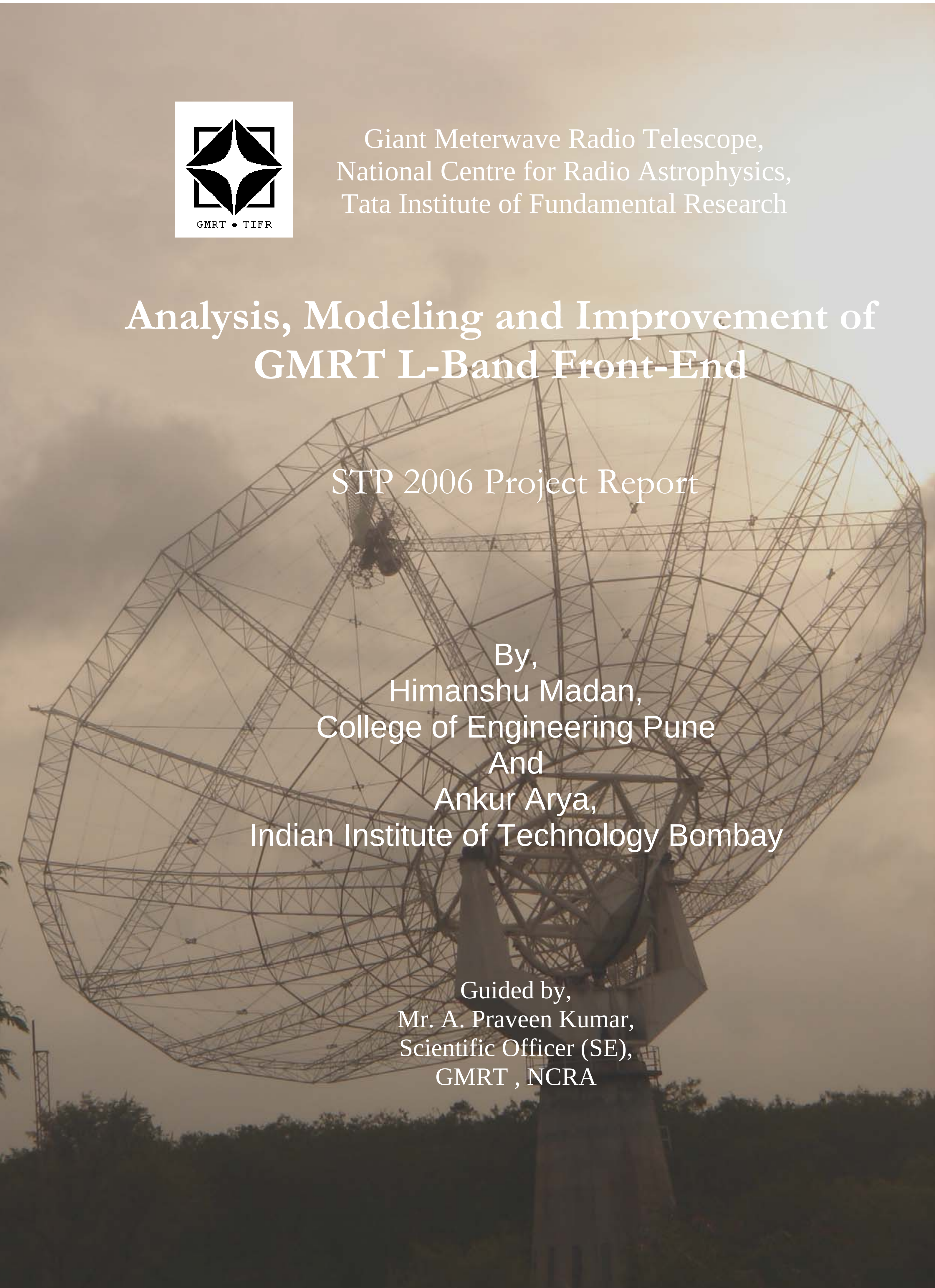

Giant Meterwave Radio Telescope,
National Centre for Radio Astrophysics,
Tata Institute of Fundamental Research

Analysis, Modeling and Improvement of GMRT L-Band Front-End

STP 2006 Project Report

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Ankur Arya
and
Himanshu Madan

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Chapter One

Introduction

1.1 GMRT Overview

The GMRT receiver chain is shown schematically in Figure 1.1. The Ist block is the multi-frequency front-end. It currently operates at five frequency bands centered at 150 MHz, 235 MHz, 327 MHz, 610 MHz and L-Band extending from 1000 to 1450 MHz. The L-Band is split into four sub-bands centered at 1060 MHz, 1170 MHz, 1280 MHz and 1390 MHz, each with a bandwidth of 120 MHz. The 150 MHz, 235 MHz and 327 MHz bands have about 40 MHz bandwidth and the 610 MHz band has about 60 MHz bandwidth.

The low noise receiving system of GMRT has been designed to receive dual polarization. Lower frequency bands from 150 to 610 MHz have dual circular polarization channels (Right Hand Circular and Left Hand Circular polarization) which have been named as CH1 and CH2, respectively. The higher frequency L-Band has dual linear polarization channels (Vertical and Horizontal polarization) and they have been named CH1 and CH2 respectively.

The receiver system has flexibility to be configured for either dual polarization observation at a single frequency band or single polarization observation at two different frequency bands. The polarization channels can be swapped whenever required. For observing strong radio sources like "sun", the selectable solar attenuators of 14 dB, 33 dB or 44 dB can be used. The front-end has RF termination facility also. Any band of the receiver can be switched OFF, whenever not in use, with the RF on/off facility provided in the front end. The receiver can be calibrated by injecting one of the four

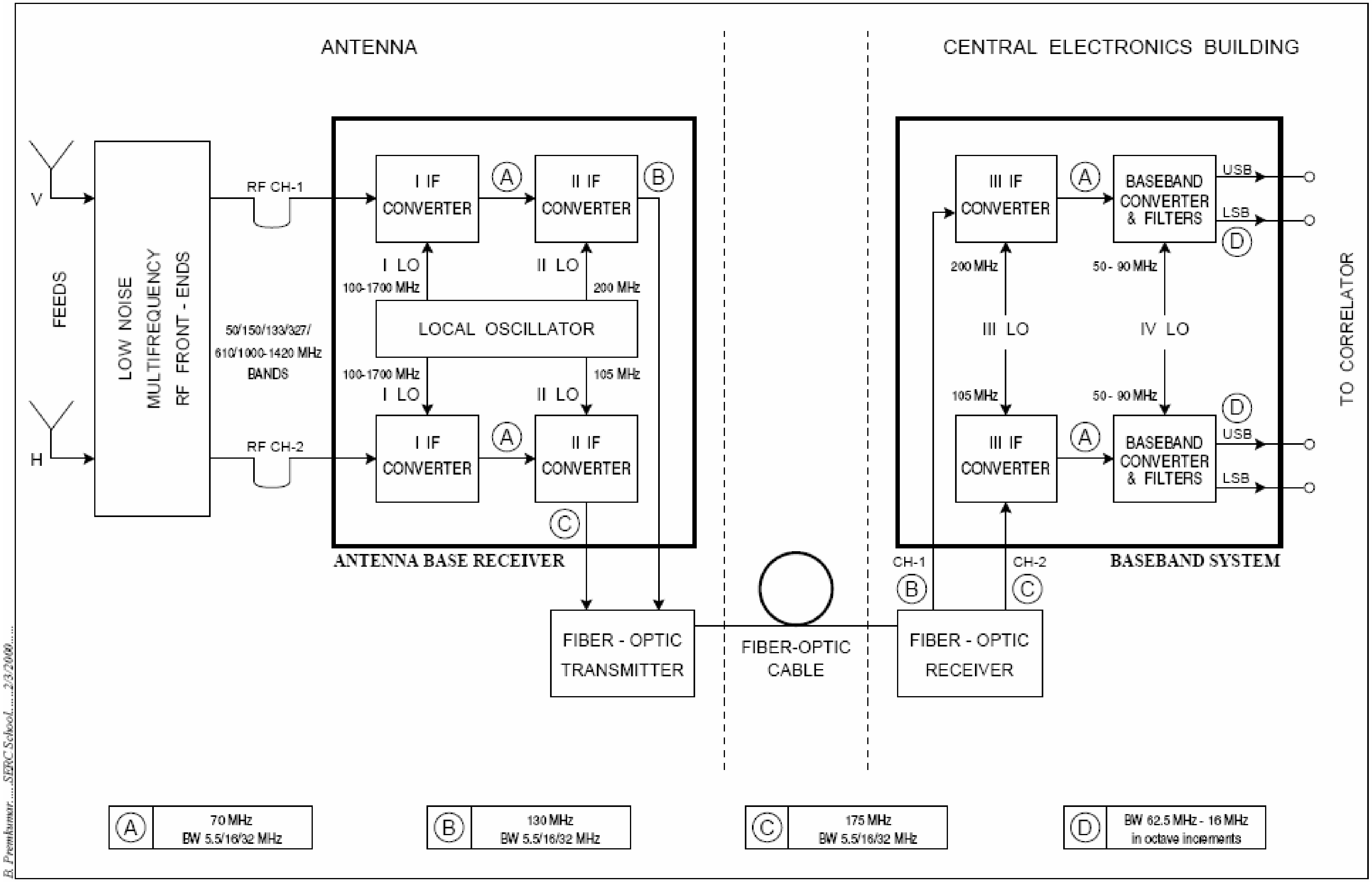

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Figure 1.1 Schematic Diagram of GMRT Receiver Chain levels of calibrated noise named Low cal, Medium cal, High cal and extra High cal depending upon the flux density of the source being observed. To minimize cross coupling between channels, a phase switching facility using Walsh functions is available at RF section of the receiver.

The first synthesized local oscillator converts the RF band to an IF band centered at 70 MHz. The synthesized local oscillator has a frequency range of 100 MHz to 1795 MHz. 100 MHz to 600 MHz is covered by synthesizer-1 and 605 MHz to 1795 MHz is covered by synthesizer-2. The local oscillator frequencies from 100 MHz to 354 MHz can be set with a step size of 1 MHz and the frequency range from 355 MHz to 1795 MHz can be set with a step size of 5 MHz. The IF bandwidths of either 5.5 MHz, 16 MHz or a full RF bandwidth of 32 MHz can be selected. The IF at 70 MHz is then translated to a second IF at 130 MHz and 175 MHz for CH1 and CH2 respectively. The maximum bandwidth available at this stage is 32 MHz for each channel. This frequency translation is done so that they can be transported to the Central Electronics Building (CEB) over a single fiber-optic cable. Two sets of 0 to 30 dB programmable attenuators are incorporated in the IF chain in each channel, which can be varied in steps of 2 dB. An automatic level control (ALC) facility is provided at the output stage of IF which can be bypassed whenever required (e.g. For Pulsar Observations).

The IF signals at 130 and 175 MHz along with telemetery and LO round trip phase carriers directly modulate a laser diode operating at 1300 nm wavelength which is coupled to a single mode fiber-optic cable link between the receiving antennas and the CEB. At the CEB, these signals are recovered with a PIN photodiode detector and suitably amplified. The 130 MHz and 175 MHz signals are then separated out and sent for baseband conversion. There is a monitor port available at the fiber-optic receiver front panel at CEB, where all the received signals can be monitored.

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

1.2 Scope

The concept of High Dynamic Range (HDR) receiver implies not only an ability to detect with low distortion but also the desired signal differing in amplitude by large amounts. The response of the entire system must remain linear over a wide range of noise temperature. More importantly the entire receiver system should remain linear even in presence of strong interference signals. The receiver must have a higher degree of immunity to spurious responses produced by non-linear interaction of multiple high level interfering signals.

The purpose of this report is to look at the present GMRT L-Band Front-end system and modify the same in order to improve the dynamic range using high dynamic range devices and change in configuration of the system. With these modifications we can expect an improvement of6 dB in Compression Dynamic Range (CDR) and 5 dB in Spurious Free Dynamic Range (SFDR). This proposal is aimed at improvements for sustaining the current phase of GMRT with the existing interference scenario.

This report also attempts to describe the nonlinear behaviour of L-Band Front-end approximately in polynomial expressions. The polynomial model referred aids in analysis of the device without actually measurement.

The tables containing the components incorporated with their specifications for the existing and proposed receiver configuration are included.

The existing and proposed designs of the L-Band front-end were simulated using MATLAB SIMULINK models. The simulated results have been included in the report. The results contain the simulated results of the dynamic range related parameters at every stage of the receiver chain.

The report also brings up the measurement of parameters of devices with examples. The various notions of nonlinearity and linearity measurement of devices are also covered in the report with a fair amount of details.

Chapter Two

L-Band Front End

The block diagram and schematic diagram of the L-Band front-end receiver are shown in figure 2.1 and 2.2, respectively. The L-Band front-end consists of corrugated horn feed to collect the radiations reflected from the parabolic dish with a quadridge orthomode transducer (OMT) in which the waveguide mode of the signal is converted into coaxial mode. In the OMT two linear components of the incoming signals are picked up in two perpendicular directions which are designated as V and H channels.

The signals are then amplified by a *low noise amplifier* (LNA). The LNA is designed using three stages of FUJITSU HEMT's FHX35LG. The gain of the LNA is $34.4 \pm 1dB$ over the frequency range of 1000 MHz to 1500MHz, with a noise temperature of $35 \pm 3K$ over the same frequency range.

The LNA is followed by the *post amplifier* consisting of a stage of Minicircuits MMIC amplifier MAR-3. The gain of the post amplifier is almost 8 dB in the frequency range.

The next stage of the amplifier is *phase switch* unit. The signals are phase modulated to 180 and 0 degrees with a Walsh function input, so that the common mode signals can be rejected. It consists of stages of MAR-3, 4 dB attenuator, Mini Circuits SRA-2010MH double balanced mixer and two stages of MAR-3, followed by a 6 dB attenuator. The gain of the phase switch is 11.5 dB.

The signals then pass through a set of switched filter bank where there is a provision to bypass the filters to get full band output. The switches used are SW338 and SW419, which have an insertion loss of 0.65dB and

1dB, respectively. In the switched filter mode one of the four sub-bands centered at 1060MHz, 1170MHz, 1280MHz and 1390MHz, each with a bandwidth of 120MHz can be selected. The insertion loss of the *narrow band-pass filter* is around 6dB. At the final output a 550 MHz bandwidth pass filter with the center frequency of 1250 MHz is incorporated. This *wideband BPF* has an insertion loss of 0.5dB.

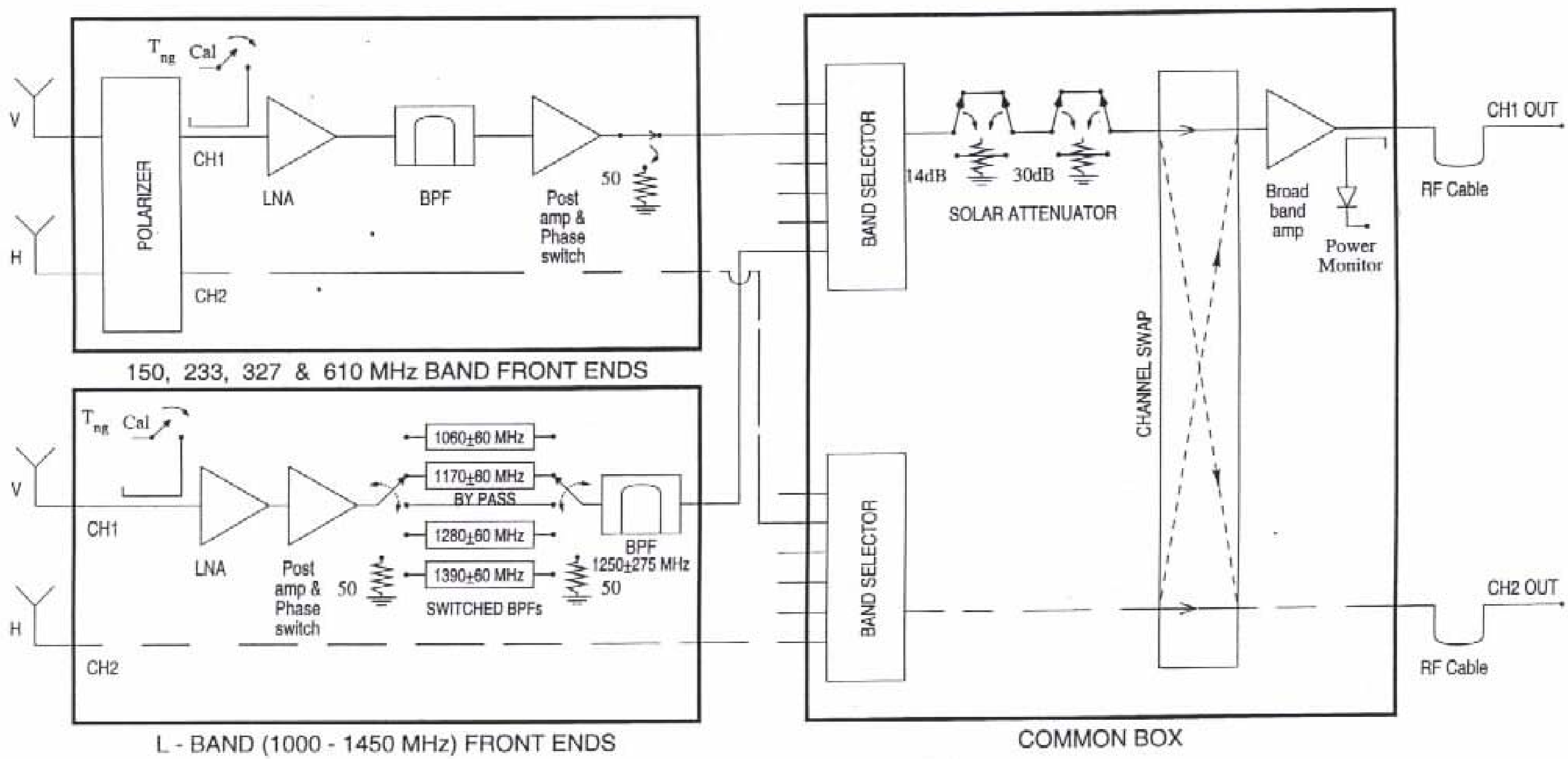

Figure 2.1 Block Diagram of L-Band Front-end receiver and Common Box

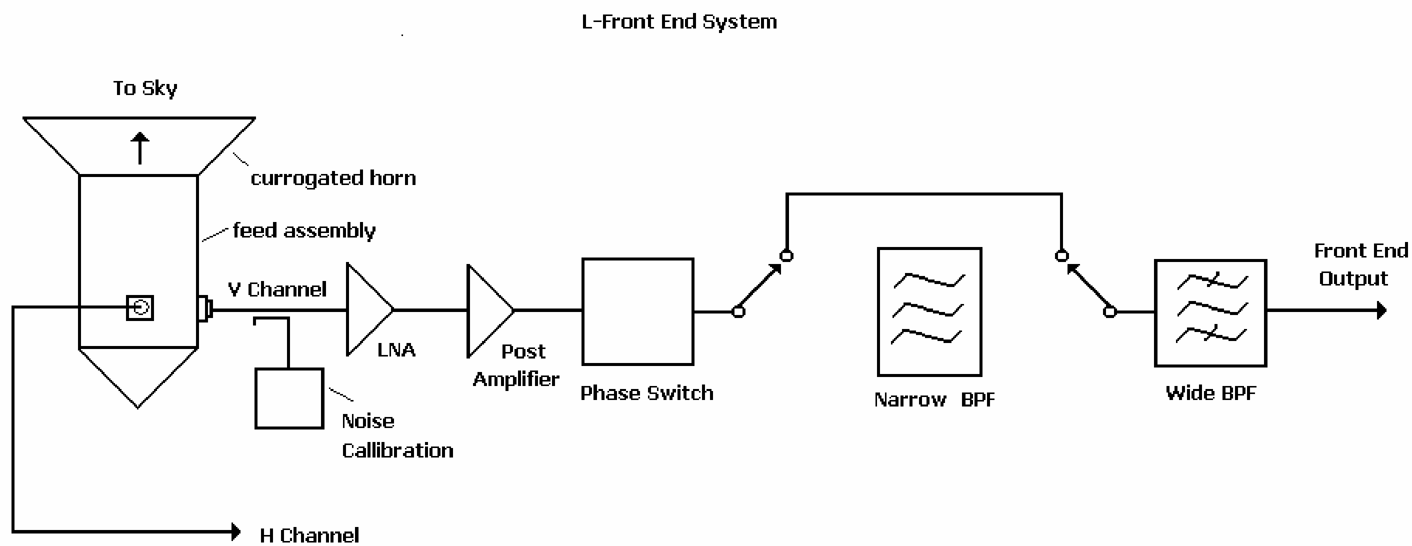

Figure 2. 2 Schematics of L-Band Front-End receiver

Chapter Three

Concepts in RF Receiver

3.1 Signal

The variations of the electric field, received from a source of radio waves can be called a signal. The signals from celestial sources are of interest to astronomers. These signals usually get coupled with other unwanted signals from receivers and other noise emitting sources.

3.2 Noise

Noise may be called as unwanted signals which are of no interest to astronomers. There are various sources of noise, they include thermal agitation, semiconductor noise, intermodulation noise and celestial and many man made noise sources.

3.3 Thermal Agitated Noise

The rapid and random motion of electrons in a resistive element is the source of noise generated in a component. The random thermal motion of electrons causes a current to flow in the resistor. On the average there are as many electrons moving in one direction as in the opposite direction, and the average current is zero. But the noise power is proportional to square of the current which is not zero. According to kinetic theory absolute temperature of the component is proportional to the rms value of the velocity of motion of the particles in the body. Thus it is apparent that the noise is also proportional to the absolute temperature of the device, in addition to the bandwidth over which it is measured. It can be well approximated by nyquist formula.

$$p_n = kT\delta f$$

where k is the Boltzmann Constant = $1.38 \times 10^{-23} JK^{-1}$ and δf is the bandwidth.

3.4 Signal to noise ratio

Signal to noise ratio is the parameter which tells about the usefulness of a signal. It is defined as the ratio of the desired signal power to the undesired noise power.

$$SNR = \frac{P_S}{P_n} \qquad \text{(Power in watts)}$$

3.5 Noise Figure

Noise Factor is defined for a device as the ratio of signal to noise ratio of the signal at the input to the signal to noise ratio of the signal at output.

$$F = \frac{SNR_{in}}{SNR_{out}}$$

An alternative definition of noise figure is that F is equal to the signal to noise ratio of an ideal system divided by the signal to noise ratio at the output of the device under test, both working at the same temperature over the same bandwidth and fed from the same source. The noise factor expressed in decibels is known as Noise Figure.

$$NF = 10 \log F$$

If a signal is input to an active RF device of gain *G*, the output noise power is actually a contribution from input noise signal and noise produced by the network.

$$N_{out} \neq GN_{in}$$

If we replace the device with an equivalent noise less device then the effective input noise power will be $(N_{in} + N_{ckt})$ where N_{in} is the noise power in the input signal and N_{ckt} is the noise generated by the RF device.

$$N_{out} = G(N_{in} + N_{ckt})$$

Mathematically

$$F = \frac{input\ \ s/n}{output\ \ s/n}$$

$$= \frac{P_{in}/N_{in}}{P_{out}/N_{out}}$$

$$= \frac{P_{in}/N_{in}}{GP_{in}/G(N_{in}+N_{ckt})}$$

$$= \frac{N_{in}+N_{ckt}}{N_{in}}$$

$$N_{ckt} = (F-1)N_{in}$$

Where,

F= noise figure

P_{in} = input signal power to the device

N_{in} = input noise power

P_{out} = output signal power

N_{out} = output noise power

G = gain of the device

N_{ckt} =noise generated in the device (or circuit)

The Noise Figure of an attenuating device equals the Gain (or Loss) of the device in magnitude. Hence an attenuator of gain -3dB has a noise figure of 3dB.

3.6 Noise Temperature

It is temperature at which a device must be assumed to be in thermal agitation for it to produce the same noise power as is actually produced by the device. Thus

$$T_A = \frac{noise\ \ power}{kB}$$

Considering the same example of the non ideal active RF device taken in the last section.

Let the input noise temperature be $T_{in}=P_{in}/kB$, *similarly* the device noise temperature be T_{ckt}.

So,

$$N_{in} + N_{ckt} = kB(T_{in} + T_{ckt})$$

or, $$FN_{in} = kB(T_{in} + T_{ckt})$$

$$FkBT_{in} = kB(T_{in} + T_{ckt})$$

thus, $$T_{ckt} = (F-1)T_{in}$$

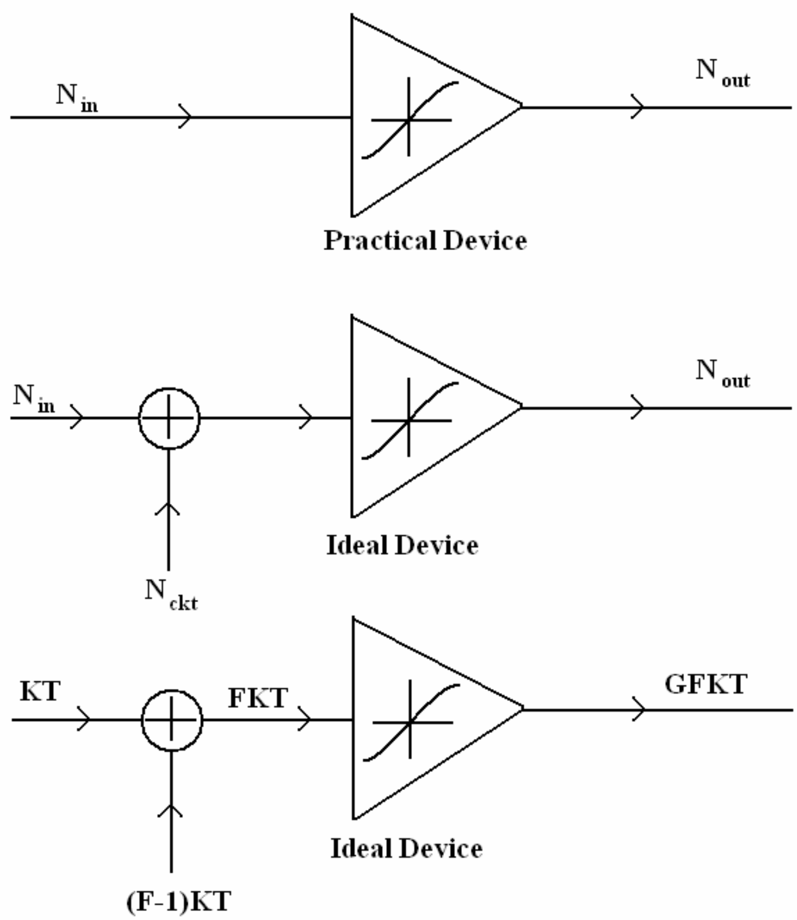

Figure 3.1 Illustration of the Noise Factor F, of the circuit

The Noise Figure (or noise factor) of any device is generally expressed for a reference temperature $T_o = 290K$. The T_{noise} is then

$$T_{noise} = (F-1)T_o \qquad \text{(Kelvin)}$$

In next sections, temperature T has been used as a source of input noise signal.

3.7 Minimum Discernible Signal (MDS)

The equivalent noise power present on the input to a receiver sets a lower limit on the smallest signal the system can detect. The smallest signal which can be detected is the *Minimum Detectable or Discernible Signal*. Its is hence the noise floor of the system. MDS is a measure of *sensitivity* of the system. In this report the terms used for MDS at the input terminal of the device is MDS_{in} and at the output of the device is MDS_{out}.

3.8 System Temperature T_{sys}

The system temperature T_{sys} for the antenna is the sum of the receiver temperature $T_{receiver}$ and the antenna temperature T_A. The system temperature is defined at the output terminal of the antenna which is same as the input terminal of LNA. The system temperature with input from an antenna looking in the sky includes sky temperature and temperature due to spill-over, temperature from ground and lossy elements in the feed path along with receiver temperature.

$$T_{sys} = T_{sky} + T_{ground} + T_{spill} + T_{loss} + T_{receiver} = T_A + T_{receiver}$$

Hence the total noise power incident on antenna is $P = kT_{sys}$ (per unit frequency) or equivalently the input noise level (dBm) will be $10\log_{10} kT_{sys}$ for 1 Hz bandwidth.

On using the T_{sys} for computing the noise power, the Noise Factor F is dropped from the expression as it includes contributing temperature from other sources. Table 3.1 shows the system temperature of various bands at the GMRT along with the bandwidth.

SYSTEM TEMPERATURES

Frequency Band [MHz]	Input Cable Loss L' [dB]	Polarizer Loss L [K]	LNA Temp. T_{LNA} [K]	Receiver Temp. (Includes cable losses) T_R [K]	Ground Temp. T_{Gnd} [K]	Sky Temp. T_{Sky} [K]	System Temp. T_{Sys} [K]	Bandwidth [MHz]
150	0.2	0.75[1]	150	260	12	308	**580**	40
235	0.55[2]	0.25	35	103	32	99	**234**	40
327	0.13	0.18	30	55	13	40	**108**	40
610	0.22[3]	0.15	30	59	32	10	**101**	60
1060	0.22[4]	—	35	53	25	5	**83**	120
1170	0.22[4]	—	32	49	24	4	**77**	120
1280	0.22[4]	—	30	47	23	4	**74**	120
1390	0.22[4]	—	28	45	23	4	**72**	120

1. Includes 2–1 combiner insertion loss.
2. Insertion loss of Balun & associated cables.
3. Insertion loss of Balun & 20 cm 0.141" semirigid cable from Balun to Probe.
4. Contains loss of OMT & OMT to LNA input cable.

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Table 3.1 System Temperatures for various frequency bands at GMRT

3.9 Gain for Cascaded Stages:

The gain of the cascade is simply the product of the gain of the individual stages.

$$G = G_1 G_2 G_3 ... G_{n-1} G_n$$

G_i is gain of the i^{th} stage in the cascade in linear units. The gains are added when expressed in decibels.

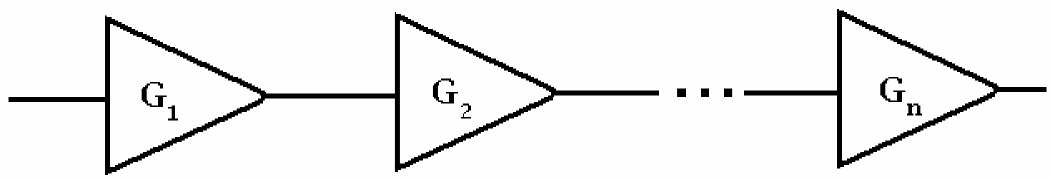


Figure 3.2 Gain of the cascade system

3.10 Noise Figure for Cascaded Stages:

Every stage in the cascade adds noise to its input signal; hence the noise figure of the cascade depends on the noise figures and gains of the constituent stages. A stage with gain G and noise factor F' referenced at temperature T adds a noise power of $(F'-1)\,kT$ to its input signal power. Thus in the cascade, a i^{th} stage adds a noise power of $(F_i-1)\,kT$, this is illustrated in figure 3.3. If the noises added in the device are referred back at the input of the first stage, the Noise Factor of the cascade then can be found to be,

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \frac{F_4 - 1}{G_1 G_2 G_3} + \cdots + \frac{F_n - 1}{G_1 G_2 G_3 ... G_{n-1}}$$

The noise figure NF is,

$$NF = 10\log_{10} F = 10\log_{10}\left(F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \frac{F_4 - 1}{G_1 G_2 G_3} + \cdots + \frac{F_n - 1}{G_1 G_2 G_3 \ldots G_{n-1}} \right)$$

3.11 Noise Temperature in cascade stages:

The equivalent noise temperature of the cascade can be obtained by substituting the Noise Figures of stages in by equivalent noise temperature of each stage in the above equation. The equivalent noise temperature for the cascade is

$$T_e = T_{e1} + \frac{T_{e2}}{G_1} + \frac{T_{e3}}{G_1 G_2} + \frac{T_{e4}}{G_1 G_2 G_3} + \cdots + \frac{T_{en}}{G_1 G_2 G_3 \ldots G_{n-1}}$$

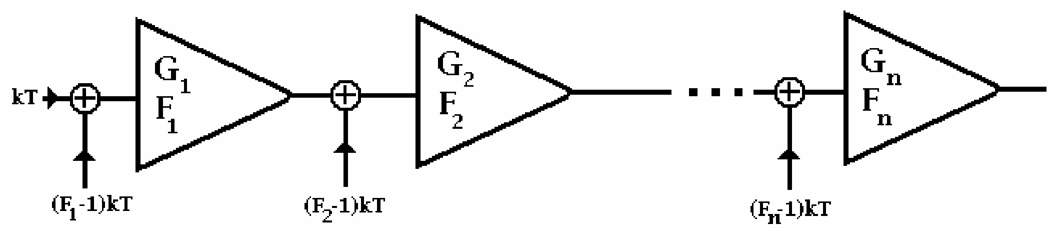


Figure 3.3 Noise Figure for Cascaded Stages

Chapter Four

Effects of Non-Linearity

While many analog and RF circuits can be approximated with a linear model; to obtain their response to small signals, nonlinearities often lead to interesting and important phenomena. For simplicity, we limit our analysis to memory-less, time invariant systems and we assume

$$y(t) \approx \alpha_1 x(t) + \alpha_2 x^2(t) + \alpha_2 x^3(t) . \qquad \text{……… (1)}$$

The effect of storage elements and higher order non linear terms must be carefully examined to ensure the above equation is plausible representation.

4.1 Harmonics

If a sinusoid signal is applied to a nonlinear system, the output generally exhibits frequency components that are integer multiples of the input frequency. In the equation.... (1),

If $x(t) = A\cos\omega t$ then,

$$y = \alpha_1 A\cos\omega t + \alpha_2 A^2 \cos^2 \omega t + \alpha_3 A^3 \cos^3 \omega t$$

$$= \alpha_1 A\cos\omega t + \frac{\alpha_2 A^2}{2}(1+\cos 2\omega t) + \frac{\alpha_3 A^3}{4}(3\cos\omega t + \cos 3\omega t)$$

$$= \frac{\alpha_2 A^2}{2} + (\alpha_1 A + \frac{3\alpha_3 A^3}{4})\cos\omega t + \frac{\alpha_2 A^2}{2}\cos 2\omega t + \frac{\alpha_3 A^3}{4}\cos 3\omega t \qquad \text{……. (2)}$$

In the above equation the term with the input frequency is called the “fundamental” and the higher-order terms the "harmonics".

Frequency	Amplitude
DC component	$\frac{\alpha_2 A^2}{2}$
Fundamental $:\omega$	$\alpha_1 A + \frac{3\alpha_3 A^3}{4}$
Second order $:2\omega$	$\frac{\alpha_2 A^2}{2}$
Third order $:3\omega$	$\frac{\alpha_3 A^3}{4}$

Table 4.1 Amplitudes of harmonics of ω frequency fundamental signal

For small values of A, amplitude of the *n*th order harmonics can be neglected as it consists of a term proportional to A^n and other terms proportional to higher power of *A*. In RF applications harmonics do not possess much threat as a frequency twice or thrice of the fundamental frequency will generally lies outside the pass band of the device.

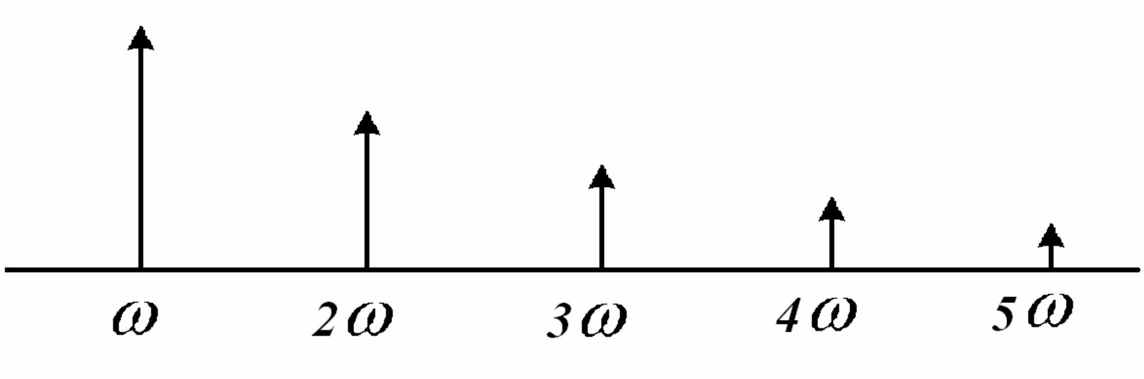


Figure 4.1 Harmonics of signal at frequency ω

4.2 Gain Compression

The small -signal gain of a circuit is usually obtained with the assumption that harmonics are negligible. However as the signal amplitude increases, the gain begins to vary. In fact, nonlinearity can be viewed as the variation of the small signal gain with the input level. This is evident from the term $3\alpha_3 A^2/4$ added to the term α_1 in equation (2). In most circuits, the output is a "compressive" or "saturating" function of the input; that is, the output signal approaches saturation for sufficiently high input levels. In equation (2) this occurs if $\alpha_3 < 0$. Written as $\alpha_1 + 3\alpha_3 A^2/4$, the gain is therefore a decreasing function of *A*.

In RF circuits, this effect is quantified by the "*1-dB compression point*", defined as the output signal level at which the gain drops by 1 dB. The gain is compressed by 1 dB at this point. Hence the P_{1dB} is the figure of merit which quantifies the upper limit to the input signal power.

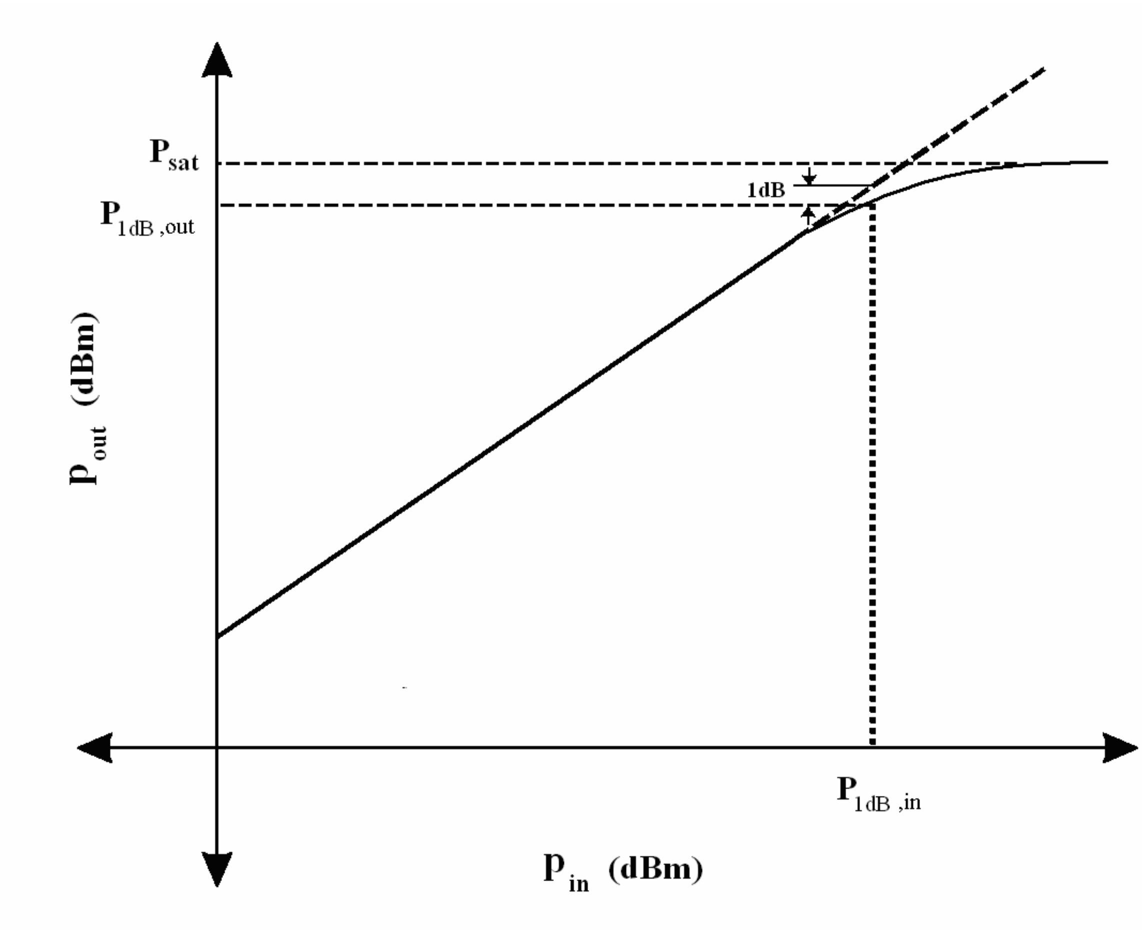


Figure 4.2 The 1dB compression point P_{1dB} ($P_{1dB,out}$) is the output power at which the output is 1dBm lesser than ideal output

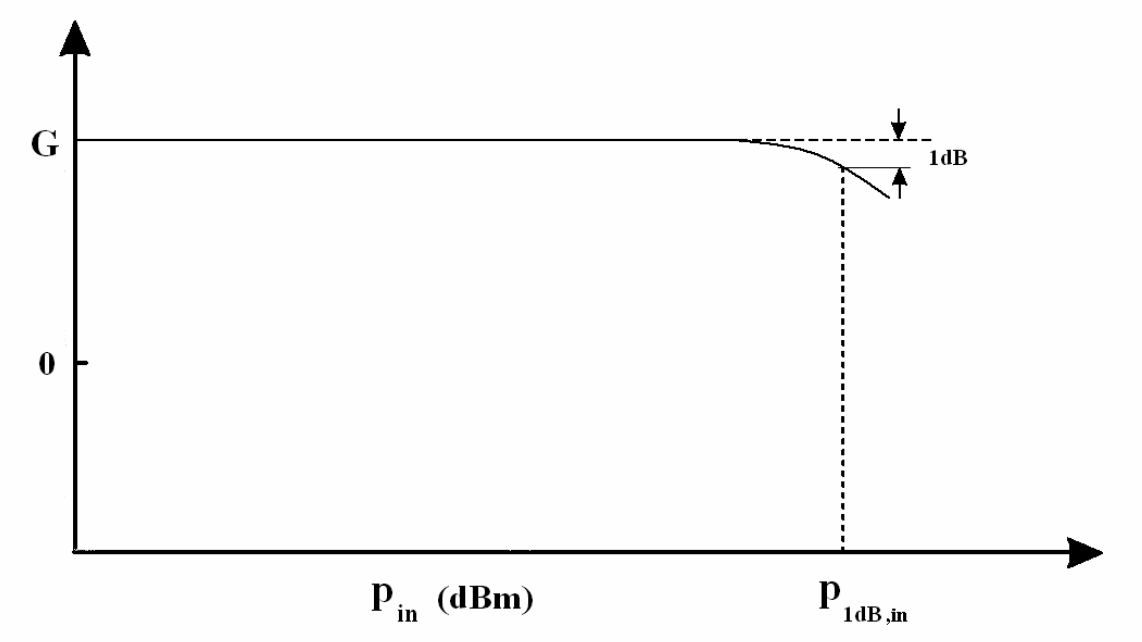


Figure 4.3 The gain falls by 1dB at 1dB compression point

For linear region,

$$P_{out} = P_{in} + G$$

for P_{in}=$P_{1dB,in}$,

$$P_{1dB} = P_{1dB,in} + G - 1 \text{ dBm}$$

and for $P_{in} > P_{1dB,in}$,

$$P_{out} \rightarrow P_{sat}$$

(P_{out},P_{in},$P_{1dB,in}$ and P_{sat} are in dBm and G is in dB)

Thus for input signal power below $P_{1dB,in}$ the device shows a linear response. P_{1dB} can be determined by feeding a single frequency signal in the device and measuring the gain for varying amplitude of the signal.

4.3 Desensitization and Blocking

In circuits showing compressive gain characteristics, a large signal tends to reduce the average gain of the circuit, the weak signal may experience a smaller gain.

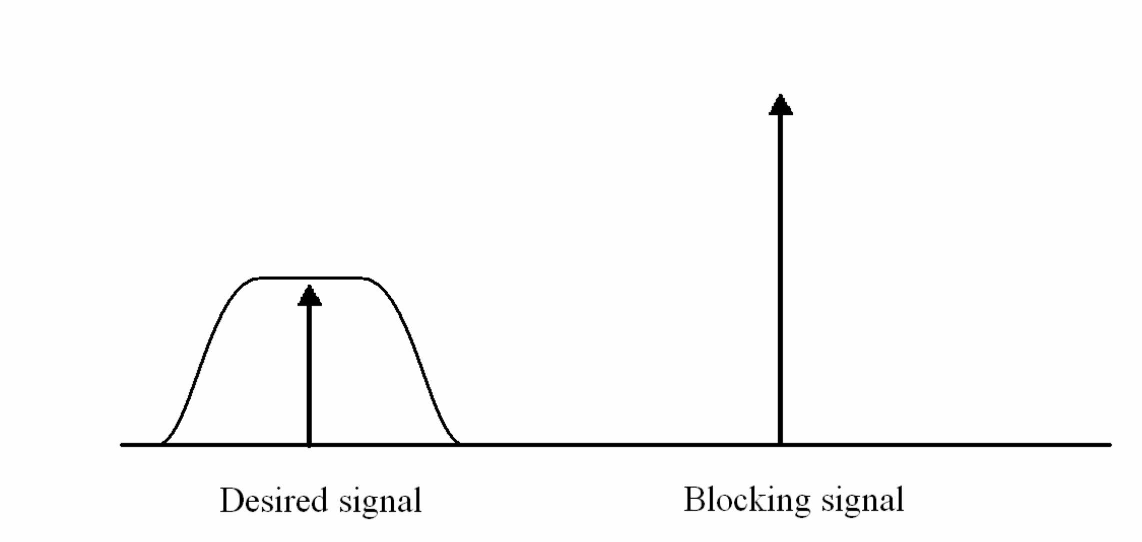


Figure 4.4 Illustration for Large signal blocking the desired weak signal

This effect is called desensitization, this effect can be analyzed for the characteristics of equation (1) by assuming

$$x(t) = A_1 \cos \omega_1 t + A_2 \cos \omega_2 t \qquad\qquad A_1 << A_2$$

where $A_1 \cos \omega_1 t$ is the desired signal and $A_2 \cos \omega_2 t$ is the strong interfering signal. The output is

$$y(t) = (\alpha_1 A_1 + \frac{3}{4}\alpha_3 A_1^3 + \frac{3}{2}\alpha_3 A_1 A_2^2)\cos \omega_1 t + ...$$

this reduces to

$$y(t) = (\alpha_1 + \frac{3}{2}\alpha_3 A_2^2) A_1 \cos \omega_1 t + ... \qquad\qquad A_1 << A_2$$

Thus, the gain for the desired signal is equal to $(\alpha_1 + \frac{3}{2}\alpha_3 A_2^2)$, a decreasing function of A_2 if $\alpha_3 < 0$ (i.e. the gain is compressed for large signals). For sufficiently large A_2, the gain drops to zero, and we say the signal is "blocked". The term “blocking signal” usually refers to interferers that desensitize a circuit.

4.4 Cross Modulation

Another phenomenon that occurs when a weak signal and a strong interferer pass through a nonlinear system is the transfer of modulation (or noise) on the amplitude of the weak signal. Called “cross modulation”, this effect is evident from the equation (3), where the variation in A_2 affect the amplitude of the output component at ω_1.

For example, if the amplitude of the interferer is modulated by sinusoid $A_2(1+m\cos\omega_m t)\cos\omega_2 t$, where m is the modulation index and less than unity, then equation (3) assumes the following form:

$$y(t)=\left[\alpha_1 A_1+\frac{3}{2}\alpha_3 A_1 A_2^2\left(1+\frac{m^2}{2}+\frac{m^2}{2}\cos 2\omega_m t+2m\cos\omega_m t\right)\right]\cos\omega_1 t+......$$

Thus the desired signal at the output contains the amplitude modulation at ω_m and $2\,\omega_m$.

A common case of cross modulation arises in amplifiers that must simultaneously process many independent signal channels. Nonlinearities in the amplifier corrupt each signal with the amplitude variation in other channels.

4.5 Intermodulation

While harmonic distortion is often used to describe nonlinearities of analog circuits, certain cases require other measures of nonlinear behavior. When two signals with different frequencies are applied to a nonlinear system, the output in general exhibits some components that are not the harmonics of the input frequencies. Called “intermodulation”, this phenomenon arises from “mixing” (multiplication) of the two signals when their sum is raised to a power greater than unity.

To understand how equation (1) leads to intermodulation, assume

$x(t)=A_1\cos\omega_1 t+A_2\cos\omega_2 t$.

Thus

$$y(t)=\alpha_1(A_1\cos\omega_1 t+A_2\cos\omega_2 t)+\alpha_2(A_1\cos\omega_1 t+A_2\cos\omega_2 t)^2+\alpha_3(A_1\cos\omega_1 t+A_2\cos\omega_2 t)^3$$

Expanding the left side and discarding dc terms and harmonics, we obtain the following intermodulation products:

$$\omega=\omega_1\pm\omega_2:\alpha_2A_1A_2\cos(\omega_1+\omega_2)t+\alpha_2A_1A_2\cos(\omega_1-\omega_2)t$$

$$=2\omega_1\pm\omega_2:\frac{3\alpha_3A_1^2A_2}{4}\cos(2\omega_1+\omega_2)t+\frac{3\alpha_3A_1^2A_2}{4}\cos(2\omega_1-\omega_2)t$$

$$=2\omega_2\pm\omega_1:\frac{3\alpha_3A_1A_2^2}{4}\cos(2\omega_2+\omega_1)t+\frac{3\alpha_3A_1A_2^2}{4}\cos(2\omega_2-\omega_1)t$$

And these fundamental components

$$\omega=\omega_1,\omega_2:(\alpha_1A_1+\frac{3}{4}\alpha_3A_1^3+\frac{3}{4}\alpha_3A_1A_2^2)\cos\omega_1 t+(\alpha_1A_2+\frac{3}{4}\alpha_3A_2^3+\frac{3}{4}\alpha_3A_2A_1^2)\cos\omega_2 t$$

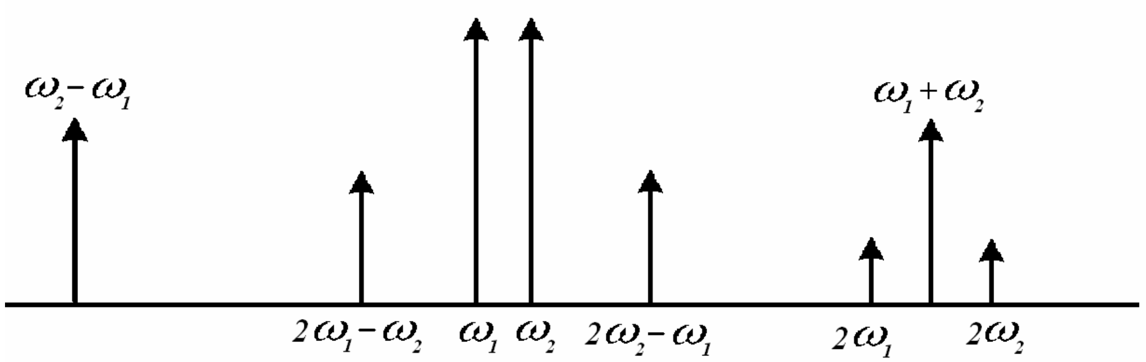


Figure 4.5 Harmonics and intermodulation products for two tone signal input

Of particular interest are the third order IM products at $2\omega_1-\omega_2$ and $2\omega_2-\omega_1$, illustrated in the fig….. The key point here is that if the difference between ω_1 and ω_2 is small, the components at $2\omega_1-\omega_2$ and $2\omega_2-\omega_1$ appear in the vicinity of ω_1 and ω_2, which may appear in the pass band of the device

Frequency	Amplitude
ω_1	$\alpha_1 A_1 + \frac{3}{4}\alpha_3 A_1^3 + \frac{3}{4}\alpha_3 A_1 A_2^2$
ω_2	$\alpha_1 A_2 + \frac{3}{4}\alpha_3 A_2^3 + \frac{3}{4}\alpha_3 A_2 A_1^2$
$\omega_1 \pm \omega_2$	$\alpha_2 A_1 A_2, \alpha_2 A_1 A_2$
$2\omega_1 \pm \omega_2$	$\frac{3\alpha_3 A_1^2 A_2}{4}, \frac{3\alpha_3 A_1^2 A_2}{4}$
$2\omega_2 \pm \omega_1$	$\frac{3\alpha_3 A_1 A_2^2}{4}, \frac{3\alpha_3 A_1 A_2^2}{4}$

Table 4.2 The amplitudes of the non linear products produced by two tone signal input

.

Signals of frequencies $f_0, f_1, f_2 ...$ mix and yield components at frequencies given by

$$f_{IM} = \sum_{k=0}^{\infty} m_k f_k \quad , \; m_k \in Z$$

The intermodulation product is of order n if

$$n = \sum_{k=0}^{\infty} |m_k|$$

Thus in dual tone test, signals of frequencies f_1 and f_2 produce components as tabulated below:

Order (n)	Products	
1	f_1, f_2	fundamental
2	$2f_1, 2f_2$	harmonics
	$f_1 + f_2$, $f_2 - f_1$	IMD
3	$3f_1, 3f_2,$	harmonics
	$2f_1 \pm f_2,\ 2f_2 \pm f_1,$	IMD
..	..	

Table 4.3 Various non linear products and the frequencies at which they are produced

The resulting products at frequencies near the fundamentals are illustrates in the figure 4.6.

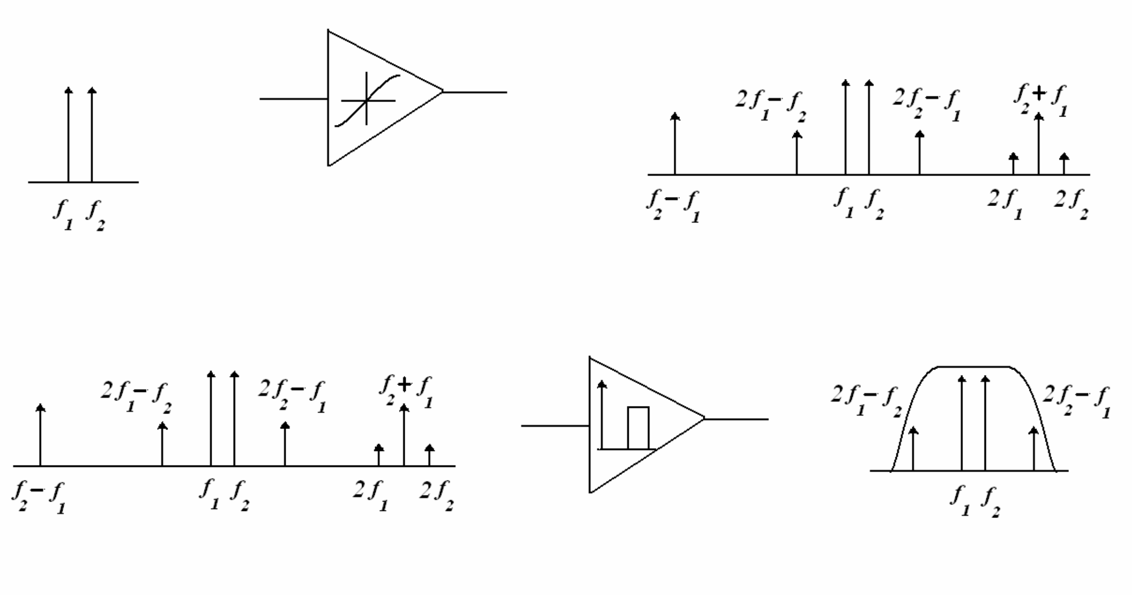


Figure 4.6 Third order IMD components occur at frequencies close to desired ones and are difficult to filter.

The products at frequencies $f_1 \pm f_2$ generally lie out side the pass band. Products at frequencies $2f_1 - f_2$ and $2f_2 - f_1$ are closer to fundamental frequencies f_1 and f_2 and hence may appear in the pass band. These intermodulated products are of primary interest as they are relatively larger in magnitude and difficult to filter. Higher order products have lower magnitude and are neglected for simplicity.

Chapter Five

Measurement of Linearity

Active RF devices show nonlinear responses for large power signals. For lower power signals the gain of the device remains almost constant, as the power of input signal is increased the output power begins to roll off and the device is said to have saturated. For large amplitude signals, the spurious signals also appear at the output owing to intermodulation of signals. The lower limit to the power of input signal is set by the minimum discernible signal (MDS) which is generally chosen to be the Noise floor or a few dBms above Noise floor. Thus these devices behave linearly for a particular range of input frequency. The measure of linearity is given in dynamic ranges of frequency of signals in terms such as *Compression Dynamic Range (CDR) and Spurious Free Dynamic Range (SFDR)*.

5.1 1dB Compression Point P_{1dB} :

When the gain drops off by 1dB on increasing the power of the signal, the power in dBm of the output signal is called as 1 dB Compression point. For large power input signal the loss in the device increases. Thus for output signal power below P_{1dB} the device shows a linear response. P_{1dB} can be determined by feeding a single frequency signal in the device and measuring the gain for varying amplitude of the signal. The output signal as observed in spectrum analyzer is shown in figure 5.1, showing the frequency and power of output signal.

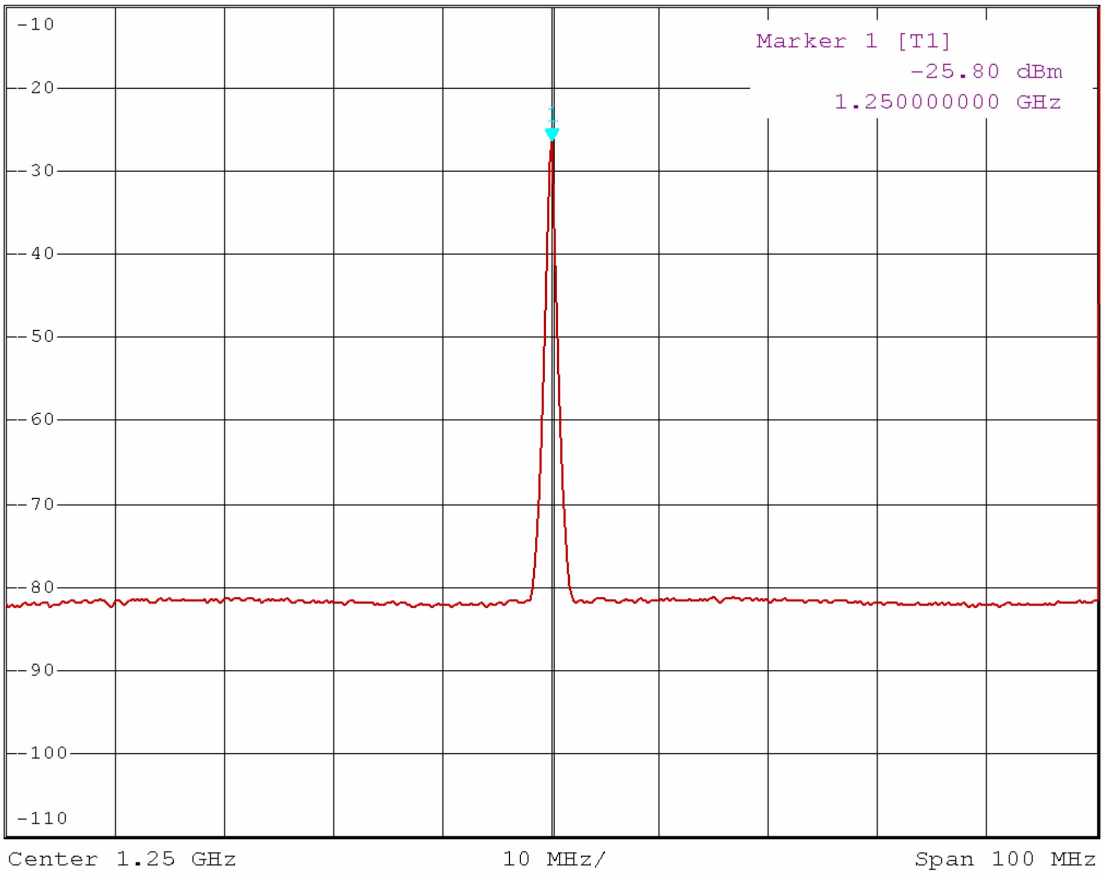


Figure 5.1 Spectrum analyzer Display of a signal at 1.25GHz

5.2 Compression Dynamic Range:

Compression Dynamic Range (CDR) is the range of signals with power level from MDS to P_{1dB}. The device has a constant gain in CDR except near P_{1dB} where it falls by 1dB. Compression dynamic range is a useful measure of linearity of device which is characterized by a constant gain. The CDR in dB is given by

$$CDR = P_{1dB} - MDS_{OUT}$$

$$= P_{1dB} - NoiseFloor_{out}$$

or,

$$CDR = P_{1dB} - 10\log_{10}(kTBFG)$$

$$CDR = P_{1dB} - 10\log_{10}(kT) - 10\log_{10} B - 10\log_{10} F - 10\log_{10} G$$

for T=290K ,

$$CDR = P_{1dB} + 174 - 10\log_{10} BG - NF \qquad \text{(in dB)}$$

or,

$$CDR = \frac{P_{1dB}}{kTBFG} \qquad \text{(in linear scale)}$$

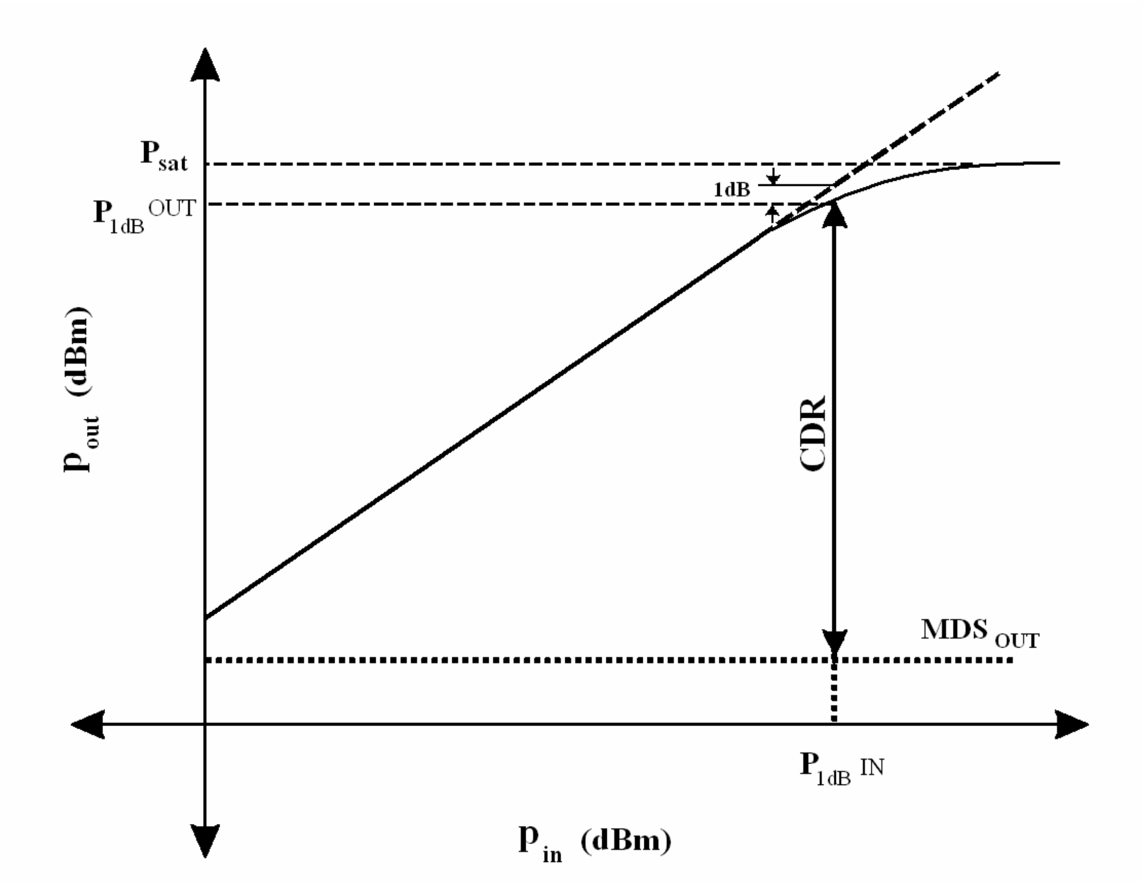


Figure 5.2 Graphical representation of CDR

5.3 Intercept Point:

Intercept point (IP) is used as a figure of merit for the intermodulated (IMD) product suppression. A high intercept point indicates high suppression of IMD products. A third order intercept point (IP3) is theoretical point at which the desired output signal power equals the third order IMD product power. The output power in dBm at IP3 is denoted as OIP_3 and input power at that point is IIP_3. These are shown in figure 5.3. For a system with third order nonlinearity the OIP_3 is generally 10 to 15 dBm above P_{1dB}.

The measurement of the IIP3 described is two tone test, which provides sufficiently accurate results as compared to multitone approximation [1]. To measure IP3, two input signals of equal power P_{IN} and offset by a small frequency range is input to the device, the power level of one of the linear response P_{OUT} and the third order IMD product P_{IM}, are measured. These are shown in figure 5.4. The IP3 points can then be determined as follows

Consider fundamental power line,

$$P_{OUT} = P_{IN} + G \qquad \ldots(1)$$

$$OIP_3 = IIP_3 + G \qquad \ldots(2)$$

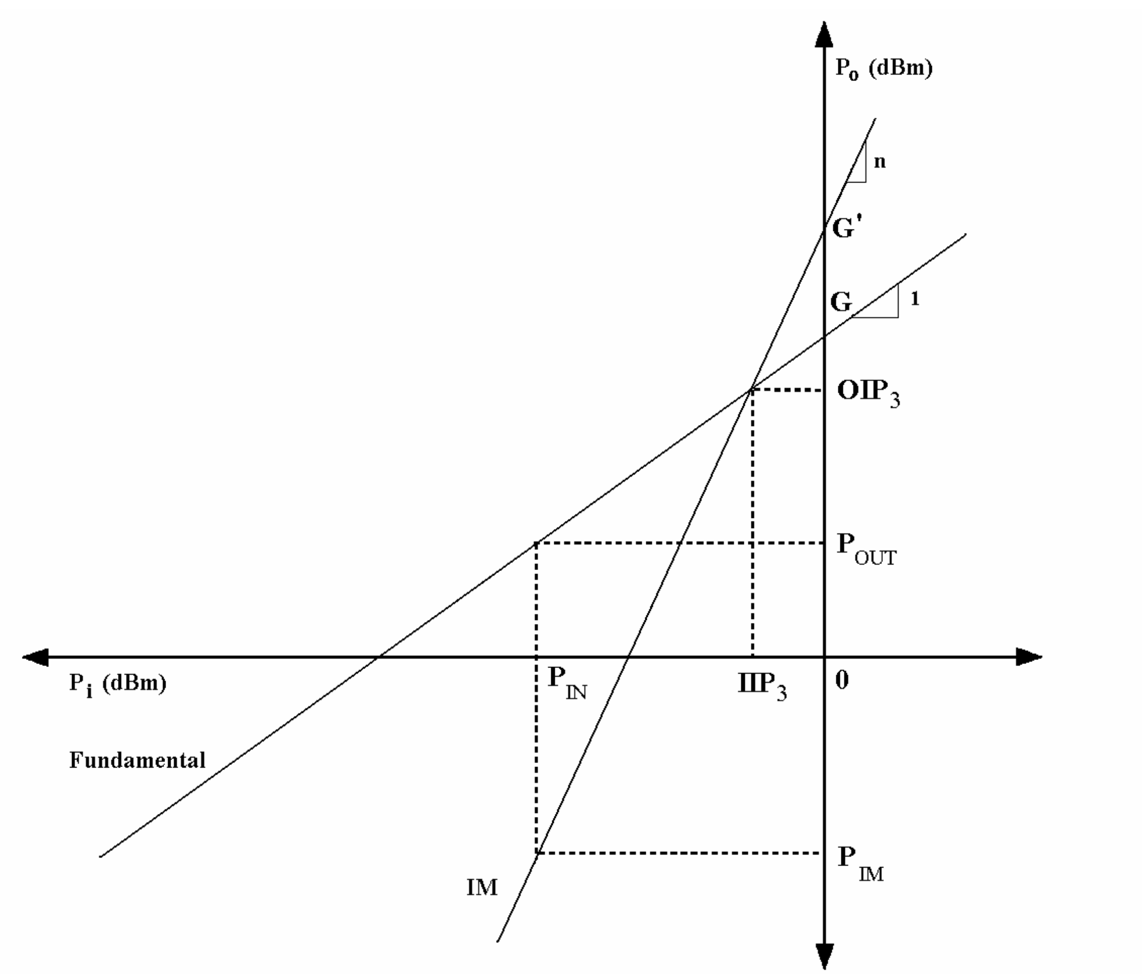


Figure 5.3 Graphical representation of nth order intercept points

and IM power line,

$$P_{IM} = nP_{IN} + G' \qquad \ldots(3)$$

$$OIP_3 = nIIP_3 + G' \qquad \ldots(4)$$

from (1) and (3),

$$G' - G = (n-1)P_{IN} + P_{OUT} - P_{IM}$$

from (2) and (4) ,

$$G' - G = (n-1)IIP_3$$

The above two equations yield,

$$IIP_3 = P_{IN} + \frac{1}{n-1}(P_{OUT} - P_{IM})$$

or, $$OIP_3 = P_{OUT} + \frac{1}{n-1}(P_{OUT} - P_{IM})$$

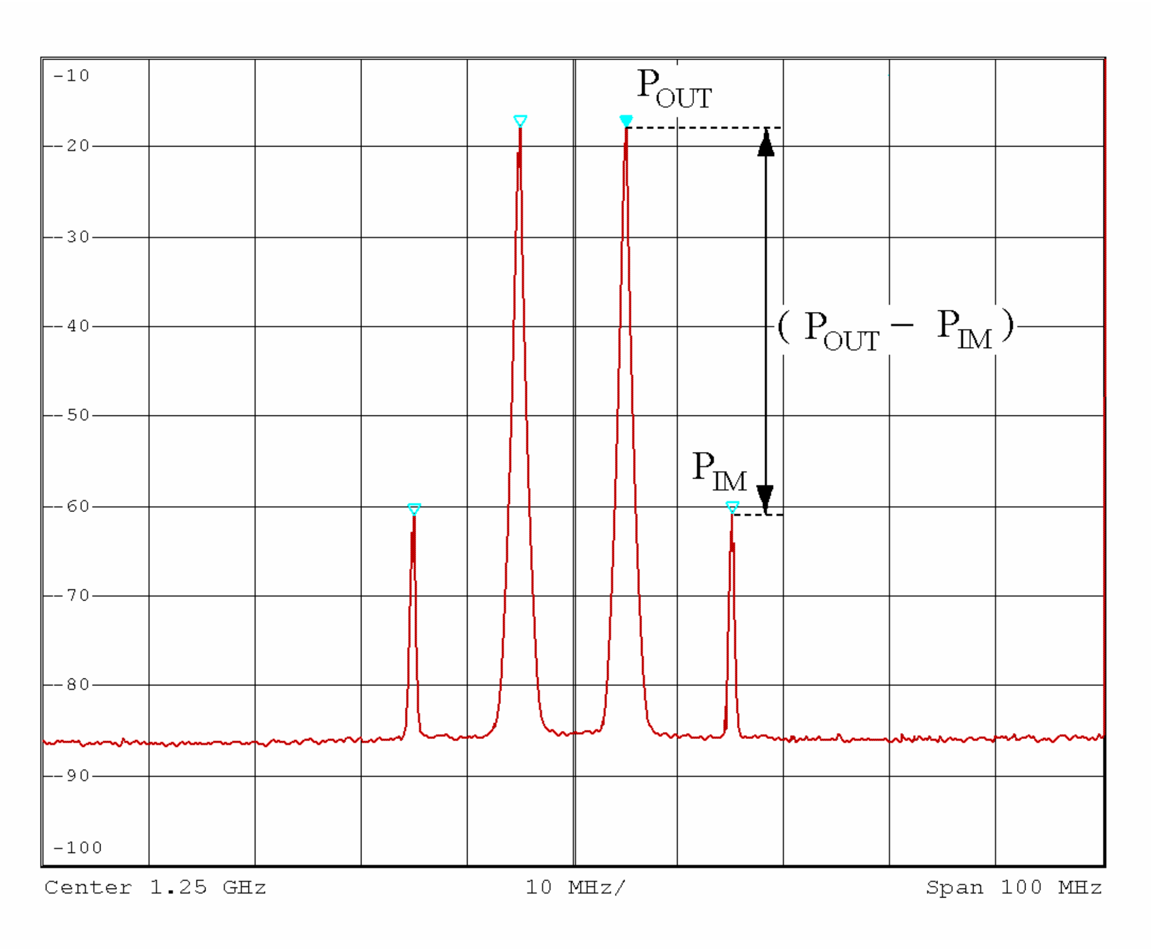


Figure 5.4 The fundamental and third order IMD products displayed in spectrum analyzer

Consider third order intermodulation or $n = 3$,

$$IIP_3 = P_{IN} + \frac{1}{2}(P_{OUT} - P_{IM})$$

$$OIP_3 = P_{OUT} + \frac{1}{2}(P_{OUT} - P_{IM}) \qquad \text{(all in dBm)}$$

Note: The Gain G (and G') is in dB in this section, However it was in linear scale in the previous section

5.4 Spurious Free Dynamic Range :

SFDR is that range of input power for which the fundamental output is obtained in the presence of noise but without non-linear interference. Third order IMD products are often used instead of 1dB compression point to impose an upper bound on Spurious Free Dynamic Range (SFDR). Lower bound is taken to be the receiver noise floor or MDS level. As the power of fundamental tone is increased the power of IMD products also increases. At the upper bound, the power of IMD product equals the MDS level. For input signals larger than this upper bound, the IMD products appear distinctively from noise along with fundamental response at the output. Thus a larger SFDR is desired for any receiver.

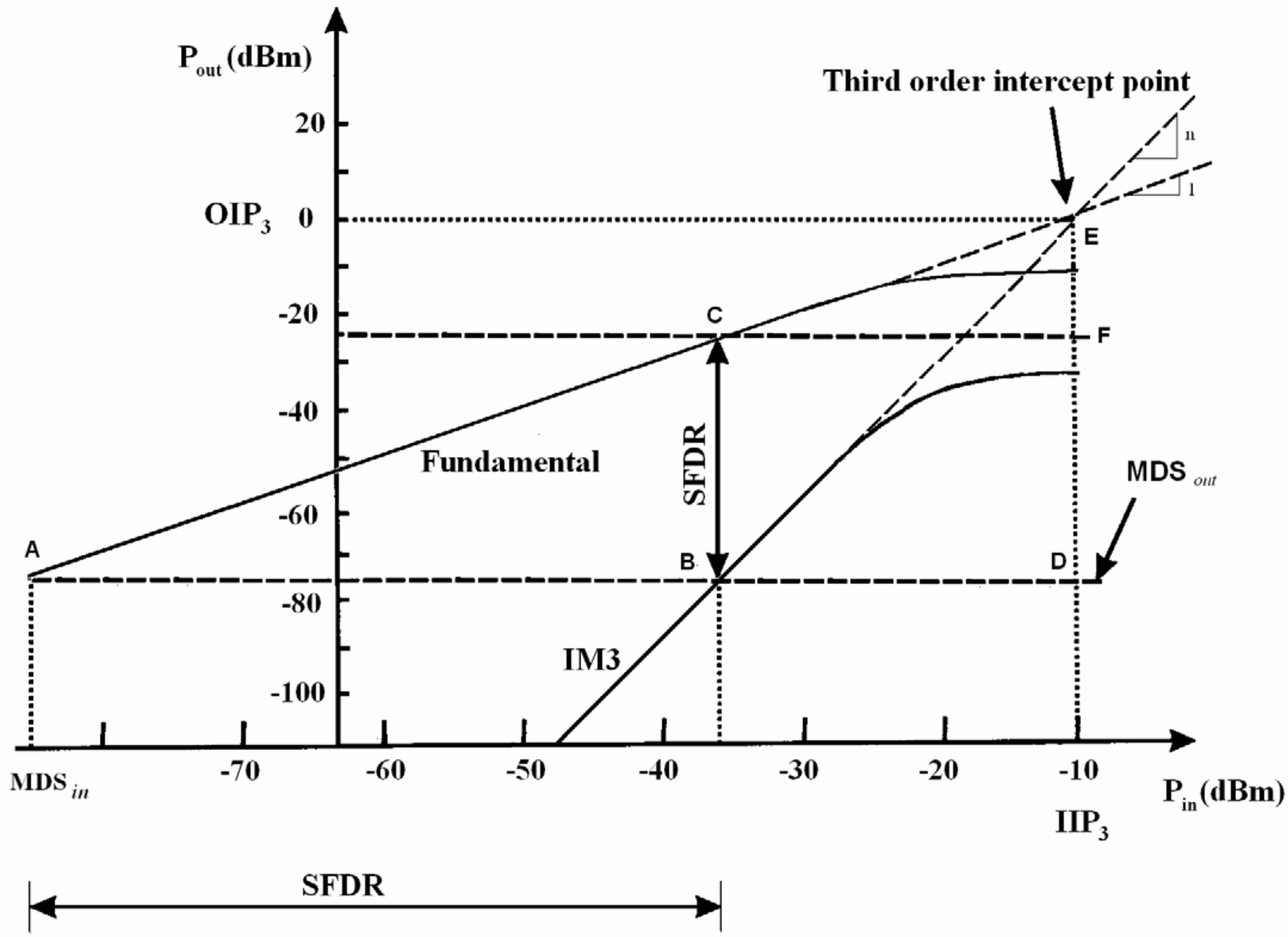


Figure 5.5 Graphical representation of Spurious Free Dynamic Range SFDR

The power of fundamental and nth order intermodulated products at the output of a device are plotted as a function of power of input signal in figure __ .s The tangential line to the fundamental (linear) response and non-linear response when extrapolated intersect at IP (point E). At output level at A is the minimum discernible signal level (MDS_{out}) which is usually the noise floor at output. The power of third order IMD products equals MDS_{out} at B. The nth order intercept point is at E. The SFDR is then the range of input power for A to B or equivalently the output power for B to C. The expression for SFDR (in dBms) can be deduced as shown.

$In\Delta BDE,$

$$ED = nBD$$

$$ED = DF + FE$$

or , $$ED = BC + BD$$

$$nBD = BC + BD$$

$$BC = (n-1)BD$$

$$BC = \frac{n-1}{n} DE$$

thus, $$SFDR = BC = AB = \frac{n-1}{n} DE$$

for $n=3$ i.e. third order intermodulation (IMD),

$$SFDR = \frac{2}{3}(IIP_3 - MDS_{IN}) = \frac{2}{3}(OIP_3 - MDS_{OUT}) \qquad \textit{(in log scale)}$$

If $MDS_{OUT} = 10\log_{10} kTBFG$ (Noise floor) , *then*

$$SFDR = \frac{2}{3}(IIP_3 - 10\log_{10} kTBF) = \frac{2}{3}(OIP_3 - 10\log_{10} kTBFG) \qquad \text{(in log scale)}$$

or,

$$SFDR = \left(\frac{P_{I,IP3}}{kTBF}\right)^{2/3} = \left(\frac{P_{O,IP3}}{kTBFG}\right)^{2/3} \qquad \textit{(in linear scale)}$$

Note: In this section, G has no units in linear scale and denotes gain.

The third order intermodulation products increase by 3dB for every 1dB increase in two input signals. For every 1dB reduction in output power, there is a 3dB reduction in the third order intermodulation products. The result is a 2dB improvement in SFDR.
SFDR overlooks several important factor given below, that influence the dynamic range.

1. It attempts to model interference by just using two interfering signals , but real signal environment is usually populated of multiple signals.

2. It does not reveal the effect of reciprocal mixing or compression range like Desensitization Dynamic Range.
3. It does not effectively test the effects of receiver input filtering.
4. It considers only the third order distortion.

Although it has such limitations, SFDR has become a very popular specification used in comparing the overall dynamic range performance of the competing receivers.

5.5 Desensitization Dynamic Range (DDR)

The measurement of degradation effect to a large out of band blocking signal is DDR. A large signal produces some components at the desired frequency and pushes the system near saturation, thus reducing the average gain of the system. The weak signal experiences a vary small gain.

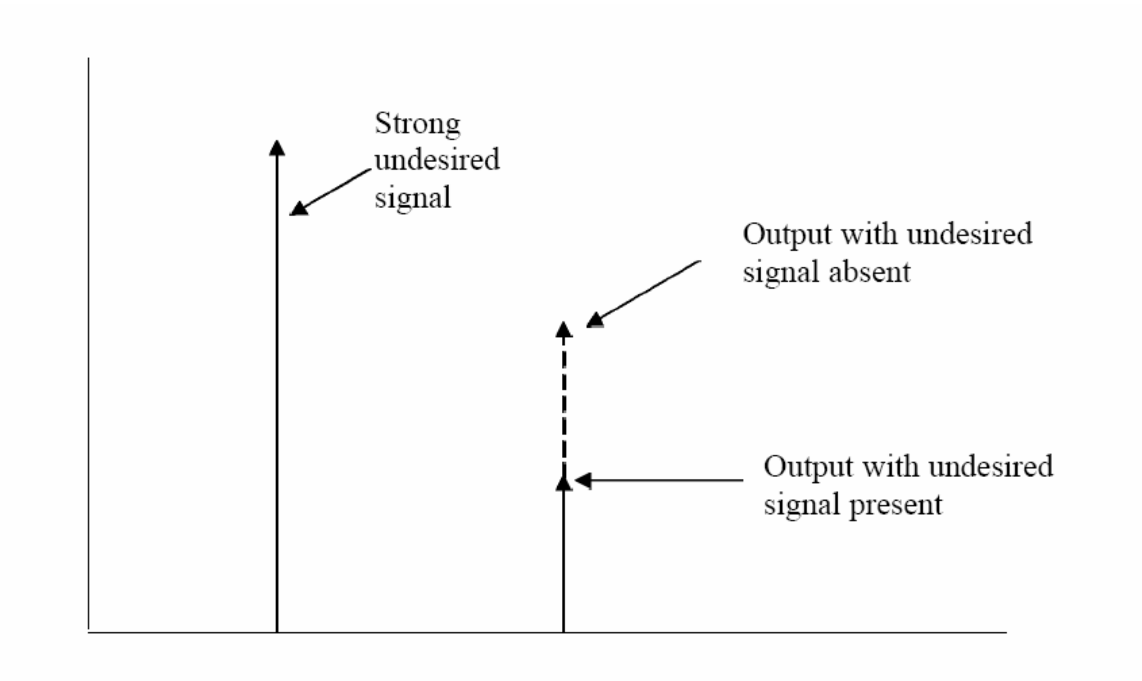


Figure 5.6 Large blocking signal and desired weak signal in frequency domain

To measure the DDR, a signal that produces an output SNR (signal to noise ratio) of 10 dB is fed at the receiver input and interfering sinusoidal signal is added to the input at a particular frequency offset from the desired frequency and its magnitude is increased until the output SNR degrades by 1dB. DDR is then the power ratio in dB of the undesired signal power in dBm to the receiver noise floor in dBm/Hz. DDR is expressed as,

$$DDR = P_{in} + 174 - NF$$

where, P_{in} is the interfering signal power in dBm, NF is the noise figure in dB

The DDR is strongly affected by the frequency offset to the interfering signal. At small frequency offsets, DDR is dominated by effects of phase noise and reciprocal mixing. At larger offsets, 1dB compression due to overload may occur.

5.6 Examples of measurements for LNA :

For measurement of P_{1dB} a single frequency signal at 1250MHz from a signal generator is input to the LNA and the power of the output signal is displayed in a spectrum analyzer. The amplitude of the input signal is varied and the corresponding output power is recorded till the output power saturates. The values measured are given in table 5.1 .

input power (dBm)	output power (dBm)
-72.9	-38.5
-70.9	-36.5
-68.9	-34.5
-66.9	-32.4
-64.9	-30.5
-62.9	-28.5
-60.9	-26.4
-59.9	-25.4
-58.9	-24.4
-57.9	-23.4
-56.9	-22.4
-55.9	-21.5
-54.9	-20.5
-53.9	-19.5
-52.9	-18.5
-51.9	-17.5
-50.9	-16.5
-49.9	-15.5
-48.9	-14.5
-47.9	-13.5
-46.9	-12.5
-45.9	-11.5
-44.9	-10.6
-43.9	-9.6
-42.9	-8.7
-41.9	-7.7
-40.9	-6.7
-39.9	-5.8
-38.9	-5
-37.9	-4.1
-36.9	-3.3
-35.9	-2.5
-34.9	-1.8
-33.9	-1
-32.9	-0.3
-31.9	0.3
-30.9	0.8

Table 5.1 The output signal power with power input signal 1250MHz to the LNA

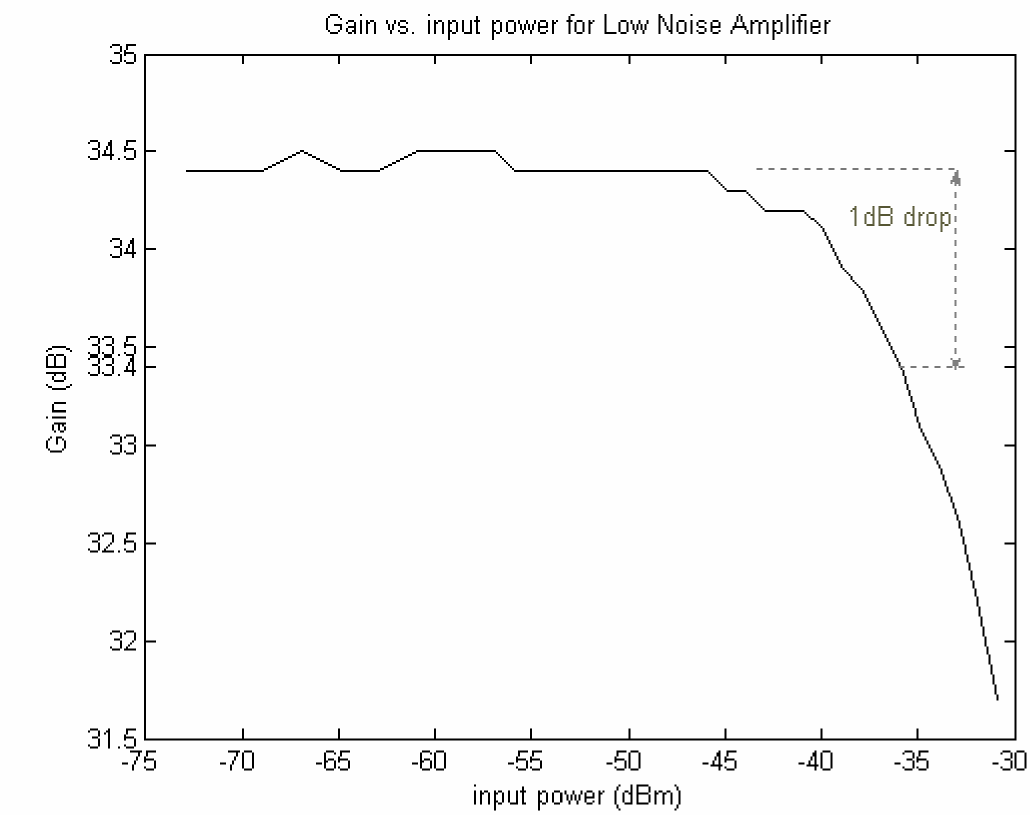
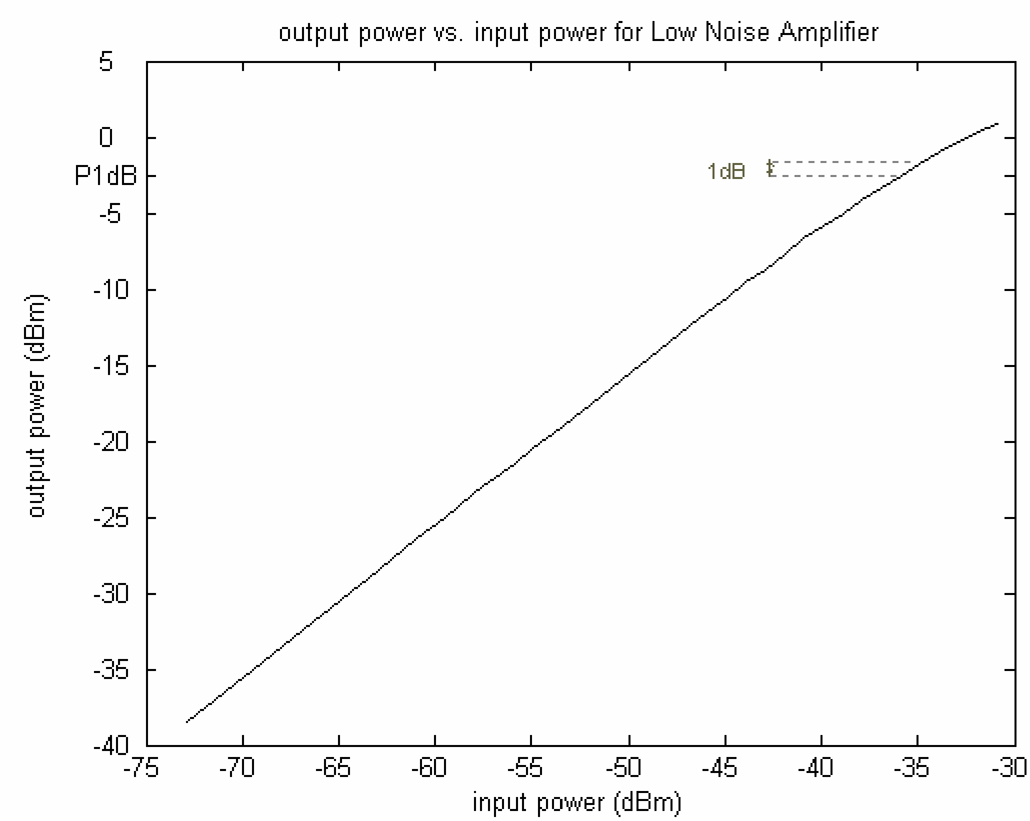


Figure 5.7 (top) : The Gain of LNA shows a non linear response for varying amplitude of input signal. Figure 5.8 (below) : The output power as a function of input power for LNA



From the above data, it is found that with increase in power of input signal the output signal increases but after certain point it begins to saturate to a value close to 1dBm. The gain drops off by 1 dB at input signal of -35.9 dBm. Thus as evident from the recorded data, the **1dB compression point** P_{1dB} of LNA is **-2.5 dBm**.

. The Noise floor, at the input side of the LNA, is given by $10\log_{10} kT_{sys}B$. For L-Band frequency range T_{sys} is taken as 80K, the noise floor then turns out to be

$$Noise\ Floor = 10\log_{10} kT_{sys} + 10\log_{10} B$$

$$\cong -180dBm \quad \text{for 1Hz bandwidth}$$

$$= -105dBm \quad \text{for 32MHz bandwidth}$$

$$= -99dBm \quad \text{for 120MHz bandwidth}$$

The **Compression Dynamic Range (CDR)** is hence found to be

$$CDR = P_{1dB} - NoiseFloor$$

$$= -2.5 - (-105 + 34.4)dB$$

CDR = 68.1 dB *for 32 MHz bandwidth*

To determine intercept points, two signals of equal power P_{IN} at 1245MHz and 1255MHz are input to the LNA. The output power P_{OUT} of the desired signal and P_{IM}, power of one of the third order IMD products at 1235MHz and 1265MHz is measured.

$P_{IN} = -56.2dBm$, input signal power

$P_{OUT} = -22dBm$, fundamental signal power

$P_{IM} = -76dBm$, Third order IMD product signal power

As per the equations derived in previous section, the output third order intercept point **OIP3** is computed as ***5dBm***

$$OIP_3 = P_{OUT} + \frac{1}{2}(P_{OUT} - P_{IM})$$

$$= -22 + 27dBm$$

$$= 5dBm$$

Similarly the input third order intercept point is obtained as

$$IIP_3 = OIP_3 - Gain$$
$$= 5 - 34.4dBm$$
$$= -29.4dBm$$

The **spurious free dynamic range SFDR** is then obtained

$$SFDR = \frac{2}{3}(IIP_3 - MDS_{IN}) = \frac{2}{3}(IIP_3 - 10\log_{10} kTBF)$$
$$= \frac{2}{3}(-29.4 - (-105))dB$$
$$= \frac{2}{3}(75.6)dB$$

SFDR* = *50.4dB *for 32MHz bandwidth*

The values of CDR, SFDR obtained, depend on the noise floor or the minimum discernible signal. The noise floor chosen here is the input from the OMT in L-Band Front-End receiver.

Note: The loss due to connecting cables and connectors was found to be negligible and has been assumed to be invariable for L-band frequency range.

Chapter Six

Linearity Modeling

6.1 Introduction

Almost every nonlinear system can be modeled as a linear system over a narrow operating range. It is also possible to model the non-linear behaviour to a certain range. The mathematical model describing the system can be employed to make simulation, hence analyze easily. Novel attempts have been made to model devices in L-Band front-end receiver system and analyze using the models.

RF devices such as amplifier show almost linear response for low power signals, but deviate for higher power signals due to various non linear effects examined earlier. The behaviour of such system can be described in a function which is a truncated Maclaurin series (a polynomial). Consider the input voltage P_{in} at the system and the output voltage P_{out} as,

$$P_{out} = \alpha_0 + \alpha_1 P_{in} + \alpha_2 P_{in}^2 + \alpha_3 P_{in}^3 + \alpha_4 P_{in}^4 + ... + \alpha_k P_{in}^k$$

At low values of input signal power P_{in}, the higher powers of P_{in} become negligible and the transfer function is reduced to a linear equation in P_{in}. For large values of P_{in} the system behave nonlinearly due to higher powers of P_{in}.

The response of a device can be measured for input signal amplitude range of interest. The techniques such as curve fitting can be used for approximating the response with polynomial. It involves evaluating or fixing the values $\alpha_0, \alpha_1, \alpha_2, ...\alpha_k$, coefficients of the Maclaurin series or coefficients of powers of argument of the function. Polynomials of different degrees can be obtained but the accuracy of the model increases with the degree of the equation. Figure 6.1 illustrates polynomials of degrees one, two and three along with the measure response in the same plot. The power has been measured in log scale, so the equation obtained hold for values in log scale.

The approximate polynomials obtained for a device are as follows:

One degree (linear) equation:

$$P_{out} = 26 + 0.9P_{in}$$

Two degree (quadratic) equation:

$$P_{out} = 20 + 0.64P_{in} - 0.0024P_{in}^2$$

Three degree (cubic) equation:

$$P_{out} = -3.9 - 0.92P_{in} - 0.033P_{in}^2 - 0.00018P_{in}^3$$

Note: All the equations are valid for power values in dBm units

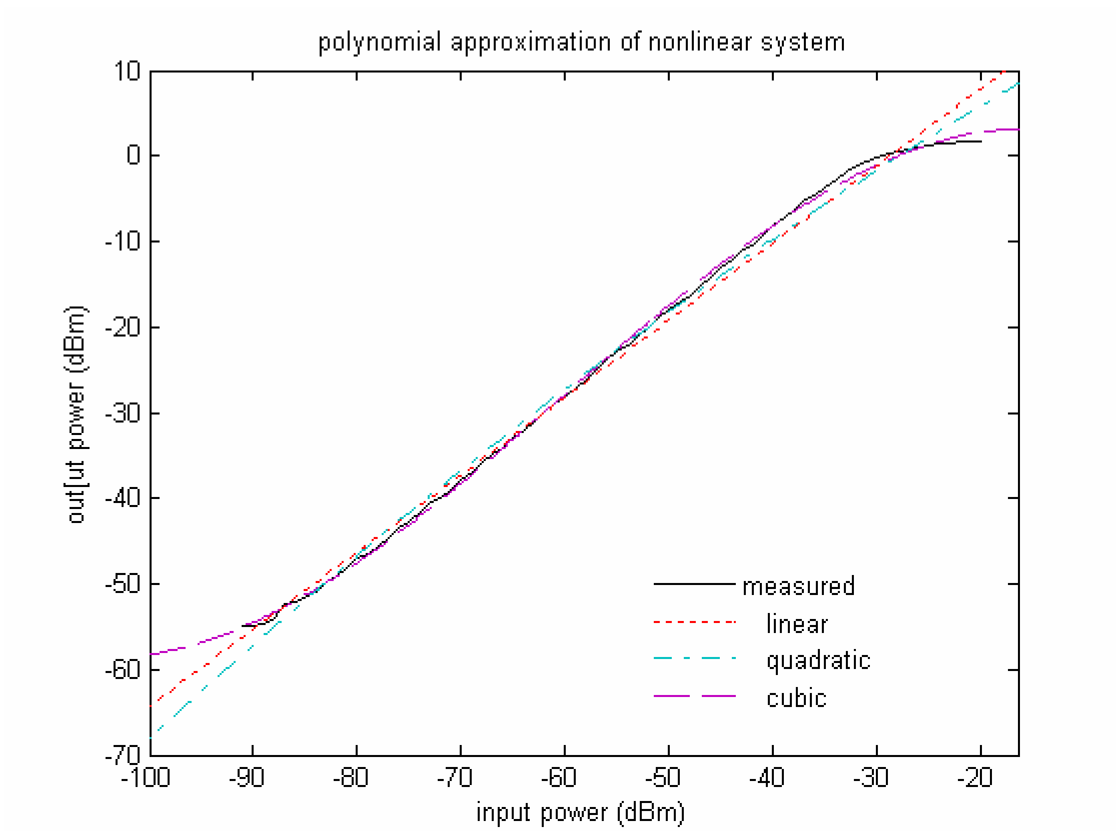


Figure 6.1 Polynomial approximation of nonlinear system

The model differs depending on the number of tones input and the frequency. The non linear products must also be considered to accurately model the nonlinear behaviour of the device. The model can only then be suitable for simulation and subsequent analysis. In this section, the 3^{rd} order intermodulation produced on two tone signal input to an amplifier has been analyzed and the cascading of devices has also been examined with models.

6.2 Analysis of nonlinear response of amplifier

This is an attempt to model the linear as well as non linear response of devices such as Low Noise Amplifier. The gain of the LNA is 34 dBm for L-Band frequency range and a noise temperature of 35K over the same frequency range. This Low noise amplifier is being used in the L-Band Front-end Box at GMRT.

Two tones of equal power at 1245MHz and 1255MHz are input to the LNA, resulting in 3^{rd} order IMD products at 1235MHz and 1265MHz. The amplitude of the two tones is increased and remained equal throughout. The power levels of fundamental and IM3 product are recorded for varying input power to the tones. The recorded data is then used to model the linear and nonlinear behaviour of the device.
The linear response of LNA has been modeled as polynomial in terms of input power (dBm) of the signal fed into the LNA. The 3 degree polynomial model is obtained by measuring the response of the LNA and then by curve fitting technique. The model describes the behaviour of the LNA with deviation less than a dBm from the actual measured values in the input power range shown in figure 6.2.

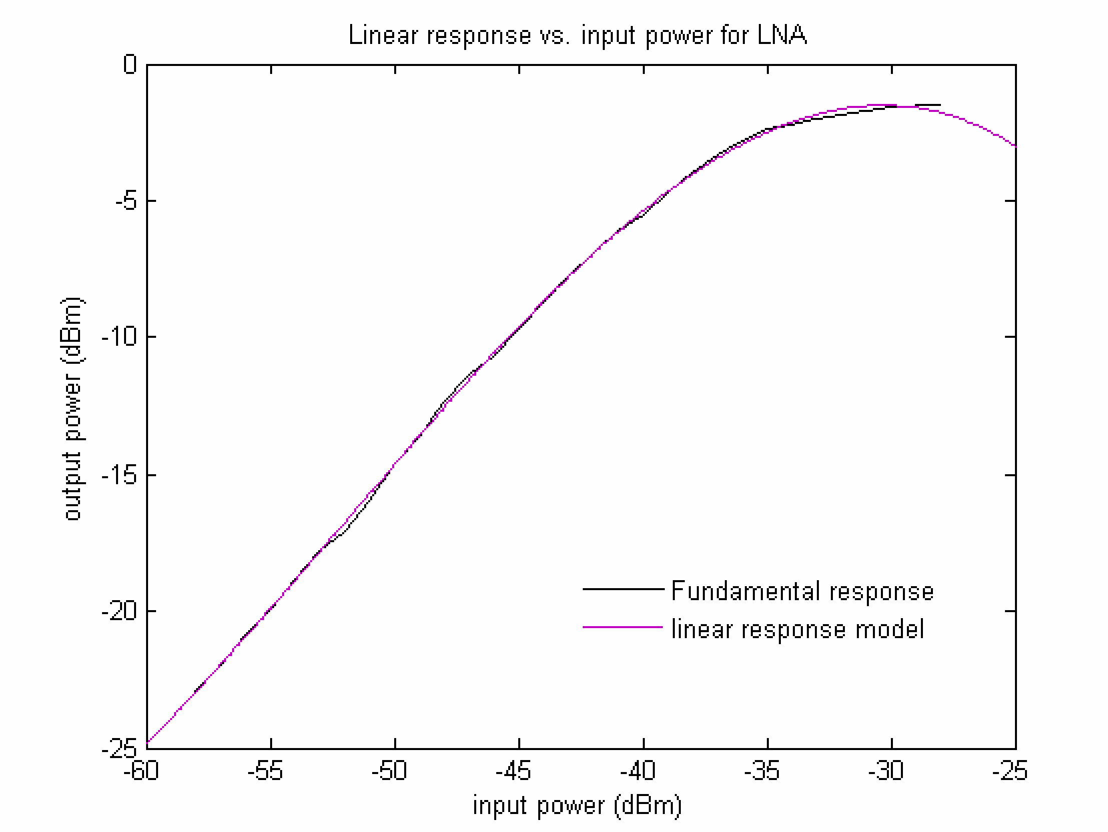


Figure 6.2 The linear response model function for LNA

The non-linear response of LNA has also been modeled as polynomial in terms of input power (dBm) of the signal fed into the LNA. The 3 degree polynomial model is obtained by measuring the power of third order IMD products as a result of two tones with different frequencies input to the LNA and then by curve fitting technique. The model describes the power of the third order IMD products owing to nonlinear behaviour of the LNA with a maximum deviation of 1dBm from the actual measured value for range of values shown in the figure 6.3. It is noteworthy that the accuracy of the model for some points was more with higher degree polynomial, but three degree polynomial is chosen for simplicity.

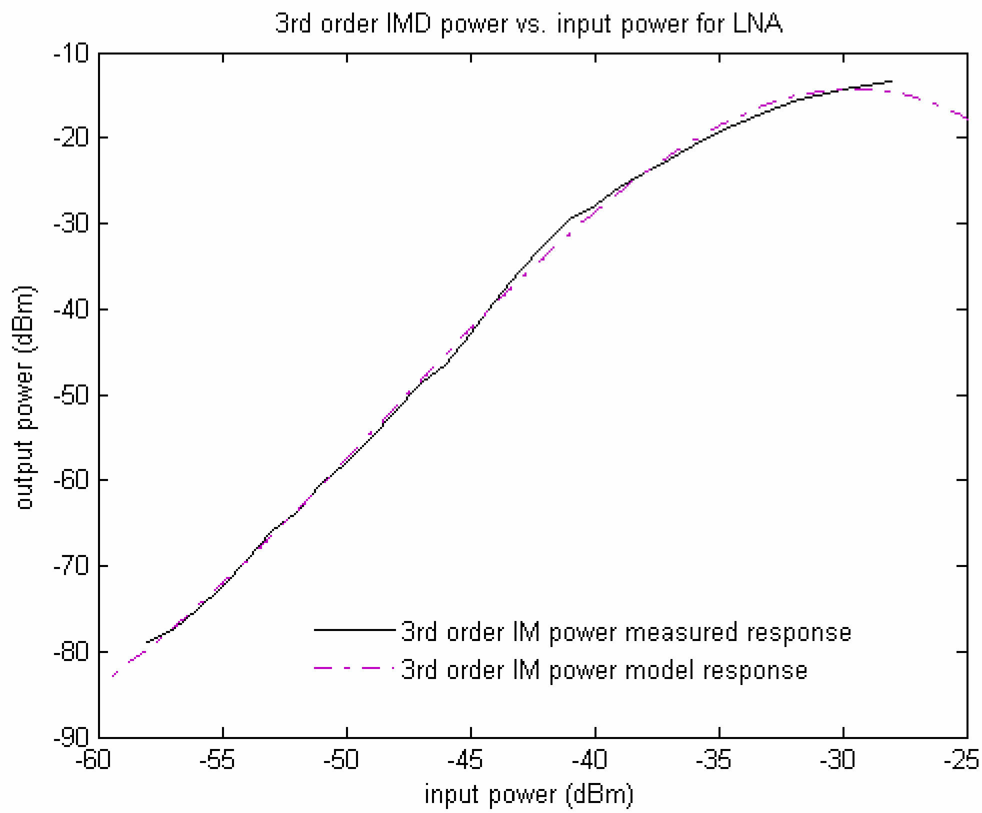


Figure6.3 Third order IMD product model for LNA

The polynomial functions describing the responses of the amplifier are as follows:

Fundamental response -

Third degree model : $P_{out} = f(P_{in}) = -66.847 - 4.9933P_{in} - 0.11616P_{in}^2 - 0.00074331P_{in}^3$

Linear model (tangent): $P_{out} = 32 + 0.94P_{in}$

3rd order IMD response –

Third degree model : $P_{out} = g(P_{in}) = -225.94 - 16.792P_{in} - 0.40868P_{in}^2 - 0.002804P_{in}^3$

Linear model (tangent): $P_{out} = 84.552 + 2.839P_{in}$

Note: the slopes of tangents to the responses are close to 1 and 3 for fundamental and nonlinear response, respectively.

The tangential line to the fundamental (linear) response of the LNA and non-linear response producing 3rd order IMD products when extrapolated in figure 6.4 gives *OIP3* as 6.27 dBm (*IIP3* = -26.67 dBm) which is close to the measured value of *OIP3* =5 dBm (*IIP3* = 29 dBm). The measurement of the IP3 points has been done by two tone test as it provides sufficiently accurate results as compared to multitone approximation[1].

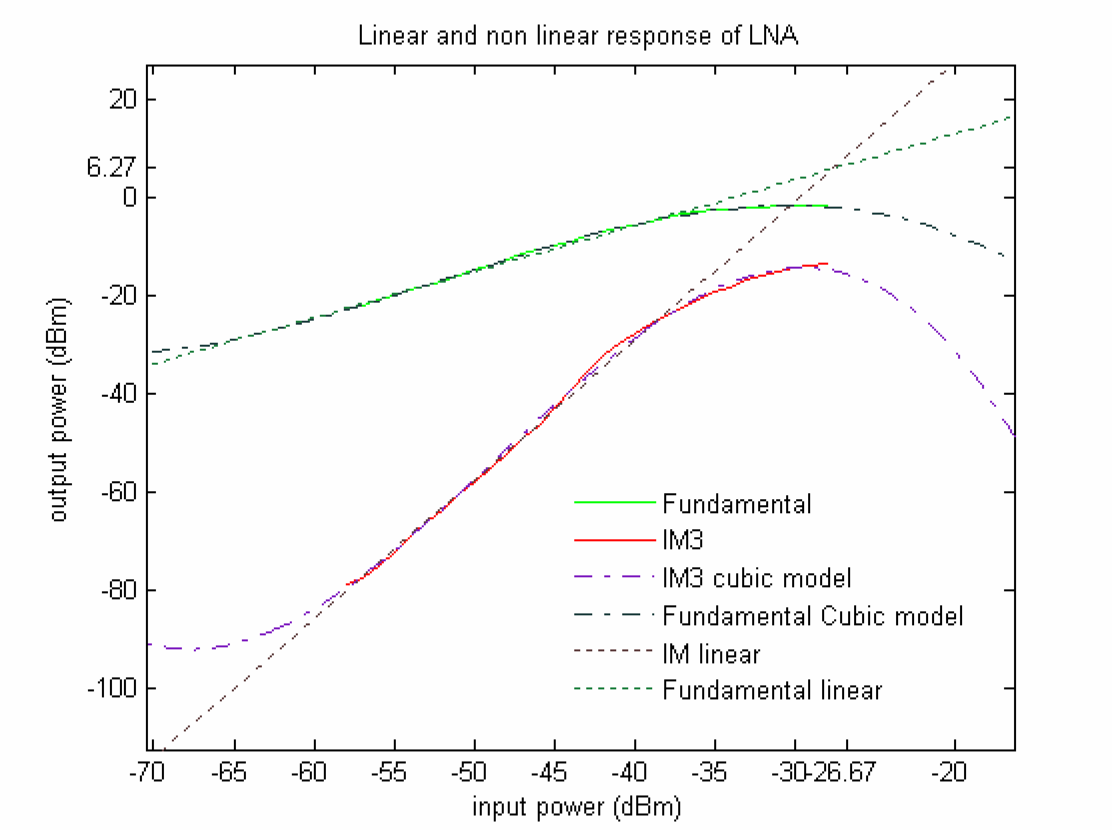


Figure 6.4 Analysis of LNA using models for linear and non linear product

The mathematical model of the device can thus be used for determining the behaviour of the device for input powers for which the deviation from measured value is tolerable. Note that the model is memory-less, hence it cannot entirely model a memory device such as LNA. Neglecting the higher order intermodulation products and other nonlinear products also adds to deviation of the model from real device.

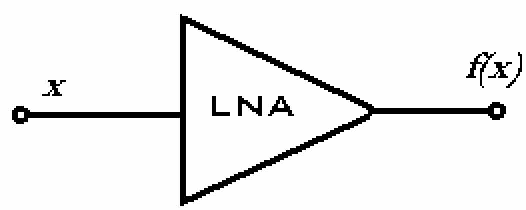
6.3 Analysis of cascaded stages using models

The linearity models of device can be utililized to describe the behaviour of the system cascaded with such devices. In this section, models for two different devices are determined and their cascaded stage is then analyzed and compared with measured result. The usefulness of models explained in this report is evident with the following analysis.

The devices, Low Noise Amplifier and Phase Switch from the proposed L-Band Front-end receiver are put under examination. The LNA has three stages of FUJITSU HEMT's FHX35LG and phase switch box has cascades stages of ERA-6, attenuator, SRA 2010MH mixer, 2 stages of ERA-6 and an attenuator. The models for the devices are determined using the basic curve fitting technique. The model functions obtained are,

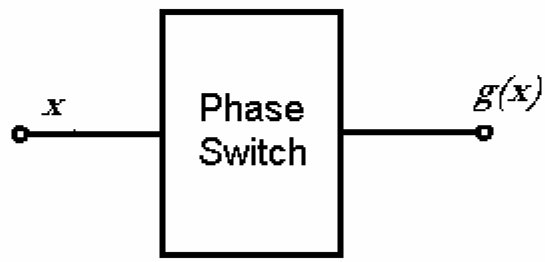
for Low Noise Amplifier, LNA the model function *f* is

$$f(x) = -0.00015x^3 - 0.028x^2 + 0.66x - 0.26$$



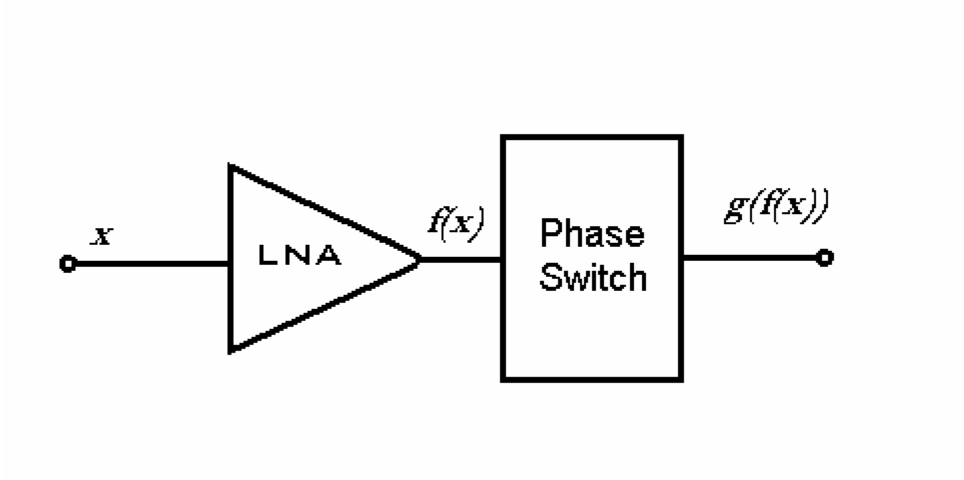
And for Phase switch, the model function *g* is

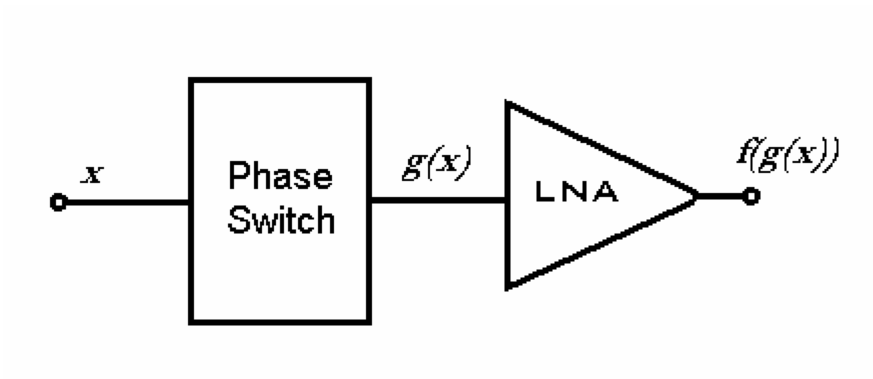
$$g(x) = -6.2\times10^{-8}x^5 - 1.3\times10^{-5}x^4 - 0.001x^3 - 0.037x^2 + 0.34x + 11$$



The plots of the functions with the measured values can be seen in figure 6.5. A 5 degree polynomial was chosen for phase unit to show that better accuracy is obtained with higher degree of polynomial. The 3 degree polynomial modeling for LNA deviates more near saturation.

The devices LNA and phase unit can be cascaded in two ways. In way I, the LNA is the first stage , followed by the phase switch and in way II the LNA follows the phase switch. The model functions for the two cascades are then simply a composition of model functions of cascading devices. Therefore, for cascade I , the model function is $g \circ f$, while for cascade II the model function is $f \circ g$.





Images of the composition functions at x are given as,

$$g \circ f(x) = -6.2\times10^{-8} f(x)^5 - 1.3\times10^{-5} f(x)^4 - 0.001 f(x)^3 - 0.037 f(x)^2 + 0.34 f(x) + 11$$

$$f \circ g(x) = -0.00015 g(x)^3 - 0.028 g(x)^2 + 0.66 g(x) - 0.26$$

The composition functions above have not been expanded as they become too lengthy and complex to be included in the report. Using these two functions the output values are computed as values of functions at the inputs of the cascades. The functions were used to find the gain and P1dB of the cascades. The measure response of the LNA, phase switch and cascade I are plotted in figure 6.5. It is observed from the plots that

For LNA,

$$Gain = 32dB,$$

$$P_{1dB} = -1dBm$$

For phase switch,

$$Gain = 16dB,$$

$$P_{1dB} = 8.5dBm$$

For cascade II, i.e., phase switch followed by LNA,

$$Gain = 48.3dB,$$

$$P_{1dB} = -1.7dBm$$

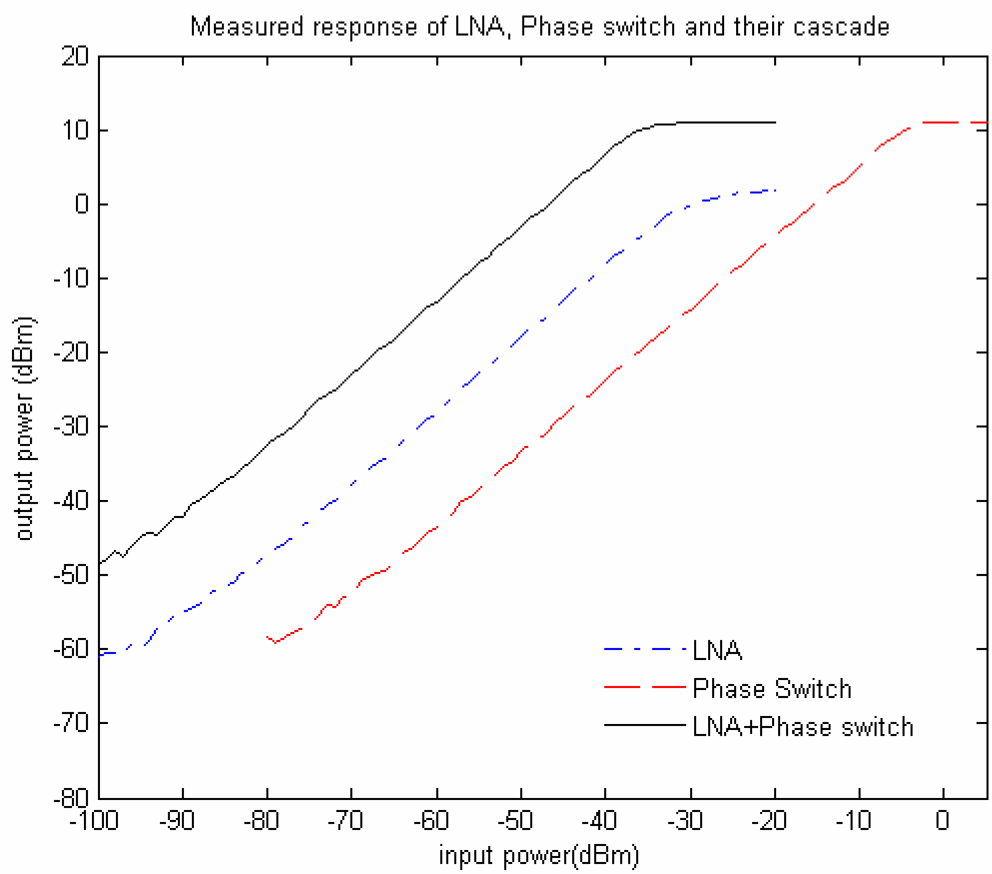


Figure6.5 Measured response of LNA, phase switch and cascade I

The model functions for LNA , phase switch and cascade I are plotted in the figure 6.6. From the plots, following results can be deduced,

For LNA,

$$Gain = 31dB\,,$$

$$P_{1dB} = -5dBm$$

For phase switch,

$$Gain = 15.5dB\,,$$

$$P_{1dB} = 5.5dBm$$

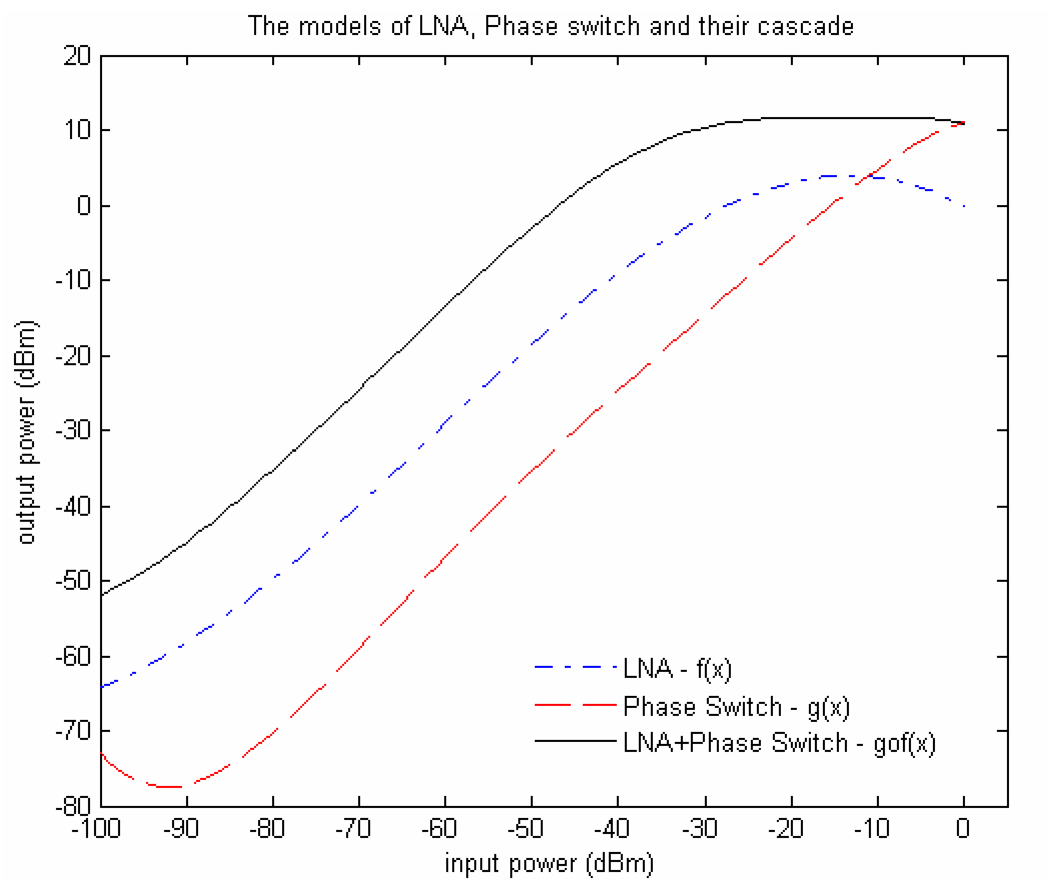


Figure 6.6 Models of LNA, Phase Switch and Cascade I

It thus noted that the values from model functions differ by few units from the actual value. The values obtained with 5 degree polynomial for phase switch are closer as accuracy is improved with higher degree of polynomials. The results for cascade I, i.e., LNA followed by phase switch, are obtained as

$$Gain = 46.5dB\,,$$

$$P_{1dB} = 5.5dBm$$

which are near to measured values of 46.5dB and 5.5 dBm.

Chapter Seven

Analysis, Measurement and Modeling of Existing L-Band Front End

The linearity measurement is done for Existing L-Band Front End. The values obtained are also summarized in this chapter. Model Functions for gain and output for the device is also obtained. Measurement of the existing Front-End system with narrow BPF bypassed and with the input given to the LNA and the out put taken at the output of the Front-End is done.

7.1 1dB compression point

The output power signal of the Front-End was measured for varying amplitude of input signal at 1250 MHz. The values obtained by measurement are in table 7.1.

Input power (dBm)	output power (dBm)
-95.3	-42.1
-94.3	-41.1
-93.3	-40.1
-92.3	-39.1
-91.3	-38.1
-90.3	-37.1
-89.3	-36.1
-88.3	-35.1
-87.3	-34.1
-86.3	-33.1
-85.3	-32.1
-84.3	-31.1
-83.3	-30.1
-82.3	-29.1
-81.3	-28.1
-80.3	-27.1
-79.3	-26.1
-78.3	-25.1
-77.3	-24.1
-76.3	-23.1
-75.3	-22.1
-74.3	-21.1
-73.3	-20.1
-72.3	-19.1
-71.3	-18.1
-70.3	-17.1
-69.3	-16.1
-68.3	-15.1
-67.3	-14.1
-66.3	-13.2

-65.3	-12.2
-64.3	-11.2
-63.3	-10.2
-62.3	-9.2
-61.3	-8.2
-60.3	-7.3
-59.3	-6.3
-58.3	-5.4
-57.3	-4.5
-56.3	-3.8
-55.3	-3
-54.3	-2.4
-53.3	-1.9
-52.3	-1.4
-51.3	-1
-50.3	-0.7
-49.3	-0.4
-48.3	-0.3
-47.3	-0.3
-46.3	-0.3

Table 7.1 Output power values for varying power of input at 1250 MHz for L-Band Front End

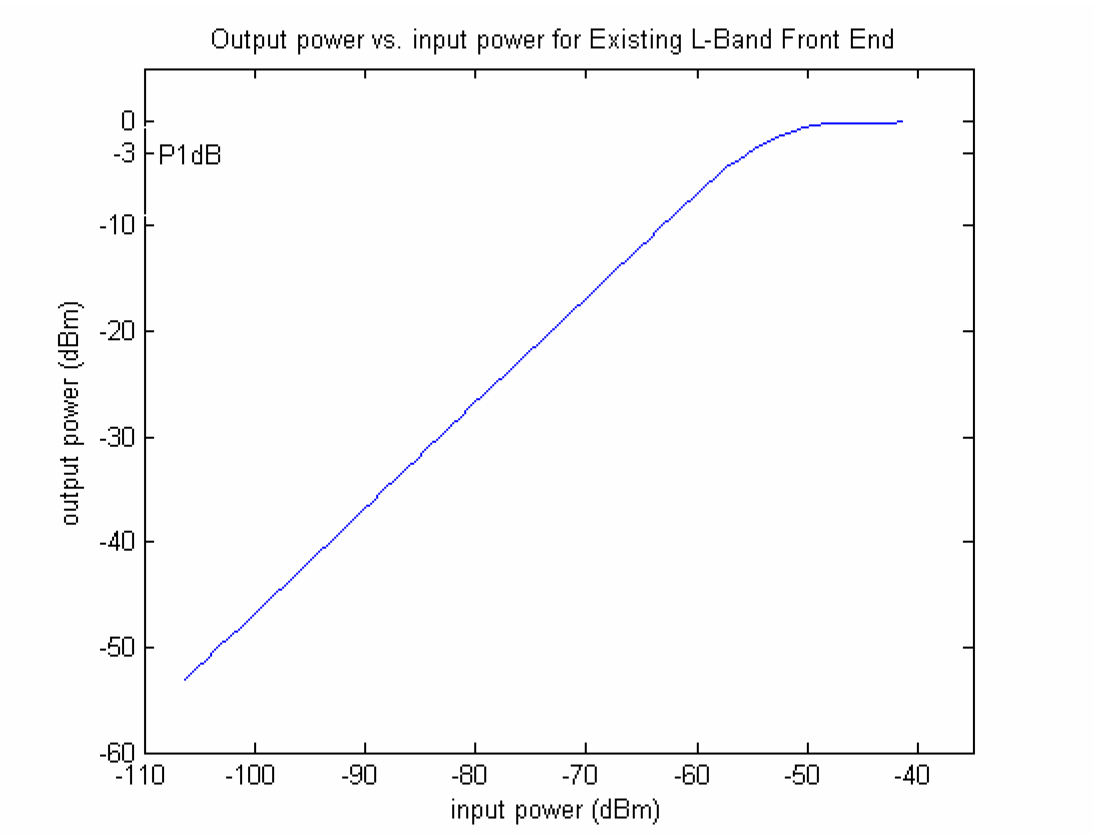


Figure 7.1: The output power as a function of input power for L-Band front-end

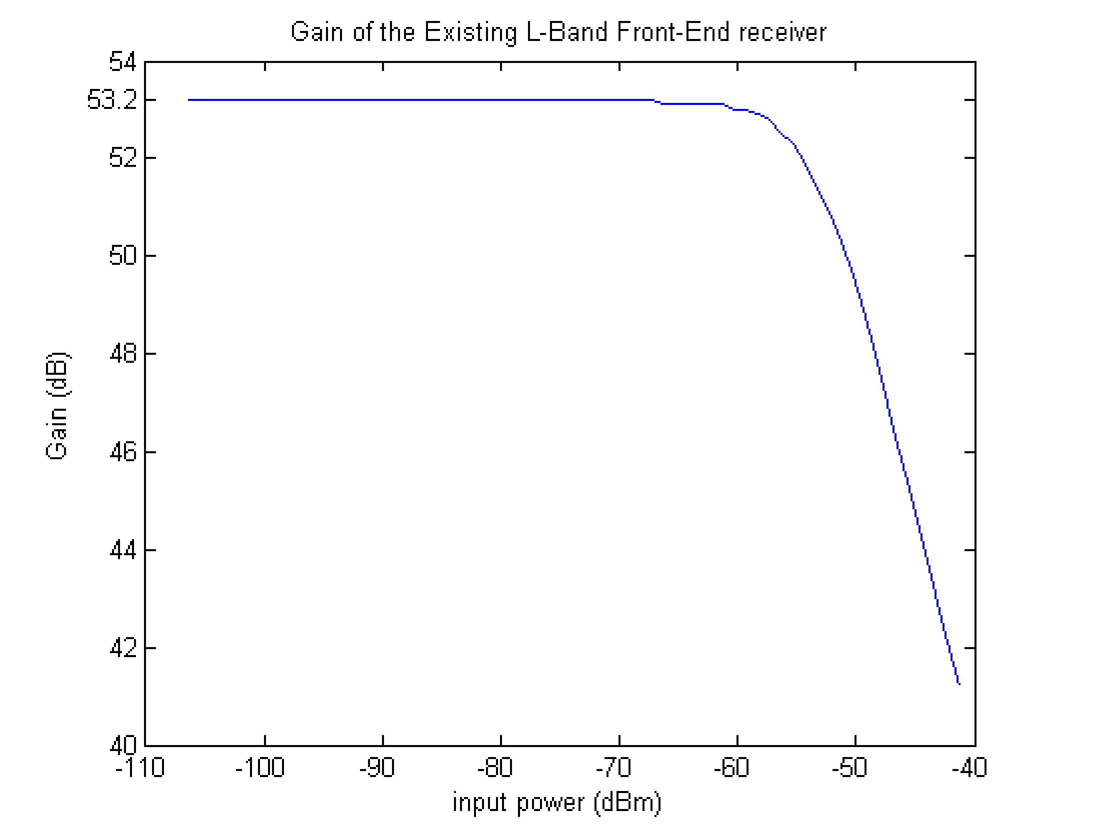


Figure 7.2: The Gain for varying amplitude of input signal.

From the above data, it is found that with increase in power of input signal the output signal increases but after certain point it begins to saturate to certain value. The gain drops off by 1 dB at input signal of -55.2 dBm. Thus as evident from the recorded data, the **1dB compression point** P_{1dB} of existing front-end is **-3 dBm**.

7.2 Third Order Intercept Pont

To determine intercept points, two signals of equal power P_{IN} at 1245MHz and 1255MHz are input to the LNA. The output power P_{OUT} of the desired signal and P_{IM}, power of one of the third order IMD products at 1235MHz and 1265MHz is measured.

$P_{OUT} = -11dBm$, fundamental signal power

$P_{IM} = -45dBm$, Third order IMD product signal power

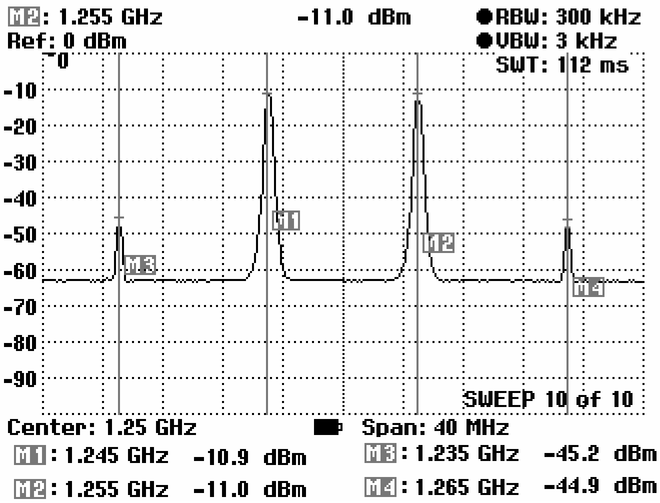


Figure 7.3 IP3 measurement of Existing Front-End

As per the equations derived in previous section, the output third order intercept point **OIP3** is computed as ***6 dBm***

$$OIP_3 = P_{OUT} + \frac{1}{2}(P_{OUT} - P_{IM})$$

$$= -11 + 17 dBm$$

$$= 6 dBm$$

7.3 Noise floor

At the input of the receiver,

$$\text{Noise Floor} = 10\log_{10} kT_{sys} + 10\log_{10} B$$

$$\cong -180 dBm \quad \text{for 1Hz bandwidth}$$

$$= -105 dBm \quad \text{for 32MHz bandwidth}$$

$$= -99 dBm \quad \text{for 120MHz bandwidth}$$

7.4 Compression Dynamic Range (CDR)

The Compression Dynamic Range (CDR) is hence found to be

$$CDR = P_{1dB} - NoiseFloor$$

$$= -3 - (-105 + 53.2)dB$$

$$\textbf{\textit{CDR = 48.8 dB}} \quad \textit{for 32 MHz bandwidth}$$

7.5 Spurious Free Dynamic Range (SFDR)

The spurious free dynamic range (SFDR) is then obtained

$$SFDR = \frac{2}{3}(OIP_3 - MDS_{OUT})$$

$$= \frac{2}{3}(6 - (-105 + 53.2))dB$$

$$= \frac{2}{3}(57.8)dB$$

$$\textbf{\textit{SFDR}} \quad = \textbf{\textit{38.5dB}} \quad \textit{for 32MHz bandwidth}$$

Note: The loss due to connecting cables and connectors was found to be negligible and has been assumed to be invariable for L-band frequency range.

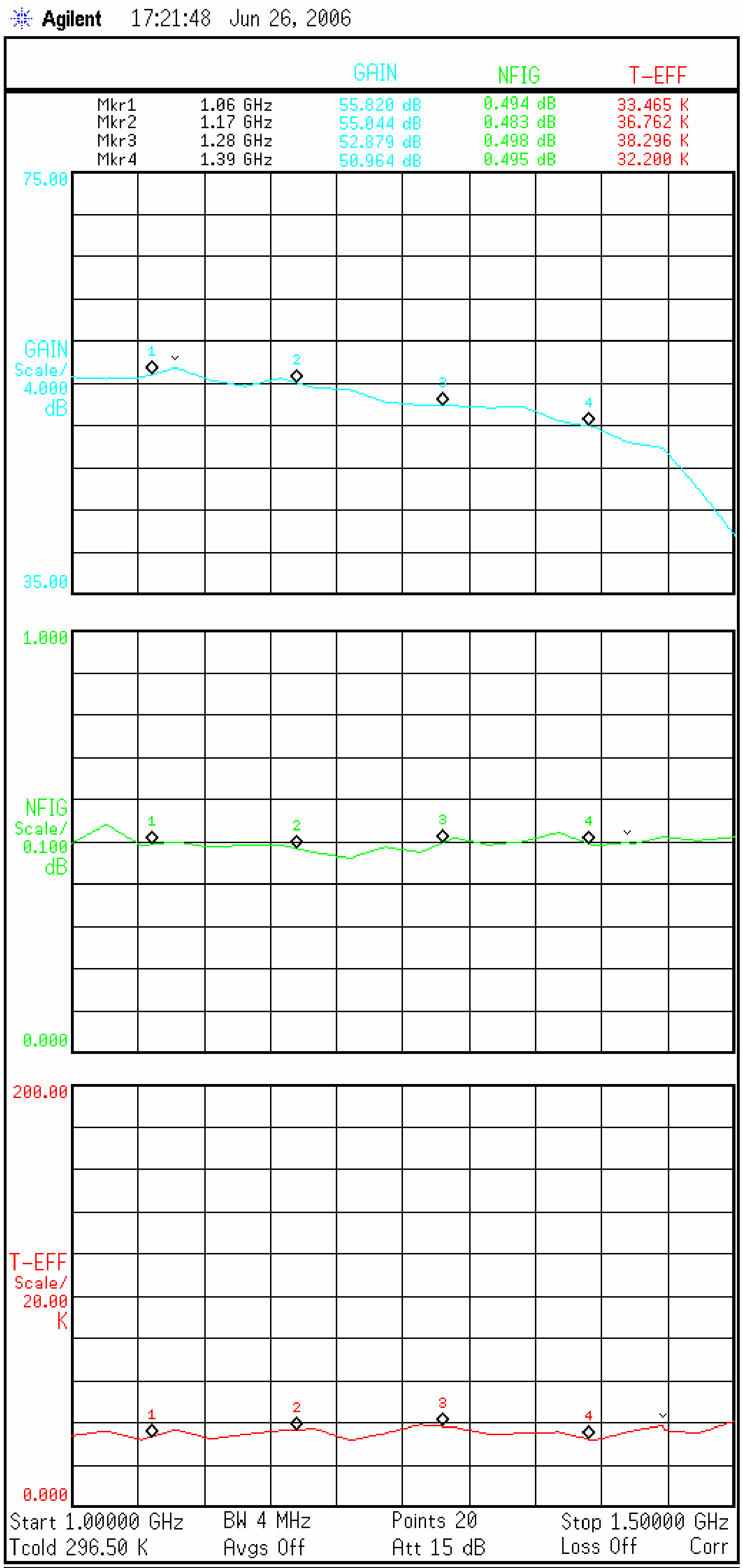


Figure 7.4 The Noise Figure meter display for the Existing Front-End system

7.6 RESULT

The measurement for the existing L-Band Front-End receiver is summarized in the following table

Table 7.2 Measurement of L-Band Front-End

Parameter	Value
Gain	53.2 dB
P1dB	-3 dBm
Noise Temperature	47 K *(includes cable+OMT losses)*
OIP3	6 dBm
CDR	48.8 dB
SFDR	38.5 dB

7.7 MODELLING OF EXISTING DESIGN

The recorded data for existing front-end receiver is then used to obtain model functions for the devices. The model functions obtained are as follows

Gain Model : $g(x) = -3.9\times10^{-8}x^5 - 2\times10^{-5}x^4 - 0.0039x^3 - 0.38x^2 - 18x - 270$

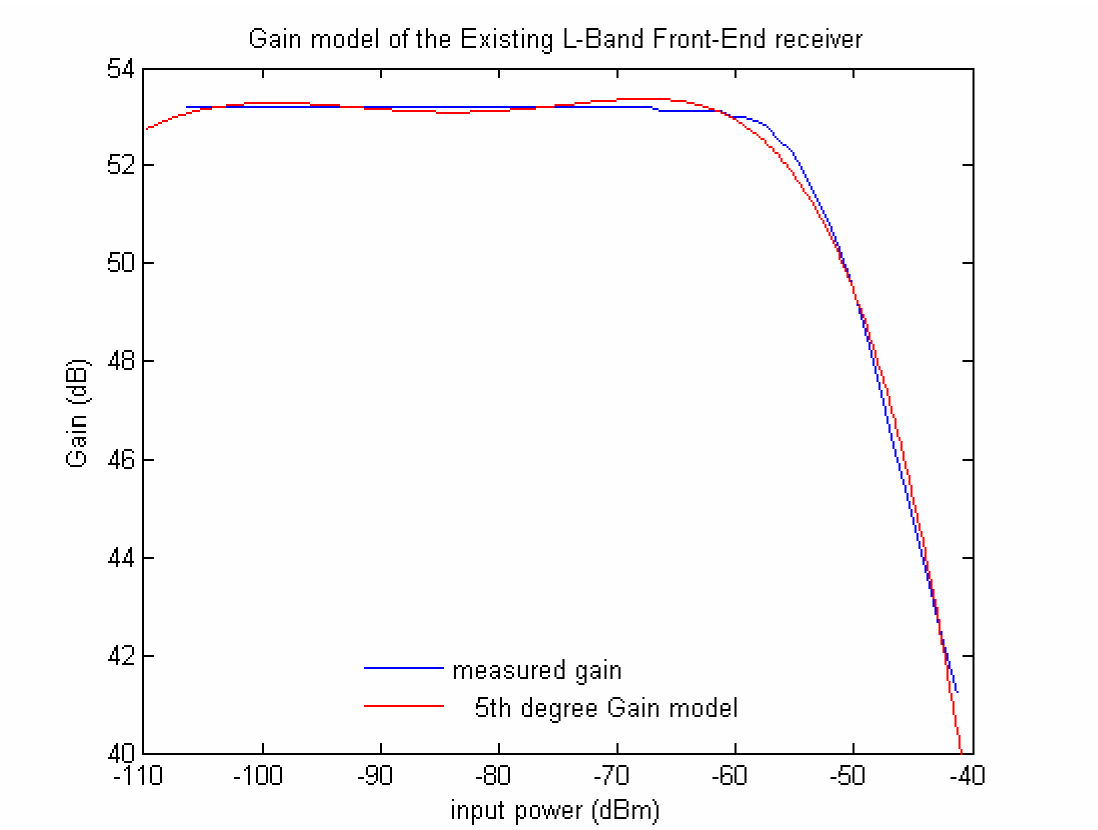


Figure 7.5 Model function for gain of the Front End Receiver

Fundamental Output Model : $f(x) = -2.1\times10^{-4}x^3 - 0.053x^2 - 3.3x - 59$

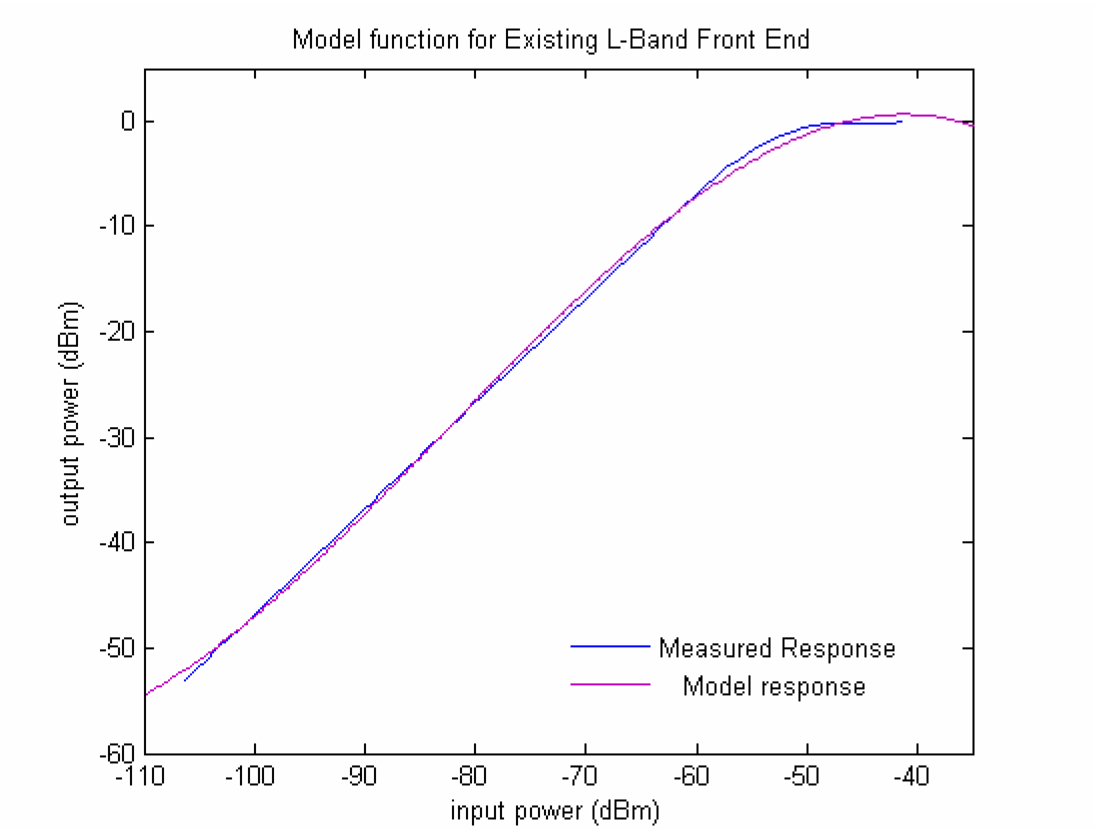


Figure 7.6 Model for fundamental output of L-Band Front End Receiver

Chapter Eight

Comparison Between MAR 3 and ERA 6 Amplifier

MAR 3 and ERA 6 are both mini circuit MMIC amplifier. MAR 3 is incorporated in the existing L-band front-end design in the post amplifier and the phase switch network with one stage of MAR 3 in post amplifier and 3 stages of MAR 3 in post amplifier. The following are datasheet parameters of MAR 3 and ERA 6 amplifiers.

Model	Freq GHz	Gain (dB) Typical at GHz						Maximum power dBm		Dynamic range		Maximum rating (25 ° C)		Dc Power at pin 3	
	f_L f_U	.1	1	2	3	4	min	o/p 1dB comp	i/p no damage	NF dB Typ	IP3 dBm Typ	I (ma)	P (mw)	I ma	V volts
MAR3	dc - 2	12.5	12	10.5			8	+1.5	+20	6.0	+23.0	70	400	35	5.0
ERA6	dc - 4	12.6	12.5	12.2	12.7	11.8	11	+17.9	+20	4.5	+36.0	120	650	70	5.0

Table 8.1 Datasheet values for ERA 6 and MAR 3 monolithic amplifiers.

As we can see ERA 6 amplifiers have a comparatively very high 1 dB compression point and IP3 point and the gain of both the devices are comparable. To compare the characteristics of MAR 3 and ERA 6 amplifiers we have designed an ERA 6 amplifier and compared the result with a MAR 3 amplifier. The circuit diagram for a MAR 3 amplifier is given in the figure 8.1.

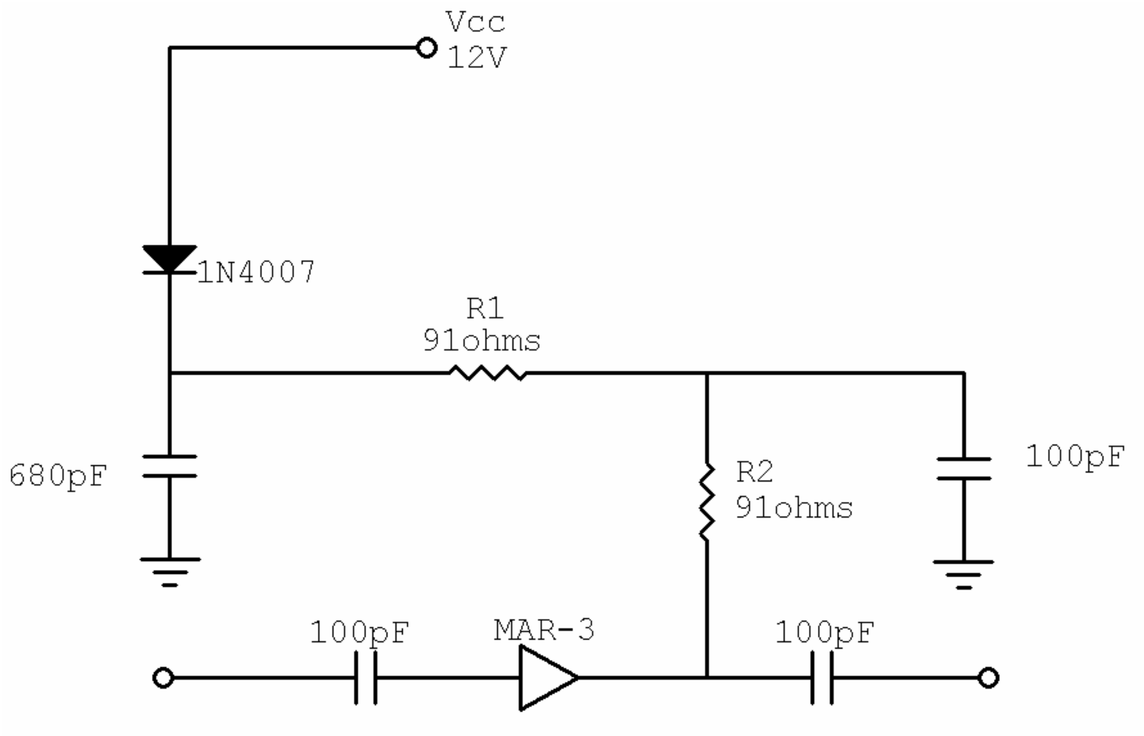


Figure 8.1 Circuit Diagram for Amplifier with MAR 3

8.1 Design of ERA 6 amplifier

The specification of ERA 6 is as follows

Typical Values for

current=70 mA

voltage=5 V

i/p voltage=12V

considering the diode drop to be 0.7 the calculated value of the biasing resistance is R=90ohms, 0 .441watt. Including the safety margin two half watt 180 ohms resisters are used in parallel. The resulting circuit of the amplifier is seen in figure 8.2

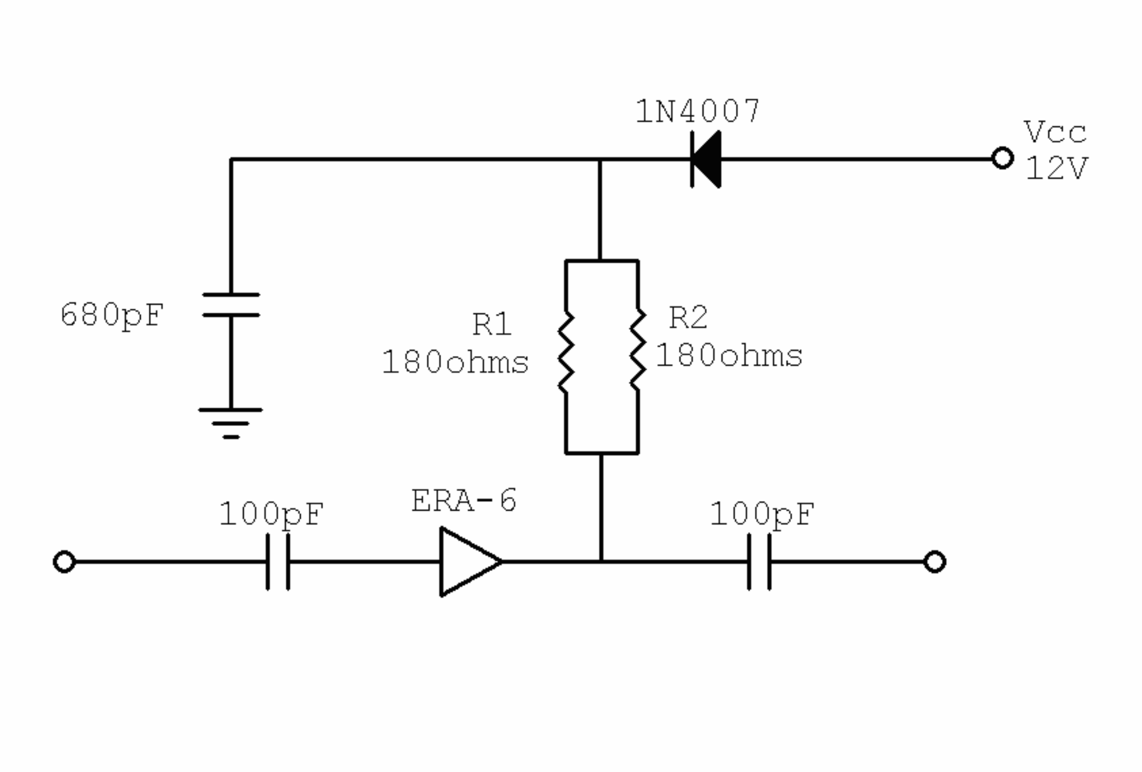


Figure 8.2 Circuit Diagram of Amplifier with ERA-6

The parameters of the amplifiers were measured and compared. The gain, noise temperature and noise figure of the amplifiers for L-Band frequency range are seen in the following plots.

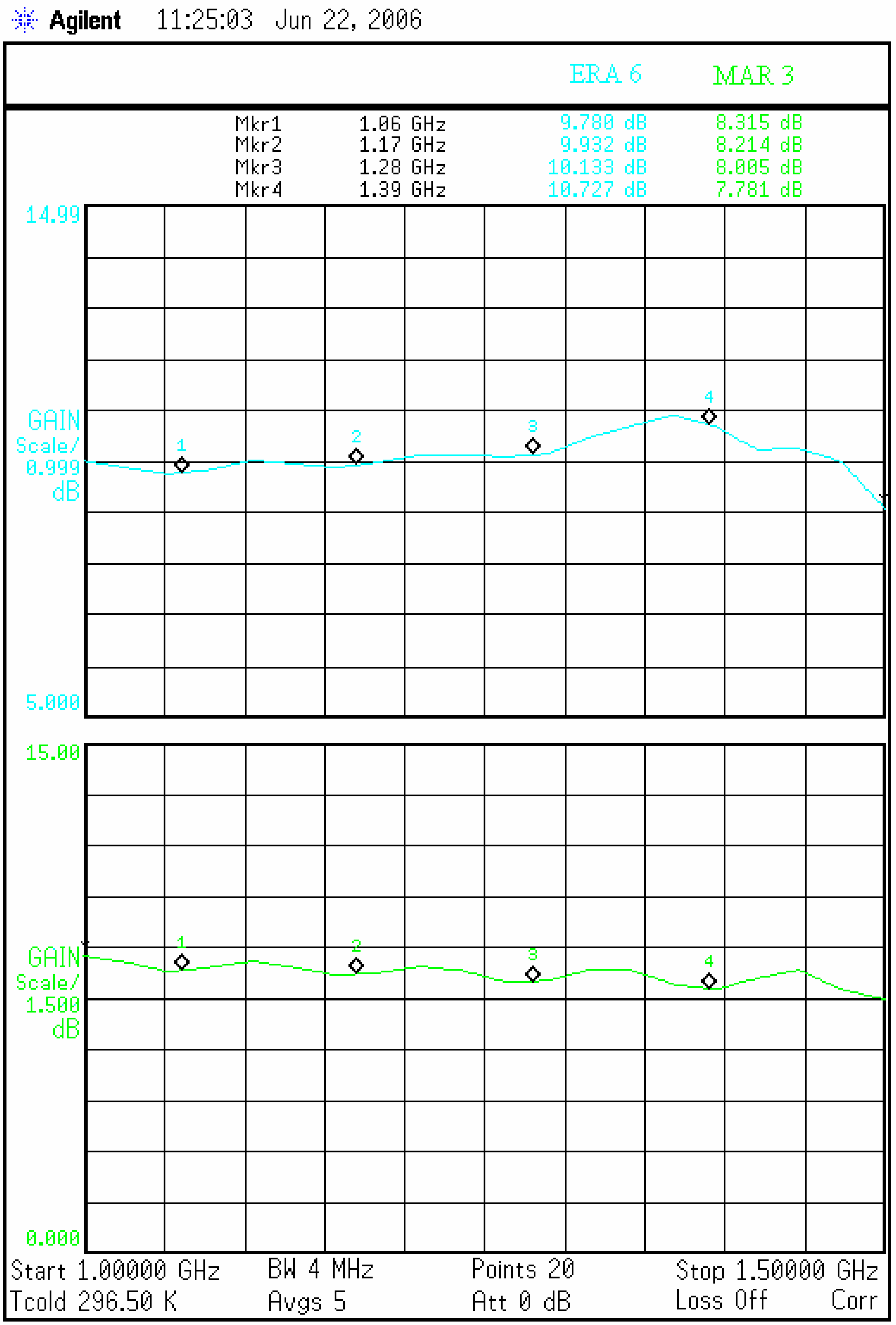


Figure 8.3 Gain of the amplifiers of MAR 3 and ERA 6 over L-Band frequency range

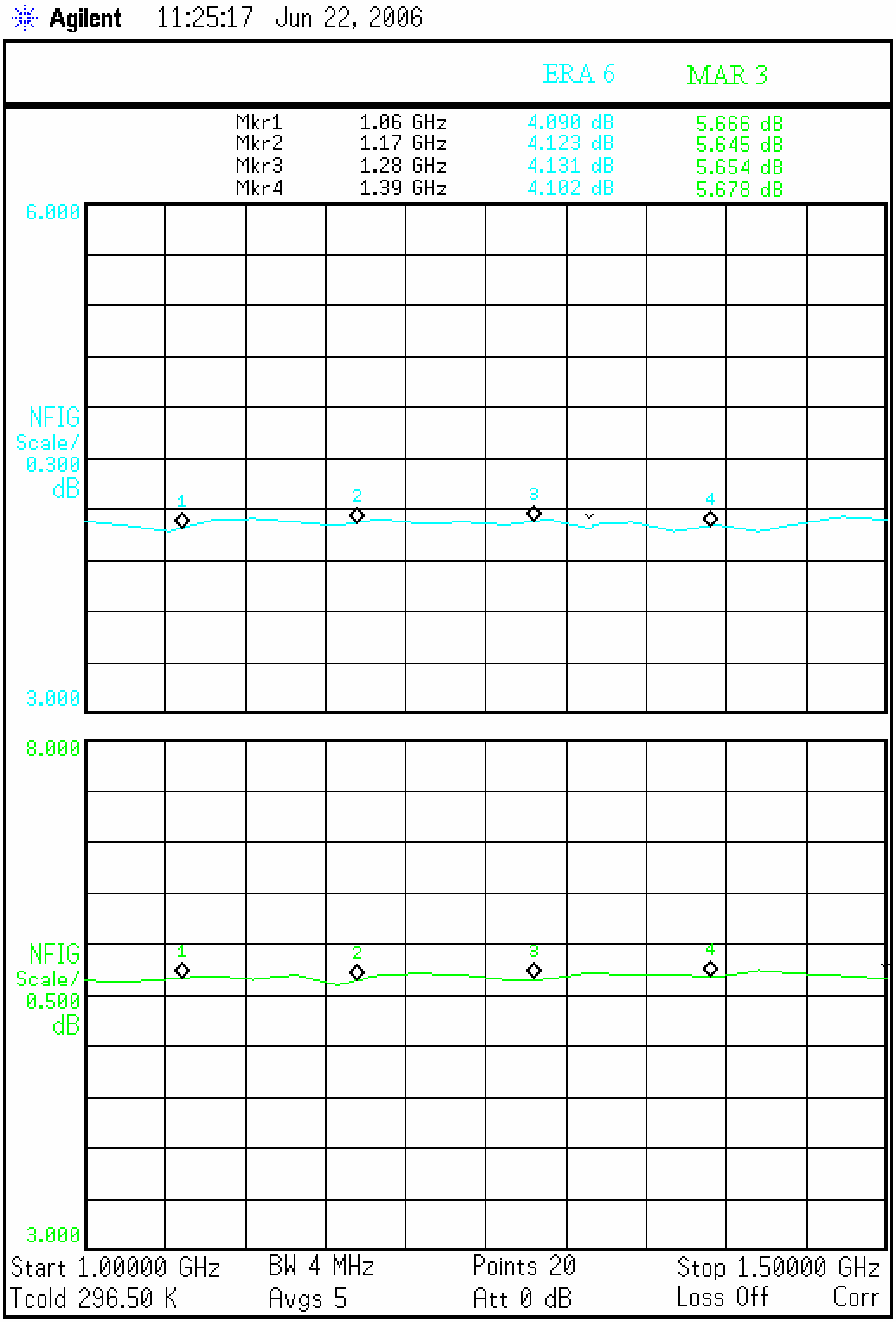


Figure 8.4 Noise Figure of the amplifiers of MAR 3 and ERA 6 over L-Band frequency range

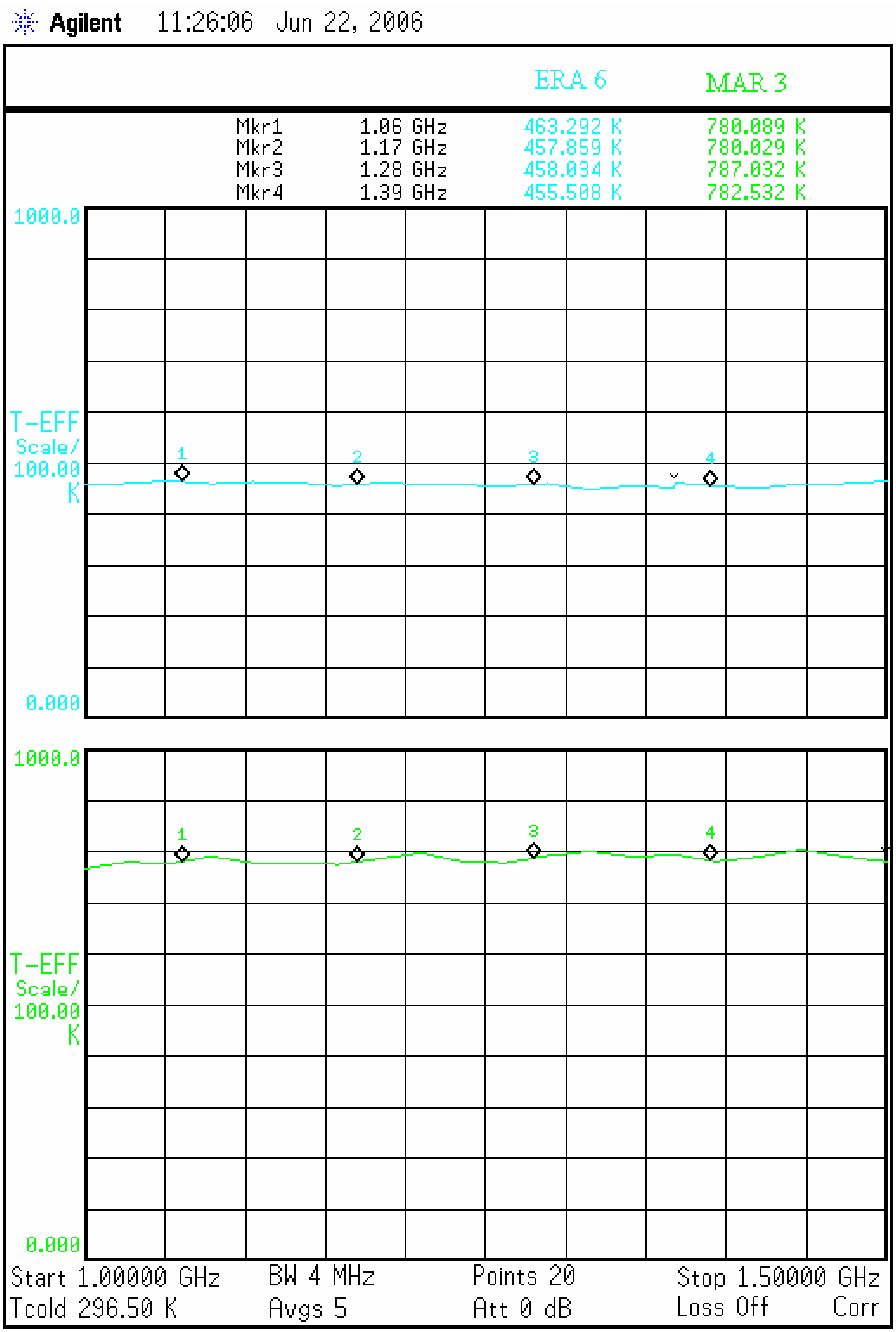


Figure 8.5 Noise Temperature of the amplifiers of MAR 3 and ERA 6 over L-Band frequency range

8.2 Calculation of 1dB compression point

For measurement of P1dB a single frequency signal at 1250MHz from a signal generator is input to the LNA and the power of the output signal is measured in a spectrum analyzer. The amplitude of the input signal is varied and the corresponding output power is recorded till the output power saturates. The values measured for ERA 6 and MAR 3 amplifiers are given in Table 8.2.a and 8.2.b,respectively

Table 8.2.a

Input power (dBm)	Output power (dBm)
-12.2	-2.3
-11.2	-1.3
-10.2	-0.3
-9.2	0.7
-8.2	1.7
-7.2	2.7
-6.2	3.7
-5.2	4.7
-1.2	8.8
-0.2	9.7
0.8	10.7
1.8	11.7
2.8	12.7
3.8	13.6
4.8	14.5
5.8	15.3
6.8	15.8
7.8	16
8.8	16.1
9.8	16.2
10.8	16.3

Table 8.2.b

Input power (dBm)	Output power (dBm)
-21.3	-13.3
-19.3	-11.3
-17.3	-9.3
-15.3	-7.4
-13.3	-5.3
-11.3	-3.3
-9.3	-1.3
-8.3	-0.4
-7.3	0.6
-6.3	1.6
-5.3	2.6
-4.3	3.5
-3.3	4.4
-2.3	5.3
-1.3	6.2
-0.3	6.9
0.7	7.5
1.7	7.9
2.7	8.3
3.7	8.8
4.7	9.1
5.7	9.3

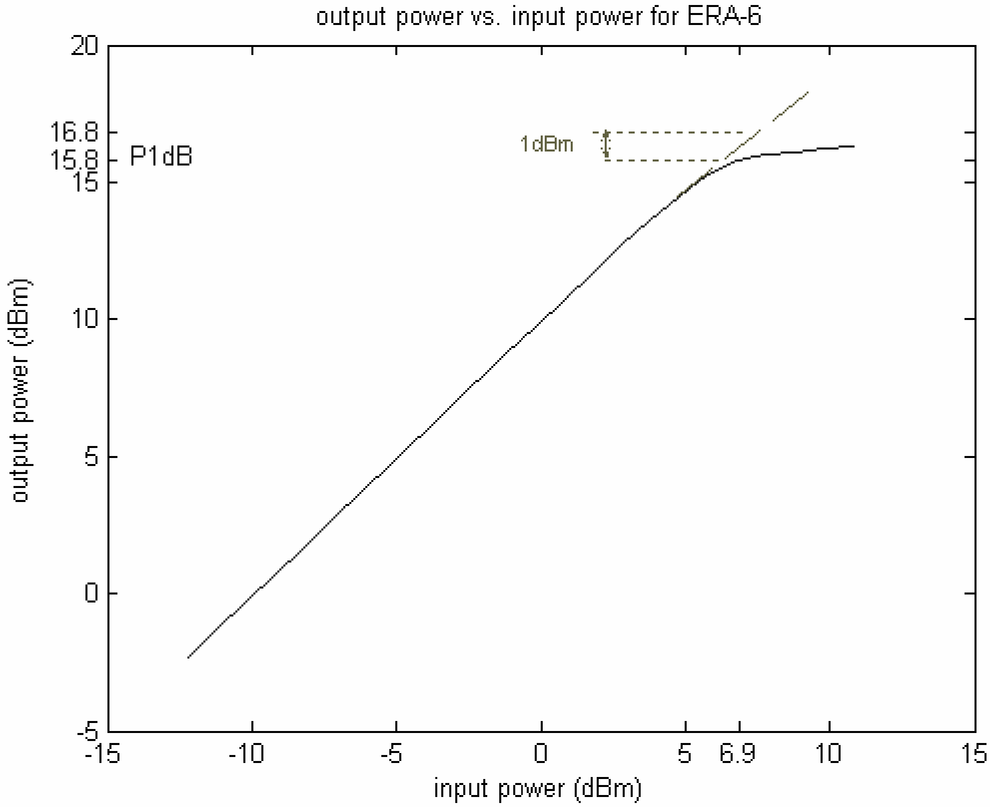
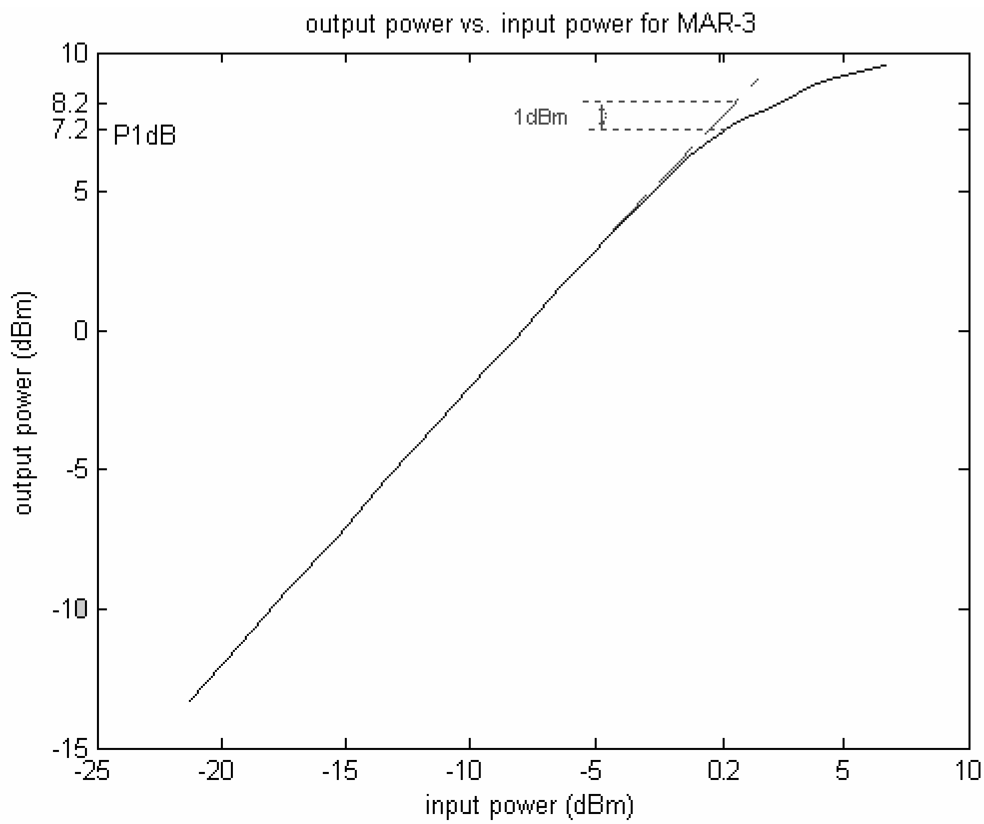


Figure 8.6 (top) The output power as a function of input power for ERA 6 amplifier. Figure 8.7 (below) The output power as a function of input power for ERA 6 amplifier



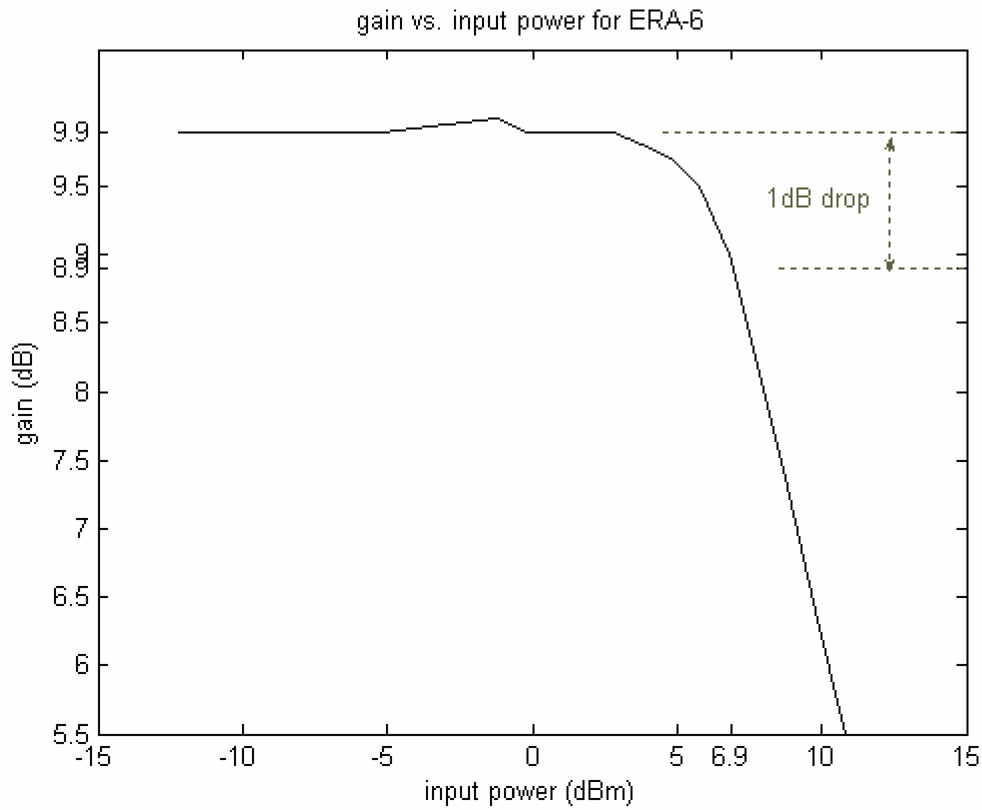


Figure 8.8 The gain as a function of input power for ERA 6 amplifier

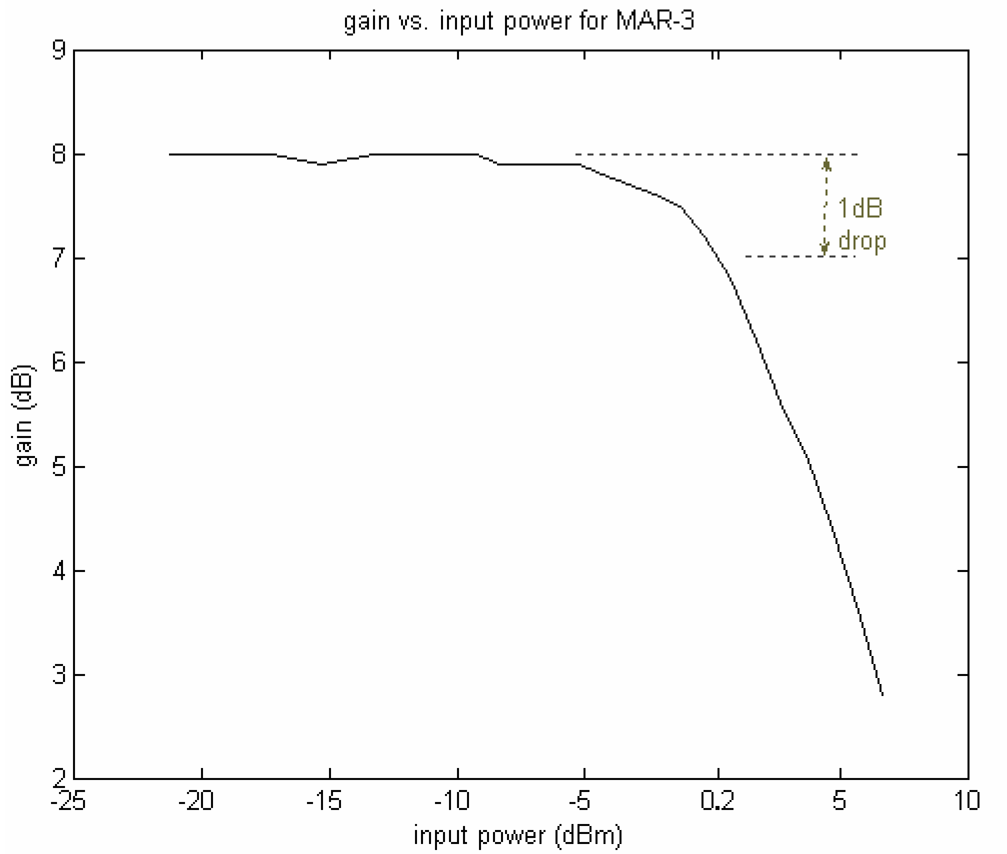


Figure 8.9 : The gain as a function of input power for MAR 3 amplifier

From the above plots, 1dB compression point is obtained as

*P1dB for **ERA 6** amplifier = **15.8 dBm***

*P1dB for **MAR 3** amplifier = **7.2 dBm***

To determine intercept points, two signals of equal power P_{IN} at 1245MHz and 1255MHz are input to the amplifiers. The output power P_{OUT} of the desired signal and P_{IM}, power of one of the third order IMD products at 1235MHz and 1265MHz is measured.

8.3 Calculation of OIP_3 for ERA 6 amplifier

$P_{OUT} = -6dBm$, fundamental signal power

$P_{IM} = -63dBm$, Third order IMD product signal power

As per the equations derived in previous section, the output third order intercept point OIP_3 for ERA 6 amplifier is computed as

$$OIP_3 = P_{OUT} + \frac{1}{2}(P_{OUT} - P_{IM})$$

$$= -6 + 28.5dBm$$

=***22.5 dBm*** for ERA 6 amplifier

for MAR 3 amplifier

$P_{OUT} = -1dBm$, fundamental signal power

$P_{IM} = -37dBm$, Third order IMD product signal power

As per the equations derived in previous section, the output third order intercept point OIP_3 for MAR 3 amplifier is computed as 5dBm

$$OIP_3 = P_{OUT} + \frac{1}{2}(P_{OUT} - P_{IM})$$

$$= -1 + 18\,\mathrm{dBm}$$

= **17 dBm** for MAR 3 amplifier

8.4 Comparison of Values

The measured parameters for ERA 6 and MAR 3 amplifiers are compared in the table below. It is thus evident that ERA 6 can be chosen to replace MAR 3 in L-Band Front-End receiver to achieve higher dynamic range. The changes in the front-end are measured and compared are included in next chapter in the report.

Device Parameter	MAR 3 amplifier	ERA 6 amplifier
Gain	8 dB	10 dB
P1dB	7.2 dBm	17.8 dBm
Noise Figure	5.6 dB	4 dB
OIP_3	17 dBm	22.5 dBm

Table 8.3 Comparison of MAR 3 and ERA 6 amplifiers

Chapter Nine
Modified Front-End Design

As we have seen the ERA 6 amplifier is a better high dynamic range device as compared to the MAR 3 amplifier, so the existing MAR 3 amplifier stages in post amplifier and phase switch circuit are replaced in the modified Front-End with ERA 6 amplifier stages. The circuit diagrams for the existing and modified phase switch circuit are shown in figure 9.1 and figure 9.2, respectively.

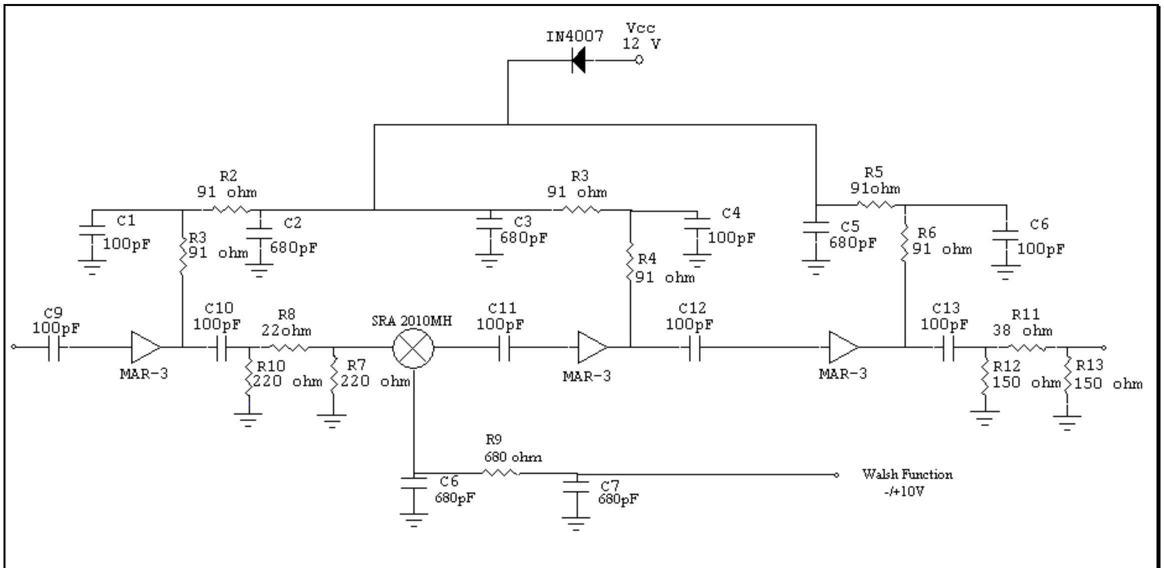


Figure 9.1: Existing Phase Switch circuit diagram

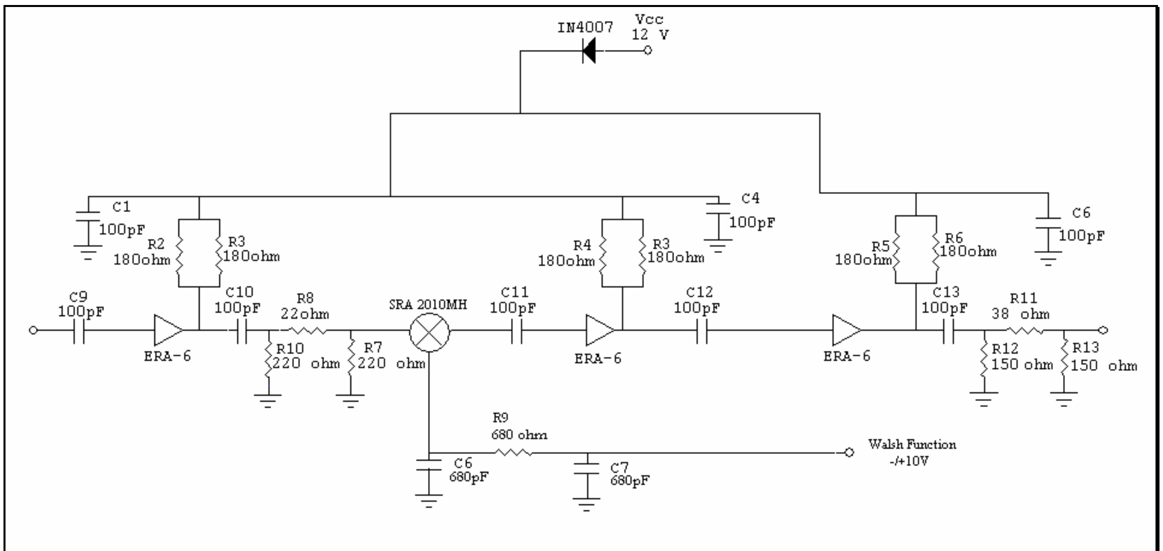


Figure 9.2: Modified Phase Switch circuit

Note: Measurement of the modified Front-End system with narrow BPF bypassed and with the input given to the LNA and the out put taken at the output of the Front-End is done.

9.1 P1db Measurement

The output power signal of the Front-End was measured for varying amplitude of input signal at 1250 MHz. The values obtained by measurement are in table 9.1.

Input power (dBm)	Output power (dBm)
-77.3	-18.9
-76.3	-17.9
-75.3	-16.9
-74.3	-15.9
-73.3	-14.9
-72.3	-13.9
-71.3	-12.9
-70.3	-11.9
-69.3	-10.9
-68.3	-9.9
-67.3	-8.9
-66.3	-7.9
-65.3	-6.9
-64.3	-5.9
-63.3	-4.9
-62.3	-3.9
-61.3	-2.9
-60.3	-1.8
-59.3	-0.8
-58.3	0.2
-57.3	1.2
-56.3	2.1
-55.3	3.1
-53.3	4.1
-52.3	5
-51.3	5.9
-50.3	6.7
-49.3	7.4
-48.3	8

-47.3	8.6
-46.3	8.8
-45.3	9
-44.3	9.2
-43.3	9.2
-42.3	9.2
-41.3	9.2
-40.3	9.2

Table 9.1 Output power values for varying power of input at 1250 MHz for modified L-Band Front End

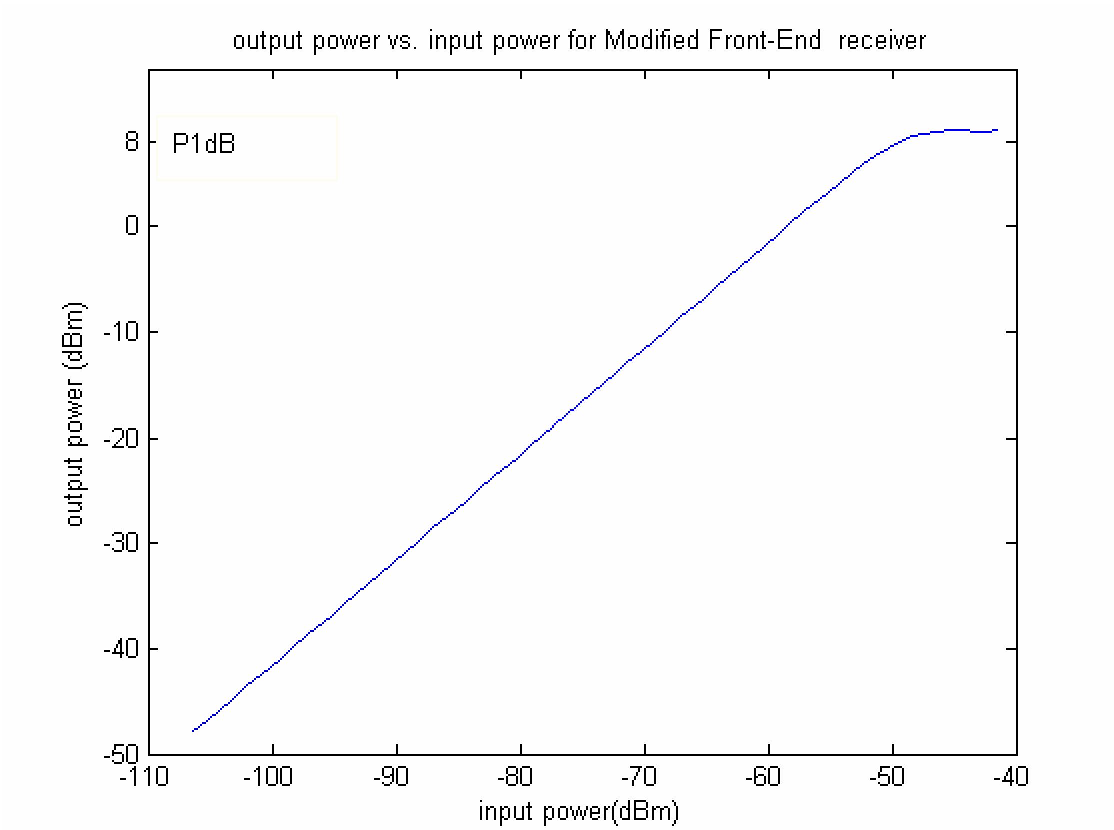


Figure 9.3The output power as a function of input power for modified L-Band front-End

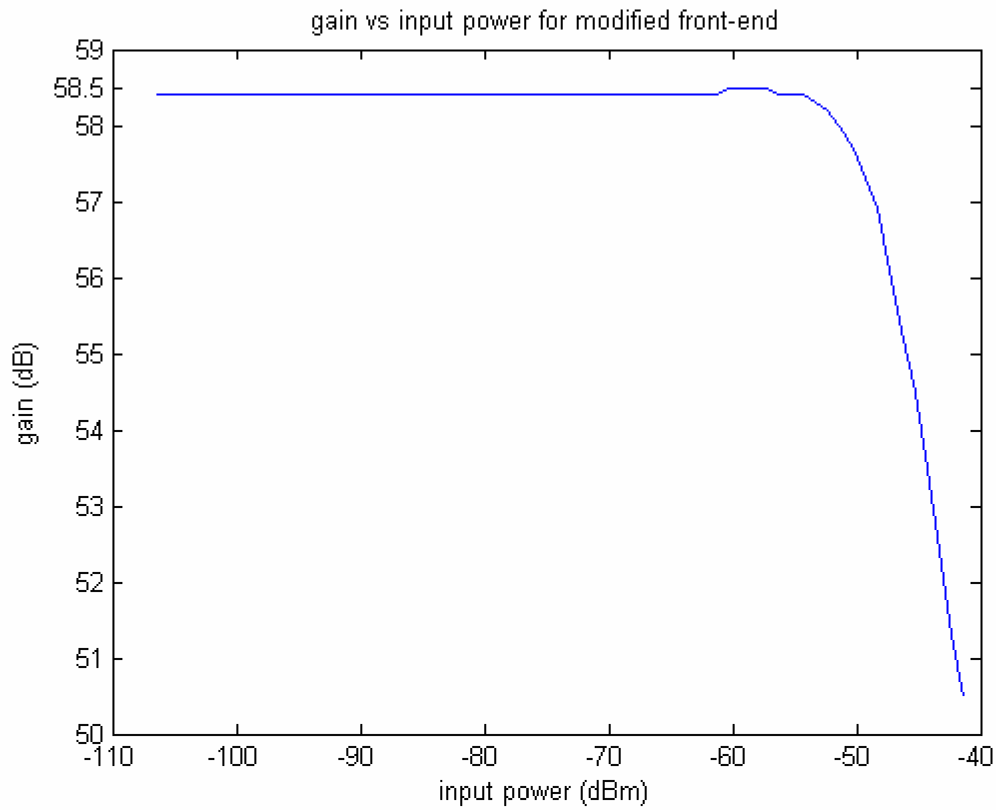


Figure 9.4: The Gain for varying amplitude of input signal.

From the above data, it is found that with increase in power of input signal the output signal increases but after certain point it begins to saturate to certain value.Thus as evident from the recorded data, the **1dB compression point** P_{1dB} of existing front-end is **8 dBm**.

9.2 OIP3 Measurement

To determine intercept points, two signals of equal power P_{IN} at 1245MHz and 1255MHz are input to the LNA. The output power P_{OUT} of the desired signal and P_{IM}, power of one of the third order IMD products at 1235MHz and 1265MHz is measured.

$P_{OUT} = -2.2dBm$, fundamental signal power

$P_{IM} = -45dBm$, Third order IMD product signal power

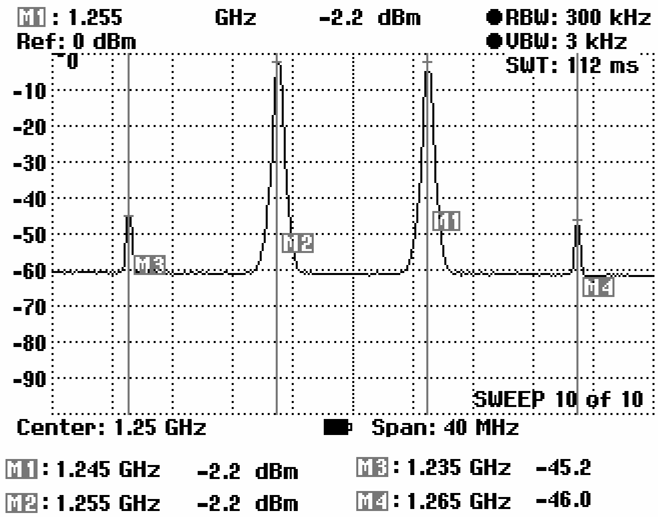


Figure 9.5 IP3 measurement of Modified Front-End

As per the equations derived in previous section, the output third order intercept point **OIP3** is computed as ***19.2 dBm***

$$
\begin{aligned}
OIP_3 &= P_{OUT} + \frac{1}{2}(P_{OUT} - P_{IM}) \\
&= -2.2 + 21.4 dBm \\
&= 19.2 dBm
\end{aligned}
$$

9.3 Noise floor

At the input of the receiver,

$$
\begin{aligned}
Noise\ Floor &= 10\log_{10} kT_{sys} + 10\log_{10} B \\
&\cong -180 dBm \quad \text{for 1Hz bandwidth} \\
&= -105 dBm \quad \text{for 32MHz bandwidth} \\
&= -99 dBm \quad \text{for 120MHz bandwidth}
\end{aligned}
$$

9.4 Compression Dynamic Range (CDR)

The Compression Dynamic Range (CDR) is hence found to be

$$CDR = P_{1dB} - NoiseFloor$$

$$= 8 - (-105 + 58.5)dB$$

CDR = 54.5 dB *for 32 MHz bandwidth*

9.5 Spurious Free Dynamic Range (SFDR)

The spurious free dynamic range (SFDR) is then obtained

$$SFDR = \frac{2}{3}(OIP_3 - MDS_{OUT})$$

$$= \frac{2}{3}(19.2 - (-105 + 58.5))dB$$

$$= \frac{2}{3}(65.7)dB$$

SFDR = ***43.8 dB*** *for 32MHz bandwidth*

Note: The loss due to connecting cables and connectors was found to be negligible and has been assumed to be invariable for L-band frequency range.

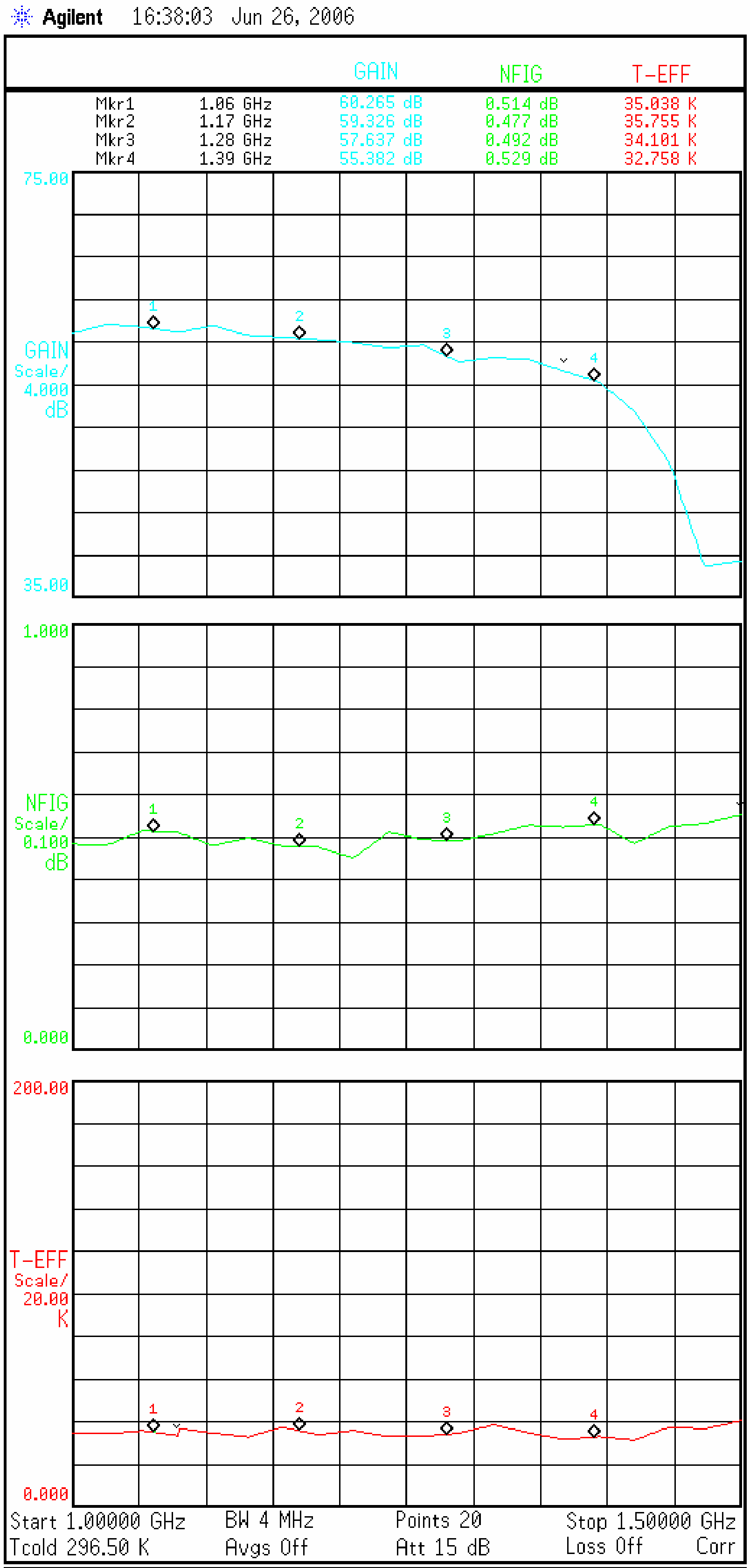


Figure 9.6 The Noise Figure meter display for the modified Front-End system

9.6 RESULT

The measurement for the modified L-Band Front-End receiver is summarized in the following table

Parameter	Value
Gain	58.5 dB
P1dB	8 dBm
Noise Temperature	47 K *(includes cable+OMT losses)*
OIP3	19.2 dBm
CDR	54.5 dB
SFDR	43.8 dB

Table 9.2 Measurement of modified L-Band Front-End

9.7 Comparison of design of Front-End receivers

Parameters	Existing	Modified	Difference
Gain	53.2 dB	58.5 dB	5.3 dB
Noise Temperature	47 K	47 K	No change
CDR	48.8 dB	54.5 dB	5.7 dB
SFDR	38.5 dB	43.8 dB	5.3 dB
Noise Floor	-51.8 dBm	-46.5 dBm	5.3 dBm

Table 9.3: comparison between existing and modified L-Band Front-End

With the modifications in the L-Band Front-End, there is significant increase in the figures of gain Compression Dynamic Range and Spurious Free Dynamic Range. The modification doesn’t produce any change in noise temperature.

Chapter Ten

Proposed Design of GMRT L-Band Front End

We noticed that because of the increase in gain of the modified design the noise floor has increased at the output. Secondly the increase in gain might drive the further stages of the receiver into saturation. Thirdly because of the increase in noise floor the high dynamic range obtained by high p1db and oip3 is not achieved. So using matlab simulation various designs for the phase switch were simulated and the following front-end design gives better and very high dynamic range result. The proposed Phase Switch includes a first 4 dB attenuator then SRA 2010mh mixer followed by two stages of ERA 6 amplifier. For simulation refer appendix A.

Parameters	Existing	Modified	Proposed
Output Noise	**-58.9**	**-31.7**	**-55.5**
Noise Temperature	**35.80**	**35.59**	**35.84**
Gain	**46.1**	**53.3**	**49.5**
CDR	**52.3**	**53.7**	**63.5**
P 1dB	**-6.6**	**2**	**8**
SFDR	**40.56**	**43.47**	**48.3**
OIP3	**1.93**	**13.5**	**16.95**

Table 10.1: Simulation comparison of existing, modified and proposed L-Band Front-End receiver

As we can see the proposed design gives the best result. So it requires redesigning of the existing Phase Switch

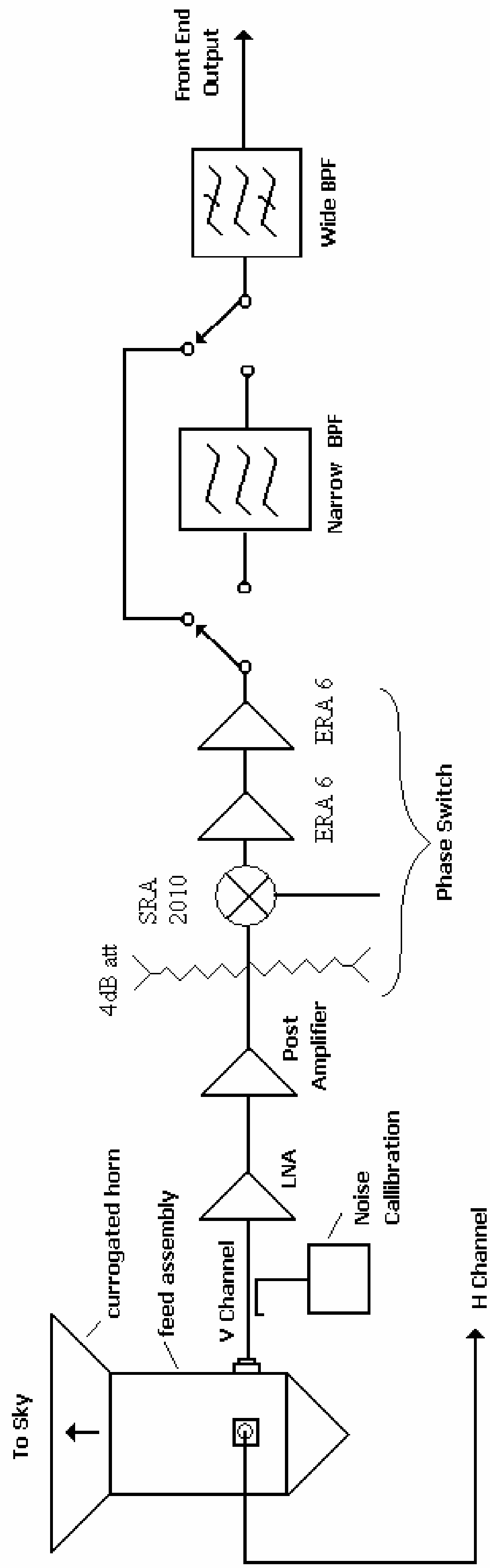


Figure 10.1 Proposed L-Band Front-End Receiver

Conclusion

The Existing Front-End receiver parameters were measured and analysed. Various concepts in RF design have been dealt with good amount of detail. The model for the front-end is also obtained, which can be utilized for simulation of the system. The Scope of the report has been fulfilled with Modified and Proposed Front-End receivers. The compression dynamic range has been improved by 5.7 dB and spurious free dynamic range increased by 5.3 dB. The Proposed Front-End receiver shows a significant increase the values of CDR and SFDR. For the proposed model CDR is 63.5 and the SFDR is 48.3. For the proposed model new phase switch layout has to be designed. However the important factors such as VSWR needs to be considered for the good receiver design.

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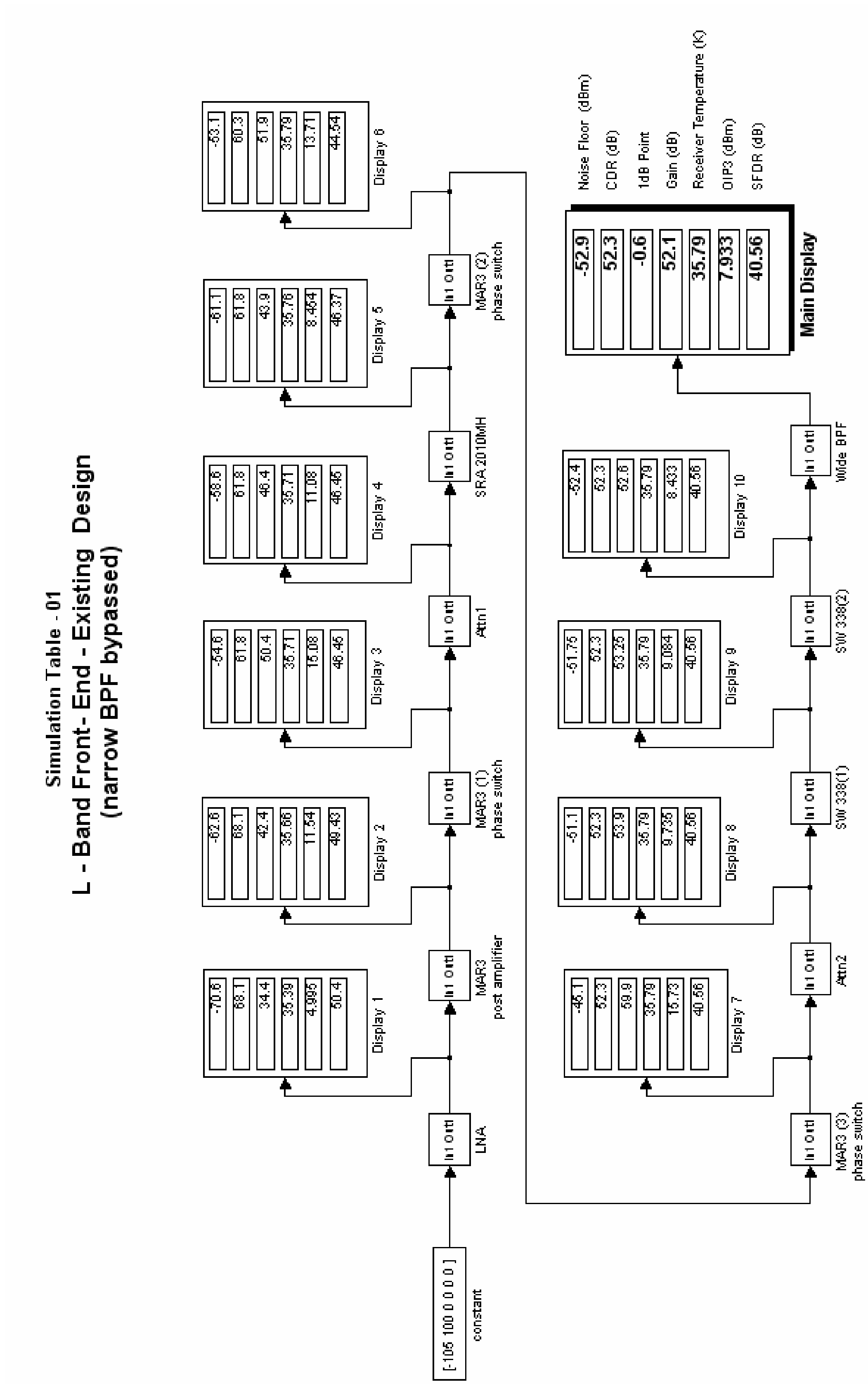
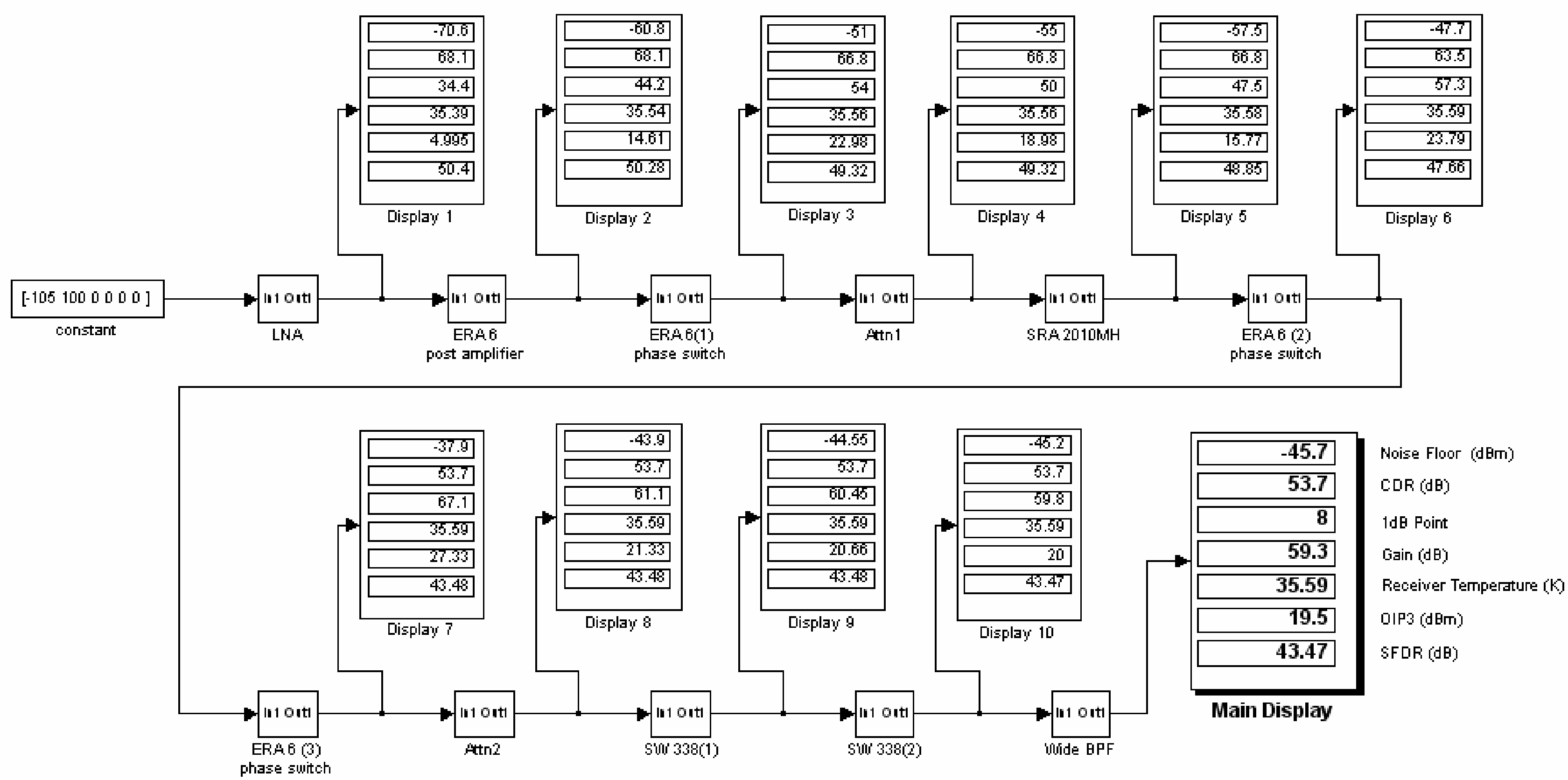
[9] A. Praveen Kumar and M Srinivas, *Signal Flow Analysis of the GMRT Receiver System*, GMRT Internal Technical Report, Oct. 1996.

Appendix A
Matlab Simulation



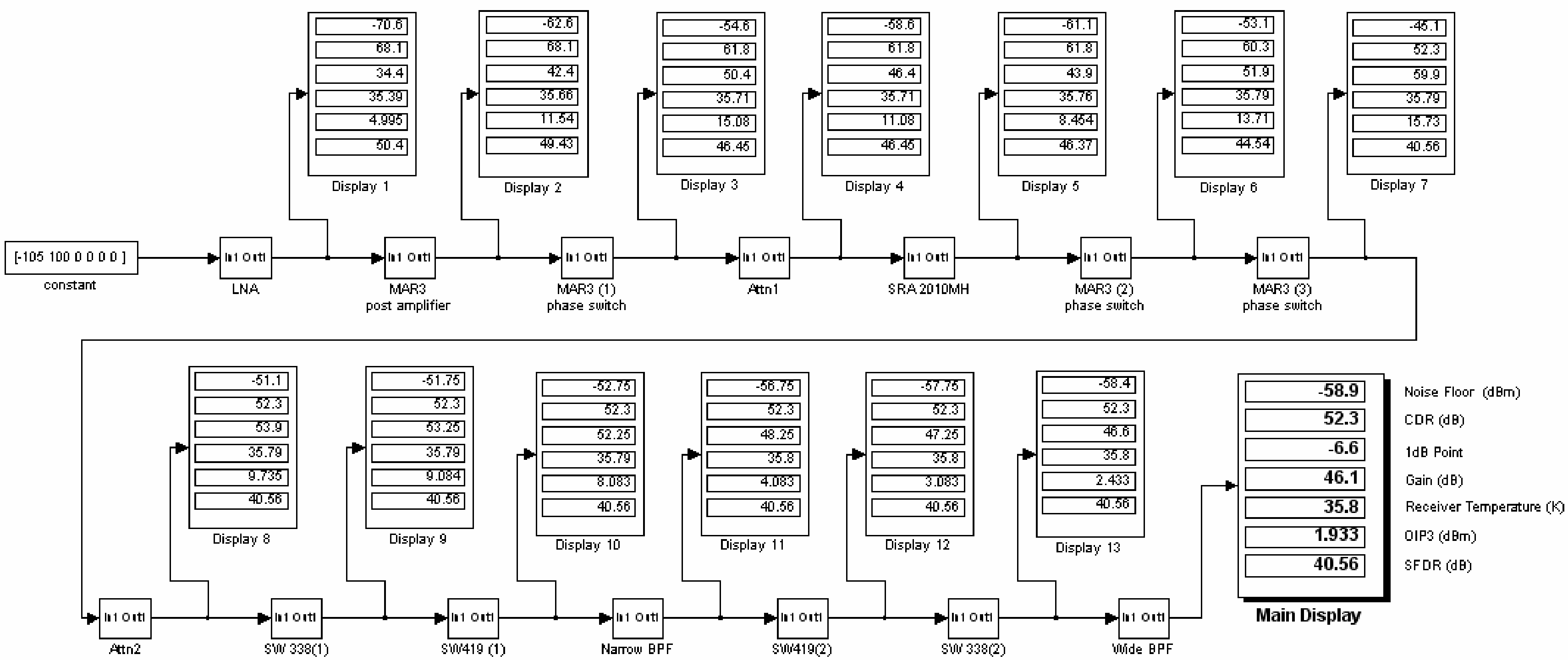
Simulation Table - 02

L - Band Front- End - Modified Design
(narrow BPF bypassed)



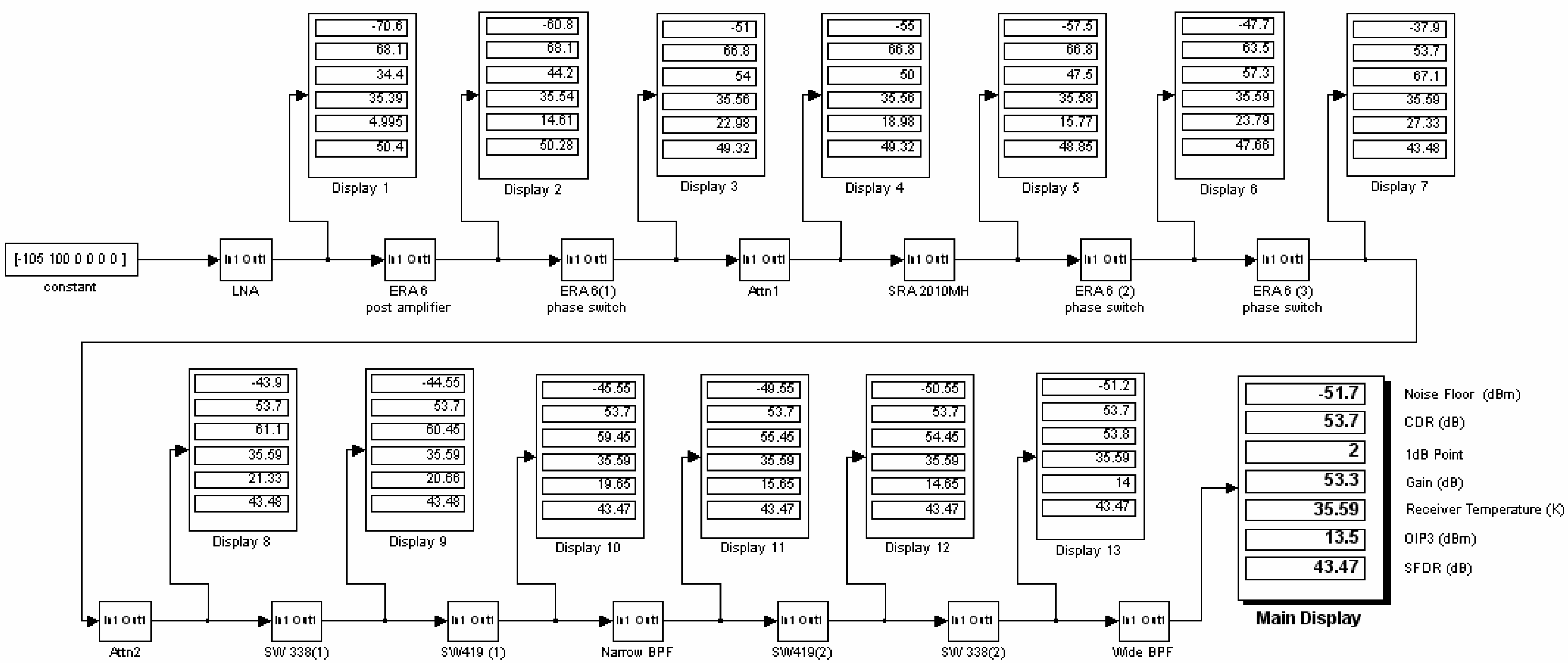
Simulation Table - 03

L - Band Front-End - Existing Design



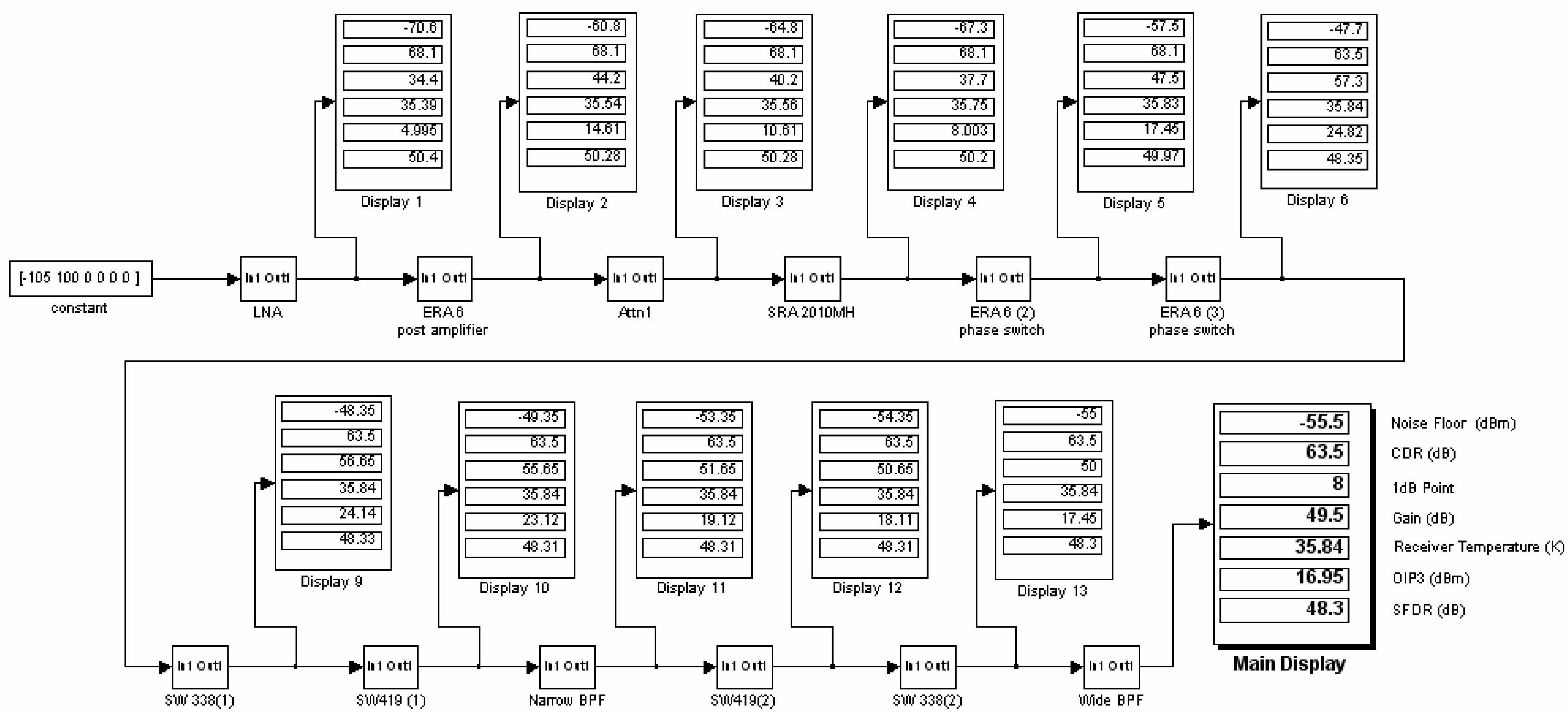
Simulation Table - 04

L - Band Front-End - Modified Design



Simulation Table - 05

L - Band Front-End Proposed Design

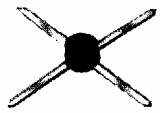


up to 56 mW (+17.5 dBm) output
dc to 2 GHz

case style selection

outline drawings Table of Contents

RAM

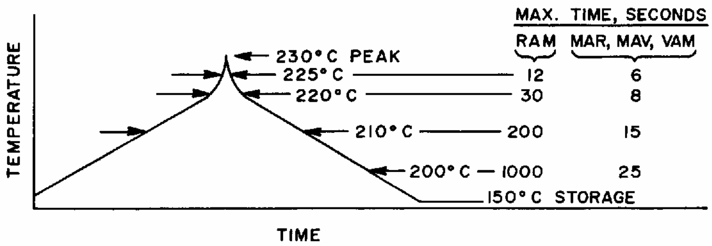


MAR

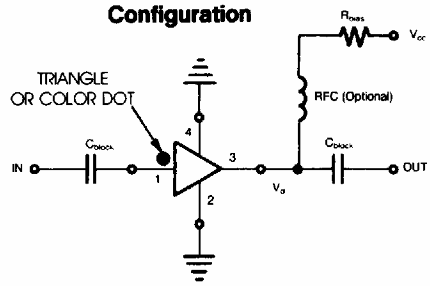
MAR–SM

	MODEL NO.	•FREQ. MHz f_L f_U	GAIN, dB Typical (at MHz) 100	500	1000	2000	(Note 4) MIN	MAXIMUM POWER, dBm Output (1dB Comp.)	Input (no damage)	DYNAMIC RANGE NF dB Typ.	IP3 dBm Typ.	VSWR In	Out	MAXIMUM RATING (25°C) I(mA)	P(mW)	DC POWER at Pin 3 Current (mA)	Volt. Typ.	PRICE $ Qty. (25)	DISTRIBUTOR FACTORY	LOCAL
□ MAR case VV105	MAR-1	DC-1000	18.5	17.5	15.5	—	13.0	+ 1.5	+20	5.5	+14.0	1.3:1	1.3:1	40	200	17	5.0	0.99	●	●
	MAR-2	DC-2000	12.5	12.3	12.0	11.0	8.5	+ 4.5	+20	6.5	+17.0	1.3:1	1.4:1	60	325	25	5.0	1.35	●	●
	MAR-3	DC-2000	12.5	12.2	12.0	10.5	8.0	+10.0	+20	6.0	+23.0	1.5:1	1.7:1	70	400	35	5.0	1.45	●	●
	MAR-4	DC-1000	8.3	8.2	8.0	—	7.0	+12.5	+20	6.5	+25.5	1.6:1	2.0:1	85	500	50	5.25	1.55	●	●
	MAR-6	DC-2000	20.0	18.5	16.0	11.0	9.0	+ 2.0	+20	3.0	+14.5	1.5:1	1.4:1	50	200	16	3.5	1.29	●	●
	MAR-7	DC-2000	13.5	13.1	12.5	11.0	8.5	+ 5.5	+20	5.0	+19.0	1.4:1	1.5:1	60	275	22	4.0	1.75	●	●
	MAR-8	DC-1000	32.5	28.0	22.5	—	19.0	+12.5	+20	3.3	+27.0	#	#	65	500	36	7.8	1.70	●	●
o□ MAR–SM case WW107	MAR-1SM	DC-1000	18.5	17.5	15.5	—	13.0	+ 1.5	+20	5.5	+14.0	1.3:1	1.2:1	40	200	17	5.0	1.04	●	●
	MAR-2SM	DC-2000	12.5	12.3	12.0	11.0	8.5	+ 4.5	+20	6.5	+17.0	1.5:1	1.4:1	60	325	25	5.0	1.40	●	●
	MAR-3SM	DC-2000	12.5	12.2	12.0	10.5	8.0	+10.0	+20	6.0	+23.0	1.5:1	1.7:1	70	400	35	5.0	1.50	●	●
	MAR-4SM	DC-1000	8.3	8.2	8.0	—	7.0	+12.5	+20	7.0	+25.5	1.5:1	1.9:1	85	500	50	5.25	1.60	●	●
	MAR-6SM	DC-2000	20.0	18.5	16.0	11.0	9.0	+ 2.0	+20	3.0	+14.5	1.7:1	1.7:1	50	200	16	3.5	1.34	●	●
	MAR-7SM	DC-2000	13.5	13.1	12.5	11.0	8.5	+ 5.5	+20	5.0	+19.0	1.7:1	1.7:1	60	275	22	4.0	1.80	●	●
	MAR-8SM	DC-1000	32.5	28.0	22.5	—	19.0	+12.5	+20	3.3	+27.0	#	#	65	500	36	7.8	1.75	●	●
o RAM case AF190	RAM-1	DC-1000	19.0	17.5	15.5	—	13.0	+ 1.5	+20	5.5	+14.0	1.3:1	1.3:1	40	200	17	5.0	4.95	●	●
	RAM-2	DC-2000	12.5	12.0	11.8	11.0	8.5	+ 4.5	+20	6.5	+17.0	1.2:1	1.4:1	60	325	25	5.0	4.95	●	●
	RAM-3	DC-2000	12.5	12.2	12.0	10.5	8.0	+10.0	+20	6.0	+23.0	1.6:1	1.7:1	80	425	35	5.0	4.95	●	●
	RAM-4	DC-1000	8.5	8.2	8.0	—	7.0	+12.5	+20	6.5	+25.5	1.4:1	1.9:1	100	540	50	5.25	4.95	●	●
	RAM-6	DC-2000	20.0	19.0	16.0	11.0	9.0	+ 2.0	+20	2.8	+14.5	1.4:1	1.3:1	50	200	16	3.5	4.95	●	●
	RAM-7	DC-2000	13.5	13.1	12.5	11.0	8.5	+ 5.5	+20	4.5	+19.0	2.0:1	1.8:1	60	275	22	4.0	4.95	●	●
	RAM-8	DC-1000	32.5	28.0	23.0	—	19.0	+12.5	+20	3.0	+27.0	#	#	65	420	36	7.8	4.95	●	●

REFLOW - SOLDERING PROFILE

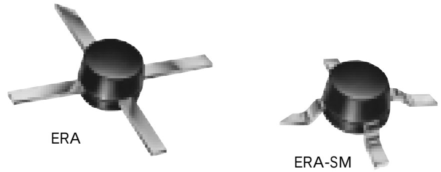


Typical Biasing Configuration



MONOLITHIC AMPLIFIERS 50Ω

BROADBAND DC to 8 GHz



low power, up to +13.5 dBm output

all specifications at 25°C

MODEL◆ NO.	✪ FREQ. GHz	GAIN, dB Typical over frequency, GHz								MAXIMUM POWER (dBm) at 2 GHz*			DYNAMIC RANGE at 2 GHz*		VSWR (:1) Typ.				ABSOLUTE MAX. RATING[3]		DC OPERATING POWER[4] at Pin 3				THERMAL RESISTANCE	CASE STYLE	CONNECTION	PRICE $
	f_L - f_U	0.1	1	2	3	4	6	8	Min.@ 2 GHz	Output (1 dB Comp.) Typ.	Output (1 dB Comp.) Min.	Input (no dmg)	NF (dB) Typ.	IP3 (dBm) Typ.	In DC-3 GHz	In 3-f_U** GHz	Out DC-3 GHz	Out 3-f_U** GHz	I (mA)	P (mW)	Current (mA)	Device Volt. Typ	Device Volt. Min	Device Volt. Max	θjc Typ. °C/W	Note B		Qty. (30)
ERA-1	DC-8	12.3	12.1	11.8	10.9	9.7	7.9	8.2	9	12.0	10.0	15	4.3	26	1.5	1.8	1.5	1.9	75	330	40	3.4	3.0	4.1	178	VV105	cb	1.37
ERA-2	DC-6	16.2	15.8	15.2	14.4	13.1	11.2	—	13	13.0	11.0	15	4.0	26	1.3	1.4	1.2	1.6	75	330	40	3.4	3.0	4.1	155	VV105	cb	1.52
ERA-3	DC-3	22.1	21.0	18.7	16.8	—	—	—	16	12.5	9.0	13	3.5	25	1.5	—	1.4	—	75	330	35	3.2	3.0	4.1	154	VV105	cb	1.67
ERA-1SM	DC-8	12.3	12.1	11.8	10.9	9.7	7.9	8.2	9	12.0	10.0	15	4.3	26	1.5	1.8	1.5	1.9	75	330	40	3.4	3.0	4.1	183	WW107	cb	1.42
ERA-21SM	DC-8	14.2	13.9	13.2	12.2	10.8	8.7	8.9	11.2	12.6	10.6	15	4.7	26	1.1	1.4	1.3	1.9	75	330	40	3.5	3.0	4.1	194	WW107	cb	1.57
ERA-2SM	DC-6	16.2	15.8	15.2	14.4	13.1	11.2	—	13	13.0	11.0	15	4.0	26	1.3	1.4	1.2	1.6	75	330	40	3.4	3.0	4.1	160	WW107	cb	1.57
ERA-33SM	DC-3	19.3	18.7	17.4	15.9	—	—	—	15	13.5	11.5	13	3.9	28.5	1.6	—	1.25	—	75	330	40	4.3	3.8	4.8	140	WW107	cb	1.72
ERA-3SM	DC-3	22.1	21.0	18.7	16.8	—	—	—	16	12.5	9.0	13	3.5	25	1.5	—	1.4	—	75	330	35	3.2	3.0	4.1	159	WW107	cb	1.72
ERA-8SM	DC-2	31.5	25.0	19.0	15.0	12.0	—	—	17	12.5	—	13	3.1	25	1.4	1.8	1.8	2.2	65	250	36	3.7	3.2	4.2	140	WW107	cb	1.22

see suggested PCB layout PL-075 for ERA models

features
- low thermal resistance
- miniature microwave amplifier
- frequency range, DC to 8 GHz, usable to 10 GHz
- up to 18.4 dBm typ. (16.5 dBm min) output power

absolute maximum ratings
operating temperature: -45°C to 85°C
storage temperature: -65° to 150°C

model identification

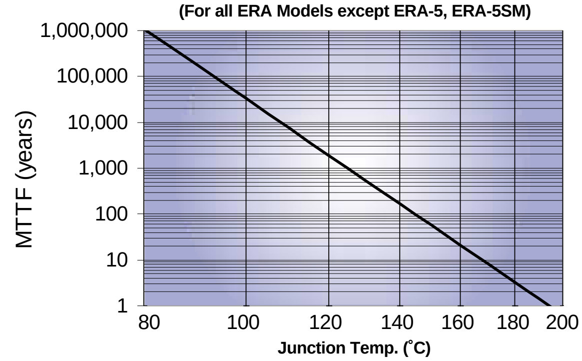
Model	marking (see note below)
ERA-1, ERA-1SM	1
ERA-2, ERA-2SM	2
ERA-21SM	21
ERA-3, ERA-3SM	3
ERA-33SM	33
ERA-4,-4SM	4
ERA-4XSM	4X
ERA-5, ERA-5SM	5
ERA-50SM	50
ERA-51SM	51
ERA-5XSM	5X
ERA-6, ERA-6SM	6
ERA-8SM	E8

Note: Prefix letter (optional) designates assembly location. Suffix letters (optional) are for wafer identification.

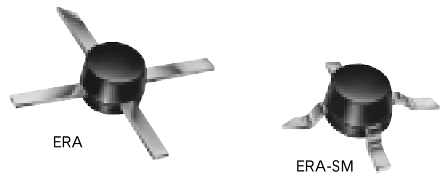
prefix letter
number
suffix letter

NOTES:
- ◆ Aqueous washable
- * at 1 GHz for ERA-4,5,6, 4SM,4XSM ,5SM,5XSM, 50SM, 51SM, 6SM, 8SM.
- ** f_u is the upper frequency limit for each model as shown in the table; for ERA-8SM VSWR (In & Out) is specified at DC-1GHz & 1-4 GHz.
- *** Gain and VSWR are specified at 1.5 GHz.
- ✪ Low frequency cutoff determined by external coupling capacitors.
- A. Environmental specifications and re-flow soldering information available in General Information Section.
- B. Units are non-hermetic unless otherwise noted. For details on case dimensions & finishes see "Case Styles & Outline Drawings".
- C. Prices and Specifications subject to change without notice.
- D. For Quality Control Procedures see Table of Contents, Section 0, "Mini-Circuits Guarantees Quality" article. For Environmental Specifications see Amplifier Selection Guide.
- 1. Model number designated by alphanumeric code marking.
- 2. ERA-SM models available on tape and reel.
- 3. Permanent damage may occur if any of these limits are exceeded. These ratings are not intended for continuous normal operation.
- 4. Supply voltage must be connected to pin 3 through a bias resistor in order to prevent damage. See "Biasing MMIC Amplifiers" in minicircuits.com/application.html. Reliability predictions are applicable at specified current & normal operating conditions.

MTTF vs. Junction Temp.
(For all ERA Models except ERA-5, ERA-5SM)



040618



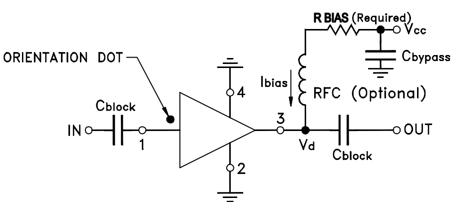
medium power, up to +18.4 dBm output

all specifications at 25°C

MODEL◆ NO.	✪ FREQ. GHz f_L - f_U	GAIN, dB Typical over frequency, GHz 0.1	1	2	3	4	Min.@ 2 GHz	MAXIMUM POWER (dBm) at 2 GHz* Output (1 dB Comp.) Typ.	Min.	Input (no dmg)	DYNAMIC RANGE at 2 GHz* NF (dB) Typ.	IP3 (dBm) Typ.	VSWR (:1) Typ. In DC-3 GHz	In 3-f_U** GHz	Out DC-3 GHz	Out 3-f_U** GHz	ABSOLUTE MAX. RATING[3] I (mA)	P (mW)	DC OPERATING POWER[4] at Pin 3 Current (mA)	Device Volt. Typ	Min	Max	THERMAL RESISTANCE θjc Typ. °C/W	CASE STYLE Note B	CONNECTION	PRICE $ Qty. (30)
ERA-6	DC-4	12.6	12.5	12.2	11.7	11.3	10.5	17.9	16	20	4.5	36	1.3	1.2	1.6	1.8	120	650	70	5.0	4.6	5.6	170	VV105	cb	3.85
ERA-4	DC-4	14.3	14.0	13.4	12.7	11.8	11	17.3	15	20	4.2	34	1.2	1.2	1.3	1.8	120	650	65	4.5	4.2	5.5	163	VV105	cb	3.85
ERA-5	DC-4	20.2	19.5	18.5	16.7	14.3	16	18.4	16.5	13	4.3	32.5	1.3	1.3	1.2	1.3	120	650	65	4.9	4.2	5.5	278	VV105	cb	3.85
ERA-6SM	DC-4	12.6	12.5	12.2	11.7	11.3	10.5	17.9	16	20	4.5	36	1.3	1.2	1.6	1.8	120	650	70	5.0	4.6	5.6	175	WW107	cb	3.90
ERA-4SM	DC-4	14.3	14.0	13.4	12.7	11.8	11	17.3	15	20	4.2	34	1.2	1.2	1.3	1.8	120	650	65	4.5	4.2	5.5	168	WW107	cb	3.90
NEW ERA-4XSM	DC-4	14.7	14.2	13.5	12.0	11.8	12	17.0	15.0	20	4.2	35	1.2	1.2	1.2	1.4	100	650	65	4.5	4.2	5.5	196	WW107	cb	1.69
NEW ERA-5XSM	DC-4	20.5	19.5	17.6	15.5	13.7	16	17.8	16.5	13	3.5	33	1.2	1.3	1.2	1.4	120	650	65	4.9	4.2	5.5	133	WW107	cb	1.69
ERA-51SM	DC-4	18.0	17.4	16.1	14.8	12.5	14	18.1	16.5	13	4.1	33	1.1	1.2	1.2	1.9	120	650	65	4.5	4.2	5.5	154	WW107	cb	3.90
ERA-5SM	DC-4	20.2	19.5	17.6	15.6	14.0	16	18.4	16.5	13	4.3	32.5	1.3	1.3	1.2	1.3	120	650	65	4.9	4.2	5.5	283	WW107	cb	3.90
ERA-50SM***	DC-1.5	20.7	19.4	18.3	—	—	16	17.2	16.0	13	3.5	32.5	1.3	—	1.2	—	120	650	60	4.4	4.0	4.9	177	WW107	cb	2.95

see suggested PCB layout PL-075 for ERA models

typical biasing configuration

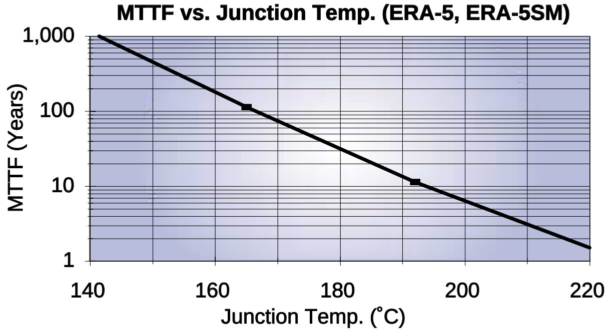


R BIAS

"1%" Resistor Values (ohms) for Optimum Biasing of ERA Models

Vcc	ERA-1 ERA-1SM	ERA-2 ERA-2SM	ERA-21SM	ERA-3 ERA-3SM	ERA-33SM	ERA-4, 4SM,4XSM	ERA-5, 5SM	ERA-50SM, -51SM,5XSM	ERA-6, ERA-6SM	ERA-8SM
7	90.9	88.7	88.7	107	69.8	38.3	33.2	40.2	30.1	88.7
8	113	113	113	133	93.1	52.3	48.7	53.6	43.2	118
9	137	137	137	162	115	66.5	63.4	68.1	56.2	143
10	162	162	162	191	140	80.6	78.7	82.5	69.8	174
11	187	187	187	221	165	95.3	95.3	97.6	84.5	200
12	215	215	215	249	191	110	110	113	97.6	226
13	237	237	237	280	215	127	124	127	113	255
14	261	261	261	309	243	143	140	143	127	280
15	287	287	287	340	267	158	158	158	140	309
16	309	309	316	365	287	174	174	174	154	340
17	332	332	340	392	316	187	187	191	169	365
18	357	365	365	422	340	205	205	205	182	392
19	383	392	392	453	365	221	221	221	196	422
20	412	412	412	475	392	237	232	237	210	453

MTTF vs. Junction Temp. (ERA-5, ERA-5SM)



NSN GUIDE

MCL NO.	NSN
ERA-1SM	5962-01-459-9075
ERA-3SM	5996-01-516-5438
ERA-4SM	5962-01-459-7410
ERA-5SM	5962-01-459-9314

pin connections

PORT	cb
RF IN	1
RF OUT	3
DC	3
CASE GND	2,4
NOT USED	—
DEMO BOARD	ERA-TB

designers kits available

KIT NO.	Model Type	No. of Units in Kit	Description	Price $ per kit
K1-ERA	ERA	30	10 of each 1,2,3	49.95
K2-ERA	ERA	20	10 of each 4,5	69.95
K1-ERASM	ERA-SM	30	10 of each 1SM, 2SM,3SM	49.95
K2-ERASM	ERA-SM	20	10 of each 4SM, 5SM	69.95
K3-ERASM	ERA-SM	30	10 of each 4SM, 5SM, 6SM	99.95

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