which are then mounted in sub racks and the sub racks in 19" racks. These racks are designed to provide high levels of RFI protection and the shielding is improved on the door side with shielding gaskets [18]. Each of these racks will have one dc power supply unit and 4 sub racks to house various system plug in modules. All cables inside the rack are run inside PVC channel guides provided on the side walls and also behind each sub rack. All connectors are located on the top of the rack and all interconnections between racks are done by cables running on a channel above the racks. Proper air circulation inside the racks is ensured by three fan trays installed inside the rack. The racks are then mounted on a air duct which brings in cool air from a central air conditioning plant.

Based on the type of cards installed the Baseband plug in modules may be classified into 7 different categories. The IF separation circuit is installed in the optical receiver unit and the IF signals are brought over to the baseband units.

*1. **Baseband Mixer Unit*** - This includes all the IF circuits and frequency conversion circuits including the SSB mixer. Thus the input to this unit is the IF from the antennas and the outputs are the sidebands. All necessary circuits for one IF channel is mounted in one unit

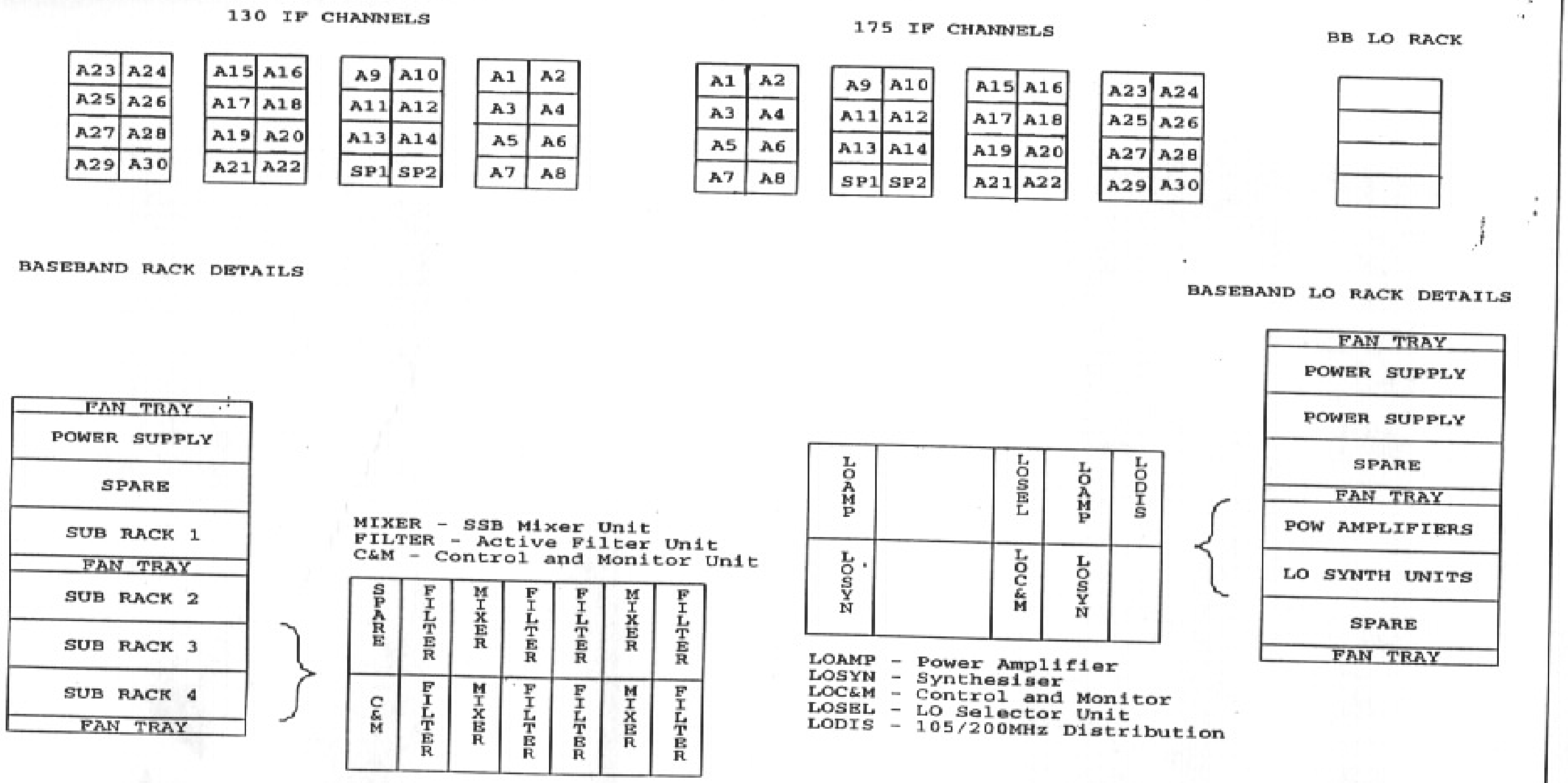
*2. **Baseband Filter Unit*** - This includes all the Baseband processing circuits for one sideband.

*3. **Baseband C & M Unit*** - This includes all control and monitor and related interfacing circuits. One C & M unit can directly control baseband units for 4 antennas of same IF channel.

*4. **LO Synthesizer Unit*** - This includes the synthesizer modules and related control circuits.

*5. **LO Amplifier Unit*** - This includes the amplifier circuits for the LO signals.

*6. **LO Select Unit*** - This includes the circuits for switching in the selected synthesizer output to the baseband system.

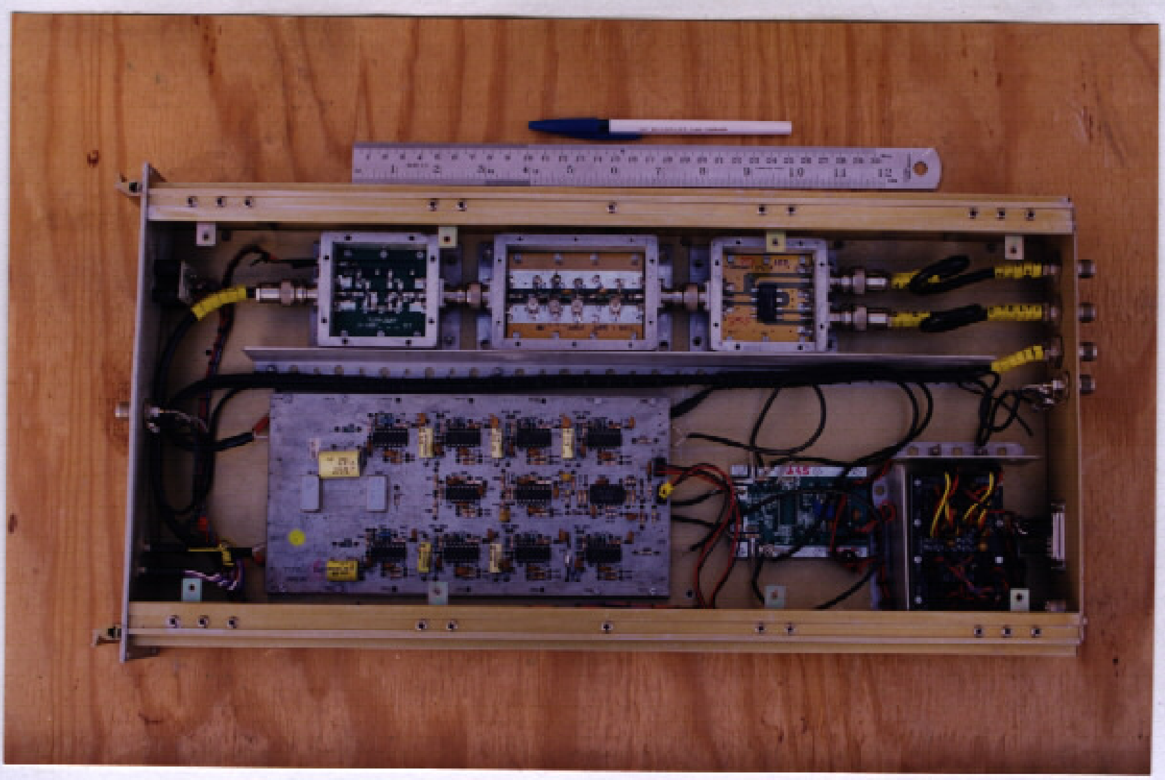


NOTES :

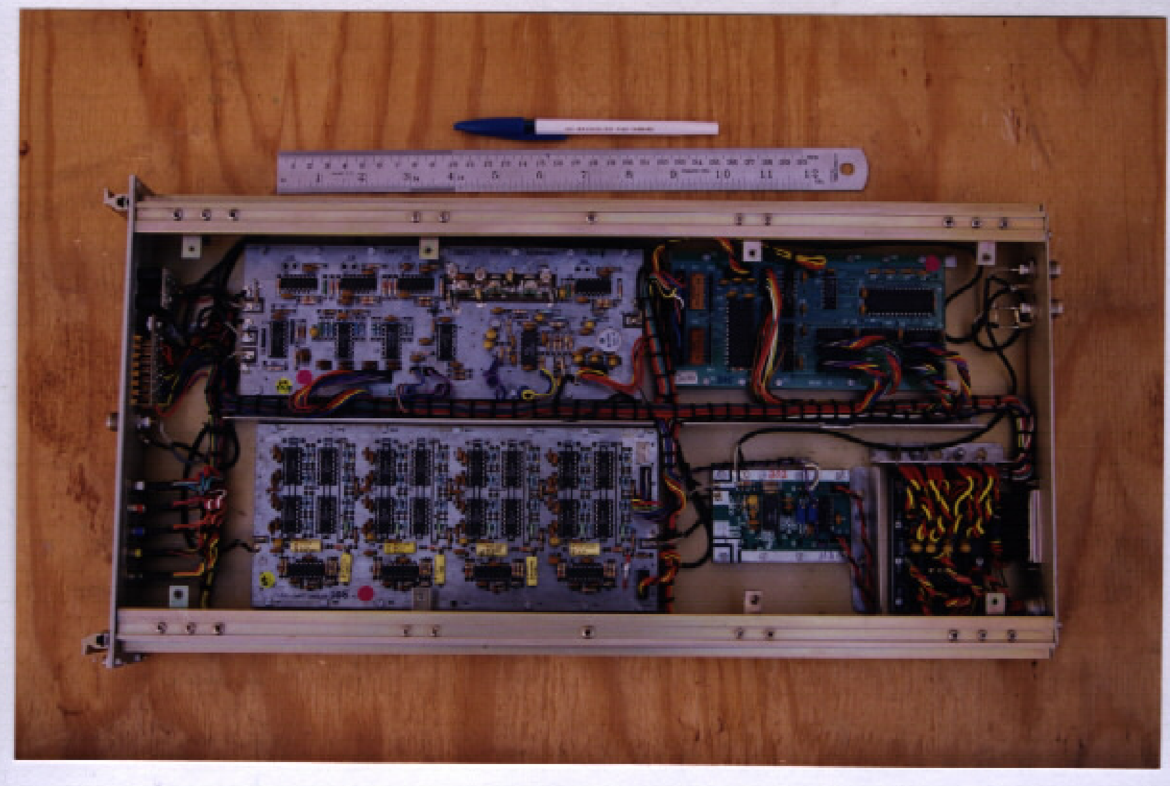
1. A1..A30 Baseband System for the Antenna
2. All Standard 19" Racks with RFI Shielding

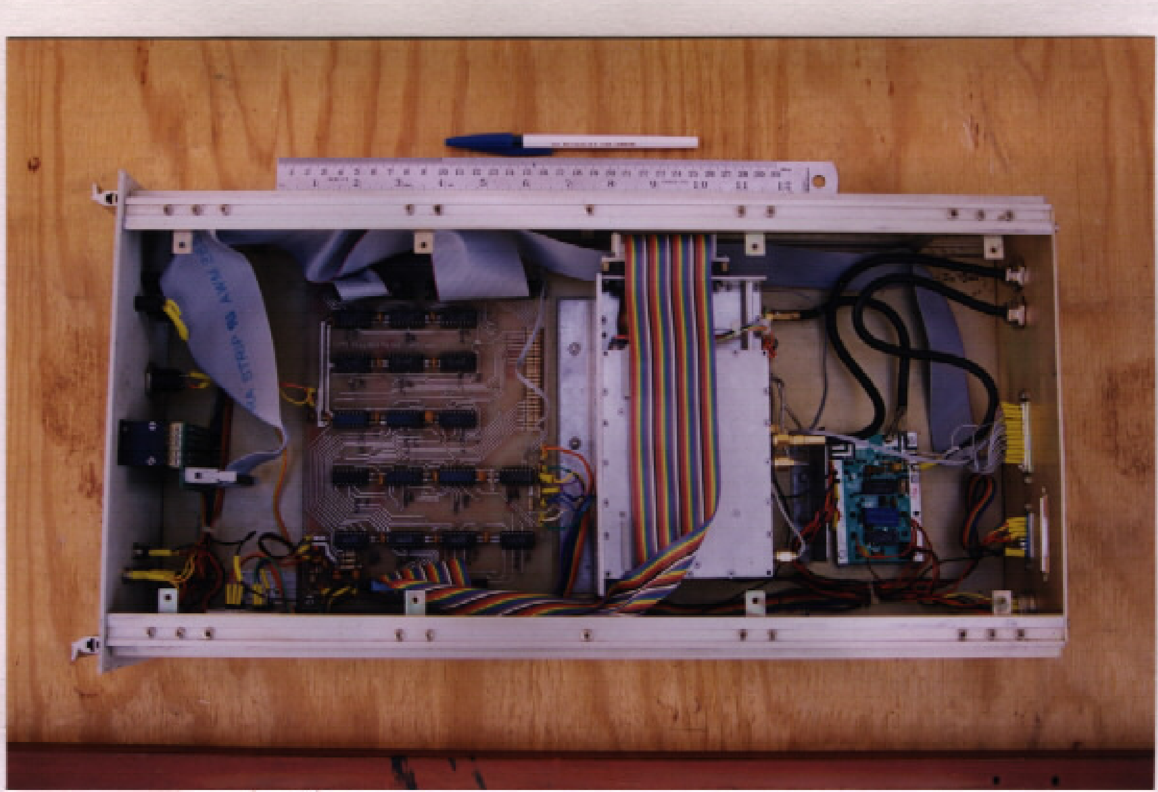
PROJECT	GMRT		NATIONAL CENTRE FOR RADIO ASTROPHYSICS TIFR CENTRE, PUNE	
GROUP	RECEIVER			
SUB-GRP	BASEBAND			
DRAWN			RACK MOUNTING SCHEME	
CHECKED				
APPRVD				
	SIGN	DATE	SIZE	DRAWING NO.	REV

Fig 9.1 - Rack Mounting Scheme for Baseband Circuits

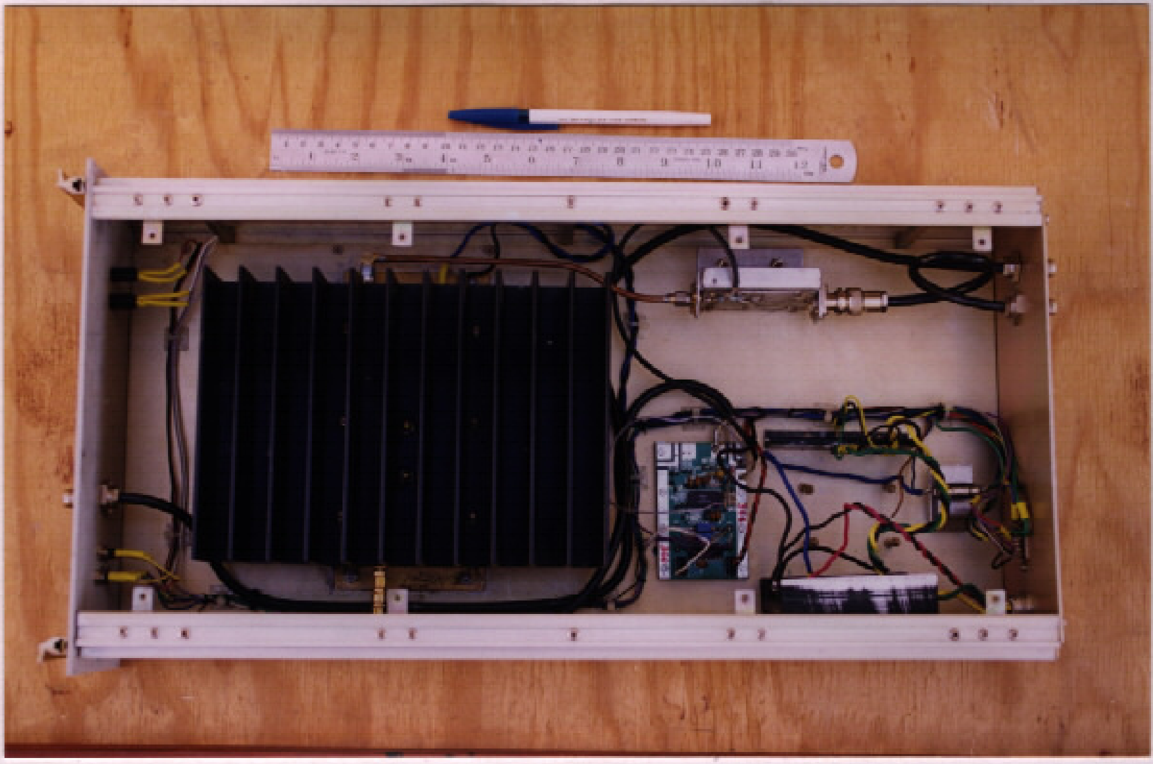


Ph-1 Baseband Mixer and Filter Units - Photograph





Ph-2 LO Synthesiser and Amplifier Units - Photograph



*7. **LO C& M Unit*** - This includes all the control and monitor circuits for the LO synthesizer and selector units.

For simplifying the LO cable wiring the units for same polarisation channel is kept in the same group of racks. Thus 4 racks are used for 130 Mhz channel system and another set of 4 racks for 175 Mhz channel. Each rack has 4 sub racks and each sub rack contains the baseband system for two antennas same IF channel. Each rack has one free sub-rack space meant for future upgradation. All LO system units are mounted in a separate rack and the LO signals are distributed to the individual racks using RF cables. The Fig 9.1 shows the details of the mounting scheme used in various racks.

Photographs of the plug in modules in the Baseband system is also enclosed.

The ac power is fed through an isolation transformer to reduce RFI and noise pick up. Each rack has an ac power distribution and a dc power distribution. The ac distribution provides power to the dc power supply and fan trays. Each Baseband rack takes a current of 4.2A (approx.) from the ac mains. This includes 0.9A current for each fan tray and the current drawn by the CVT based power supply. The CVT based power supply provides a constant dc voltage at the output irrespective of the input voltage variations. The dc distribution unit consists of protective circuit breakers and individual switches to control power to each sub-rack.

*** *** ***

Chapter 10

SYSTEM INTEGRATION AND MEASUREMENTS

This chapter describes the overall performance of the baseband system after the circuits are mounted in the plug in modules and the units are integrated to form the final system. Various system parameters are measured and the results are checked so as to verify whether the system satisfies the original specifications as described in chapter 3. Overall System response given in this chapter does not include IF Separation circuit. This circuit is installed along with the fiber optic receiver unit.

Several equipments are used in these measurements. These include instruments like Network Analyzer, Spectrum Analyzer, Vector voltmeter, Signal Generator and Power Meter. Apart from these a noise generator unit (home made) is also used to test parameters of the integrated system. The noise generator unit is made so that it can inject a wideband noise power with a frequency response similar to that from the antenna. It simulates the IF signal from the antenna with a centre frequency of 130/175 Mhz and a bandwidth of 32 Mhz defined by SAW filters. This unit is equipped with variable attenuators so that the actual power fed to the Baseband system can be varied and the performance can be studied.

10.1 OVERALL SYSTEM RESPONSE

This measurement is done so as to find the overall response of the Baseband system and to verify the gain and other parameters. This is done using an Network Analyzer set in mixer measurement mode. The test set up is shown in Fig 10.1. The input frequency from the Analyzer is set to the IF frequency of 130 Mhz and 175 Mhz and wideband detector is selected. The output power is measured in each case and the response is plotted. The results are given in Fig 10.2.

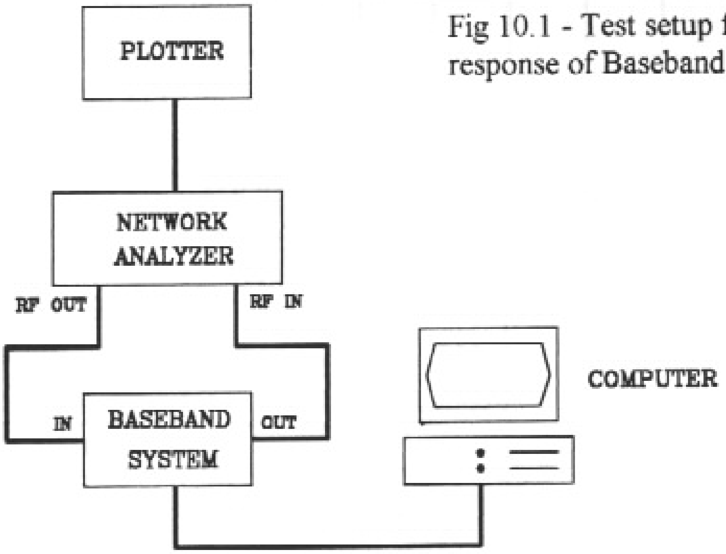


Fig 10.1 - Test setup for measuring overall response of Baseband System

The response of the system can be calculated from these measurements, also the image rejection of the overall system and also the gain are checked. These measurements are done with the ALC in the off mode.

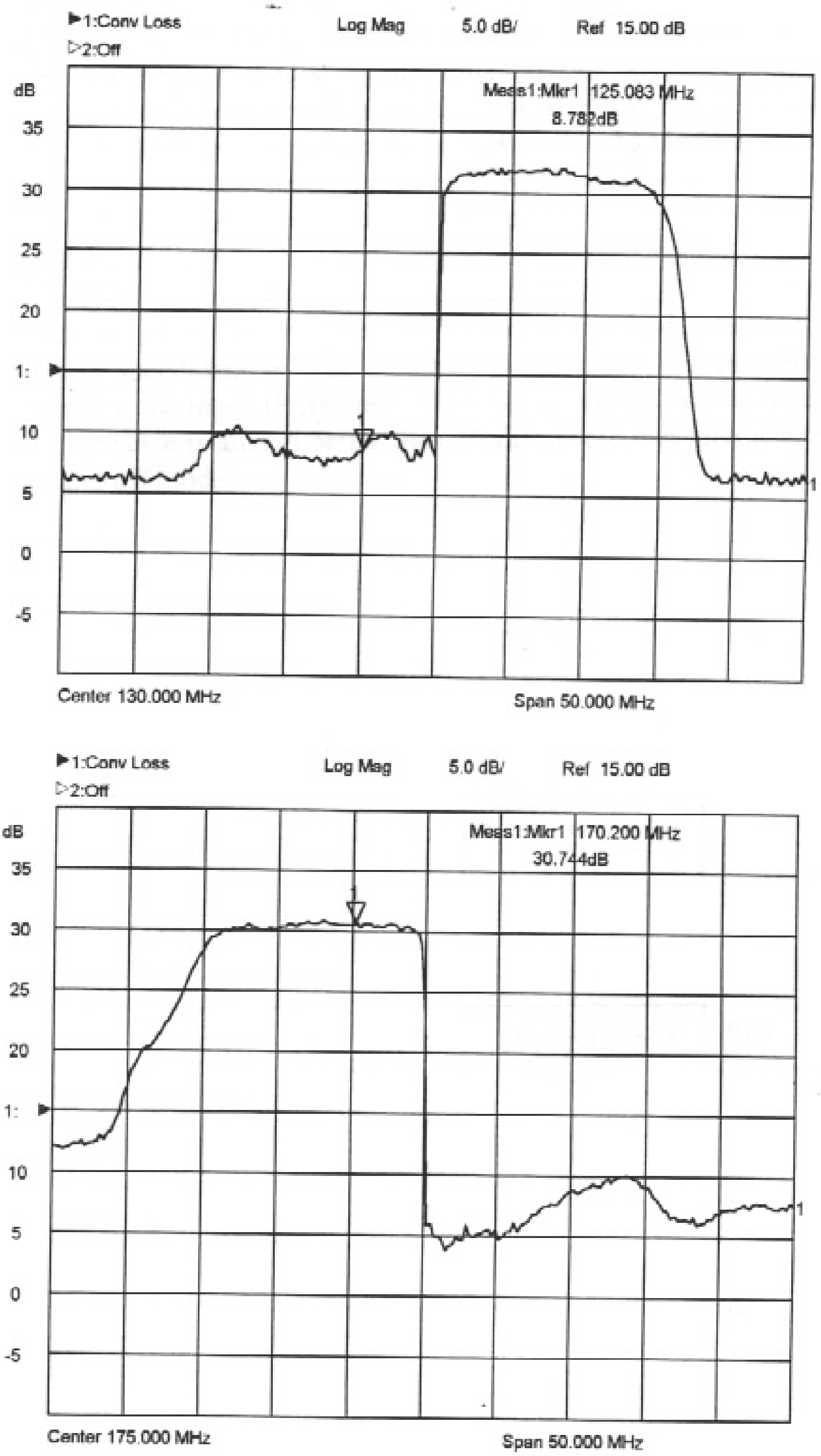


Fig 10.2 - Response of Baseband System, 130 LSB (above), 175 LSB (below)

10.2 HARMONIC PERFORMANCE

The harmonic performance of the system shows the non-linearity in the response of the circuits. The test setup for this measurement is shown in Fig 10.3. The signal generator used can generate a CW signal with adjustable power output. The signal generator feeds the Baseband system with a signal at the IF frequency, 180 Mhz with enough power so that the Baseband output power is 0 dBm. The bandwidth of the Baseband system is kept at 16 Mhz. The baseband output is then measured on a spectrum analyzer. The input from the signal generator will cause an output of 5 Mhz at the USB output with a power level of 0 dBm. Now in the spectrum analyzer the harmonics of this frequency are also seen. The spectrum seen in the analyzer is shown in Fig 10.4. It may be seen that the second harmonic at 10 Mhz is at least 37 dB below the fundamental. This is acceptable as per the system specifications. The harmonic performance is measured at various frequencies by changing the input IF frequency and measuring the output in each case. It is seen that the performance is always better than 35 dB. These measurements are repeated both in ALC on and off modes.

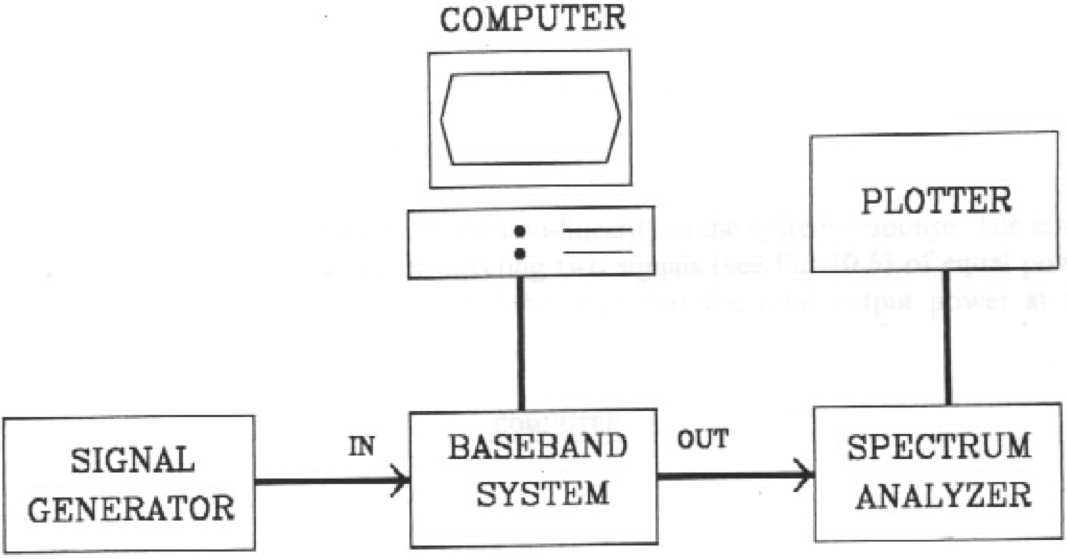


Fig 10.3 - Test setup for Harmonic measurements

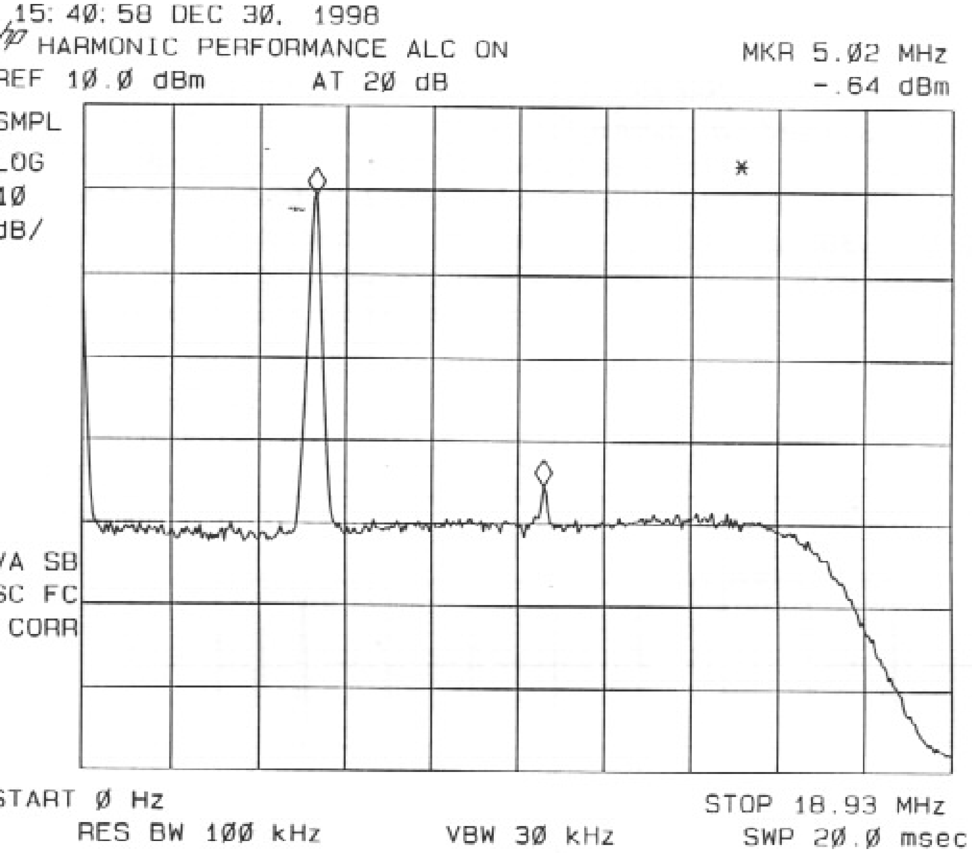


Fig 10.4 - Harmonic Levels at Baseband Output

10.3 INTERMODULATION DISTORTION

Intermodulation distortion is caused by the non-linearity in the system response. The effect in the Baseband system is studied by injecting two signals (see Fig 10.5) of equal power level at IF frequency, 178Mhz and 179 Mhz such that the total output power at the Baseband is 0 dBm.

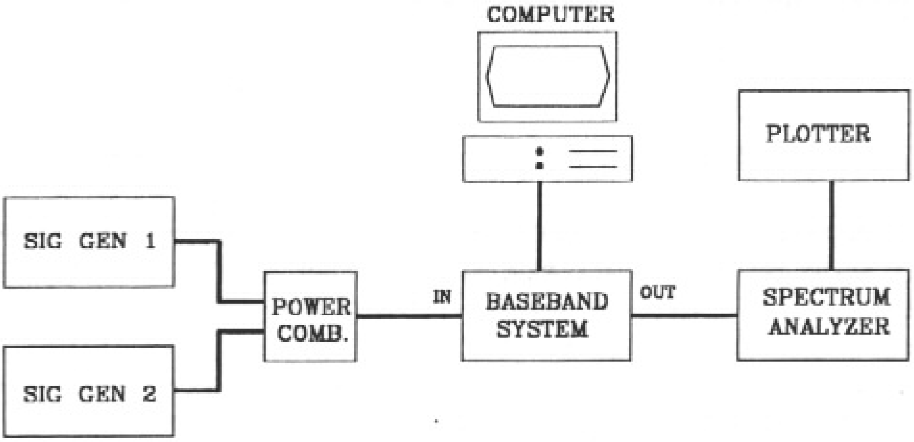


Fig 10.5 - Test setup for measuring the intermodulation distortion

The output spectrum is studied using a spectrum analyzer. The spectrum is plotted in Fig 10.6. The IF frequencies at the input will generate two baseband signals at 3 Mhz and 4 Mhz respectively. Apart from this one can see signals at 1 Mhz and at 7 Mhz generated due to intermodulation. It is seen that the power at 1 Mhz is at least 37 dB below the power at the fundamental and the 7 Mhz signal is atleast 32 dB below. The intermodulation performance is tested at different frequencies keeping the ALC on/off and is found to be within the acceptable limits.

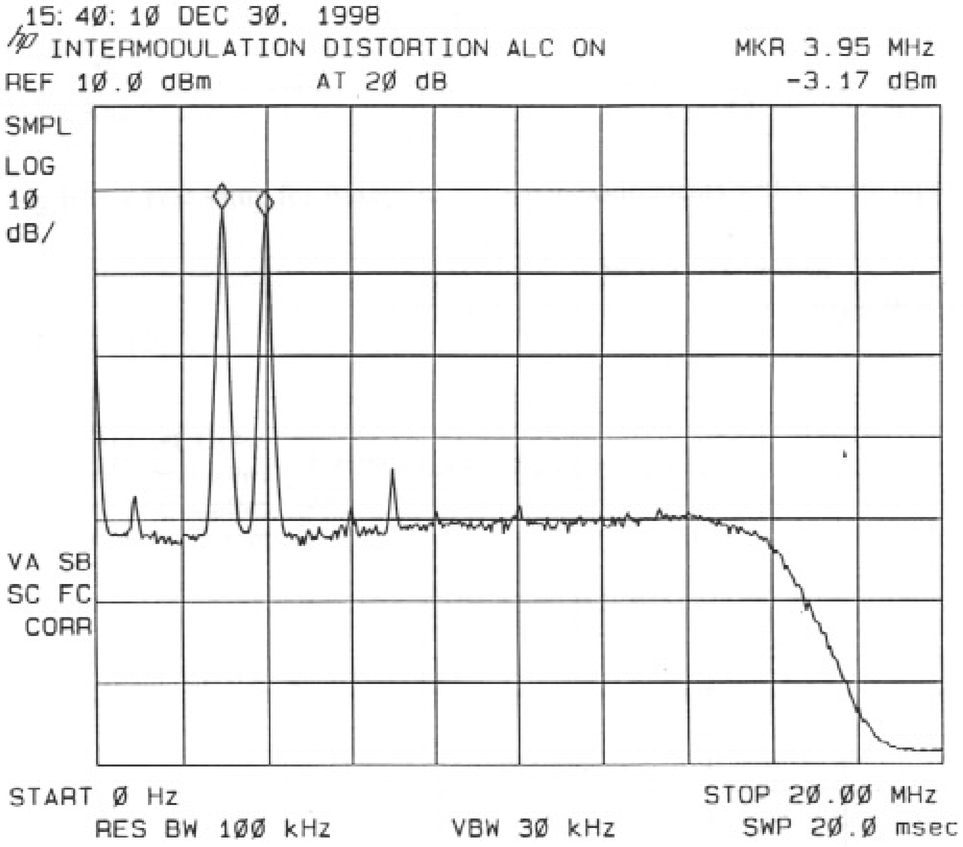


Fig 10.6 - Intermodulation Distortion at Baseband Output

10.4 MEASUREMENTS WITH THE NOISE GENERATOR

A wideband noise signal is used in these measurements to simulate the IF signal. The output Baseband signal is studied on a spectrum analyzer. The test setup is shown in Fig 10.7. The measurements are done with different bandwidth, and gain settings with the ALC in off mode. The performance of the ALC is studied by measuring the output spectrum with a normal noise given to the IF input and then by increasing the power by 8 dB. These measurements are done for ALC on and off modes. The results are plotted in Fig 10.8. The results show that in the ALC off mode the output power increases by the same amount as the increase in the input noise, whereas in the ALC on mode the output power remains more or less constant irrespective of the increase in the IF input power.

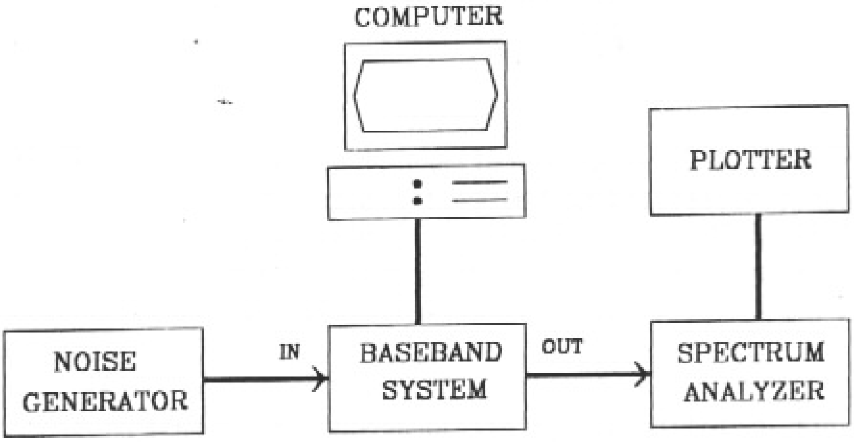


Fig 10.7 - Test setup for Baseband System measurements with a noise input

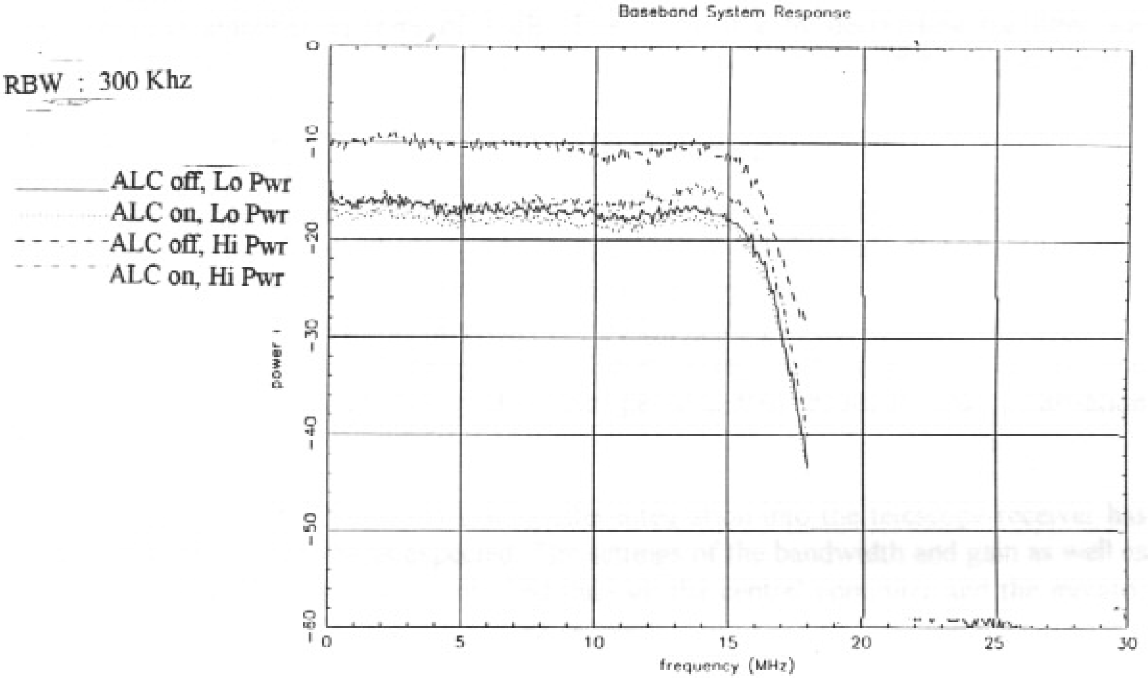


Fig 10.8 - ALC Response with Noise Input

All other parameters of the system are also measured and found to be as per the specifications. The stability of the system is checked by using a noise input and then recording the output power level for a long period of time. This gives any variations in power level over a time period. The system is found to be stable.

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Chapter 11
SUMMARY

The Baseband system has been designed to meet the stringent requirements of the GMRT receiver system. The system specifications has been laid down so as to improve the overall performance of the receiver system and to provide extra facilities during observations. The system has been tested thoroughly for various parameters and is found to meet the specifications and hence integrated with the GMRT receiver and used in the regular observations.

The Baseband system adds many facilities in the receiver system. With the Baseband filter the user has a wide choice of bandwidths possible, the setting can be decided based on the observational requirements. Bandwidth compensation is done to take care of power level variations, but the actual gain in the system can be varied if required using the bandwidth compensation amplifier in steps of 3 dB. This is possible by decoupling the filter and compensation gain settings in the software. The Baseband ALC has been designed so that it does not affect the normal observations as the time constant is designed to be 1 sec. Also the ALC on/off allows pulsar observations without any difficulty.

Another facility introduced in the baseband is variation of Baseband LO frequency in steps of 100 Hz. By varying this frequency one can scan over the available IF band. Using this LO along with the other local oscillators in the system any frequency range in the available RF band can be observed. The step size of 100 Hz is very useful as the main LO at the antenna base has a step of 1 Mhz upto 500 Mhz and then a step of 5 Mhz.The baseband system provides facilities for having different operating frequencies for the polarisation channels.

The performance of the Baseband system after integration into the telescope receiver has been tested and found to be as expected. The settings of the bandwidth and gain as well as the baseband LO frequency are controlled through the central computer and the monitor of various prameters is also working as per design.

A few additions will improve the usability of the Baseband system further. These include a computer selectable variable attenuator at the input of the first IF conversion circuit so as to adjust for any variation in gain of optical fiber or other circuits in the antenna electronics. This attenuator may be adjusted so as to maintain the Baseband ALC operating point at 2 dB above the knee irrespective of gain variations in the preceeding systems. Many monitoring points are provided to measure the voltage/current in various units of the system and the measured data is available at the central computer. A program is to be developed to read these and directly indicate the health of each unit, this will simplify troubleshooting and the down time in case of any maintenance work.

*** *** ***

APPENDIX - A

Performance Characteristics of Important Devices Used

AD 5539 Details

GBW MHz	Slew V/µS	FPR MHz	Noise nV/Hz$^{1/2}$	CMRR dB	PSRR µV/V	Z_{in} Ω	Z_{out} Ω	O/P swing V
1200	600	48	4	80	200	10	100	-2.2 to +2.7

FPR - Pull Power Response

SD 5002 Details

I/P Range V	Threshold Max V	Cross Talk dB	On Resistance Ω	Source Capacitance pF
± 7.5	2.0	-107	30	3.5

HOS 200 Details

Freq Range MHz	Slew Rate V/µS	Output Swing V	Prop Delay nS	Harmonic Dist %
0 to 200	1500	4.25	1.5	0.1

PSC2-1 Details

Freq Range (Mhz)	Isolation (dB)	Insertion Loss (dB)	Amp Unbalance (dB)	Phase Unbalance (deg)
0.1 - 400	25	0.4	0.2	3

PSC5-1 Details

Freq Range (Mhz)	Isolation (dB)	Insertion Loss (dB)	Amp Unbalance (dB)	Phase Unbalance (deg)
1 - 300	23	0.6	0.3	4

PSC12-1 Details

Freq Range (Mhz)	Isolation (dB)	Insertion Loss (dB)	Amp Unbalance (dB)	Phase Unbalance (deg)
1 - 200	27	0.8	0.4	4

PSCQ2-90 Details

Freq Range (Mhz)	Isolation (dB)	Insertion Loss (dB)	Amp Unbalance (dB)	Phase Unbalance (deg)
55 - 90	30	0.3	1.2	3

SCM-1NL Details

Freq Range (MHz)	Conversion Loss (dB)	LO-RF Isolation (dB)	LO-IF Isolation (dB)
LO/RF 1-500 IF DC-500	5.72	45	45

MAR-3 Details

Freq Range (Mhz)	Gain (dB)	Output 1dB (dBm)	NF (dB)	IMD O/P (dBm)	VSWR	DC Voltage (V)	DC Current (mA)
DC-2000	12.5	+10.0	6.0	+23.0	1.5	5.0	35

SAW Filter Details

	Centre Freq MHz	Insertion Loss dB	Delay µS
FBA 059	130	28.12	0.583
FBA 060	175	26.32	0.518

TOH 54 Details

Freq GHz	VSWR	Loss dB	Isolation dB	Power Watts	Switch Time ms	Coil R Ω	Coil V Volts	Coil Current mA
dc-2.5	1.2	0.11	45	20	20	60	12	230

ZHL 5W-1 Details

Freq Range (Mhz)	Gain (dB)	Output 1dB (dBm)	NF (dB)	IMD O/P (dBm)	VSWR	DC Voltage (V)	DC Current (A)
1-500	40	+37.0	8.0	+49.0	2.5	24.0	3.3

Notes :
Refer to the data manuals of the corresponding devices for detailed information about the characteristics and the test conditions.

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