An Economical do - it - yourself Ground Station for School Pupils

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Thesis presented in partial fulfilment of the requirements for the degree of Master of Science (Electronic Engineering) at the University of Stellenbosch.

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Declaration

I, the undersigned, hereby declare that the work contained in this thesis is my own original work and that I have not previously in its entirety or in part submitted it at any university for a degree.

Signature

Date

Abstract

The aim of the thesis is the design of an economical do - it - yourself ground station for school pupils to communicate with SUNSAT I. The ground station should also be more economical than a hand - held transceiver radio. The do - it - yourself requirement is there to arouse an interest in electronics, radio frequency electronics and satellite communications in school pupils.

A system-level design was done for a ground station consisting of modules which may be bought individually as do - it - yourself kits to eventually produce a full set. The modules are a VHF receiver, a VHF transmitter, a UHF down converter and a modem. Each module has functions which aid in the process of communications (data as well as voice) between the satellite and the ground station.

A VHF receiver was designed and implemented to be capable of receiving RF signals from SUNSAT I. A crystal controlled oscillator was designed that oscillates with a frequency tolerance of less than or equal to $\pm$ 0.003 % when aligned without the necessary RF equipment. An economical Broadband Signal Generator was implemented with a 74ACT14 logic IC, which may be used to align the receiver. The higher harmonics of a square wave with a fundamental frequency of 4 kHz are used as a RF source.

A sound card was utilised as a modem to receive 1200 baud AFSK (AX.25 protocol) data and the software was also used to display the data on PC. The data was transmitted from another ground station

Opsomming

Die doel van die tesis is die ontwerp van 'n ekonomiese doen - dit - self - grondstasie vir skoolkinders om met SUNSAT I te kan komunikeer. Die grondstasie moet ook meer ekonomies wees as 'n handstelradio. Die doel van die doen - dit - self - beginsel is om die belangstelling in elektronika, RF elektronika and satelliete by skoolkinders aan te moedig.

'n Stelsel ontwerp van 'n grondstasie is gedoen wat bestaan uit modules wat afsonderlik as doen - dit - self - modules aangeskaf kan word om so tot 'n totale grondstasie op te bou. Die modules is die "VHF" - ontvanger, "VHF" - sender, UHF - afmenger en 'n modem. Elke module verskaf funksies wat bydra om met SUNSAT I te kan kommunikeer.

'n VHF - ontvanger wat in staat is om RF - seine vanaf SUNSAT I te ontvang is ontwerp en gebou. 'n Kristal beheerde ossillator is ontwerp met 'n frekwensie toleransie van kleiner en gelyk aan $\pm$ 0.003 % wanneer dit ingestem word sonder die nodige RF toerusting. 'n Ekonomiese wyeband - seingenerator is gemplementeer met 'n 74ACT14 logiese vlokkie om as 'n RF - bron gebruik te word om die ontvanger in te stel. Die boonste (hoër) harmonieke van die 4 kHz vierkantsgolf word as 'n "RF bron" gebruik.

'n Klankkaart is suksesvol gebruik as 'n modem om 1200 baud AFSK data (AX.25 protokol) te ontvang en die data met die nodige sagteware op 'n skerm te vertoon. Data is uitgestuur vanaf 'n ander grondstasie.

This thesis is dedicated to my father who would have enjoyed this project as much as I did.

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- Ron Olivier for his advice on thesis writing and his encouragement.

Contents

List of Acronyms

AC	Alternating Current
AFSK	Audio Frequency - Shift - Keying
b/s	bits per second
BPF	Bandpass Filter
CW	Continuous Wave
DC	Direct Current
DUT	Device Under Test
FM	Frequency Modulation
GSSP	Ground Station for School Pupils
IMD	Intermodulation Distortion
IC	Integrated Circuit
IF	Intermediate Frequency
LO	Local Oscillator
LED	Light Emitting Diode
NBFM	Narrow Band Frequency Modulation
PC	Personal Computer
PCB	Printed Circuit Board
PSD	Power Spectral Density
PTT	Push to Talk
RF	Radio Frequency
RSSI	Real Signal Strength Indicator
RX	Receive/Receiver
TNC	Terminal Node Controller
THD	Total Harmonic Distortion
TX	Transmit/Transmitter
SUNSAT	Stellenbosch University Satellite
SUNSTEP	Stellenbosch University Schools Technology Electronics Program
SINAD	Signal plus Noise, Distortion to Noise and Distortion
SNR	Signal to Noise Ratio
SWR	Standing Wave Ratio
UHF	Ultra High Frequency
VHF	Very High Frequency

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Chapter 1

Introduction

When most people think of electronics and technology, words like internet, IT, cell phones and satellites come to mind as these technology fields are rapidly growing.

Engineers, and especially electronic engineers, are part of the process of growing technology but school children in South Africa have not always had the opportunity to gain experience in electronic fields. The proper motivation for taking subjects such as maths and science was not offered at all schools in South Africa, especially in the previously disadvantage communities. School pupils should therefore be properly motivated to take maths and science. Experience in electronic related fields can encourage school pupils to take subjects like maths and science.

SUNSAT I (Stellenbosch University Satellite) was launched on 23 February 1999, and is the first South African satellite to be launched. One of the initial goals of the SUNSAT Programme in 1992 was to create an interest in science and technology amongst school children [38].

SUNSAT I was declared non-operational on 1 February 2001 [38]. The ground station was mainly designed for operating the SUNSAT I micro-satellite. Other Amateur Radio satellites can also be utilised by the ground station if they have similar capabilities to SUNSAT I.

SUNSAT I hosted school experiments which are part of creating an interest in electronics among school children. A Parrot Repeater which was intended for school pupils was also on board SUNSAT I . The up link frequency on 145.825 MHz was recorded for 10 s and it was then played back on the down link frequency of 145.825 MHz [38].

The MTN - SUNSTEP (Stellenbosch Schools Technology in Electronics Programme) project was launched in 1997. The project realised SUNSAT's aim to promote science and technology amongst school children. The focus was mainly on schools from previously disadvantaged communities. Nearly 40,000 children have been reached. School Pupils are taught the basics of electronics and are given electronic kits to assemble [38].

SUNSTEP does not have a kit that is able to receive RF signals from a micro-satellite or to transmit RF signals to a micro-satellite e.g. SUNSAT I.

Hand-held radios could be used to communicate with SUNSAT I and to operate the parrot repeater mode. The prices for dual-band hand-held radios can be anything from $ 329 Alinco DR605TQ to $ 489.99 for ICOM IC 2800H [31]. Dealers in South Africa quoted R 2000 and more for a dual band transceiver radio. The quote for a single band VHF hand-held transceiver was approximately R 1200.

Hand-held radios are not affordable and accessible to a great part of the population, especially school children. As part of the drive to stimulate interest in electronics and satellite communications amongst all school children, irrespective of the economic status of their parents, it was thought expedient to use the wide publicity of SUNSAT I to investigate more economical ways of providing school children with access to the satellite.

The most expensive kit at SUNSTEP is R 36. Only 1000 of these kits are sold a year according to Marinda Myburgh at SUNSTEP. Ten thousand children were reached in 2000 [39]. Therefore the higher the price of a ground station to access the satellite, the fewer school pupils will be able to afford the ground station.

This project was conceived as being the 'top end' of the SUNSTEP school kit range, and is expected only to be built by those wishing to follow a career in electronics.

1.1 Task Definition

The aim of this project is the design of a ground station that will communicate with an operational SUNSAT I, but which is more economical than a hand-held radio with the same capabilities (e.g. frequency, output power, sensitivity). The school pupils have to assemble and align the ground station themselves. This thesis completes the system level design of the ground station as well as the implementation of the first module of the ground station.

The following communication functions between the ground station and an operational SUNSAT I have to be achieved in this thesis:

- Listen to an Amateur Radio downlink frequency from SUNSAT I. The default downlink frequencies on an operational SUNSAT I were 145.825 MHz, 436.25 MHz and 436.3 MHz . Other downlink frequencies in the Amateur Radio frequency band were attainable by programming the synthesizers on the satellite. Narrow band Frequency Modulation (NBFM) was implemented on SUNSAT I.

- Supply the necessary software and/or hardware to display 1200 baud AFSK data (AX.25 protocol) on a PC.

- Transmit on one of the Amateur Radio uplink frequencies of an operational SUNSAT I. The uplink frequencies were 145.825 MHz, 145.850 MHz, 436.291 MHz and 436.24 MHz.

- Operate the Parrot Repeater mode. In this mode in SUNSAT I the uplink on 145.825 MHz was recorded for 10 seconds. The recorded audio was then transmitted on the downlink of 145.825 MHz.

- Operate one of the voice/data repeater modes (Mode J or B) on an operational SUNSAT I. The uplink was on 436.291 MHz and the downlink on 145.825 MHz for Mode B. The uplink was on 145.825 MHz and the downlink on 436.250 MHz for Mode J.

The functions listed above were selected to introduce various aspects of the satellite and RF communications. These include listening to voices of Radio Amateurs on a downlink, being able to operate the Parrot Repeater mode which was intended for the school pupils, and operating the Mode J Repeater mode (data or voice). Although the 9600 baud G3RUH modems were active on SUNSAT I, it was decided not to set their use as a goal for the ground station due to the fact that 1200 baud modem is easier to implement. A smaller IF bandwidth is required for 1200 baud than for 9600 baud. The SNR increases as the bandwidth is decreased for a fixed RF input, giving better performance at low

1.2 Thesis Presentation

The presentation of the thesis is outlined below:

- The need for a Ground Station for School Pupils (GSSP) and the task definition are discussed in chapter one.
- The system level design and specifications of the ground station are given in chapter two.
- The implementation of the VHF receiver which is, by default, a fixed frequency receiver, the method of applying a "RF source" to the receiver IC and the design of an oscillator that oscillates with a frequency tolerance of less than or equal to 3 kHz when it is tuned without the necessary equipment, are discussed in chapter three.
- The utilisation of a sound card as a modem to receive 1200 baud AFSK data is explained in chapter four.
- The quantitative evaluation of the VHF receiver (e.g. sensitivity, two signal spurious responses and other measurements) is described in chapter five.
- The final conclusions and recommendations for the future development of the ground station are given in chapter six.

Chapter 2

System Level Design and Specifications

2.1 Basic Block Diagram

Figure 2.1 shows a basic block diagram of the ground station.

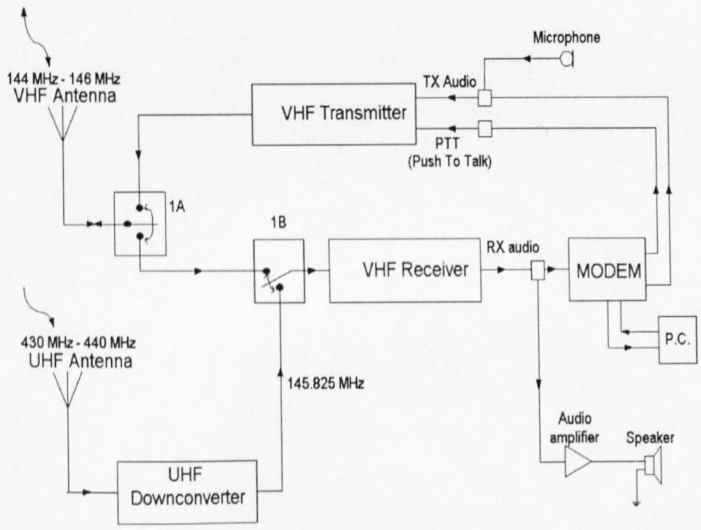

Figure 2.1: Basic blockdiagram of Ground Station for School Pupils

The system level design was done so that communication functions, set out in chapter 1, are met as well as helping to keep the ground station an economical and a do - it - yourself one. The system level design is discussed in the paragraphs below.

The ground station is divided into four modules to establish a building block concept. The modules can be bought individually as do - it - yourself kits to eventually make up a full set. They are a VHF receiver, a VHF transmitter, a UHF down-converter and a modem. A license which allows the scholar to listen on the 2 m and 70 cm Amateur Radio Frequency bands is required for the VHF receiver and UHF down-converter. A Radio Amateur license to transmit on VHF is required before the VHF transmitter can be bought.

The VHF receiver is the first module to be bought. The receiver is by default a fixed frequency receiver which can be expanded to a variable frequency receiver. The fixed frequency is 145.825 MHz, which was the operational downlink VHF frequency on SUNSAT I. The VHF antenna should be obtained with the VHF receiver.

The sequence of the addition of the UHF down-converter, the modem and the VHF transmitter module can vary. The purpose of each module is discussed in the paragraphs below.

The UHF down-converter gives the advantage of reducing unnecessary redundancy in the ground station by converting the UHF down to VHF so that it can be fed to the VHF receiver. The output frequency of the UHF down-converter is on 145.825 MHz. The UHF antenna should be acquired at the same time as the UHF down-converter.

Switch 1B is part of the UHF down-converter module. Switch 1B can be just a connector with a 50 Ω dummy load (resistive) so that the cable from the UHF antenna or switch 1A can be connected to switch 1B. The goals set in chapter 1 do not include the option of receiving both at VHF and UHF simultaneously or switching in between VHF and UHF during the passing of the satellite.

The modem converts the analogue signals which are received from the VHF receiver so that they can be utilised by the applicable software on the Personal Computer (P.C.), and converts the data from the PC into an analogue signal and feeds it to the VHF transmitter to be transmitted.

The VHF transmitter, in co-operation with the VHF receiver or UHF down-converter, is essential in operating the Parrot Repeater mode and the Mode J Repeater (voice and data) mode.

Switch 1A is part of the VHF transmitter module which should connect the VHF antenna by default to the VHF receiver. The VHF transmitter should be connected to the VHF antenna during transmission on VHF and the VHF receiver should be protected from the VHF transmitter.

A transceiver radio cannot be bought in separate modules to build up a transceiver radio such as the ground station. The option to first buy a scanner and later a transceiver when the pupil has gained a Radio Amateur License to transmit is not an economical manner of accessing the Amateur Radio World. This is just another example of illustrating the advantage of the ground station.

2.2 General System Specifications

The communication environment between the satellite and the ground station and the specifications of the satellite will determine the communication specifications. The goal for maximum time of communication is to give the specifications that will allow communication from zero degree elevation (arrival of signal) to zero degree elevation (loss of signal), when the satellite and ground station are in line of sight of each other.

The different factors in the communication environment are discussed and/or mentioned in the paragraphs below.

We start at the satellite where 5 W (37 dBm) is transmitted on VHF. The received signal strength at the receiver of the ground station is influenced by the loss in signal due to the distance the radio waves have to travel (path loss, free space attenuation), Faraday rotation, multiphase due to ground reflections, atmospheric absorption, spin modulation and gain of antennas (satellite and ground station) [20].

The signal to noise ratio at the ground station is determined by the ratio of the received signal power to the noise power at the ground station. Noise is contributed by the receiver, man made sources and natural noise sources. Equation 2.1 [20] shows how the noise at the R.F. input of the receiver can be calculated when it is connected to an antenna.
T_a is the noise received by the antenna, $T_{receiver}$ is the noise generated by the receiver, B the IF bandwidth and k is Boltzmann's constant.

$$Pa = k(T_a + T_{receiver})B \tag{2.1}$$

The SNR needed for communications with minimal data errors and a clear audio signal is discussed in the paragraphs below.

Narrow band frequency modulation (NBFM) is used on the ground station for the receiver because it was used on an operational SUNSAT I. The FM Discriminator should operate above the threshold level to give a clear audio output and minimize the error probabilities for non coherent FSK data communications. The appearance of spike noise is an indication that the discriminator is operating in the threshold region. There is no appearance of spike noise in the discriminator output for a predetection signal-to-noise ratio (signal-to-noise ratio in IF) of 10 dB according to [28].

Equation 2.2 shows the relation between the SNR at the IF at the predetection section and the audio output of the discriminator. SNR_{yd} represents the SNR at the audio output and SNR_{e1} represents the SNR at the predetection section. In equation 2.2 A_c is the amplitude of the carrier, N_o is the noise power, B_T is the IF bandwidth, W is the audio bandwidth, D is the frequency deviation for the maximum value of $|m_n|$. The SNR at the audio output is greater than or equal to the SNR at the predetection section for an IF bandwidth equal to 12 kHz, an audio bandwidth equal to 3 kHz and $3D^2\overline{m_n^2(t)} \geq 0.250$ [28]. A 10 dB SNR at the IF before the demodulator will therefore produce a SNR of greater than or equal to 10 dB SNR at the the lowpass filter of the audio output of the receiver.

$$\frac{SNR_{yd}}{SNR_{e1}} = \frac{3D^2\overline{m_n^2(t)}A_c^2/2N_oW}{A_c^2/2N_oB_T} = \frac{3D^2\overline{m_n^2(t)}B_T}{W} \tag{2.2}$$

SINAD (Signal plus noise, distortion to noise and distortion) is used as a means of expressing the demodulated signal quality of FM receiver systems. SINAD is measured at the output of the filter at the audio output of the receiver. A 12 dB SINAD is approximately equal to a 12 dB Signal to Noise Ratio (SNR) and a total harmonic distortion (THD) of 25% [10]. A 12 dB SNR at the IF section before the demodulation section will thus assure a clear audio signal and minimal errors for data communication.

Equation 2.3 shows how the noise factor is calculated for e.g. amplifiers or receivers, where F is the noise factor, G is the gain of the amplifier or receiver, N_o is the noise contributed by the amplifier or receiver, N_i is the available input noise power and S_i is the signal power at the input of the receiver or amplifier. Equation 2.4 gives the equation to calculate the noise factor for a cascaded configuration such as a dual conversion receiver.

$$F = (S_i/N_i)/(GS_i/(GN_i + N_o) = (GN_i + N_o)/N_i \tag{2.3}$$

$$F_t = F_1 + (F_2 - 1)/G_1 + (F_3 - 1)/(G_1G_2) \tag{2.4}$$

The SNR at the IF stage before the FM demodulator can be concluded to be approximately the same as the SNR at the input of the receiver if $GN_i >> N_o$ in equation 2.3 or $F_1 >> (F_2 - 1)/G_1 + (F_3 - 1)/(G_1 G_2)$ in equation 2.4.

According to the conclusion made and the discussions in previous paragraphs on SNR in the FM receiver the conclusion was made that a 12 dB SNR at the input of the receiver will produce a 12 dB SNR at the output of the audio lowpass filter. As already mentioned 12 dB SINAD at the audio output is equal to 12 dB SNR and a total harmonic distortion of 25%.

The received signal at the ground station is discussed in the paragraphs below.

The received signal at VHF at the ground station is greater than - 120 dBm for 85 % of the pass with an elevation from 0 degrees to 180 degrees according to the calculations of [9]. The calculations according to [9] were made for the VHF turnstile antenna on SUNSAT I, monopole antenna at the ground station, maximum path loss at VHF, minimal transmitted power at VHF of 1 W and maximum losses due other losses such as polarisation mismatch, atmospheric absorption and ground reflection. The percentage of the pass for which the received signal is greater than or equal to - 120 dBm will increase to 95% when the transmitted power at the satellite is 5 W instead of 1 W at VHF. A receiver with a 12 dB SNR for a - 120 dBm RF input in a 12 kHz measuring bandwidth will be sufficient for 95% of a pass for communication when SUNSAT I transmits 5 W at VHF.

The received signal at UHF at the ground station was determined with the software from [9] and information from [20]. At the satellite a UHF turnstile was used as an antenna and a monopole was used as an antenna at the ground station. The received signal at the ground station is greater than - 125 dBm for 75 % of the pass with an elevation of 0 degrees to 180 degrees for 2 W transmitted at the satellite. The received signal at the ground station is greater than - 125 dBm for 85 % and is greater than - 120 dBm for 75 % of the pass with an elevation of 0 degrees to 180 degrees for 8 W transmitted at the satellite. A UHF receiver with a 12 dB SNR for a - 120 dBm RF input in a 12 kHz measuring bandwidth will be sufficient for 75% of a pass for communication when SUNSAT I transmits 8 W at UHF.

2.3 Antennas

The specifications for the VHF and UHF antennas for the ground station are discussed in the paragraphs below.

The antennas should be constructed by the school pupils themselves and should therefore be as simple as possible without the need to tune the antenna with any equipment.

A 50 cm length of wire or a longer length of wire connected to the receiver may work as an antenna for the VHF receiver or UHF down-converter but it may not be effective to transmit RF signals. An antenna with a SWR of 1.9 will ensure that only 10% or less of the power is reflected when it is transmitted or received. It would be possible to build an antenna with an SWR of less than or equal to 1.9 without the necessary equipment to tune it for the lowest SWR.

The school pupil should not have to point the antenna in the direction of the satellite as it passes over. An antenna which has an omnidirectional radiation pattern in the horisontal plane can be used.

The VHF and UHF antennas on SUNSAT I are linear and right hand circular polarised [20]. The antennas at the ground station should therefore be vertically polarised or circularly polarised.

Table 2.1 gives the specifications for the antennas (VHF as well as UHF). The reasons for the selection of the specifications were discussed in the previous paragraphs.

Table 2.1: Specifications for the VHF and UHF antennas

Radiation Pattern	Omnidirectional
Polarisation	Vertical or Right hand circular
SWR	≤ 1.9

The recommendation for VHF and UHF is monopole antennas. The construction of a monopole should be easy if the person knows how to work with brass rods, coaxial cables and other types of wire rods which are used to construct a monopole antenna. Experiences with a monopole (1/4 wave) led to the conclusion that the SWR can be less than 1.9 if instructions for building the antenna and for bending the wire rods which act as a "ground plane" are given.

There are different monopole antennas. The vertical mobile antenna of [36] is recommended for a VHF monopole antenna. The vertical mobile antenna performs better than a monopole (1/4 wave) antenna according to [36] and does not have a null at 35 degrees like a monopole (5/8 wave) antenna.

The recommendation for a monopole antenna at UHF is either a 1/4 wave when a greater number of passes have a maximum elevation of less than 30 degrees or a 3/4 wave when a greater number of passes have of maximum elevation of higher than 30 degrees. The recommendations for the UHF antenna were made from the data obtained from [37].
A monopole (1/4 wave) should be selected as a first option for an antenna at UHF when the scholar is uncertain what the number of passes are for which the maximum elevation of the satellite is higher than 30 degrees.

2.4 VHF Receiver

This section describes the receiver developed for the project. Table 2.2 lists the main specifications and they are motivated in the rest of this section. The specifications were selected in order for school pupils to cope easily with them, yet simultaneously ensuring that the necessary requirements for receiving RF signals on VHF from an operational SUNSAT I are not lacking.

Table 2.2: Specifications for the VHF receiver

Type	VHF FM (Narrow band) Dual Conversion Receiver
Frequency Fixed (Default) Variable	145.825 MHz 144 - 146 MHz
Frequency Resolution	25 kHz
Intermediate Frequencies (IF)	10.7 MHz 455 kHz
IF Bandwidth (3 dB)	12 kHz
Audio Bandwidth	3 kHz
Sensitivity	12 dB SINAD for a - 117 dBm FM Modulated RF signal, 3 kHz deviation
Noise Figure	4 dB
Image Rejection	80 dB
LO Radiation at Antenna	$\leq$ - 80 dBm
Spurious Rejection	80 dB

2.4.1 Receiver Type

Figure 2.2 shows a block diagram of a dual conversion receiver which meets the needs of this project. The RF section, the first IF section and the second IF section consist of amplifiers and/or filters to attenuate the unwanted signals and amplify the desired signals. The choice of a dual conversion receiver instead of a single conversion receiver is discussed in section 2.4.4 .

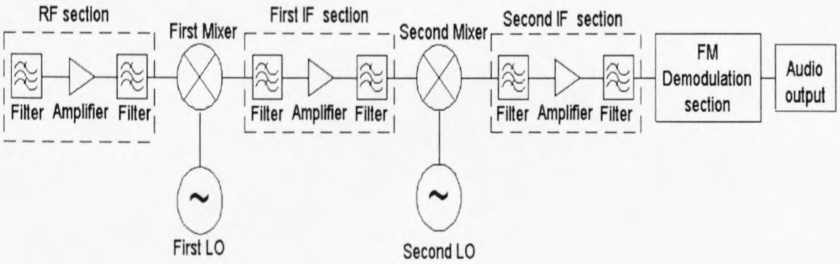

Figure 2.2: Configuration of a dual conversion receiver

2.4.2 Frequency

A fixed frequency design was selected to eliminate the need for the scholar to set the receiver to the correct frequency. The option is included in the receiver to tune to variable frequencies to allow the pupil to listen to other Amateur Radio frequencies from 144 to 146 MHz.

2.4.3 Sensitivity and Noise Figure

Table 2.3 shows the noise floor and noise figure to ensure a 12 dB SNR at the input of the receiver for the indicated RF input and a 12 kHz measured bandwidth. Equation 2.5 was used to calculate effective noise temperature (T_e) so that the noise figure could be determined with equation 2.6. The values for T_0 = 290 K, k = 13.0807 $\times$ 10^{-24} and BW = 12 kHz were used in equations 2.5 and 2.6.

$$Noise\ Floor\ = k(T_0 + T_e)BW\,[\text{dBm}] \tag{2.5}$$

$$NF = 10 * log_{10}(F) = 10 * log_{10}(1 + T_e/T_0) \tag{2.6}$$

Table 2.3: Noise Floor and Noise Figure values

Received Signal Strength [dBm]	T_e[K]	Noise Figure [dB]	Noise Floor [dBm]
- 116.36	586	4.8	- 128.36
- 120	90.8	1.18	- 132

The value of $T_0 = 290$ K for the calculations in Table 2.3 is not always valid for the noise temperature of an antenna (T_a). The noise temperatures for an omnidirectional antenna can be anything from $T_a = 1000$ K due to the average sky noise (galactic noise) temperature at VHF according to [24] to $T_a = 30\ 000$ K due to suburban man-made sources [12].

The noise factor indicates to what extent the (Si/No)/(So/No) degrades due to the noise factor of the receiver. Figure 2.3 shows the degradation in SNR vs. NF of the receiver for various values of T_a. The noise figure of the receiver on the x axis in Figure 2.3 is calculated with $T_0 = 290$ K. The effect T_e has on the degradation of the SNR decreases as T_a increases and the opposite is also true.

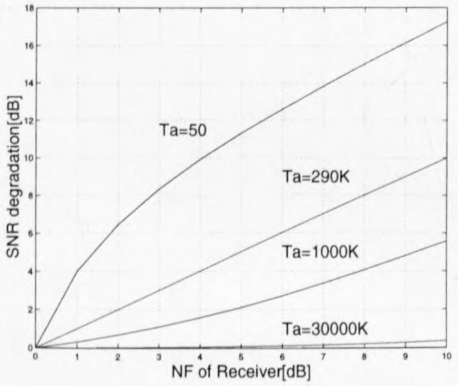

Figure 2.3: Effective Noise Figure

Figure 2.3 shows that the SNR only degrades by 1.6 dB for $T_a = 1000K$ for a noise figure of 4 dB. A degradation of 1.6 dB will not be heard by school pupils when they receive signals from an operational SUNSAT I. A noise figure of 4 dB is easily achieved with a monolithic amplifier from [14] and a bandpass filter to suppress the unwanted frequencies (spurious frequencies). These are the reasons why a noise figure of
4 dB was selected for the VHF receiver.

2.4.4 Intermediate Frequency (IF)

The reasons for the choice of dual conversion receiver, as well as the IF bandwidth and the audio bandwidth are discussed in this section.

Figure 2.4 shows the frequency diagram for dual conversion from 145.825 MHz (RF) to 455 kHz (IF2). The Radio Frequency (RF) is down converted to the first Intermediate Frequency (IF1) by the first Local Oscillator (LO1). The first Image Frequency (IF1) is a spurious response which mixes down to the first IF. The first Intermediate Frequency is down converted to the second Intermediate Frequency (IF2) by the second Local Oscillator (LO2).

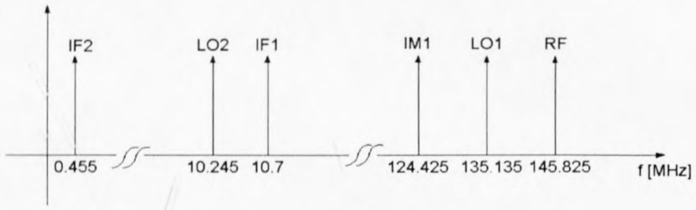

Figure 2.4: Frequency diagram for dual conversion

Figure 2.5 shows the frequency diagram of single conversion from 145.825 MHz (RF) to 455 kHz (IF1). The acronyms used in Figure 2.5 are the same as those defined for Figure 2.4.

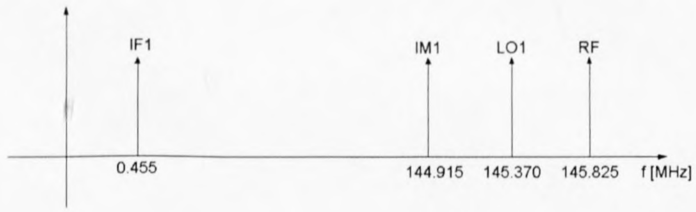

Figure 2.5: Frequency diagram for single conversion

The image frequency for single conversion from 145.825 MHz down to 455 kHz will be in the pass band of the RF section of Figure 2.2 for a pass band of 2 MHz but not for a pass band of 150 kHz. A loaded Q = f_0/BW = 972 is required for a pass band of 150 kHz. LC filters will not be able to have an unloaded Q of such high value. Crystal filters will be able to produce such narrow bandwidths as 150 kHz, but crystal filters are more expensive than LC filters.

A dual conversion receiver was selected due to the fact that a 2 MHz pass band of the RF section will exclude the image frequencies and that LC filters are less expensive than crystal filters.

A first IF (IF1) and second IF (IF2) were selected as 10.7 MHz and 455 kHz due to the availability of bandpass filters and IC's which function at these frequencies.

The highest frequency for an 1200 baud AFSK data rate for the CCITT V.23 standard is 2.1 kHz and for the BELL 202 standard it is 2.2 kHz [27]. An audio bandwidth of $BW_{3dB} = 3$ kHz was selected to include the possibility of an 1200 baud data rate as well as to hear voice communications. An IF bandwidth of $BW_{3dB} = 12$ kHz was calculated for a modulation frequency (fm) of 3 kHz and a deviation (D) of 3 kHz with the Carson's rule for FM bandwidth ($BW = 2(d + fm)$) [28].

The Doppler shift at VHF is ± 3.6 kHz [20]. The various ways of compensating for the Doppler shift in the VHF receiver are as follows:

- Select an IF bandwidth of 18 kHz .
- Select or implement a quadrature detector at the FM demodulator which can be tuned ± 3.6 kHz around the center frequency of the second IF.
- Implement a local oscillator at the second mixer which is able to be tuned ± 3.6 kHz.

The option of an IF bandwidth of 18 kHz was not selected due the degradation in SNR of 1.76 when the IF is changed from 12 kHz to 18 kHz. The option for compensating the Doppler shift at the second LO or quadrature detector was therefore selected.

2.4.5 Spurious Reponses

A spurious response rejection of 80 dB was selected to ensure that for a RF input of - 40 dBm there would be no spurious responses. A received signal strength of - 40 dBm at the receiver is equivalent to 35.96 dBm (4 W) at a distance of 1km (line of sight).

The following out-of-band spurious responses should be suppressed by the filters in the RF section of the VHF receiver: image frequencies, a RF signal equal to 10.7 MHz, a RF signal equal to 455 kHz and harmonic mixing of LO and RF.

The spurious responses due to image frequencies and harmonic mixing are discussed in the sections below. A RF signal equal to 10.7 MHz can pass through the first mixer to the first IF section. A RF signal equal to 455 kHz can pass through the first and second mixer to the second IF section.

The image frequency of the second mixer is 9.79 MHz and is produced by a signal 910 kHz below the desired received signal in the RF section. When the received frequency in the RF section is 145.825 MHz the frequency response that will produce a first IF of 9.79 MHz is 144.915 MHz. Harmonic mixing between the RF signals and LO signal can also produce a first IF equal to 9.79 MHz. The first IF equal to 9.79 MHz will produce a signal at 455 kHz if it is not attenuated enough in the first IF section.

2.4.6 Image Frequencies

Equation 2.7 shows the relation between the IF, RF and LO in a receiver. The RF frequencies (wanted signals) are from 144 MHz to 146 MHz and the LO frequencies are from 133.3 MHz to 135.3 MHz for a first IF = 10.7 MHz . The image frequencies (unwanted responses) which are from 122.6 MHz to 124.6 MHz, cause an IF of 10.7 MHz for the LO frequencies previously defined. The value for m and n in equation 2.7 is one.

$$IF = |mRF \pm nLO| \tag{2.7}$$

2.4.7 Harmonic Mixing - RF and LO1

The harmonics of the spurious responses before the first mixer can mix with the harmonics of the first LO to cause an IF of 10.7 MHz or an IF of 9.79 MHz. Equation 2.8 was derived from equation 2.7. Tables 2.4 to 2.7 list the spurious responses due to harmonic mixing calculated by equation 2.8.

$$RF = |(IF \mp nLO)/m|, \tag{2.8}$$

Table 2.4: Spurious responses ($RF = |(IF - nLO)/m|$ where IF = 10.7 MHz)

m/n	1	2	3	4	5
1	124.425 MHz	259.550 MHz	394.675 MHz	529.800 MHz	664.925 MHz
2	62.213 MHz	129.775 MHz	197.338 MHz	264.900 MHz	332.462 MHz
3	41.475 MHz	86.517 MHz	131.558 MHz	176.600 MHz	221.642 MHz
4	31.106 MHz	64.888 MHz	98.669 MHz	132.45 MHz	166.231 MHz
5	24.885 MHz	51.910 MHz	78.935 MHz	105.960 MHz	132.985 MHz

Table 2.5: Spurious responses ($RF = |(IF + nLO)/m|$ where IF = 10.7 MHz)

m/n	1	2	3	4	5
1	145.825 MHz	280.950 MHz	416.075 MHz	551.200 MHz	686.325 MHz
2	72.913 MHz	**140.475 MHz**	208.037 MHz	275.600 MHz	343.162 MHz
3	48.608 MHz	93.650 MHz	138.692 MHz	183.733 MHz	228.777 MHz
4	36.456 MHz	70.238 MHz	104.019 MHz	137.800 MHz	171.581 MHz
5	29.165 MHz	56.190 MHz	83.215 MHz	110.240 MHz	137.265 MHz

Table 2.6: Spurious responses ($RF = |(IF + nLO)/m|$ where IF = 9.79 MHz)

m/n	1	2	3	4	5
1	125.335 MHz	260.460 MHz	395.586 MHz	530.710 MHz	665.835 MHz
2	62.668 MHz	130.230 MHz	197.792 MHz	265.355 MHz	332.917 MHz
3	41.778 MHz	86.820 MHz	131.862 MHz	176.903 MHz	221.948 MHz
4	31.334 MHz	65.115 MHz	98.896 MHz	132.678 MHz	166.459 MHz
5	25.067 MHz	52.092 MHz	79.117 MHz	106.140 MHz	133.167 MHz

Table 2.7: Spurious responses ($RF = |(IF - nLO)/m|$ where IF = 9.79 MHz)

m/n	1	2	3	4	5
1	144.915 MHz	280.040 MHz	415.165 MHz	550.290 MHz	685.415 MHz
2	72.458 MHz	**140.020 MHz**	207.582 MHz	275.145 MHz	342.708 MHz
3	48.305 MHz	93.347 MHz	138.388 MHz	183.430 MHz	228.472 MHz
4	36.229 MHz	70.010 MHz	103.791 MHz	137.573 MHz	171.354 MHz
5	28.983 MHz	56.008 MHz	83.033 MHz	110.058 MHz	137.083 MHz

The value for n less than or equal to 5 was selected on the assumption that the higher harmonics of the LO will not have an effect on harmonic mixing. The value for $m > 1$ is effective when the RF level is such that the amplifiers in the RF section are near saturation. A maximum value of m = 5 was selected to allow for strong RF signals which can saturate the RF section. The nearest out-of-band spurious responses were determined as **140.475 MHz** and **140.020 MHz** from Tables 2.4 to 2.7.

2.4.8 Harmonic Mixing - RF, LO1 and LO2

The out of band RF responses can convert down to the first IF and mix with the harmonics of the second LO to produce a signal at 455 kHz at the second IF section. The out of band RF response that can cause a signal at 455 kHz, as previously described, is equal to RF = $|(|nLO2 \pm 455\, kHz| \mp mLO1)/p|$ where m is equal to one, n is less than or equal to three and p is equal to one. The value for m equal to one was selected on the assumption that the higher harmonics of the first local oscillator will not have an effect. The value of n less than or equal to 3 was selected on the assumption that the higher harmonics of the second local oscillator will not have an effect. The value for p greater than one will resemble harmonics which occurred due to saturation in the RF section. The signals produced by these harmonics in the first IF section were assumed to have no effect.

2.4.9 LO Radiation

The LO radiation at the antenna must be suppressed so that the level of radiation meets the requirement set in section 2.4 .

2.5 UHF Down-Converter

This section describes the down-converter specifications selected for the ground station. Table 2.8 lists the main specifications and they are motivated in the rest of this section. The specifications were selected in order for school pupils to cope easily with them, yet simultaneously ensuring that the necessary requirements for receiving RF signals on UHF from an operational SUNSAT I are not lacking.

Table 2.8: Specifications for UHF Down-Converter

Type	UHF single conversion down-converter
Frequency	430 - 440 MHz
Synthesizer Resolution	5 kHz
IF Frequency	145.825 MHz
Conversion Gain	15 dB
Noise Figure	$\leq$ 2 dB
Image Rejection	70 dB
Spurious Rejection	70 dB

2.5.1 Down-Converter Type

Figure 2.6 shows a basic block diagram of the UHF down-converter. The RF and IF sections consist of filters and/or amplifiers to attenuate the spurious responses and amplify the desired signals.

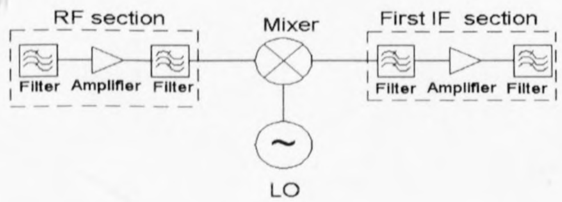

Figure 2.6: Block diagram of UHF down-converter

2.5.2 Frequency

A frequency range of 430 to 440 MHz was selected so that the school pupils could listen to other Amateur Radio frequencies in the UHF frequency band.

The frequency resolution was selected as 5 kHz in order to allow the pupil to tune the down-converter to accommodate the Doppler shift at UHF. The maximum Doppler shift is ± 10k Hz at UHF [20].

2.5.3 Conversion Gain and Noise Figure

A conversion gain of 15 dB was selected to ensure that the noise factor of the VHF receiver contributes less than or equal to 0.1 of the total noise factor of the system for a noise factor of 4.16 or noise figure of 6.20 dB for the VHF receiver.

The noise of an omnidirectional antenna at UHF can be anything from 300K, due to galactic noise [24], to 3000K, due to suburban man - made noise [12], and even higher due to other sources. Table 2.9 shows the noise figure and noise floor to produce a 12 dB SNR at the input of the receiver for a 12 kHz measured bandwidth. Equations 2.5 and 2.6 were used to determine the noise figure and noise floor in Table 2.9. The value of $T_a = 300$ K was used for the values in Table 2.9 and T_e is the effective noise temperature of the down-converter. The bandwidth was equal to 12 kHz which was used in equation 2.5.

Table 2.9: Noise Floor and Noise Figure values

Received Signal Strength [dBm]	T_e [K]	Noise Figure [dB]	Noise Floor [dBm]
- 118	303	3.1	- 130
- 119	170	2	- 131
- 120	80	1.1	- 132

The RF input required to cause a 12 dB SNR at the input of the UHF down-converter for a 12 kHz measured bandwidth increases as the noise figure of the UHF down-converter increases. A noise figure of 2 dB was selected to keep the down-converter as sensitive as possible but not selecting a value which will be unnecessarily strict in relation to the noise received by the antenna.

2.5.4 Intermediate Frequency (IF)

The choice of an IF equal to 145.825 MHz is discussed in this section. The idea concerning the down-converter is to select an IF output which is to be fed to a RF input or an IF input of the VHF receiver. The image frequency should be outside the pass band of 10 MHz in the RF section of the down-converter and it should be suppressed by 70 dB or more. This is valid for an IF equal to 145.825 MHz or an IF equal to 10.7 MHz. A first order bandpass Butterworth filter is needed to suppress the image frequency with 70 dB instead of an seventh or higher order Butterworth bandpass filter when the IF of the UHF down-converter is at 145.825 MHz and not at 10.7 MHz [30]. The 3 dB bandwidth for the Butterworth filters are 10 MHz. This is the reason why an IF equal to 145.825 MHz was selected for the down-converter.

2.5.5 Spurious Responses

A spurious response rejection of 70 dB was selected to ensure that for a RF input of - 50 dBm there will be no spurious responses. A received signal strength of - 50 dBm at the receiver is equivalent to 35.1 dBm (3.22 W) at a distance of 1 km (line of sight).

The following spurious responses must be suppressed by the filters in the RF section of the UHF down-converter: image frequencies and the harmonic mixing of LO and RF.

The image frequencies are from 138.350 MHz to 148.350 MHz. The closest out-of-band spurious frequencies due to harmonic mixing are 427.510 MHz, 440.840 MHz, 441.68 MHz and 442.530 MHz.

2.6 VHF Transmitter

This section describes the transmitter specifications selected for the ground station. Table 2.10 lists the main specifications which are motivated in the rest of this section. The specifications were selected in order for school pupils to cope easily with them, yet simultaneously ensuring that the necessary requirement for transmitting RF signals on VHF to an operational SUNSAT I are satisfied.

Table 2.10: Specifications for the VHF transmitter

Frequency	144 - 146 MHz
Modulation	FM , 3 kHz deviation
RF output power	2 W (33 dBm)
Frequency resolution	25 kHz
Power output of third harmonic at antenna	$\leq$ - 60 dBm
Audio Input	Microphone Modem

2.6.1 Frequency

The frequency range from 144 to 146 MHz with a 25 kHz resolution will allow the school pupil to transmit on the Amateur Radio frequencies in the 2 m frequency band, as well as the VHF receiver frequencies on an operational SUNSAT I.

2.6.2 RF Output Power

As already indicated in section 2.2, the received signal is greater than or equal to - 120 dBm for 85% and 95% of the pass when 1 W and 5 W are transmitted respectively. A RF output power equal to 2 W will increase the percentage of the pass from 85% to 90% for which the received signal at the satellite is greater than or equal to - 120 dBm. A transmitter with an output power of 2 W is less expensive than a transmitter of 5 W. These are the reasons why an output power of 2 W was selected for the VHF transmitter.

2.7 Modem

Table 2.11 lists the specifications for the modem to receive and transmit data.

Table 2.11: Specifications for the MODEM

Data rate	1200 b/s
Modulation Standard	CCITT V.23
Base band modulation	AFSK

The specifications are those which SUNSAT I used for 1200 AFSK data communication.

Chapter 3

Implementation of the VHF Receiver

3.1 Basic Block Diagram

Figure 3.1 depicts the block diagram of the superheterodyne VHF fixed frequency receiver, which receives FM modulated signals at 145.825 MHz.

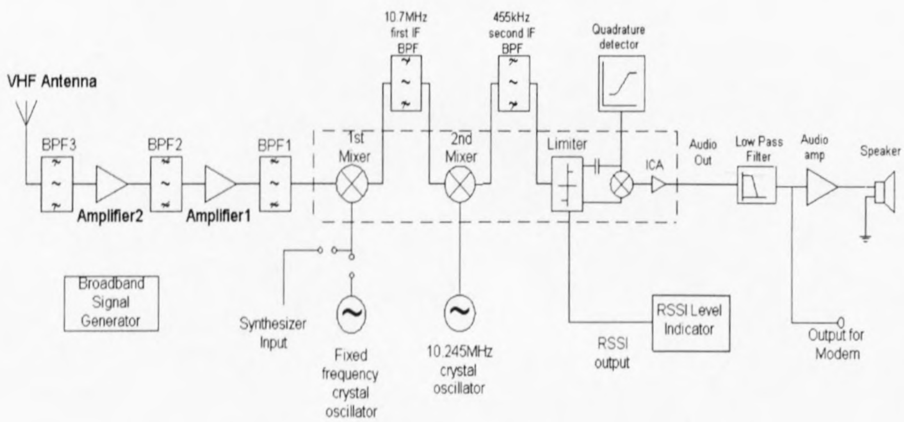

Figure 3.1: Block diagram of the VHF Fixed frequency receiver

The RF (Radio Frequency) signal received by the antenna is fed to BPF3 (Bandpass Filter 3). BPF3 has a center frequency of 145 MHz and a 3 dB bandwidth of 13 MHz. The bandwidth was selected to limit the insertion loss to 1 dB and to reject the image frequencies. The RF signal then passes through Amplifier 2 which consists of a MAR6 monolithic amplifier (Mini Circuits) followed in series by a 1.75 dB resistive pad. The resistive pad helps in ensuring that the reflection coefficient (S11) of the cascaded configuration of the bandpass filters and amplifiers is less than 0 dB. Amplifier 2 functions as a low noise amplifier with a gain of approximately 18.25 dB and a noise figure of 3 dB (noise figure of MAR6) [14]. BPF2 (Bandpass Filter 2), which is next in line for the RF signal to pass through, has a center frequency of 145 MHz and a 3 dB bandwidth of 5.2 MHz. The bandwidth was selected so that the filter does not have a loss of greater than 2.5 dB to help keeping the effective noise figure of the RF section approximately equal to 4 dB. The filter aids in suppressing the spurious (unwanted) responses such as the image frequencies. The RF signal then passes through Amplifier 1, which is identical to Amplifier 2. The RF signal then passes through BPF1 (Bandpass Filter 1). BPF1 is identical to BPF2. The RF signal is fed to the RF input of the first mixer of the MC13136 receiver IC. Everything included between the dash lines in Figure 3.1 is part of the MC13136 receiver IC.

The RF signal is down-converted to the first IF (Intermediate Frequency) of 10.7 MHz at the first mixer of the receiver IC. The first IF passes through a 10.7 MHz bandpass filter to the input of the second mixer where it is down-converted to a second IF of 455 kHz. The 10.7 MHz bandpass filter has a 3 dB bandwidth of 180 kHz. The second IF passes through a 455 kHz bandpass filter to the Limiter (Limiting IF Amplifier) [16]. The 455 kHz bandpass filter has a 6 dB bandwidth of 15 kHz. The output of the Limiter is fed to the Quadrature Detector where the audio is recovered through FM demodulation. The recovered audio signal passes through an internal audio amplifier (ICA) (part of the receiver IC) before it is available as an external audio output signal of the receiver IC. The external audio output signal passes through a second-order lowpass filter after which it is fed to a speaker through an audio amplifier. An output after the lowpass filter is provided for a modem connection.

The Received Signal Strength Indicator (RSSI) output of the receiver IC is fed to the RSSI level indicator circuit which indicates when the RSSI is above a certain level.

The local oscillator input for the first mixer is a fixed frequency crystal oscillator at 135.125 MHz. The local oscillator of the second mixer is a Colpitts oscillator which frequency is controlled by a 10.245 MHz crystal.

A variable frequency synthesizer can be applied as a local oscillator for the first mixer on the receiver. A simple jumper setting has to be made to use the synthesizer input as a local oscillator instead of the fixed frequency crystal oscillator.

Bandpass filters 3, 2 and 1 should reject the spurious frequencies such as the image frequency and spurious responses due to harmonic mixing between the first local oscillator and RF input at the first mixer with 80 dB. Amplifier 1 and 2 provide enough gain to provide a noise figure of approximately 4 dB.

BPF3, 2 and 1 and amplifier 1 and 2 are numbered in reverse due to the fact that the receiver should be assembled from the power supply backwards.

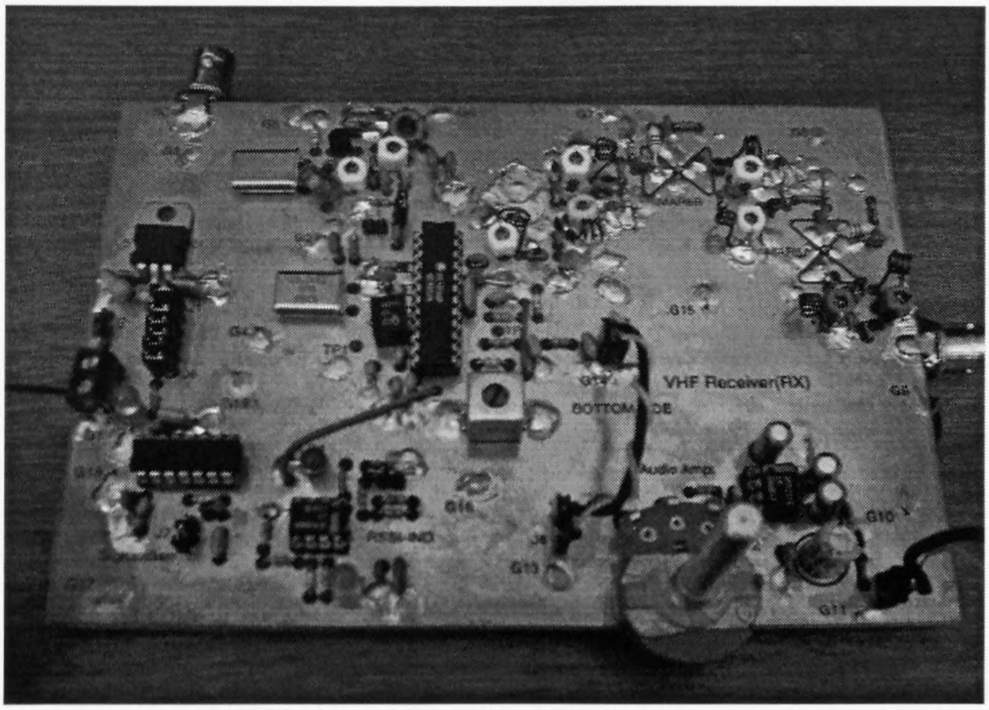

Figure 3.2: Photo of VHF fixed frequency receiver (Prototype)

The Broadband Signal Generator in Figure 3.1 is included to assist the school pupil to align and test the receiver. It can also help to determine if a certain section of the receiver is operating or not. Appendix E describes how the VHF fixed frequency receiver can be aligned.

Figure 3.2 shows a photo of the VHF fixed frequency receiver. Appendix B gives the block diagram and schematic of the VHF receiver.

The implementation of each of the different sections in the receiver will be discussed in the following sections.

3.2 Bandpass Filter 3

A 2nd order coupled resonator filter design (Butterworth response) was selected as a filter design for bandpass filters 3, 2 and 1. The design (equations included) of the f is discussed in Appendix A.

The advantage of the Butterworth response is that there is only one maximum peak in the pass band response. This will help to make the alignment of the receiver much simpler with the aid of the Broadband Signal Generator and RSSI Level Indicator.

3.2.1 Design

A central frequency of $f_0 = 145$ MHz was selected so that the central frequency is at the centre of the 2 m (144 MHz to 146 MHz) Amateur Radio Frequency band. The insertion loss should be I.L. ≤ 1 dB to keep the effective noise figure of the receiver approximately equal to 4 dB. A loaded Q equal to 15.22 and a bandwidth equal to 13.47 MHz were calculated for an unloaded Q (Q_U) of 140 and an insertion loss of 1 dB. The inductor $L_1 = 40$ nH was selected due to the fact that the coupling capacitor C_k was calculated bigger than 1 pF. An air core inductor of 40 nH with an unloaded Q = 140 was implemented with a 65 mm length of wire with a 0.7 mm diameter wound with 4 turns around a diameter of 3 mm. Three millimetre on each side of the inductor were soldered, or left to be soldered, to the printed circuit board. The Q was measured with the TF1247 and TF 1245 Marconi Instruments.

The component values that were calculated for Bandpass Filter 3 are $C_I = 30.12$ pF, $R_p = 555\ \Omega$, $C_k = 2$ pF, $C_s = 6.9$ pF and $C_1' = 21.85$ pF. Figure 3.3 shows the schematic of Bandpass Filter 3 with the calculated values.

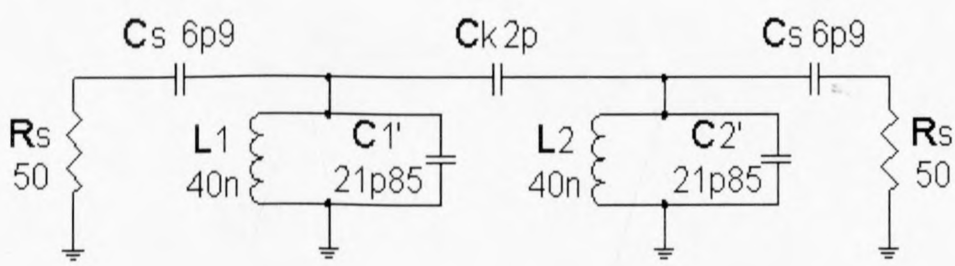

Figure 3.3: Bandpass Filter 3 with calculated values

3.2.2 Computed Response

The transmission coefficient (S21) was simulated to determine the 3 dB bandwidth of Bandpass Filter 3 and how much the image frequency (124.425 MHz) is suppressed. Figure 3.5 shows the computed transmission coefficient (S21) of Bandpass Filter 3 for the values indicated in Figure 3.4. The values in Figure 3.4 are the closest standard values for capacitors to the design values calculated in section 3.2.1 . A 10 pF capacitor and a 20 pF trimmer capacitor were used to implement the design capacitors C_1' and C_2'.
The 20 pF trimmer capacitors was tuned/set to 12 pF. The coupling capacitor C_k was selected as 1.8 pF rather than 2.2 pF to ensure that there is only one maximum amplitude response in the transfer function of the filter. The tolerance analysis is discussed in section 3.2.3.

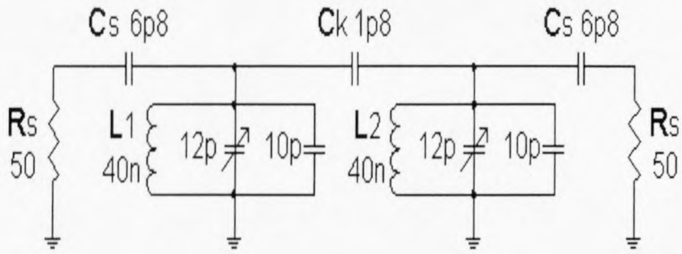

Figure 3.4: Bandpass Filter 3 with practical values

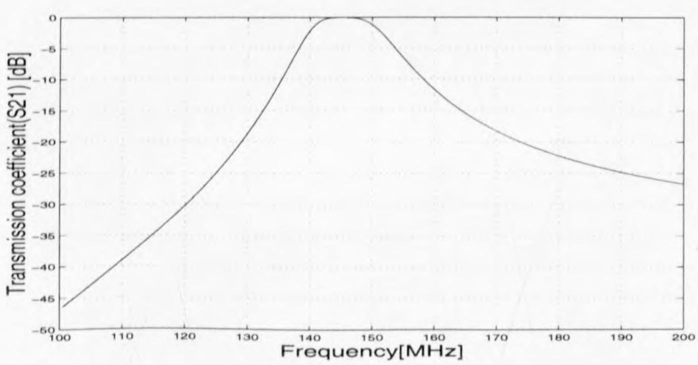

Figure 3.5: Computed responses of Bandpass Filter 3

The 3 dB bandwidth is 13 MHz. The image frequency (124.425 MHz) is suppressed by 25.5 dB. The difference between the 3 dB bandwidth of the computed S21 responses for Bandpass Filter 3 for the design values and the values which are available is less than 1.1 MHz. The difference between the amount the image frequency is suppressed between the design values and the standard values used for Bandpass Filter 3 is less than 1.5 dB.

The small difference between the results of the design values and the values available led to the conclusion that the values available for capacitors can be used to realise the filter.

The implementation of the filter will be discussed next.

3.2.3 Implementation

Figure 3.6 shows how Bandpass Filter 3 was implemented.

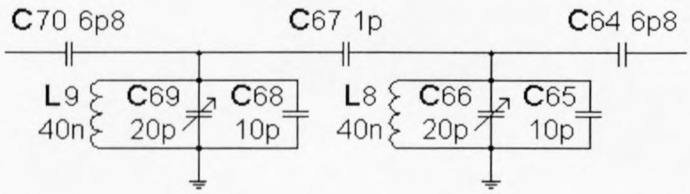

Figure 3.6: Implemented Bandpass Filter 3

The tolerance on the capacitors used is ± 10% according to the RS Components catalogue [21]. The tolerance of inductors as school pupils may wind them were determined at ± 15%. A focus group consisting of five school pupils were requested to wind an inductor of 40 nH and were provided with a set of clear instructions. The inductors so produced were then measured and it was concluded that the tolerance of inductors wound by school pupils with the necessary instruction is ± 15%. The tolerance specified in these fashion were used in the rest of the thesis for capacitors and inductors.

A 20 pF (4 to 20 pF) trimmer capacitor was selected to compensate for a tolerance of ± 10% in the capacitors and ± 15% in the inductor. The trimmer capacitor can compensate for a tolerance of ± 10% in the capacitors and ± 16.25 % in the inductor. The tolerance analysis was determined by calculating the frequency $f_0 = 1/\sqrt{(2\pi(C70 + C69 + C67)\ L9)}$ for $C70 = 6.8$ pF, $C69 = 10$ pF, $C67 = 1.8$ pF and $L9 = 40$ nH. The percentage tolerance in $C70$, $C69$, $C67$ and $L9$ for which the trimmer capacitor $C67$ can compensate was then determined so that f_0 remained the same.

The pins of the capacitor, trimmer capacitor and inductor in the resonant sections must be soldered on the same place or close together on the ground plane if the places for the pins are not indicated on the ground plane of the PCB. The circulating current in the resonant sections is I_{circ} = Q (Q_U = unloaded Q) x I_{line} [26]. This is indicated in Figure 3.7. Thus if the pins of the components are soldered apart, ground currents can occur. These ground currents in each resonant section can cause voltage differences on the ground planes which can add extra losses to the filter response. This is also applicable to Bandpass Filters 2 and 1.

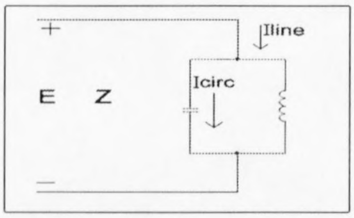

Figure 3.7: Schematic to indicate I_{circ} and I_{line}

3.3 Amplifier 2

The MAR6 monolithic amplifier from Mini Circuits was selected as the amplifier for the following reasons:

- Matched for 50 Ω input and output
- Easily available
- Adequate noise figure
- Out-of-band 50 Ω matching may reduce the effect of uncertanty in a cascading topology with band pass filters

Figure 3.8 shows a schematic of the MAR6 with an 1.75 dB resistive pad on its output.

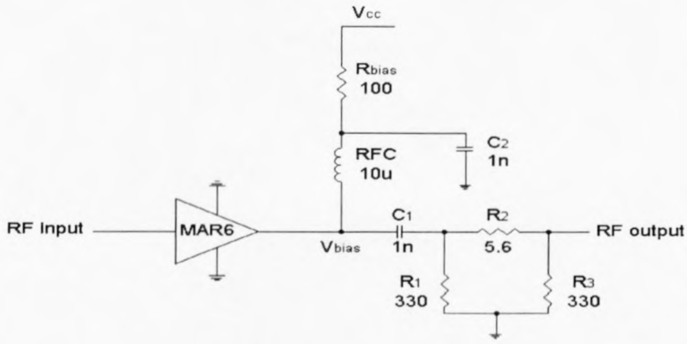

Figure 3.8: Amplifier 2

The gain of the MAR6 is in the order of 20 dB at 145 MHz [14]. The output matching capacitor of Bandpass Filter 3 also acts as the DC blocking capacitor on the input of Amplifier 2.

Resistors $R31$ to $R33$ in Figure 3.8 were selected so that they act as an 1.75 dB resistive pad. The input and output impedance of the 1.75 dB pad are approximately 50 Ω. Simulations in Touchstone for the cascaded configuration of bandpass filters 1 to 3 and amplifier 1 and 2 showed that S11 $>$ 0 dB for a resistive attenuation pad of 1.7 dB and less. The RF section can start to oscillate when S11 $>$ 0 dB of the RF section.

3.3.1 Biasing

A bias resistor $R_{bias} = 93.75\ \Omega$ will realise a bias current of 16 mA for $V_{CC} = 5$ V and $V_{bias} = 3.5$ V. A value of 100 Ω was selected for the bias resistance R_{bias}. The R.F. choke (RFC) equal to 10 μH was selected to act as a high impedance path for RF on the output of the amplifier to the power supply. The 1 nF capacitor, C2 in Figure 3.8 is for DC decoupling purposes. The 1nF capacitor C1 was placed as a DC blocking capacitor on the output of the amplifier before the resistor pad.

3.4 Bandpass Filter 2

3.4.1 Design

The aim was to limit the insertion loss of the filter to less than or equal to 2.5 dB and to keep the 3 dB bandwidth as close as possible to 2 MHz. The insertion loss was selected less than or equal to 2.5 dB to assist in keeping the effective noise figure of the receiver approximately equal to 4 dB. Below 144 MHz and above 146 MHz are other mobile services which have to be filtered out [22]. Using the equations in Appendix A the bandwidth was selected equal to 5.86 MHz to limit the insertion loss of the filter to less than or equal to 2.5 dB. The unloaded Q of the inductor was selected as 140. The loaded Q was calculated equal to 35. L_1 was selected equal to 30 nH so that C_k is bigger or equal to 1 pF. An air core inductor of 30 nH with an unloaded Q = 140 was implemented with a 55 mm length of wire with a 0.7 mm diameter wound with 3 turns around a diameter of 3 mm. Three millimetre on each side of the inductor were soldered, or left to be soldered, to the printed circuit board. The Q was measured with the TF1247 and TF 1245 Marconi Instruments.

The calculated values for Bandpass Filter 2 are $C_I = 40.16$ pF, $R_p = 957\ \Omega$, $C_k = 1.14$ pF, $C_s = 5.15$ pF and $C_1' = C_2' = 34.14$ pF. Figure 3.9 shows the schematic of Bandpass Filter 2 with the calculated values.

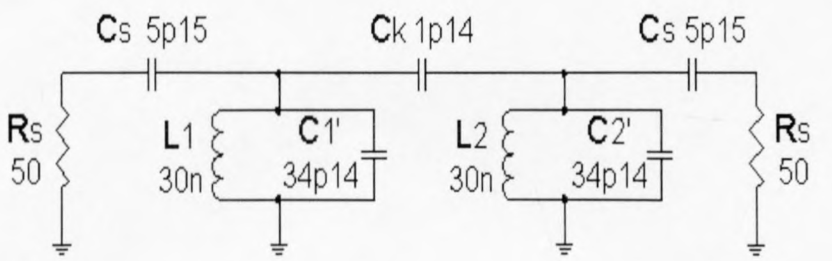

Figure 3.9: Bandpass Filter 2 with designed values

3.4.2 Computed Response

The transmission coefficient (S21) was plotted to determine the 3 dB bandwidth and how much the image frequency (124.425 MHz) is suppressed. Figure 3.11 shows the computed transmission coefficient (S21) for Bandpass Filter 2 with the values indicated in Figure 3.10. The values in Figure 3.10 are the closest standard values for capacitors to the design values. An 18 pF capacitor in parallel with a 30 pF trimmer capacitor was selected to realise the design values of the capacitors C_1' and C_2' in the design section of Bandpass Filter 2. The trimmer capacitors were set equal to 16.4 pF for the simulations. The tolerance analysis is discussed in section 3.4.3.

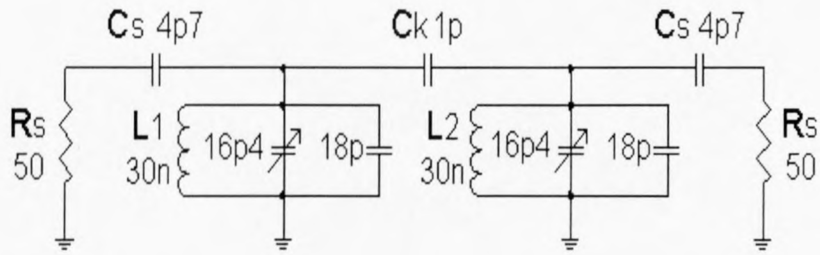

Figure 3.10: Bandpass Filter 2 with practical values

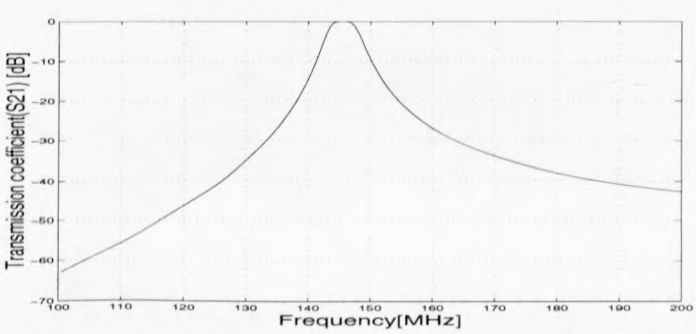

Figure 3.11: Computed response of Bandpass Filter 2

The 3 dB bandwidth of the computed responses of Bandpass Filter 2 of Figure 3.10 is 5.2 MHz. The image rejection is 41 dB. The computed responses for the design values for Bandpass Filter 2 resulted in a 3 dB bandwidth of less than 5.7 MHz and the image rejection of 39 dB. The difference between the results with the design values and standard values is adequate for standard values to be used.

3.4.3 Implementation

Figure 3.12 shows a schematic of Bandpass Filter 2 as implemented. Tolerance analysis was done in the same manner as for Bandpass Filter 3. A 30 pF trimmer capacitor was selected to compensate for the required tolerance of ± 10% in the capacitors and ± 15% tolerance in the inductor. The 30 pF trimmer capacitor compensates for a maximum tolerance of ± 21% in the inductor.

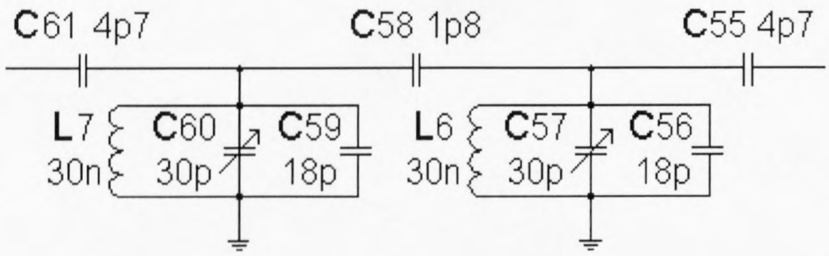

Figure 3.12: Circuit which was implemented as Bandpass Filter 2

3.5 Amplifier 1

Amplifier 1 is an exact copy of Amplifier 2. An effective noise figure of approximately 4 dB was calculated for a gain of 18.25 dB and a noise figure of 3 dB for Amplifiers 1 and 2, a loss of 1 dB for Bandpass Filter 3, a loss of 2.5 dB for Bandpass Filter 1 and 2 and noise figure of 19 dB cascaded noise figure of the MC13136 receiver IC.

3.6 Bandpass Filter 1

Bandpass Filter 1 is the exact copy of Bandpass Filter 2.

The noise created at the image frequency by Amplifier 1 will cause a 3 dB increase at the noise at the first IF if it is not attenuated before the first mixer. This will then decrease the SNR at the first IF.

To limit the increase of the noise figure by less or equal than 0.1 dB the image noise should be attenuated to less than 2.3 % of its full value. Equation 3.1 and equation 3.2 were used to determine that the image noise at the output of Amplifier 1 should be suppressed by 16.4 dB before the first mixer [20].

$$2.3\% = 10^{0.1/10} - 1 \tag{3.1}$$

$$-\ 16.4\ \mathrm{dB} = 10log_{10}(0.023) \tag{3.2}$$

The computed responses showed that the image frequency is suppressed by 41 dB which is 23.6 dB more than which is required to suppress the image noise.

3.7 Computed Response of the RF section

Figure 3.13 shows the plot of the transmission coefficient of the RF section of the VHF receiver. The RF section consists of the series connection of Bandpass Filter3, Amplifier 2, Bandpass Filter 2, Amplifier 1 and Bandpass Filter 1. The transmission coefficient (S21) was plotted to determine the 3 dB bandwidth and how much the image frequency at 124.425 MHz is suppressed.

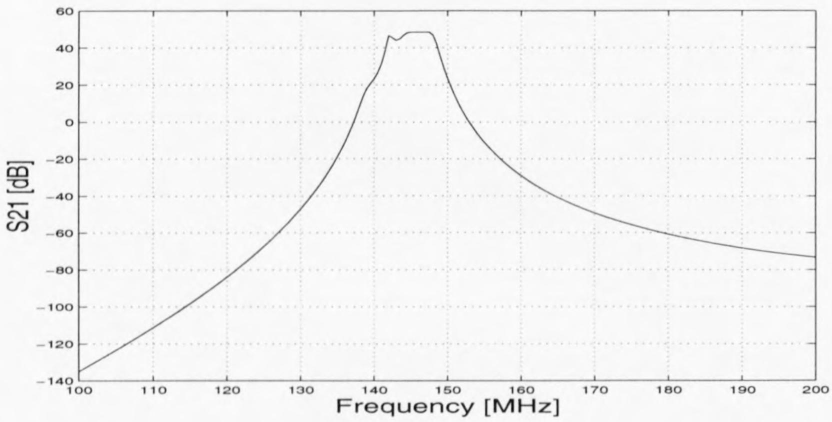

Figure 3.13: Computed response of the RF section of the VHF recevier

The 3 dB bandwidth is approximately 6 MHz. The image frequency is suppressed by 115 dB which is good. The measurements still have to be taken to be compared with the computed results. The gain of 48 dB is not a true reflection of the transmission coefficient (S21) of the RF section due to the fact that the transmission coefficient (S21) of the MAR 6 was measured equal to 19.6 dB at VHF. The losses in the filters were not accounted for in the simulation.

3.8.2 Bandpass Matching Network

The input impedance at the RF input of the Receiver IC is equivalent to a 722 Ω resistor in parallel with a 3.3 pF capacitor at 50 MHz [16]. The impedance was assumed to also be a 722 Ω resistor in parallel with a 3.3 pF capacitor at 145 MHz. A mismatch loss of 6.24 dB (mismatch loss = $10log_{10}(1/(1 - S11^2)))$ will occur in a 50 Ω system at approximately 145 MHz. A bandpass matching network was implemented to reduce the loss in RF signal due to mismatch. The design of the matching network was done in two steps.

The first step was to match 50 Ω to 722 Ω. This was accomplished by the implementation of a capacitive C - tap bandpass matching network. The formulas on page 80 of [4] were used to determine the values for the matching network as indicated in Figure 3.15. A bandwidth of 10 MHz and a central frequency f_0 = 145 MHz were selected so that the Q_t ($Q_t = f_0/BW$) of the network is bigger or equal to 10 and so that the 2m Amateur Radio frequency band is included. The formulas on page 80 of [4] require that $Q_t = f_0/BW \geq 10$. The image frequency was suppressed by 7.8 dB according to simulations with Touchstone and the 3 dB bandwidth was 20 dB. Bandpass Filter 1 is responsible for suppressing the image noise with 16.4 dB and more and therefore 7.8 dB of image rejection is acceptable.

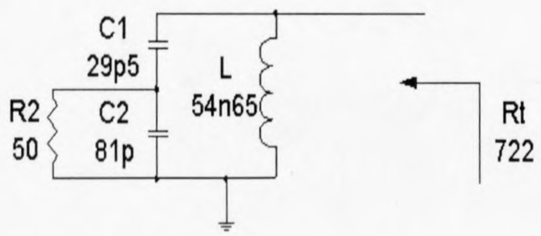

Figure 3.15: Bandpass matching network

The second step was to determine the reactance to resonate with the 3.3 pF capacitor at the RF input of the receiver IC. An inductor of value $L_r = 365$ nH was calculated. The effective inductance then was calculated determined as $L_r//L =$ 47.5 nH. Figure 3.16 shows a schematic of the bandpass matching network with the values which were selected to be implemented. The trimmer capacitor (4 - 20 pF) was selected in parallel with the 35 nH inductor to tune for the effective inductor value. The trimmer capacitor compensates for a ± 10% tolerance in the capacitors and ± 15% in the inductor. An air core inductor of 35 nH with an unloaded Q = 140 was implemented with a 60 mm length of wire with a 0.7 mm diameter wound with 3 turns around a diameter of 3 mm. Three millimetre on each side of the inductor were soldered, or left to be soldered, to the printed circuit board.

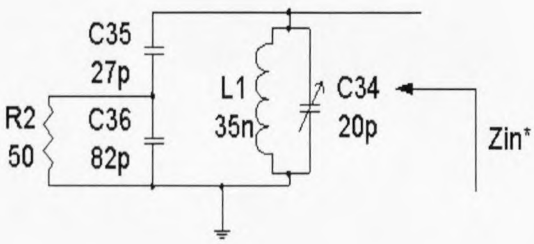

Figure 3.16: The implemented bandpass matching network

The components labeled $C34$, $C35$, $C36$ and $L1$ are part of the matching network at the RF input of the receiver IC in Figure 3.14.

3.8.3 1st Local Oscillator

A fixed frequency crystal oscillator (135.125 MHz) was implemented as an external local oscillator. This was recommended by [16] for operating a crystal oscillator above 60 MHz. A variable frequency synthesizer may be added at the synthesizer input and be selected with a jumper on the PCB.

3.8.4 1st IF Bandpass Filter

A 10.7 MHz SFE10.7MS3-A ceramic Murata bandpass filter was selected as the first IF bandpass filter. The filter has a 3 dB bandwidth of approximately 180 kHz and a 20 dB bandwidth of approximately 470 kHz. The spurious response at 9.79 MHz was suppressed by 52 dB according to measurements that were made. The second image, due to 144.915 MHz, will therefore only be rejected by less than or equal to 52 dB and not by 80 dB as the requirements were established in chapter two. A crystal filter will be able to reject the second image by 80 dB but costs R 32.80, which is ten times the price of a ceramic filter. The ceramic filter should be replaced with the crystal filter if the total price of the fixed frequency receiver is increased by less than 15% when the ceramic filter is replaced with the crystal filter.

The input and output impedance of the filter is 330 Ω [35]. The output impedance of the first mixer is 330 Ω and the input impedance of the second mixer is 4 kΩ according to [16]. A 330 Ω and 33 Ω resistor in series were used to match the filter to the IF input of the second mixer.

3.8.5 2nd Mixer

The 2nd mixer is a double balanced mixer. The input impedance is 4 kΩ and the output impedance is 1.8 kΩ [16].

A test point labeled $TP2$ was added at the input of the mixer. This was done so that the school pupil can apply a signal with the Broadband Signal Generator.

3.8.6 2nd Local Oscillator

A 10.245 MHz crystal was selected to control the frequency of the Colpitts oscillator.

3.8.7 2nd IF Bandpass Filter

A 455 kHz CFU455E ceramic Murata bandpass filter was selected as the 2nd IF Bandpass Filter. The filter has a 6 dB bandwidth of 15 kHz and a 40 dB bandwidth of 30 kHz. A stop band attenuation of 27 dB occurs at 455 kHz $\pm$ 100 kHz. The input and output impedances of the filter are 1.5 kΩ [34]. The output impedance of the second mixer is 1.8 kΩ [16].

3.8.8 Limiter and Detector

Figure 3.17 shows the block diagram of the Limiting IF Amplifier (Limiter) and FM detector layout in the MC13136 FM receiver IC.

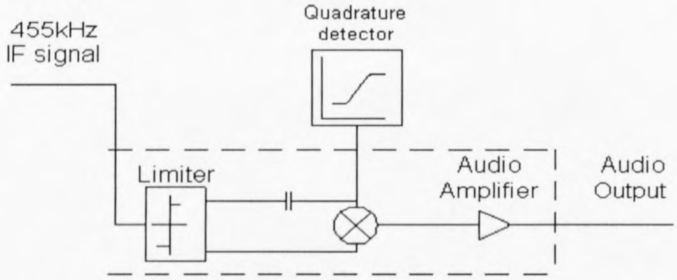

Figure 3.17: Limiter and Detector section of the receiver IC

The MC13136 receiver IC can either use a ceramic discriminator or a LC quadrature detector [16]. The ceramic discriminator is a fixed frequency resonator . A LC quadrature detector was selected as a detector at the receiver IC to allow for the Doppler shift at VHF which is $\pm$ 3.6 kHz.

The resonant section of the detector was implemented with the internal capacitor and inductor of a YRCS11098 Toko Sub - Miniature I.F. Transformer. A 68 kΩ resistor was placed in parallel with the Sub - Miniature I.F. Transformer. The frequency of the detector is determined by the capacitor and inductor while the resistor influences the Q of the detector. The Q of the detector determines the phase change per frequency at and around the frequency of resonance of the detector. The higher the value of the resistor the higher is the phase change per frequency.

Figure 3.18 shows a phase vs. frequency plot for an external resistor of 68 kΩ. The external resistor of 68 kΩ in parallel with the Sub - Miniature I.F. Transformer produces an unloaded Q of 25.2 and the effective phase change per frequency is linear over 12 kHz of the detector frequency range. The phase change should be linear over the IF bandwidth which is 12 kHz to assure that the output signal at the audio output is not distorted.

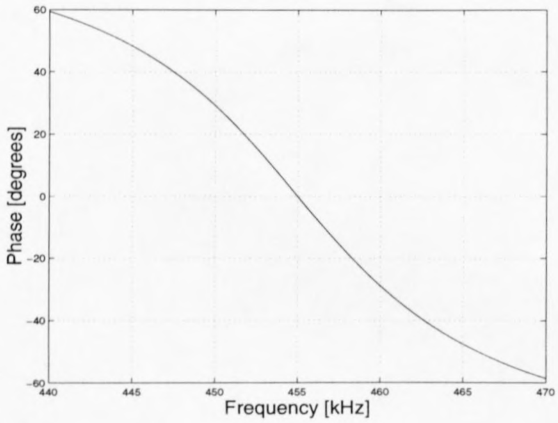

Figure 3.18: Phase vs. Frequency

The Limiter amplifies the signal until only the frequency and phase information is left in it. The Limiter has two outputs. The one output is phase shifted by 90^o and applied to the detector. Frequency changes will then produce an additional leading or lagging phase shift. The signal is then fed to the mixer in Figure 3.17. The other signal is fed directly to the mixer in Figure 3.17. The output of the mixer is fed to the audio output of the receiver IC through an audio amplifier.

3.8.9 Audio Lowpass Filter

Figure 3.19 shows the schematic of a second order lowpass filter which was implemented at the audio output of the MC13136 dual conversion receiver IC.

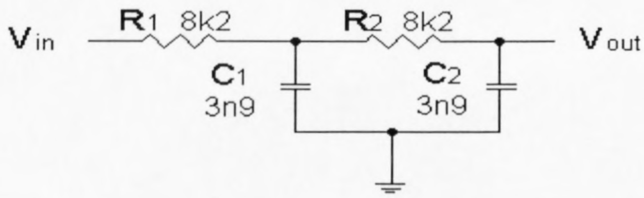

Figure 3.19: Audio filter on output of dual conversion IC

The resistors and capacitors were selected so that the 3 dB bandwidth is approximately 3 kHz. The poles of the filter are at $s = j\omega = 2\pi \times 4.98$ krad/s.

3.8.10 RSSI/Op Amp

The RSSI (Received Signal Strength Indicator) has a 70 dB range. The RSSI output at pin 12 is fed to the RSSI Level Indicator. Pin 16 of the receiver IC is a buffered RSSI output [16].

3.9 Fixed Frequency Crystal Oscillator

Figure 3.20 is the schematic of the oscillator which was implemented at the Local Oscillator input of the first mixer of the FM receiver IC. The crystal is a 135.125 MHz fifth overtone series resonant crystal. Q1 is a MPSH10 bipolar transistor.

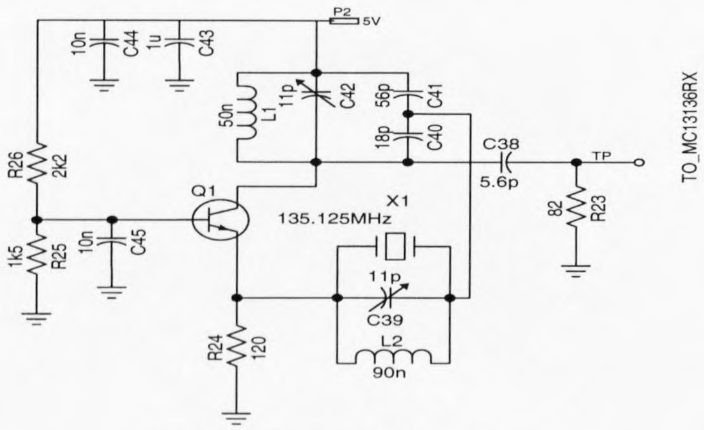

Figure 3.20: Crystal Oscillator

The design of an oscillator such as in Figure 3.20 was studied and understood to ensure the design provides a reliable oscillator that does not need extensive equipment to adjust during manufacturing. A reliable oscillator is an oscillator that will oscillate when power is applied to the oscillator and does not oscillate at any other frequency when the trimmer capacitor or tuning element is tuned.

The design and analysis of the oscillator in Figure 3.20 is discussed in the following sub sections.

3.9.1 Design: Calculations & Choices

Figure 3.20 and Figure 3.22 show the circuits of a Harmonic - Butler Common Base type oscillator using a capacitive tap. This type of circuit is appropriate for VHF frequencies between 20 - 200 MHz according to [13]. The oscillator is best analysed using the feedback oscillator model.

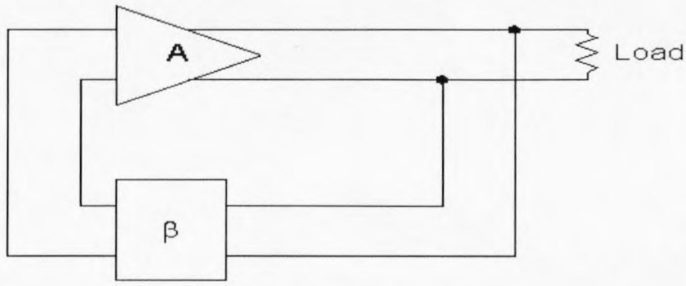

Figure 3.21: Basic oscillator feedback model

Figure 3.21 shows a schematic of a basic feedback oscillator model, where **A** is an amplifier and β the feedback network which has a gain $|A|\angle\theta_A$ and has $|\beta|\angle\theta_\beta$ separately [19, 6]. The condition for an oscillation to occur is both $|A_L| = |A\beta| = 1$ and $\angle A_L = 0^o, 360^o, n360^o$. The condition is also known as the Barkhausen criterion [6, 19]. A_L is the loop gain [6]. The loop gain is referred to as the the open loop gain in this document.

Figure 3.20 depicts the oscillator with an inductor and a trimmer capacitor in parallel with the crystal in the feedback path. The inductor and trimmer capacitor resonate with the static capacitance of the crystal to assure that the oscillator is not LC controlled but crystal controlled. The model of the crystal is discussed in section 3.9.3. The oscillator in Figure 3.20 is tuned by first turning trimmer capacitor C39 through 360^o and secondly turning trimmer capacitor C42 approximately by 30^o . This should be repeated until the oscillator oscillates. The process of tuning the oscillator in Figure 3.20 is more complex when there are two trimmer capacitors instead of one. Components have tolerance that also adds an element of uncertainty to the design. Safe margins were added to the Barkhausen criterion to ensure that the oscillator is only crystal controlled without an inductor, or an inductor and trimmer capacitor, in parallel with the crystal in the feedback path. The following criteria were added as additional safe margins in the criterion for an oscillation to occur.

- $|A_L| \geq 6$ dB at the frequency of design.
- $|A_L| <$ - 6 dB at any other frequency except the frequency of design.

Figure 3.22 shows a circuit of the Butler Common Base oscillator, and gives the values for $R1$, $R2$, Re, $C6$, $C7$ and $C8$. The reason for the values of $R1$, $R2$, Re, $C6$, $C7$ and $C8$ are discussed in section 3.9.2 . The manner in which the values for $L1$, $C2$, $C3$, $C5$, RL and Zfb were selected is discussed in the paragraphs to follow.

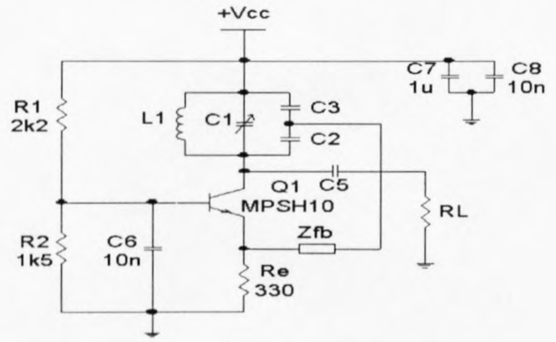

Figure 3.22: Harmonic - Butler common base oscillator - C tap

Figure 3.23 shows the AC small-signal model of Figure 3.22. The magnitude and phase of the open loop gain $A_L = I_2/I_2'$ of Figure 3.23 is determined with the MATLAB code in Appendix C.2.

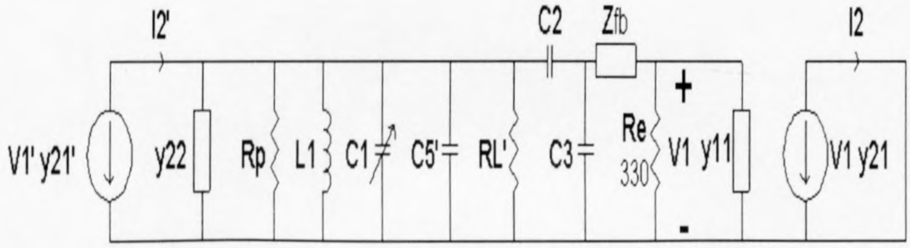

Figure 3.23: Small signal AC model

The resistor Rp represents the losses in the inductor $L1$ in Figure 3.23. RL' and $C5'$ in Figure 3.23 is the parallel conversion of RL in series with $C5$ in Figure 3.22. Zfb represents the crystal in the feedback path. The model of the crystal is discussed in section 3.9.3. The impedance of the crystal can be represented by different values for different configurations or frequencies. At resonance the assumption can be made that the crystal is equivalent to a resistor. The static capacitance dominates in the crystal at any other frequency except the resonance frequency.

The following must be considered in the selection of values for $C1$, $C2$, $C3$, $C5$, $L1$ and RL in Figure 3.22 and Figure 3.23:

- The load at the collector should be high to ensure that the gain is sufficient for the open loop gain.
- The source and load impedance of the crystal should be low to ensure a high in-circuit Q for short term frequency stability.
- The trimmer capacitor $C1$ should aid in compensating for the tolerances in $C2$, $C3$, $C5$ and $L1$
- There must be enough output power delivered to the output load R_L. An output of 5 to 7 dBm delivered in the output load should be sufficient.

Figure 3.24 shows the load seen by the collector at resonance. The derivation of R_{LX} is discussed in the paragraphs below.

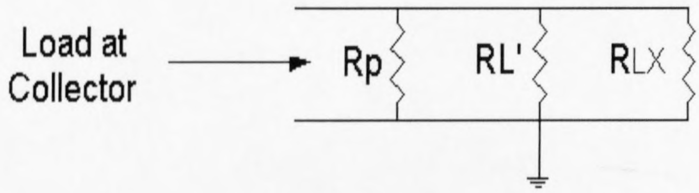

Figure 3.24: The load as seen by the collector

Figure 3.25 shows the first schematic in determining what the impedance R_{LX} is at resonance. $R11$ and $X11$ represents the input impedance of the transistor, where $X11$ is equal to j$\omega L11$. Re is the emitter resistor. The crystal can be replaced with a series resistance at the frequency of resonance. A value of 23 Ω was determined as the series resistance at 135.125 MHz (the fifth mode of the crystal) in section 3.9.3 . The capacitor $C1'$ in Figure 3.25 is the parallel combination of $C1$, $C22$ and $C5'$. The capacitor $C22$ is due to y22 in Figure 3.23 which only has an imaginary impedance at 135 MHz according to the data sheets

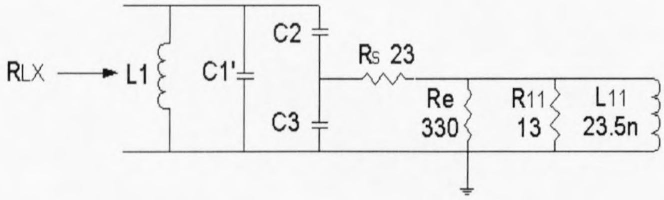

Figure 3.25: The load as seen by the collector

Figure 3.26 shows the effective impedance in parallel with $C3$ apart from $C2$, $C1'$ and $L1$.

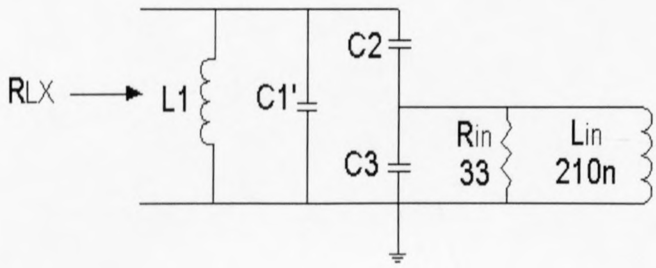

Figure 3.26: The load as seen by the collector

Figure 3.27 shows the impedance R_{LX} where $C3' = C3//Lin$. Calculations showed that $C3//Lin$ from 120 MHz to 150 MHz will have an negative imaginary impedance for $C3$ as small as 40 pF and $Lin = 210$ nH. The impedance of a capacitor is negative and imaginary which means that $C3'$ will be an effective capacitor. The reason for the 120 MHz to 150 MHz frequency range is discussed in a paragraph below.

The configuration of the inductor, capacitors and resistor in Figure 3.27 is that of a tapped resonant circuit using capacitive tapped circuit. The formulas on page 80 of [4] are used to determine the values to matched Rin to R_{LX} in Figure 3.27. The situation is just the opposite here, where every value except R_{LX} is known. The equations in [4] showed that $R_{LX}/Rin = (n')^2$ and that $n' = C3'/C2 + 1$.

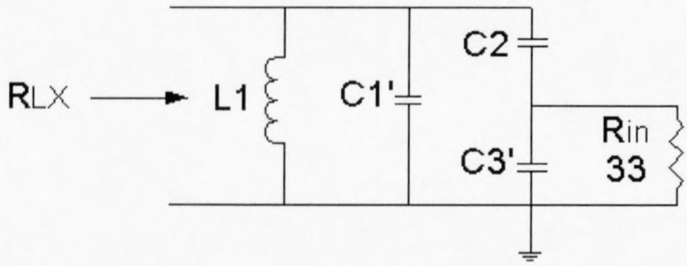

Figure 3.27: The load as seen by the collector

Figure 3.28 shows the source and load impedance as seen by the crystal. The fifth mode of the crystal, which resonates at 135.125 MHz, is shown in the schematic. The model of the crystal is discussed in section 3.9.3 . The load impedance is $Rin' + Lin' = Rin//Lin$ which is the input impedance of the transistor. The source impedance is R_{source} and C_{source} in series. Equation 3.3 was used to determine R_{source}. Equation 3.4 was used to determine C_{source}. Equation 3.3 and 3.4 were derived as the real and imaginary source impedance as seen by the crystal.

$$R_{source} = \text{Real}\left[\frac{jwL1 + RL - w^2RL(C1' + C2)}{(jwRL(C3 + C2) - w^2L1(C2 + C3) - jw^3L1RL(C3(C1' + C2) + C1'C2))}\right] \tag{3.3}$$

$$C_{source} = \text{Image}\left[\frac{jwL1 + RL - w^2RL(C1' + C2)}{(jwRL(C3 + C2) - w^2L1(C2 + C3) - jw^3L1RL(C3(C1' + C2) + C1'C2))}\right] \tag{3.4}$$

The value for R_{source} can also be determined with $(Rp//RL')/(n)^2$ where $n = C3/C2 + 1$. No mathematical proof has been done to show this but simulations and calculations showed that the same answers were obtained.

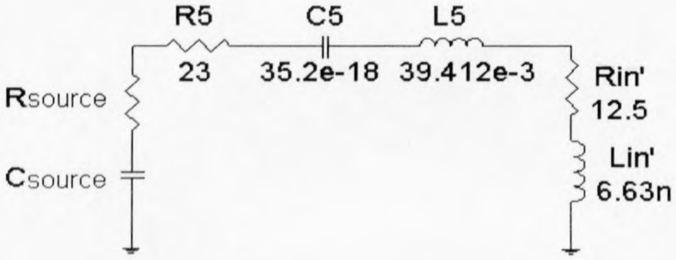

Figure 3.28: The source and load impedance of the crystal

The reactive impedances in Figure 3.28 resonate with the fifth mode of the crystal and will cause a frequency change of less than 1 kHz for the values of the fifth mode branch of the crystal in Figure 3.28. The trimmer capacitor compensates for the frequency change due to the load and source impedance of the crystal.

Design boundaries were determined to select values for components which determine the impedance seen by the collector and the crystal at resonance. The boundaries are as follows:

- $500\ \Omega \leq RL'//Rp//R_{LX} \leq 1.6\ \text{k}\Omega$
- $RL'//Rp \leq 1.5\ \text{k}\Omega$
- $6 \leq n \leq 11$ where $n = C3/C2 + 1$

The upper boundary $RL'//Rp//R_{LX} \leq 1.6\ \text{k}\Omega$ was selected to ensure that the oscillator will deliver 5 dBm or more into the load of the collector for a DC collector current of 4 mA, $V_{CC} = 5$ V, $Re = 330\ \Omega$ and $V_{CEsat} = 0.5$ V.

The lower boundary of 500 Ω was selected to limit the decrease in voltage gain by no more than 5 dB when the output load on the collector was decreased from 1.6 kΩ to 500 Ω. The value of the lower boundary equal to 500 Ω was more an initial value than a final boundary value.

The boundary value of $RL'//Rp \leq 2\ \text{k}\Omega$ was selected to decrease the in-circuit Q of the crystal with not more than a factor of 1:4 for a ratio of 1:5 for C3:C2. The higher the in-circuit Q the better the short term frequency stability, which means that the phase change per frequency change is high [13].

The upper and lower boundaries for n were selected to assist in the design boundaries for $RL'//Rp//R_{LX}$ and $RL'//Rp$.

$L1$ in Figure 3.23 was selected equal to 50 nH which produce a Rp of approximately 5000 with an unloaded Q of 120. An inductor of 50 nH was implemented with a 75 mm length of wire with a 0.7 mm diameter wound with 5 turns around a diameter of 5 mm. The following boundaries were set on RL' and R_{LX} to meet the boundaries set earlier:

- $1100 \leq RL' < 2000$
- $1100 \leq R_{LX} < 4000$

Different values for RL were selected to determine how the change in the load effected the oscillator. The values for RL that were selected are $RL = 50\ \Omega$, $RL = 100\ \Omega$, $RL = 150\ \Omega$, $RL = 500\ \Omega$ and $RL = 1000\ \Omega$. The values 50 Ω, 100 Ω and 150 Ω are input impedances of the Mini Circuits mixers [14]. The other two values were selected to see how they operate or to determine what is needed so that the oscillator operates with a load bigger than 150 Ω.

The values for C5 were selected so that the boundaries set for RL' and the frequency range of 120 MHz to 150 MHz are met.

The values for $n = 11$, $7.5 < n < 9.5$, $6 < n < 7$ and $4 < n < 5$ were selected to be evaluated where $n = C3/C2 + 1$. The values for $C2$, $C3$, $C5$ and RL for the ratios of n were selected so that the boundaries and frequency criteria were fulfilled. Some of the possibilities were simulated. Some of the simulated possibilities were built and measured to make a sensible conclusion on the ratio of $C2 : C3$ and the output load on the collector.

A frequency range of 120 MHz to 150 MHz for the LC network at the collector was selected to allow a $\pm$ 10 % tolerance on $Ct(Ct = Ceff + C5' + C1 + C22)$ and $\pm$ 15% on $L1$. A trimmer capacitor with a range of 4.2 to 20 pF was selected for $C1$. If components have a greater percentage of tolerance, a trimmer capacitor $C1$ with a greater capacitance range will be needed, but this will then also reduce the resolution of the trimmer capacitor which will complicate the tuning process.

The resonant frequency of the LC network at the collector was determined with the equation $\omega_0 = (L_1 C_T)^{1/2}$ where $C_T = C_{22} + C_{eff} + C5' + C1$. $C_{eff} = (C2C3')/(C2 + C3')$ and $C3' = C3//Lin$.

Appendix C.3 lists the values of $C2$, $C3$, $C5$ and RL for each of the values of $n = 11$, $7.5 < n < 9.5$, $6 < n < 7$ and $5 < n < 4$.

3.9.2 Biasing

The impedance of the capacitor $C6$ must be lower than the impedance at the emitter input. Choosing $C6 = 10$ nF makes the impedance equal to 0.1 Ω which is approximately 100 times smaller than |y11| of the transistor (MPSH10) at 135 MHz.

The values of $R1 = 2.2$ kΩ, $R2 =$ 1k5Ω and $Re = 330$ Ω were calculated so that $I_C = 4$ mA (DC value) for $V_{be} = 0.7$ V, $V_{CC} = 5$ V and $h_{FE} = 60$. The y-parameters in the data sheet of the MPSH10 transistor were given for a collector current of 4 mA [17].

The capacitors $C7$ and $C8$ are decoupling capacitors.

3.9.3 Computed Responses

The MATLAB code for the simulation of the oscillator (smal-signal AC model) is in Appendix C.2. Zfb was simulated for three values, $Zfb = 23$ Ω, $Zfb =$ crystal model and $Zfb = C0$ (static capacitance of the crystal). The reasons are as follows:

- $Zfb = 23\Omega$: At resonance the crystal impedance is 23 Ω. The value of 23 Ω was determined through measurements and calculations The main purpose of simulating $Zfb = 23$ Ω is to determine if $\angle A_L = 0^o, 360^o$ at the peak of the gain of $|A_L|$. The upper and lower frequency boundaries of the LC network were determined at the collector for $\angle A_L = 0^o$. This can then be correlated with the theoretical calculated values.

- $Zfb =$ crystal model. To determine if $|A_L| \approx 6$ dB so that the design will allow tolerance to ensure that the oscillator starts at all temperatures and with some detuning. Figure 3.29 shows the model of the crystal which was used in the simulations with the MATLAB code in Appendix C.2. The frequency range was only from 135.120 MHz to 135.130 MHz in the simulations so that the effect of the crystal could be seen.

- $Zfb =$ static capacitance ($C0$). In order to determine if oscillations will occur due to the static capacitance ($C0$) The value of the open loop gain was determined for $C0$ and it was decided that $|A_L| <$ - 6 dB would be a safe margin to ensure that the oscillator will not oscillate due to the static capacitance. The inductor L0 should be added in parallel with the crystal to assist in keeping $|A_L| <$ - 6 dB when $|A_L| <$ - 6 dB for C0 = 7 pF. A value of 5.5 pF was measured but 7 pF was used as a maximum value for the static capacitance. The maximum value of the static capacitance equal to 7 pF was given by Mr. Neville Bird at STC Frequency Technology.

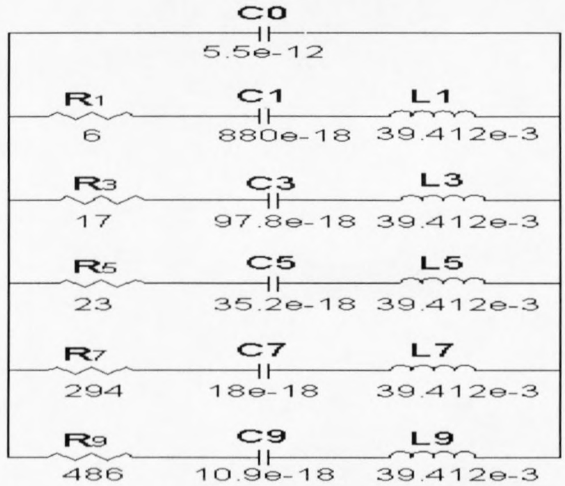

Figure 3.29: Model of crystal which was used in simulations

Figure 3.29 depicts the model of the crystal that was used in the simulations. The values for $C1$ to $C9$ and $L1$ to $L9$ were determined with $C_N = \frac{C_1}{N^2}$ and $L_N = L_1$ from [19]. $C1$ was selected as $\frac{C_0}{250*25} = \frac{5.5pF}{250*25}$ [19]. The values for $C0$, $R1$, $R3$ and $R5$ were determined through measurements. The values for R7 and R9 were determined from $R_N = R_1 N^2$ [19].

Appendix C.4 lists the lower and upper frequencies for $Zfb = 23\ \Omega$, open loop gain $|A_L|$ for $Zfb =$ crystal and the maximum value for $|A_L|$ when $Zfb = 7$ pF for the computed responses for the simulations of $n = 11$, $7.5 < n < 9.5$, $6 < n < 7$ and $4 < n < 5$. The lower and upper frequencies for $Zfb = 23\ \Omega$ were listed so that they can be compared with the calculated values. Only some of the possibilities listed in Appendix C.3 were simulated.

Figure 3.30 depicts the computed results of the magnitude and phase of the open loop gain for $Zfb = 23\ \Omega$. This is for an oscillator with $C2 = 12$ pF, $C3 = 120$ pF, $C5 = 3.9$ pF and $RL = 50\ \Omega$.

The phase of the open loop gain ($|A_L|$) was zero at the peak of the magnitude of the open loop gain ($|A_L|$) for the simulations for $n = 11$, some of the simulations for $7.5 < n < 9.5$ and $6 < n < 7$. The phase of the open loop gain (A_L) was at the right of the peak for the simulations for $4 < n < 5$, for some of the simulations for $6 < n < 7$ and $7.5 < n < 9.5$.

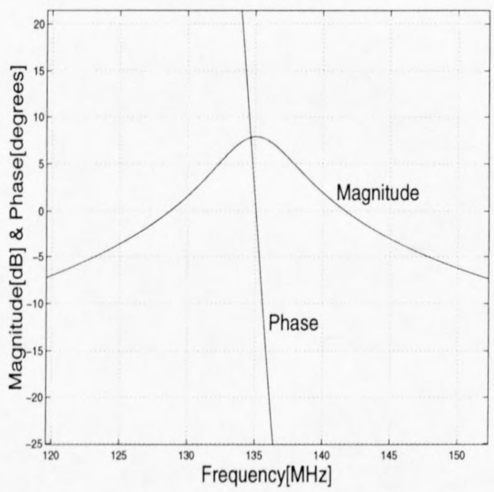

Figure 3.30: Magnitude and phase plot of A_L with $Zfb = 23\ \Omega$

Figure 3.31 depicts the computed results of the magnitude and phase of the open loop gain for Zfb = crystal. The phase of the open loop gain was zero at the peak of the magnitude of the open loop gain for all the simulations for $4 < n < 5$. The trimmer capacitor $C1$ had to be changed by 0.1 pF to 0.2 pF for $n = 11$, $7.5 < n < 9.5$ and $6 < n < 7$ so that $\angle A_L = 0^o, 360^o$ at the peak of the gain of $|A_L|$, when $Zfb = 23\ \Omega$ was replaced with Zfb = crystal in the simulations. The need to change $C1$ to ensure that $\angle A_L = 0^o, 360^o$ is at the peak of the gain of $|A_L|$ may indicate that the oscillator may not oscillate after a 23 Ω resistor is replaced with the crystal and power is applied.

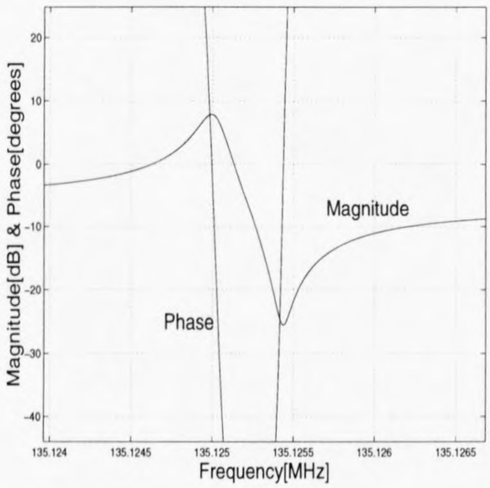

Figure 3.31: Magnitude and phase plot of A_L with Zfb = crystal

The magnitude of the open loop gain was greater than or equal to 5 dB for all the computed results. The magnitude of the open loop gain was greater than or equal to 6 dB for 33 of the 35 computed results.

The lower boundary for the output load $(RL'//Rp//R_{LX})$ of 500 Ω delivered a magnitude of 6 dB and more for $4 < n < 5$ and $6 < n < 7$. An output load $(RL'//Rp//R_{LX})$ equal to 682 Ω produced an open loop magnitude of 6.14 dB for $7.5 < n < 9.5$. An output load $(RL'//Rp//R_{LX})$ equal to 735 Ω produced an open loop magnitude of 5.2 dB for $n = 11$.

The dip in the magnitude of the open loop gain in Figure 3.31 is due to the static capacitor which resonates with the fifth order mode section of the crystal to cause a high impedance above the resonance frequency. This then causes a decrease in the magnitude of the open loop gain.

An inductor in parallel with the crystal in the feedback path is required for 34 of the 35 computed results to ensure that $|A_L| \leq$ - 6 dB. An inductor in parallel with the crystal in the feedback path is required for 30 of the 35 computed results to ensure that $|A_L| \leq$ - 5 dB. The value of the inductor to ensure that $|A_L| <$ - 6 dB is discussed in section 3.9.4.

There is a difference of less than 0.75 MHz between the calculated frequencies and those determined through simulation for the upper and lower frequency boundaries for the LC network on the collector of the oscillator circuit.

3.9.4 An Inductor (L0) in parallel with the Crystal

A value of $L0 = 310$ nH will ensure that the impedance of the inductor and static capacitor ($C0$) in parallel at 135 MHz is nearly an open circuit for $C0 = 4$ pF. The impedance for $C0 = 7$ pF in parallel with an inductor of 310 nH is 466 Ω which is still approximately 20 times the impedance of the fifth mode branch of the crystal at resonance.

A value of $L0 = 220$ nH is needed to cause an inductor in parallel with the crystal to be nearly an open circuit at 135 MHz for $C0 = 7$ pF. The problem is that although the impedance of a 4 pF capacitor and 220 nH inductor is 500 Ω at 135.125 MHz simulations showed that the magnitude of the open loop gain at 120 MHz is greater than - 6 dB and even 0 dB which led to the decision to choose $L0 = 310$ nH.

The computed results for the worst case reduced $|A_L| < 3$ dB to $|A_L| <$ - 15 dB for $C0 = 7$ pF and produced $|A_L| \leq$ - 6 dB for $C0 = 4$ pF.

3.9.5 Measurements

The frequency at which the circuit oscillates was measured with a magnetic pickup loop connected to a Hewlett Packard 8920A RF Communications Test Set. The measurements were done in the following manner:

- Zfb was implemented as a 22 Ω resistor to determine the upper and lower frequency boundaries of the LC network on the collector.

- The oscillator was then tuned to approximately 135.125 MHz with $Zfb = 22\ \Omega$.

- $Zfb = 22\ \Omega$ resistor was then replaced with $Zfb =$ crystal. Power was applied to the circuit to determine if the oscillator oscillates at the frequency of design.

- The power supply was reduced from 5 V to the level where the oscillator still oscillated. The output power was measured when RL was equal to 50 Ω. The spectrum analyzer or power meter acted as a 50 Ω load.

- The tuning range of the trimmer capacitor for which the oscillator oscillates was determined.

- The level where LC control may occur was determined by replacing $Zfb =$ crystal with $Zfb = C0$. The value for $C0$ where LC control started was then determined .

Measurements with oscillator circuits with $n = 11$ led to the decision to determine the response of the ratios for $6 < n < 7$ and $4 < n < 5$. The circuits with $n = 11$ did not always oscillate after the power was switched off and then on again. The oscillators with $n = 11$ oscillated for less than a 1/12 of the tuning range of the trimmer capacitor. The measurements with the three groups for the ratio of n which were evaluated led to the decision that the ratio for $4 < n < 5$, is the optimal ratio for n.

Appendix C.5 lists the measurements for $n = 11$, $6 < n < 7$ and $4 < n < 5$. Only two or three options of the simulations of each of the three ratios for n were built and measured.

The oscillation at the frequency of design occurred when $Zfb = 22\ \Omega$ was replaced with $Zfb =$ crystal for $4 < n < 5$. The oscillator circuits for $n = 11$ and $6 < n < 7$ had to be tuned so that the oscillator started to oscillate after $Zfb = 22\ \Omega$ was replaced with $Zfb =$ crystal.

The oscillator oscillated with a frequency tolerance of less than or equal to $\pm$ 0.003 % for $\pm$ 1/7 of the tuning range of the trimmer capacitor for $4 < n < 5$. The oscillator oscillated with a frequency tolerance of less than or equal to $\pm$ 0.003 % for less than a 1/11 of the tuning range of the trimmer capacitor for $n = 11$ and $6 < n < 7$.

The oscillator still oscillated with a 50% decrease of V_{CC} (5 V = 100%) (DC supply voltage) for $n = 11$, $6 < n < 7$ and $4 < n < 5$.

An output power of 7 dBm was measured for $6 < n < 7$ and 7.3 dBm for $4 < n < 5$ in a 50 Ω load.

Spurious (unwanted) frequencies (e.g. 27.307 MHz, 422.5 MHz, 449 MHz, 500 MHz) were observed with all the measurements with $Zfb =$ crystal when $C1$ was tuned such that the oscillator did not oscillate at 135.125 MHz. The spurious frequencies that were observed and the solution to solve the problem of spurious frequencies are discussed in section 3.9.6 .

The frequency of oscillation was controlled by the LC network on the collector output when $C0 > 10$ pF (computed results were $|A_L| = 1.4$ dB with $C0 = 7$ pF) and for $C0 > 22$ pF (computed results were $|A_L| =$ - 2.9 dB with $C0 = 7$ pF). This was for $4 < n < 5$.

3.9.6 Solution to Spurious Frequencies

The following spurious frequencies were observed when the measurements taken:

- $27.307MHz$, $\pm422.5MHz$, $\pm449MHz$,$\pm477MHz$, $\pm500MHz$
- $\pm530MHz$, $\pm558MHz$.

The spurious frequencies appeared simultaneously or started with $\pm$ 477 MHz and other frequencies mentioned. The spurious frequencies increased as the gain was increased by increasing V_{CC}. An 18 Ω resistor in series with the crystal in the feedback path or a 120 Ω resistor in series with Cb at the base of the transistor, lowered the gain so that the spurious frequencies disappeared and the oscillator still oscillated at the desired frequency. This was valid for $V_{CC} \leq 10$ V. A 12 Ω resistor in series with the crystal also solved the problem of unwanted frequencies with a maximum V_{CC} of 5.5 V.

A number of unwanted frequencies appeared simultaneously for a small percentage of the trimming range of the trimmer capacitor from 5 to 10 kHz below 135.125 MHz to 5 to 10 kHz above 135.125 MHz. The number of unwanted frequencies was not counted. This was discovered after the upper and lower frequency boundaries of the Spectrum analyser were decreased. The spurious frequencies did not disappear when the resistor was placed in the base or in the feedback path, but did so when an inductor of 180 nH was placed in parallel with the crystal. The unwanted frequencies could be heard on the speaker of the VHF receiver even though a FM modulated RF signal was not applied at the RF input of the receiver.

3.9.7 Conclusions

An oscillator that oscillated with a frequency tolerance of less than or equal to 0.003% was implemented with $C2 = 15$ pF, $C3 = 56$ pF, $RL = 50$ and $RL = 100\ \Omega$, $C5 = 4.7$ pF, $V_{CC} = 5$ V, $C1 = 4.2$ - 20 pF and $L1 = 50$ nH. The frequency tolerance was applicable when the oscillator was tuned or aligned without the necessary RF equipment. When the oscillator is tuned without the necessary RF equipment the frequency tolerance will be less than or equal to $\pm$ 3 kHz. A 120 Ω resistor in series with Cb or an 18 Ω resistor in series with the crystal and an inductor equal to approximately 180 nH in parallel with the crystal had to be included to ensure that there were no unwanted frequencies when the oscillator was tuned so that it does not oscillates at the frequency of design. A 180 nH inductor was implemented with a 190 mm length of wire with a 0.7 mm diameter wound with 10 turns around a diameter of 4 mm. RF output power of 7.3 dBm was measured with an 120 Ω resistor in series with Cb. The measuring port of the spectrum analyser acted as a fifty ohm load (RL) for the oscillator. A tolerance of $\pm$ 15% for L1 and $\pm$ 10% for Ct ($Ct = Ceff + C2 + C22 + C5'$) was accomplished with the trimmer capacitor $C1$ (4.2 - 20 pF). Figure 3.32 shows the schematic of the oscillator.

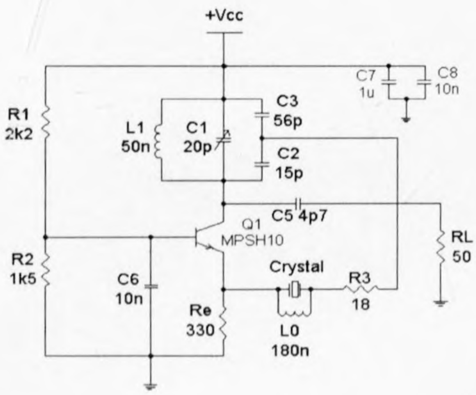

Figure 3.32: A spurious free oscillator with a frequency tolerance of $\pm$0.003%.

As the ratio of n decreased so did the output load decreased to keep $|A_L| = 6$ dB. The lower boundary value of $RL'//Rp//R_{LX} = 500\ \Omega$ to realise an open loop gain of $|A_L| \geq 6$ dB was valid for $4 < n < 7$.

There was a difference of less than 0.75 MHz between the calculated frequencies and those determined through simulation for the upper and lower frequency boundaries for the LC network on the collector of the oscillator circuit.

By replacing the 135.125 MHz crystal with the required crystal the frequency of the oscillator may be changed so that the receiver can receive any other frequency from 144 MHz to 146 MHz.

3.10 Audio Amplifier

Figure 3.33 shows the schematic of the audio amplifier. The audio amplifier with volume control is simply and economically implemented with the LM386 in the circuit, which is recommended in [18].

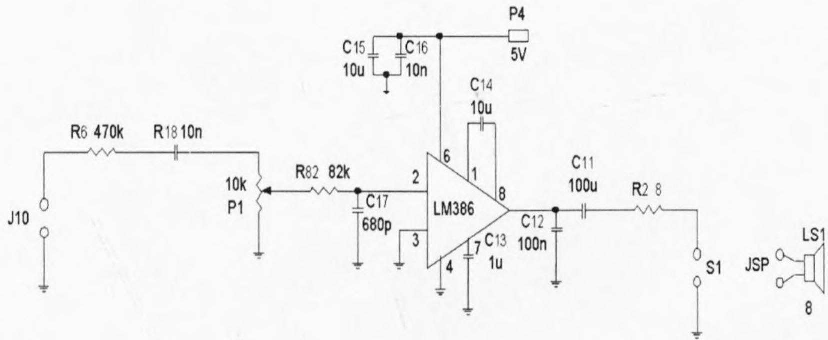

Figure 3.33: Audio Amplifier

3.11 RSSI Level Indicator

Figure 3.34 shows a schematic of the RSSI Level Indicator.

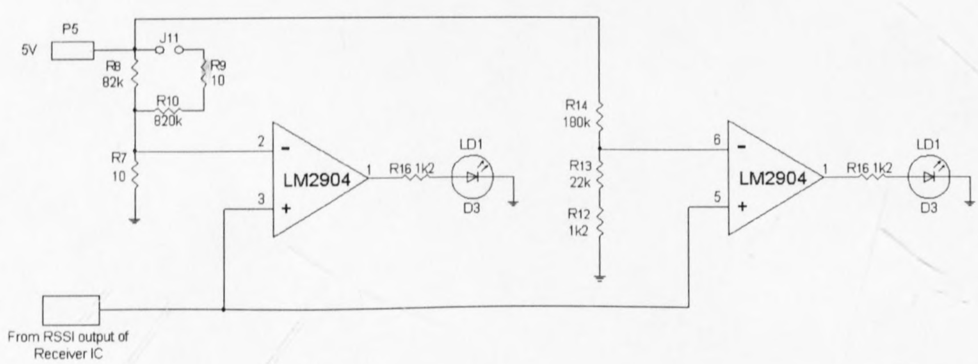

Figure 3.34: RSSI Level Indicator

The LED labeled $LD1$ in Figure 3.34 indicates (shines) when the RSSI at pin 12 of the receiver IC is above 0.57 V. The LED labeled $LD2$ in Figure 3.34 indicates (shines) when the RSSI at pin 12 of the receiver IC is above 0.45 V when $J9$ is not selected and when the RSSI is above 0.49 V when $J9$ is selected.

3.12 Broadband Signal Generator

R.F. (Radio Frequency) test and measurement equipment is not commonly available at schools or to the public. The equipment is usually limited to universities and companies that have an interest in the radio frequency field and have the money to buy such expensive equipment. It was decided to determine if there is an economical solution to apply "R.F. signals" (in the usual sense it can rather be called interference) to the fixed frequency receiver when the receiver is assembled and faults are to be determined. It is possible to generate a square wave with a fundamental frequency from 1 kHz to 20 MHz or even higher with logic IC's. The cost of these IC's is in the order of R 1.50 to R 2.50.

It was decided to investigate the matter of generating a square wave with a fundamental frequency of 1 kHz and 4 kHz. The power of the harmonics in a resistive load and the distribution of the harmonics at 455 kHz, 10.7 MHz and 145.825 MHz were analysed in section 3.12.1 .

Measurements were done to determine how effective the Broadband Signal Generator can be to apply test signals to the receiver at various places in the VHF receiver circuit. These tests were succesul. The final circuit for the Broadband Signal Generator is in Figure 3.39.

Figure 3.35 shows a schematic of the manner in which a square wave was generated with a Hex Schmitt-Trigger Inverter IC (74ACT14).

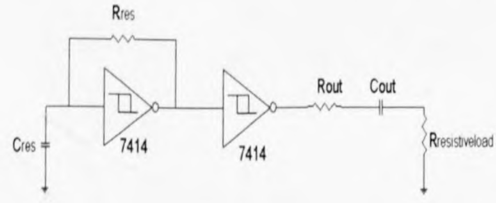

Figure 3.35: General configuration to generate a square wave

The values of $Rres$ and $Cres$ determines the frequency of the square wave and are approximately equal to $1/(0.66RresCres) = f_0$. The values Rres and Cres for a square wave with a fundamental frequency of 1 kHz were selected as $Rres$ = 100 kΩ and Cres = 15 nF. The values $Rres$ and $Cres$ for a square wave with a fundamental frequency of 4 kHz were selected as $Rres$ = 100 kΩ and $Cres$ = 3.9 nF. The values of $Rout$, $Cout$ and the load where the output is applied determine the output frequency response. A value of $Rout$ = 120 Ω was selected to limit the output current beneath 50 mA for a peak to peak voltage of 5 V for the square wave [25].

3.12.1 Calculations

The duty cycle of the of the square wave generated by the Broadband Signal Generator was measured as approximately 75 %. A duty cycle of 70% and 80% was included for calculations due to tolerance in IC's, resistors and capacitors. It was decided to change the duty cycle of the square wave to 50% only if the signal generator was not adequate to align or test the receiver.

Appendix D.1 gives the exponential Fourier components (C_N) for square waves with duty cycles of 70%, 75% and 80% to determine the distribution of the harmonics and the power of a harmonic in a resistive load. The MATLAB program Harm4.m, listed in Appendix D.4, was used to do the necessary calculations where needed.

Calculations showed that for a 12 kHz bandwidth at 455 kHz, 10.7 MHz and 145.825 MHz the difference between square waves with duty cycles of 70% , 75% and 80% were less than 0.5 dB. The power of each harmonic in a 50 Ω load was calculated and then added together for a bandwidth of 12 kHz. It was decided to work only with a duty cycle of 75 % due to the difference of less than 0.5 dB for the three duty cycles of the square waves.

Figure 3.36 depicts the Power Spectral Density (PSD) plots for two configurations A and B indicated in Figure 3.37, and a RC highpass filter labeled C. The highpass filter reduces the spectral density of the fundamental and lower harmonics to protect the receiver IC. The value of Rout, Cout and $R_{resistiveload}$ determines the - 3 dB frequency. The zeros of the envelope of the PSD plot are due to the fact that the square wave has finite rise/fall times. The frequency of the first zero of the envelope is equal to the inverse of the rise/fall time of the square wave. The zero is close or at 145.825 MHz for a rise/fall time of 6.9 ns. A zero at 145.825 MHz will result a signal generator which is not adequate to align the receiver at 145.825 MHz. A rise/fall time of less than 5 ns was measured. A rise/fall time of up to 12.5 ns is possible according to [25] which may thus include a IC that has a rise/fall time of 6.9 ns and will not be adequate as a signal generator at 145 MHz. A rise/fall time of less than or equal to 12.5 ns will produce a change of less than 2 dB in the power of the harmonics below 10.7 MHz.

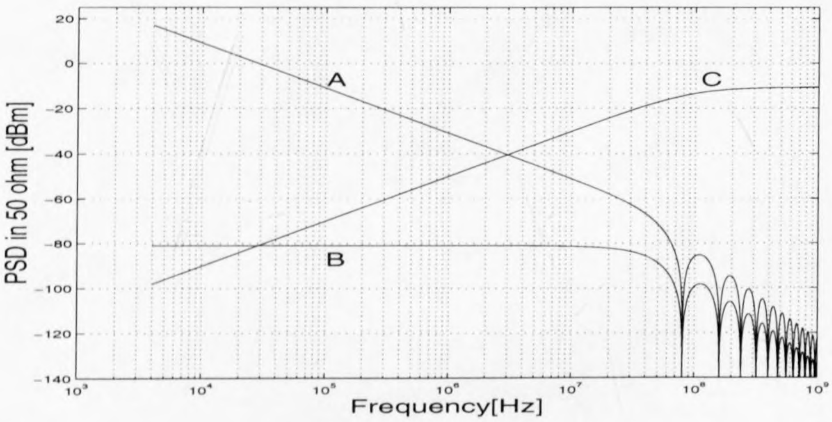

Figure 3.36: PSD plots of the Broadband Signal Generator output

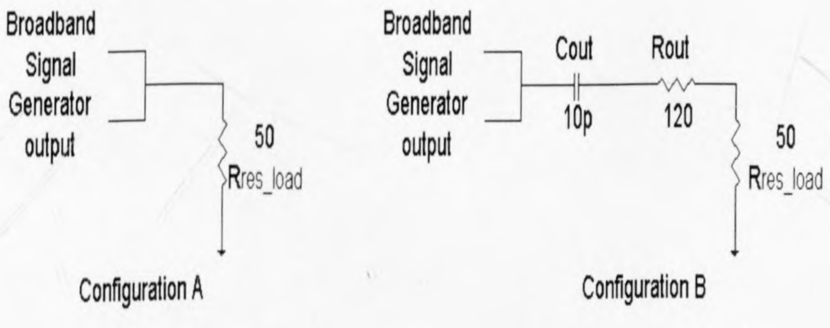

Figure 3.37: Configuration A and B for PSD plots

The calculated values for power of a harmonic in a resistive load is listed in Appendix D.3. The calculations were done for a fundamental frequency of 1 kHz and 4 kHz, $R_{resistiveload}$ = 50 Ω, 1.5 kΩ and 4 kΩ and $Cout$ = 10/100 pF at 455 kHz, 10.7 MHz and 145.825 MHz. The values for the $R_{resistiveload}$ were selected to be values of the input impedances at the VHF receiver where signals have to be applied. The value for $Cout$ = 10 pF was selected to suppress the lower harmonics when a signal is applied at the RF inputs (145.825 MHz). The value for $Cout$ = 100 pF was selected to utilise the harmonics at 455 kHz and 10.7 MHz better than $Cout$ = 10 pF. The best results were obtained with a square wave with a fundamental frequency of 4 kHz and $Cout$ = 100 pF.

Figure 3.38 shows the harmonics for square waves with a fundamental frequency of 1 kHz and 4 kHz. The harmonics are approximately 10 dB greater for a 4 kHz square wave than for a 1 kHz square wave, but the harmonics of an 1 kHz square wave are closer spaced than for a 4 kHz square wave.

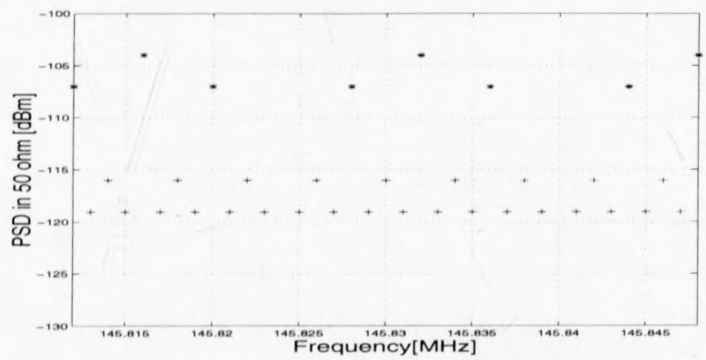

Figure 3.38: Harmonics for square waves with fundamental frequencies of 1 kHz (+) and 4 kHz (∗) separately.

An audio tone approximately equal to the difference between the frequency of two CW inputs at the receiver was heard when they were applied at the RF input of the receiver.

The three harmonics of the 4 kHz square wave at 145.828 MHz, 145.832 MHz and 145.836 MHz or any other three harmonic spectra of the 4 kHz square wave spaced 4 kHz apart, with the center line spectra the biggest amplitude, resemble the line spectra of a 4 kHz FM modulated signal (3 kHz deviation). The line spectra at 145.832 MHz is 3 dB greater than the line spectra of the harmonics at 145.828 MHz and 145.836 MHz. The center line spectrum of 4 kHz FM modulated signal (3 kHz deviation) is 6 dB greater than the line spectra left and right of the center line spectrum. The weighting functions of the line spectra of the square wave and a 4 kHz FM modulated signal (3 kHz deviation) are not the same, but the space between the line spectra and the "envelope" shape of the line spectra are similar.

The measurement and observation refer in the two preceding paragraphs led to the conclusion that an audio output other than an increase of noise would be heard at the speaker.

3.12.2 Measurements

The output of the Broadband Signal Generator was applied to various places on the VHF receiver to determine the response.

Table 3.1 lists the values for $C1$ in Figure 3.37 and the fundamental frequency of the square wave which delivered the best audio output, highest voltage level at the RSSI level indicator and the position on the VHF receiver where the signal was applied. An audio tone (clear or distorted) was heard with each of the measurements taken in Table 3.1.

Table 3.1: Best results obtained with Broadband Signal Generator at VHF receiver

Place	Frequency f_0	$C1$	RSSI level [V]
Audio speaker	4 kHz	10 pF	not applicable
Audio amplifier	4 kHz	100 pF	not applicable
455 kHz Limiter input	4 kHz	100 pF	0.47
second mixer input (10.7 MHz IF input)	4 kHz	100 pF	0.5
RF matched input receiver IC	4 kHz	10 pF	0.6
BPF1	4 kHz	10 pF	0.6
BPF2	4 kHz	10 pF	0.7
BPF3	4 kHz	10 pF	0.78

The 455 kHz Sub - Miniature IF Transformer will be referred to as the Quadrature Coil in this section. The signal generator will only affirm if the Limiter section and second mixer section of the receiver IC operate. The indication of the RSSI level indicator and the audio at the speaker do not change when the Quadrature Coil is tuned below, at or above 455 kHz.

The Quadrature Coil, Bandpass Matching network and the crystal oscillator are aligned and tuned by what is heard at the speaker and the indication of the LED's at the RSSI Level indicator. The Quadrature Coil can be tuned to 455 kHz by tuning it until the maximum noise is heard at the speaker. A distorted audio tone is heard when the oscillator oscillates and the bandpass network is tuned at optimum and the Quadrature Coil is tuned approximately 30 to 45 degrees away from the position where the maximum noise is heard. When the Quadrature Coil is tuned to 455 kHz the noise at the discriminator output dominates the audio tone of the signal generator at the speaker. The harmonics nearest to 145.825 MHz are 5 kHz below and 3 kHz above at 145.820 MHz and 145.828 MHz separately when the Quadrature Coil is tuned at 455 kHz. The spacing of the nearest harmonics 3 kHz and 5 kHz away from 145.825 MHz may be a reason why the output noise of the discriminator dominates the output at the speaker.

Both LED's of the RSSI Level Indicator shine when the Bandpass Matching network is tuned at optimum performance. Detail instructions of how to align the Quadrature Coil, Bandpass Matching network and the crystal oscillator at the first mixer are given in Appendix E.

Bandpass Filters 1, 2 and 3 are aligned by tuning the trimmer capacitors until the best indication is given by the RSSI Level Indicator. The Quadrature Coil should be tuned 30 to 45 degrees away from the place where the maximum noise is heard. This should be done to assure that the distorted audio tone is heard when the filters are aligned. Detailed instructions of how to align Bandpass Filter 1, 2 and 3 are given in Appendix E.

When the signal generator is applied at the unmatched RF input of the receiver IC a soft distorted audio tone is heard. The output at the speaker does not change while the local oscillator or the Quadrature Coil is tuned. The signal generator can thus not be used to tune the oscillator by applying a signal at the unmatched RF input of the receiver IC.

3.12.3 Conclusion

The Broadband Signal Generator can be used to test and/or align the following:

- Audio speaker
- Audio amplifier
- 455 kHz limiter input of receiver IC
- Second mixer stage of receiver IC
- Bandpass matching network, crystal oscillator and LC Quadrature Coil at the FM demodulator of the receiver IC
- Bandpass Filter 1 to Bandpass Filter 3

Figure 3.39 shows the schematic of the final Broadband Signal Generator. The jumpers $J1$ to $J3$ select which capacitor is connected to the output of the signal generator.

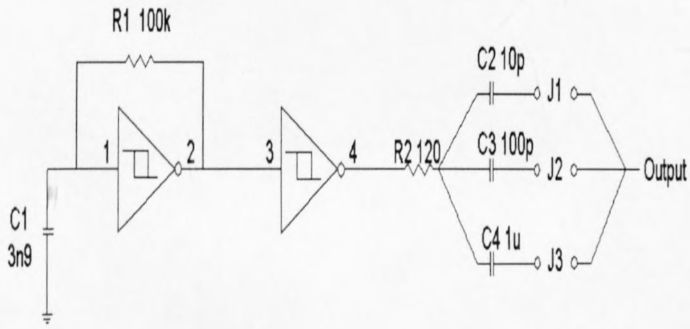

Figure 3.39: Final schematic of Broadband signal generator

3.13 Power Supply

Figure 3.40 shows a schematic of the power supply. A 5 V regulator (marked U1 in the schematic) is used to protect the electronic circuitry of the receiver and allow any DC source between 7 V and 24 V [23] as an input to the power supply. A 9 V PP3 size battery can be used to operate the receiver.

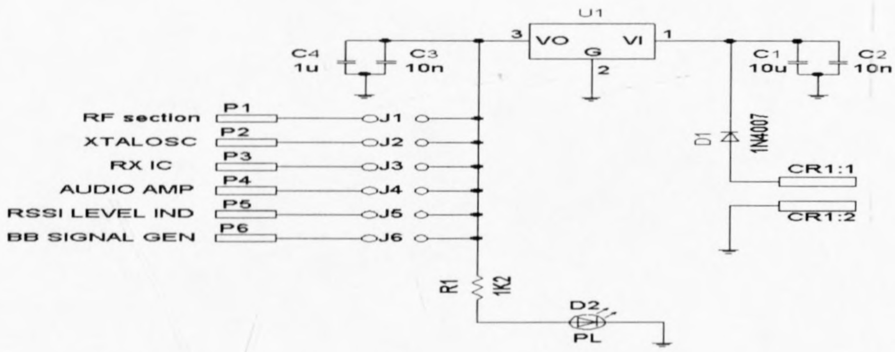

Figure 3.40: Power Supply

The diode $D1$ in Figure 3.40 is used for polarity protection. The LED labeled $D2$ indicates if there is DC power on the output of the regulator.

Jumpers $J1$ to $J6$ can be selected to activate/deactivate those circuits which are supplied with DC power through the jumpers. This will be of great value during a fault-finding process.

3.14 Prices of components

Table 3.2 lists the prices when the components were bought as well as in February 2001. The price for the ground station was approximately R 198 (in 2000) and it is approximately R 218 (February 2001). The total price excludes the price for the PCB (Printed Circuit Board). For a pupil who through SUNSTEP kits has become interested in electronics the VHF fixed frequency receiver is a more economical way of accessing the Amateur Radio world than a scanner of R 1000. The prices for 135.125 MHz crystals increased from 5 for R 55 each in 2000 to R 52 for one crystal when 26 to 49 crystals are ordered.

Table 3.2: Prices of components

No. of in RX	Component	Price when bought (in 2000)	Price 2001
1	MC13136P	R 20	R 24
1	135.125 MHz crystals	R 55	R 52
1	10.245 MHz crystals	R 12	R 2.68
1	10.7 IF filter	R 1.79	R 1.80
1	455 kHz IF filter	R 7.85	R 12.55
1	MPSH10 transistor	R 0.81	R 0.81
1	LM2904	R 3.15	R 3.15
1	LM386	R 3.30	R 3.30
34	1/8 resistors	R 0.0275	R 0.0275
1	10 kΩ Pot	R 4	R 5.40
1	L7805 (5 V Regulator)	R 1.75	R 2.08
1	8 Ω speaker	R 4.50	R 4.50
69	Capacitors	R 0.22	R 0.22
2	BNC connector (Female)	R 5.35	R 9.60
3	LED's	± R 0.30	R 0.21
1	IC Socket 24 pin	R 2.47	R 1.75
1	IC Socket 14 pin	R 1.26	R 1.12
2	IC Socket 8 pin	R 1.26	R 0.67
1	74ACT14	R 1.56	R 1.56
2	MAR6 (mini circuits)	R 12	R 13
4	30 pF trimmer capacitors	R 1.25	R 2.47
4	20 pF trimmer capacitors	R 1.25	R 2.47
1	YRCS11098 Sub - Miniature IF transformers	R 7.56	R 7.56
1	RCA inline (M)	R 1.40	R 2.70
1	RCA inline (F)	R 2.10	R 2.70
	Total	R 198	R 218

Chapter 4

Utilising a sound card as a modem.

A PC is needed in the process of data communications in Amateur Radio. Most PC's have a sound card installed in the computer when it is bought. A sound card costs less than a tenth of the price of a PC, which is why the possibility of utilising a sound card as a modem was investigated.

The process of receiving 1200 baud AFSK data, from the moment of receiving an audio signal from the receiver until it is displayed on a PC is, discussed in section 4.1.
Section 4.2 explains how the sound card was utilised as a modem and how the data is displayed on a PC.

4.1 Background

Figure 4.1 depicts the functional block diagram for 1200 baud AFSK data communication from that the audio signal is received until it is displayed on a PC.

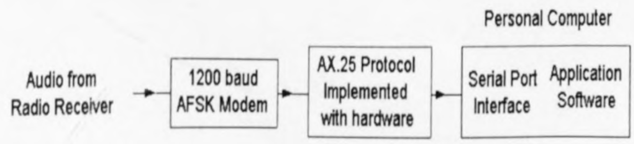

Figure 4.1: Basic blockdiagram for 1200 baud AFSK data communication

The 1200 baud AFSK modem in Figure 4.1 converts the analogue signals into digital signals. There are two modulation standards for 1200 baud AFSK. The modulation standards are the CCITT V.23 modulation standard and the BELL 202 modulation standard. The CCITT V.23 modulation standard should be used for the 1200 baud AFSK modem for optimal communication with SUNSAT 1 . The BELL 202 modulation standard can also be used in the 1200 baud AFSK modem, but due to a slight difference in standards there will be a degradation of the signal performance.

The bit stream is sent from the modem to the hardware which implements the AX.25 protocol. The AX.25 protocol was implemented for reliable data communications for Radio Amateurs. Data is sent as small blocks of information called frames. The three basic types of frames are the Information frame (I frame), the Supervisory frame (S frame) and the Unnumbered frame (U frame) [11]. Each of the three frames contains blocks of information called fields. The fields which are contained by the three frames are the flag field, the address field, the control field and the frame-check sequence (FCS) field. The I and UI frames also contain an information field and a protocol identifier field. The flag field indicates the beginning and ending of a frame. The address field indicates the source and destination of the frame. The control field identifies what type of frame has been received. The FCS field determines if the frame which has been received is a valid frame. The information field conveys user data [2]. The protocol identifier field indicates which level three protocol in the Open Systems Interconnection Reference Model (OSI - RM) is used, if any is used [2, 11]. The AX.25 protocol has the option of transmitting data in an unconnected mode which means the frame received is only detected if it is a valid AX.25 frame. There are no other acknowledgements that the transmitted frame has been received. The frames which are transmitted in an unconnected mode are UI frames. The unconnected mode is useful when data can only be received and a transmitter is not available, for example when a school pupil has only the fixed frequency receiver module of the ground station.

The data is now sent to the serial port of the PC and it is then displayed by the application software on the PC. The necessary configuration will have to be made between the AX.25 which is implemented with hardware and the application software used by the PC.

4.2 Sound Card Utilisation

Figure 4.2 shows the block diagram of how a sound card was utilised as a radio modem for receiving 1200 AFSK data from another Amateur Radio station. UI (unconnected mode) (AX.25 protocol) frames were transmitted.

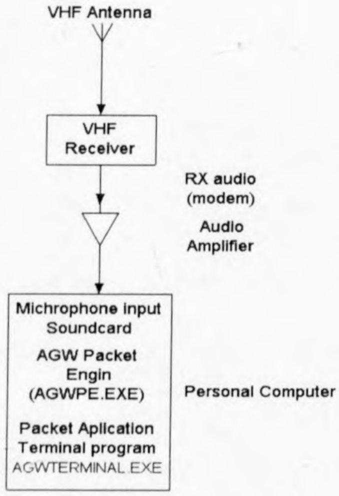

Figure 4.2: The utilisation of a sound card as a radio modem

The VHF antenna in Figure 4.2 was a monopole antenna (1/4 wave length) and the VHF Fixed Frequency Receiver was used as a VHF receiver. The audio signal at the modem output was fed to the microphone input of the soundcard via an audio amplifier. The audio amplifier was the exact copy of the audio amplifier which is indicated in chapter 3, section 3.10.

The two programs (AGWPE.exe and AGWTERM.exe) that were used were down-loaded from [32] as zip files. AGWPE is the packet engine which activated the soundcard as a modem. It also does the processing of the AX.25 protocol. The Bell modulation standard is used. AGWTERM is an application program which was used to display the data which was received.

The software operates satisfactorily. The AGWPE.hlp file gives instructions of how to install the packet engine program. The help file and packet engine file refer to the sound card as a tnc (terminal node controller) although it is only utilised as a modem. A serial port which is not used should be selected to activate the push to talk (PTT) even though data is only received. There is not an option of not selecting a serial port when data is only received. The packet engine or application software will not operate correctly if a serial port which is already occupied is selected in the properties of the agwpe.exe program. The packet engine should be installed before the application program is run. The AGWTERM.hlp file gives the instructions for the configuration of the application program.

Far less hardware is needed when a sound card and the necessary software are used for receiving 1200 baud AFSK data than with the method outlined in section 4.1.

Chapter 5

Measurements and Results

5.1 RF Section (VHF receiver)

The RF section of the VHF receiver consists of the series connection of Bandpass Filter 3, Amplifier 2, Bandpass Filter 2, Amplifier 1 and Bandpass Filter 1. The measurements that were done on the RF section are discussed in the following two sub-sections.

5.1.1 Frequency Response and Gain

Frequency response and gain measurements were done with the HP8753C Network Analyser and HP 85047A S - parameter Test set. The frequency response measurements were done separately to determine how the designed frequency responses compared with the measured responses of the bandpass filters (Bandpass Filter 3, 2 and 1) . The frequency response of the RF section (Bandpass filter 3, amplifier 2, Bandpass filter 2, amplifier 1 and Bandpass filter 1 in series) was measured to determine how much the image frequency is suppressed and what the 3 dB bandwidth is.

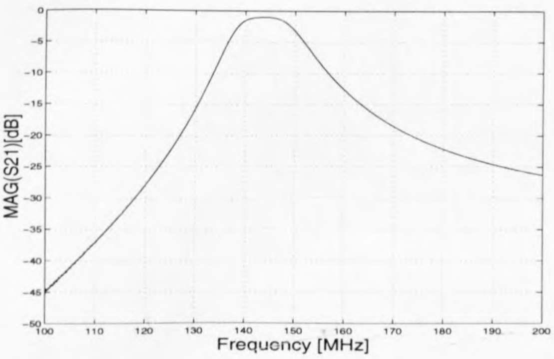

Figure 5.1: S21 response measurement of Bandpass Filter 3

Figure 5.1 shows the measurement of the transmission coefficient (S21) response of Bandpass Filter 3. The attenuation is equal to - 37 dB at 110 MHz while it is - 22 dB at 180 MHz. Both are 35 MHz away from 145 MHz but have different attenuations. The difference in attenuations is due to the coupling capacitor. The image frequency (124.425 MHz) is suppressed by 22 dB, which is 3.5 dB less than the simulated results. The insertion loss is approximately 1.1 dB at 145.825 MHz which is 0.1 dB more than the 1 dB which was set as a design goal. The 3 dB bandwidth is approximately equal to 13.6 MHz which is 0.6 MHz more than the simulated response in chapter 3.

Figure 5.2 shows the transmission coefficient (S21) response of Bandpass Filter 2. The image frequency is suppressed by 43 dB which is 2 dB more than the simulated results of chapter 3. The insertion loss is approximately 2.45 dB at 145.825 MHz and the 3 dB bandwidth is approximately equal to 5 MHz. The insertion loss and 3 dB bandwidth met the design goals and agreed with the simulated results of chapter 3 which are an insertion loss of less than or equal to 2.5 dB and a bandwidth of less than or equal to 5.2 MHz.

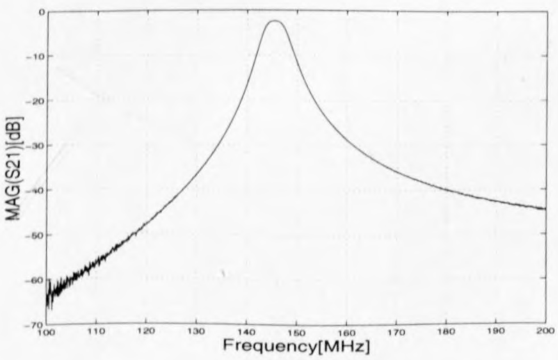

Figure 5.2: S21 response measurement of Bandpass Filter 2

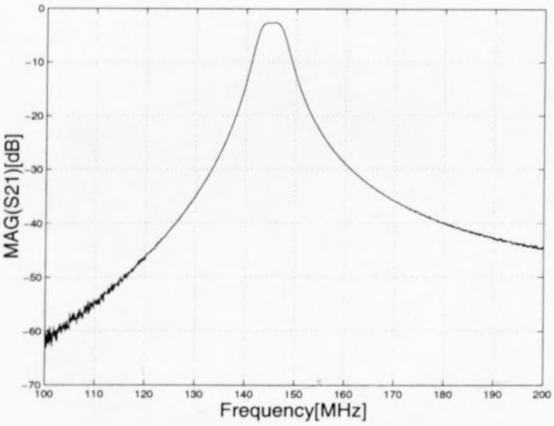

Figure 5.3: S21 response measurement of Bandpass Filter 1

Figure 5.3 shows the transmission coefficient (S21) response of Bandpass Filter 1. The image frequency is suppressed by 42 dB (1 dB more than the simulated results). The insertion loss is 2.6 dB at 145.825 MHz which is 0.1 dB more than the design goal set in chapter 3. The 3 dB bandwidth is approximately equal to 5.7 MHz which is 0.5 MHz more than the simulated results in chapter 3.

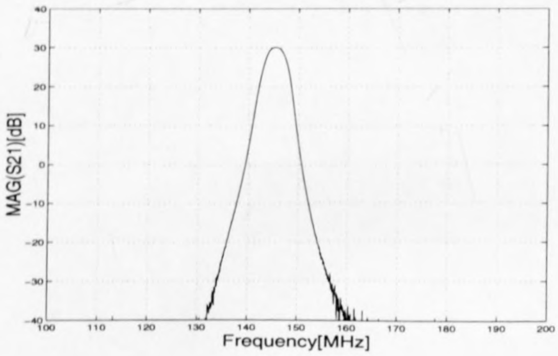

Figure 5.4: S21 response of the RF section (VHF receiver)

Figure 5.4 shows the transmission coefficient (S21) response of the RF section (Bandpass filter 3, amplifier 2, Bandpass filter 2, amplifier 1 and Bandpass filter 1 in series) of the VHF receiver. The magnitude of the transmission coefficient (S21) is 30 dB from 145 MHz to 146 MHz and the 3 dB bandwidth is 4.5 MHz. The image frequency is rejected by more than 70 dB.

Figure 5.5 shows the gain plot of the RF section. The measurement was done when the noise figure was measured. The gain of the MAR6 is 19 dB $\pm$ 1/2 dB from 100 MHz to

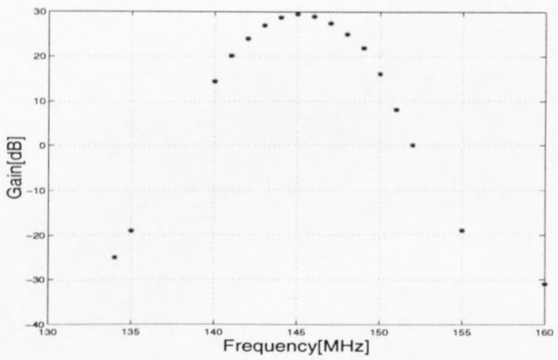

Figure 5.5: Gain measurement of the RF section (VHF receiver)

200 MHz.

5.1.2 Noise Figure

The noise figure was measured with the 8970B Noise Figure Meter and the 346B Noise Source. The Faraday room was closed when the measurements were done.

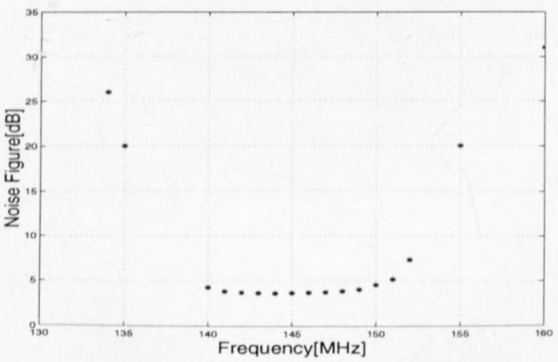

Figure 5.6: Noise Figure of RF section (VHF Receiver)

Figure 5.6 shows a plot of the noise figure of the RF section. The noise figure of the MAR6 is less than 2.85 dB from 100 MHz to 200 MHz.

5.2 Two-Signal Spurious Response Measurements of Receiver

The purpose of the the two-signal spurious measurements is to determine how well the receiver rejects spurious (unwanted) signals while receiving the desired signal.

Figure 5.7 shows the measuring configuration for the two-signal spurious measurements. The Fluke 6060B Synthesized RF Signal Generator was used as the sweep generator. The HP8660D Synthesized Signal Generator was used as the signal generator and was set at 145.825 MHz which is the desired frequency. A Mini Circuits LAPS-2-4 Combiner was used as a combiner. A 9 V battery was used as a power source for the receiver to reduce the spurious responses due to the power supply. The following three configurations were applied for as Device Under Test (DUT). They were the receiver IC not matched to 50 Ω, the receiver IC matched to 50 Ω and the input of the RF section of the VHF receiver. The Hewlet Packard 334A Distortion Analyser was used as a SINAD meter.

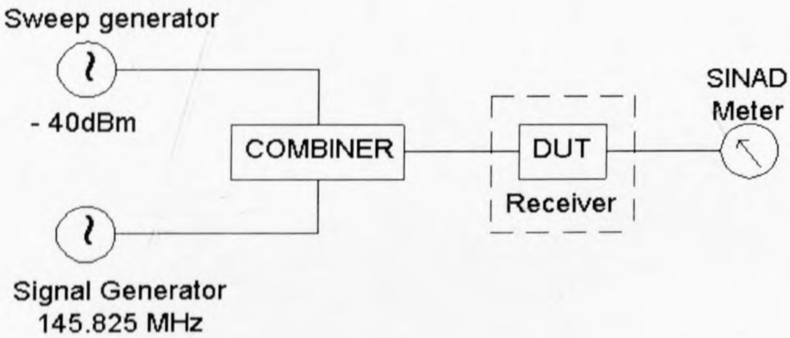

Figure 5.7: Two-signal spurious response measurements configuration

The measurements were done in the following manner. The signal generator in Figure 5.7 applied a RF signal at 145.825 MHz (1 kHz FM modulated with 3 kHz deviation) with a signal level so that a 12 dB SINAD was measured. The Sweep generator was set to apply a RF signal (400 Hz FM modulated with 3 kHz deviation) of - 40 dBm and was than swept over a frequency range of 100 to 200 MHz with a frequency step of 10 kHz. The frequencies were written down where a reduction in the SINAD was observed. The RF output at the spurious frequency was then decreased until a SINAD of 12 dB was obtained again. The difference in modulating frequencies was used to distinguish the audio output of the two RF generators.

The amount a spurious signal is rejected is equal to the level of the sweep generator minus the level of the RF generator to produce a 12 dB SINAD. Equation 5.1 shows how the amount of rejection was determined.

$$\text{Rejection} = \text{RF Level on sweep generator} - \text{Level for 12 dB SINAD at signal generator} \quad (5.1)$$

Figure 5.8 shows the plot of the spurious response measurements with the receiver IC not matched to 50 Ω. The signal generator was set to a RF output power of - 94 dBm to produce a SINAD of 12 dB.

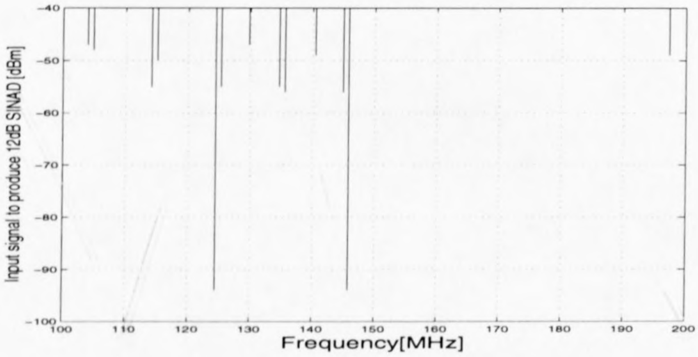

Figure 5.8: Spurious response measurements of receiver IC with no matching

There were no spurious responses for a RF input less than and equal to - 45 dBm at 455 kHz and for a RF input less than and equal to - 85 dBm at 10.7 MHz. The spurious response at 124.425 MHz is the image frequency and is rejected by 0 dB. The spurious responses at 125.335 MHz and 144.915 MHz mix down to 9.79 MHz which is the image frequency of the second mixer in the receiver IC. They are rejected by 39 dB and 38 dB separately. It was mentioned in section 3.8.4 that a crystal filter will reject the second image frequency by 80 dB but costs 10 times the price of a ceramic filter.

Table 5.1 lists the spurious frequencies in Figure 5.8 due to harmonic mixing between a harmonic of the spurious frequency and a harmonic of the local oscillator (LO). The amount that the response is rejected is also indicated in the column, Rejection Level.

Table 5.1: Spurious responses due to harmonic mixing

Frequency [MHz]	Rejection Level [dB]	Harmonic of RF Frequency	Harmonic of LO
129.775	47	2	2
140.475	45	2	2
197.340	45	2	3

Table 5.2 shows the spurious responses in the RF section which produce a signal in the first IF section which is 455 kHz above or below a second or higher harmonic of the local oscillator (10.245 MHz) at the second mixer which then mixes down to 455 kHz in the second IF.

Table 5.2: Spurious responses (harmonic mixing at second mixer)

Spurious Response [MHz]	Rejection Level [dB]	Harmonic of Spurious Response	Harmonic of First LO	Harmonic of Second LO
103.935	47	1	1	3
104.845	46	1	1	3
114.180	39	1	1	2
115.090	44	1	1	2

The spurious responses at 134.670 MHz and 135.580 MHz mix down to 455 kHz at the first IF and pass through the second mixer to the second IF. They are rejected by 39 dB and 38 dB separately.

Figure 5.9 depicts the spurious response measurement plot of the receiver IC matched to 50 Ω. The signal generator (receiver frequency) was set to a RF output power of - 99 dBm for a 12 dB SINAD.

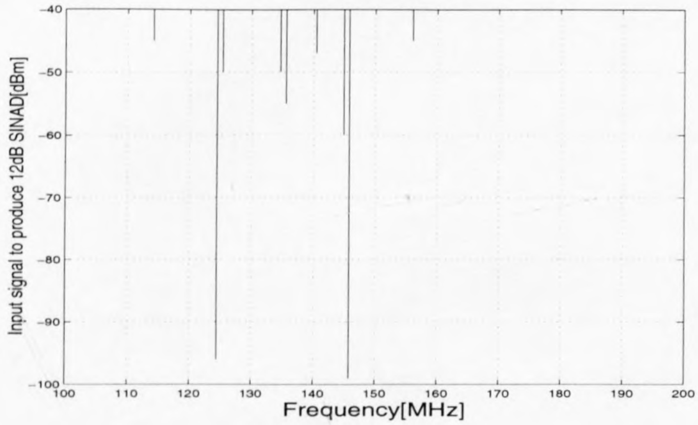

Figure 5.9: Spurious response measurements of receiver IC matched to 50 Ω

There was no spurious response measured with a RF input of - 40 dBm at 455 kHz and 10.7 MHz. There are 4 spurious frequencies less than when there was no matching at the RF input of the receiver IC. The spurious frequency at 156.070 MHz is a new spurious response. The response mixes down to 20.945 MHz at the first IF where it then mixes with the second harmonic of the 10.245 MHz Colpitts oscillator down to 455 kHz. The bandpass matching network has helped in reducing the spurious responses for a RF input of - 40 dBm. Table 5.3 lists the responses with the amount of rejection of each response.

Table 5.3: Spurious responses

Spurious Response [MHz]	Rejection Level [dB]
114.180	54
124.425	3
125.335	49
134.670	49
135.580	44
140.0	52
144.915	39
156.070	54

Figure 5.10 shows the spurious response measurements plot of the fully operated receiver (RF section included). The signal generator (receiver frequency) was set to a RF output power of - 119 dBm for a 12 dB SINAD.

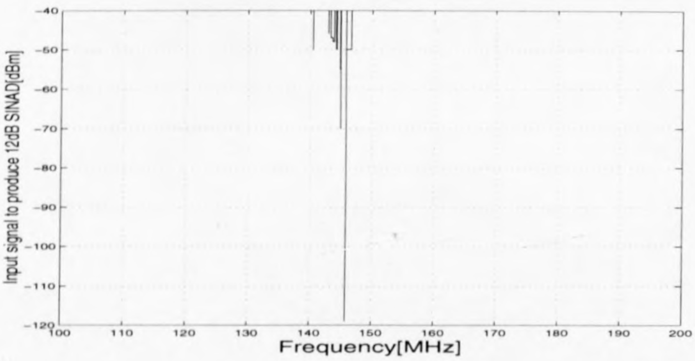

Figure 5.10: Spurious response measurements of the total Receiver

The two spurious responses which remained from the previous measurements are 140.475 MHz and 144.915 MHz which are rejected by 70 dB and 50 dB separately. The SINAD decreased by 1 to 4 dB for responses from approximately 143.290 MHz to 147 MHz. The image frequency is suppressed by 80 dB which is good. The frequency responses below 140 MHz and above 150 MHz are also suppressed by 80 dB. The spurious responses between 140 MHz and 150 MHz are due to the lack of sufficient filtering in the RF section and IF filters. At 280.950 MHz is a spurious response which is rejected for an input of - 60 dBm and less. The spurious response at 280.950 MHz mixes with the second harmonic of the first local oscillator to cause a 10.7 MHz IF response. The sweep generator in Figure 5.7 was only swept from 255 MHz to 300 MHz and from 390 MHz to 440 MHz to determine the effect of the 2nd and 3rd harmonic of the local oscillator. This was done separately from the frequency sweep from 100 MHz to 200 MHz.

A 10.7 MHz bandpass filter will increase the price of the receiver with less than 15% and will increase the rejection of 144.915 MHz from 50 dB to 80 dB.

5.3 Sensitivity

Figure 5.11 shows the configuration which was used to measure the sensitivity of the receiver. A RF signal was applied at the RF input of the receiver IC (not matched to 50 Ω), RF input of the receiver IC (matched to 50 Ω) and the RF section included. An 1 kHz FM Modulated RF signal (3 kHz deviation) was applied with the Fluke RF signal generator at the various places mentioned at the receiver.

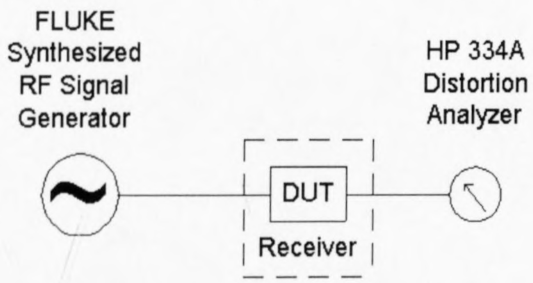

Figure 5.11: Configuration to measure SINAD

Figure 5.12 shows the SINAD plot for the three conditions previously mentioned. A 15 dB SINAD was measured for a - 120 dBm 1 kHz FM Modulated RF signal (3 kHz deviation). The specification of 12 dB SINAD was thus met and even surpassed.

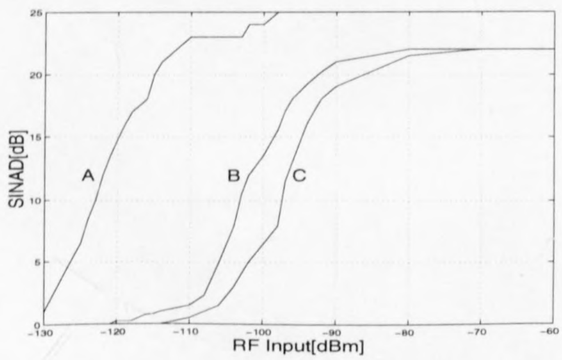

Figure 5.12: SINAD for unmatched IC (C), matched IC (B) and RF section included (A).

Figure 5.13 shows the RSSI plot for the three conditions previously mentioned.

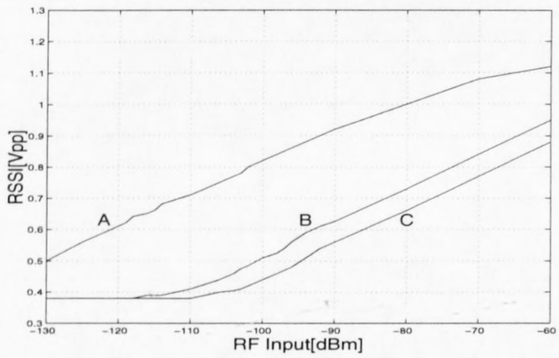

Figure 5.13: RSSI for receiver IC unmatched (C), receiver IC matched (B) and RF section included (A).

5.4 LO Radiation

A RF signal of - 70 dBm was measured at the input of the receiver for a 0 dBm input at the synthesizer input of the receiver.

A signal of - 66 dBm was measured at RF input of the receiver when the crystal was selected as a Local Oscillator.

5.5 Power Consumption (Current measurements)

Table 5.4 lists the current measurements of the receiver for various sections in the receiver.

Table 5.4: The current measurement of the receiver

Section	Current (DC) [mA]
Total Receiver (Volume zero)	63
Total Receiver (Full Volume)	91
Regulator	7.7
RF section	29.3
Local Oscillator	12
Receiver IC	5
Broadband Signal Generator	4

It clearly shows that 33% of the current at full volume is drawn by the audio amplifier when the audio amplifier is set at full volume.

5.6 Practical Measurements

On Saturday, 6 May 2000, signals were clearly heard over the speaker when RF signals (145.825 MHz) from SUNSAT I were received during a pass set out for the Radio Amateurs to use SUNSAT I as a repeater (Mode B). A monopole antenna (quarter wave) was used as an antenna with coax and BNC connectors. A 9 volt PP3 battery was used as power supply for the VHF fixed frequency receiver.

On Wednesday, 21 of February 2001, audio signals (audio tones) were heard over the speaker of the VHF fixed frequency receiver (RF frequency = 145.825 MHz). OSCAR 11 (UoSat 2) was transmitting 1200 baud PSK data on 145.825 MHz [33]. The same configuration was used as for the measurement carried out in the previous paragraph .

A SINAD of 5 dB was measured for a RF input signal (FM modulated) of - 120 dBm at the VHF receiver after the receiver was aligned with the Broadband Signal Generator. A RSSI equal to 0.78 V was measured at the RSSI output of the receiver IC. The receiver produced a 15 dB SINAD for a RF input signal of - 120 dBm when it was tuned with a network analyser and RF Signal Generator.

Chapter 6

Conclusions

6.1 Results obtained

The following were achieved in this thesis:

- The system level design of a ground station for school pupils was selected so that the Parrot Repeater mode and the voice/data Mode J repeater mode on SUNSAT I could be operated.

- A VHF fixed frequency receiver, which is capable of receiving RF signals from an operational SUNSAT I, was designed and built.

- A SINAD of 15 dB was measured for a RF input (1 kHz NBFM modulated (3 kHz deviation) signal) of -120 dBm.

- The spurious measurements showed that the receiver has no spurious responses for an RF input of - 70 dBm or weaker.

- The image rejection was 80 dB according to the two signal spurious measurements.

- A crystal controlled oscillator, which oscillates with a frequency tolerance of less than or equal to $\pm$ 0.003%, was designed and built. The frequency tolerance is applicable when the oscillator is aligned or tuned without the necessary RF equipment.

- An economical Broadband Signal Generator was implemented which can be used to align the receiver.

- A soundcard was utilised as a radio modem to receive 1200 baud AFSK data (AX.25 protocol). The VHF receiver was used as a radio receiver.

The total cost of the VHF fixed frequency receiver is approximately R 218 excluding the cost of the PCB. This is approximately 25% less than the cost of R 1000 of a scanner radio which only receives.

6.2 Additional work

The following additional work has still to be done on the VHF fixed frequency receiver:

- Replacing the 10.7 MHz ceramic bandpass filter with a crystal filter.
- The addition of an 10 kΩ variable resistor in the RSSI level inidicator circuit so that RSSI levels up to 0.78 V can be indicated. This will help to align the receiver with a 10 dB uncertainty of maximum sensitivity. That is when a RF signal (1 kHz FM modulated, 3 kHz deviation) of -120 dBm is applied.
- A manual which assists school pupils in assembling the VHF receiver step by step with schematics of a PCB included.
- A test phase where a school pupil assembles and aligns the VHF receiver.

The detailed design and implementation of the frequency synthesizer for the VHF receiver, UHF down-converter, VHF transmitter have as well as the utilisation of a soundcard to transmit data will have to be done before a ground station will be available to fulfil the communication functions set out in chapter one.

6.3 Other

The Amateur Radio satellite OSCAR 11 (UoSAT 2) transmits 1200 baud PSK data on 145.825 MHz. The receiver has received signals from OSCAR 11 (UoSAT 2) such that the audio tones of the 1200 baud PSK data could be heard [33].

The frequency of the fixed frequency receiver can easily be changed to receive any frequency from 144 MHz to 146 MHz, by replacing the 135.125 MHz crystal with the required crystal.

6.4 Future Research

Economical solutions have to be found to align the UHF down-converter and VHF transmitter. The VHF receiver can aid the process of aligning the UHF down converter and VHF transmitter.

A way of assembling and aligning surface mount components for school pupils has to be determined. The assembling of surface mount components is not simple but due to the increasing amount of surface mount products available this would be a reason to investigate the possibility.

Design or find a module to align the receiver so that it is at maximum sensitivity.

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[36] 2m Mobile vertical (3/8 wave), http://www.members.aol.com/k5oe/

[37] Monopole anteanna (3/4 wave) http://web.usna.navy.mil/ bruninga/astars.html

[38] SUNSAT Web Site, http://sunsat.ee.sun.ac.za

[39] SUNSTEP Web Site,http://www.sunstep.sun.ac.za/

Appendix A

General Design of a 2nd Order Coupled Resonator Filter (Bandpass Filter)

The Butterworth response corresponds with a critically coupled response of a double tuned circuit in [1]. Both have only one frequency where maximum power transfer can occur. Chapter 6 from [30] and chapter 8 from [1] were used in the design of the filter.

Figure A.1 shows the schematic of a general 2nd order coupled resonator filter [30].

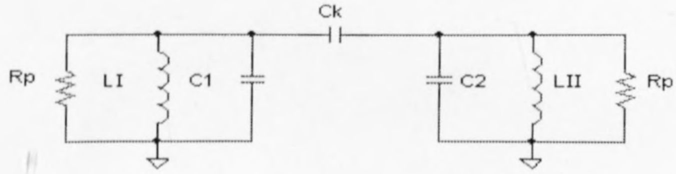

Figure A.1: General 2nd order coupled resonator filter

The following equations and relations are valid for the circuit in Figure A.1 [30]:

$$L_I = L_{II} \tag{A.1}$$

$$C_1 = C_2 \tag{A.2}$$

$$C_1 = C_I - C_k \tag{A.3}$$

$$C_2 = C_{II} - C_k \tag{A.4}$$

$$C_I = C_{II} \tag{A.5}$$

Equation A.6 [30] gives the relation between f_m (center frequency), C_I and L_I.

$$f_m = \frac{1}{2\pi\sqrt{L_I C_I}} \tag{A.6}$$

Usually two of the three variables in equation A.6 are chosen and the third is then determined. Later in this section $L_I = L_{II}$ is replaced with L_1,L_2 where $L_1 = L_2 = L_I = L_{II}$.

According to equation A.3 C_k (also known as C_{12}) has to be calculated to determine C_1. C_{12} is calculated from equation A.7 from [30].

$$C_{12} = K_{12}\sqrt{C_I C_{II}} \tag{A.7}$$

In equation A.7 K_{12} is the unnormalized coupling coefficient. K_{12} must be equal to $1/Q_L$ so that there is only one maximum power frequency (critical coupling). Q_L is the unnormalized loaded Q of the resonant section in Figure A.1.

Equation A.8 gives the relation between the unnormalized coupling coefficient (K_{12}), the normalized coupling coefficient (k_{12}), the 3 dB bandwidth (ΔF) and the center frequency (f_m). Equation A.9 gives the relation between the unnormalized loaded Q (Q_L), the normalized loaded q ($q_1 = q_2$), the 3 dB bandwidth (ΔF) and the center frequency (f_m). Equation A.10 gives the relation between the unnormalized unloaded Q ($Q_0 = Q_U$), the normalized unloaded q (q_0), the the 3 dB bandwidth (ΔF) and the center frequency (f_m) [30].

$$K_{12} = k_{12}\frac{\Delta f}{f_m} \tag{A.8}$$

$$Q_L = Q_1 = Q_2 = q_1\frac{f_m}{\Delta f} = q_2\frac{f_m}{\Delta f} \tag{A.9}$$

$$Q_0 = Q_U = q_0\frac{f_m}{\Delta f} \tag{A.10}$$

In equation A.8 $k_{12} = 0.7071$ and in equation A.9 $q_1 = 1.4142$ for a Butterworth response (critically coupled resonator filter).

When a certain bandwidth is required $Q_1 = Q_2$ and K_{12} can be calculated from A.9 and A.8. When a certain loaded Q_L is required the Δf (bandwidth) and K_{12} can be calculated from A.9 and A.8. When a certain insertion loss (I.L.) is required the normalized q_0

can be determined from the table on page 341 of [30]. The 3 dB bandwidth can then be calculated from A.10, which may then be used to determine the loaded Q (Q_L) and the unnormalized coupling coefficient (K_{12}). The values for k_{12} and q_1 must be known. C_{12}, C_1 and C_2 can be determined after K_{12} has been determined.

The bandwidth of a 2nd order coupled network (also applicable to resonator filter) according to [1] is

$$BW = \sqrt{2}b \sin \alpha \frac{\omega_0}{Q_L}, \tag{A.11}$$

where b = 1 and $\alpha = 90^o$ for critically coupled response. Equation A.11 with b = 1 and $\alpha = 90^o$ corresponds to equation A.8 as rewritten in equation A.12. $k_{12} = \sqrt{2}$ in equation A.12.

$$\Delta f = k_{12} \frac{f_m}{K_{12}} \tag{A.12}$$

R_p in Figure A.1 can be calculated from equation A.13.

$$R_p = j\Omega_0 L_1 Q_L \tag{A.13}$$

At this point in the design every component in Figure A.1 has been calculated or can be calculated. The next step is to calculate the values in Figure A.2 so that the input and output impedances match to 50 Ω.

Figure A.2 depicts the schematic of the bandpass filter (2nd order coupled resonator filter) with the input and output matched to 50 Ω.

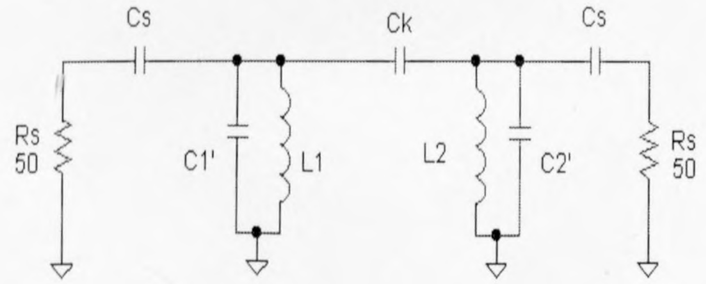

Figure A.2: Band Pass Filter for implementation in a 50 Ω system

In Figure A.2 C_s should be calculated so that C_1' can be calculated. $C_1' = C_1 - C_p$, where C_p is the parallel equivalent of C_s. R_s was chosen equal to 50 Ω. The value of Rp and Rs in Figure A.2 and Figure A.1 are known and therefore the matching Q_M can be determined from A.14.

$$Q_M = \sqrt{\frac{R_p}{R_s} - 1} \tag{A.14}$$

In A.15 and A.16 the relation between Q_M, C_s and C_p are given.

$$Q_M = \frac{R_p}{X_p} = \frac{X_s}{R_s} \tag{A.15}$$

$$Q_M = \frac{1}{j\omega Q_M R_s} \tag{A.16}$$

$$C_p = \frac{Q_M}{j\omega R_p} \tag{A.17}$$

An example now follows. Bandpass Filter 3 will be used. The specifications are center frequency (f_m) equal to 145 MHz and an insertion loss of less or equal to 1 dB.

The normalized unloaded q_0 was determined as 13.0042 from the table on page 341 of [30]. The unloaded Q (Q_U) is equal to 140. Inductors were wound with a $Q_U = 140$. The 3 dB bandwidth was calculated as 13.47 MHz with equation A.10, $Q_U = 140$ and $f_m = 145$ MHz. Equation A.9 and equation A.8 were used to determine the unnormalized loaded $Q_L = 15.22$ and the the unmoralized coupling coefficient $K_{12} = 65.687e - 3$.

The values for L_I, L_{II}, C_1 and C_{II} can now be calculated or chosen due to the fact that the bandwidth, loaded and unloaded Q are chosen or known. Values of 30 nH to 50 nH are values for which air core inductors can be wound. The value of the inductor determines the value of C_I which further influences the value of C_{12} as well as C_s. The value of C_{12} and C_s should be equal or bigger than 1 pF to be able to use standard components. $L_I = 40nH$ was chosen as a starting value. C_I was calculated as 30.12 pF with equation A.6. Capacitor $C_{12} = C_k$ was calculated as 2 pF with equation A.7. Capacitors C_1 and C_2 were calculated as 28.14 pF with equation A.3 to A.4. Resistor R_p was calculated as 555 Ω with equation A.13.

The values for C_1' and C_2' were calculated as 21.85 pF with $R_S = 50\ \Omega$ and equations A.14 to A.17.

Appendix B

Schematic of VHF Receiver

This appendix contains the block diagram and schematics of the VHF receiver which is by default a fixed frequency receiver. Figure B.1 shows the block diagram of the receiver. Figure B.2 is a schematic of the VHF receiver.

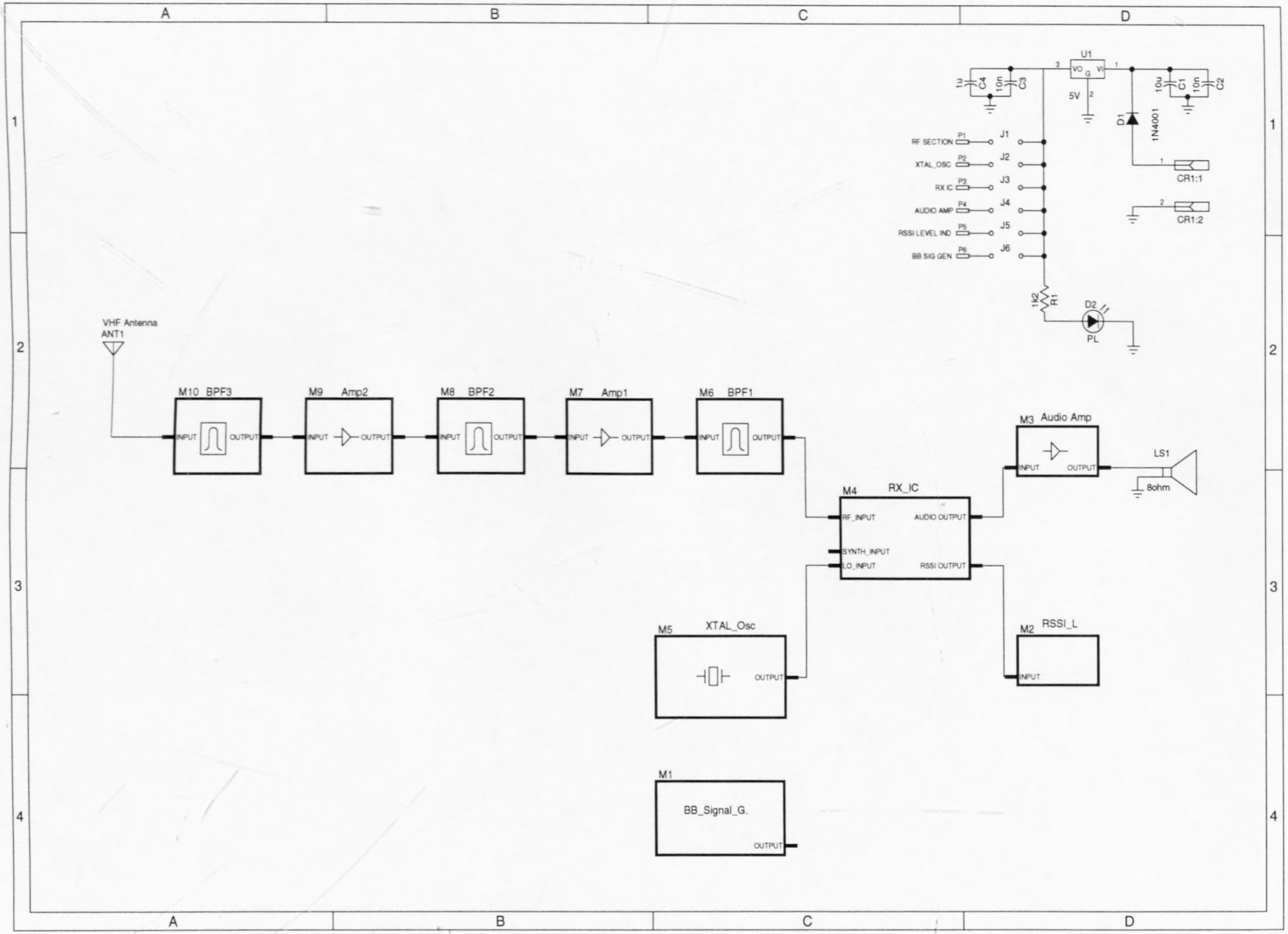

Figure B.1: Blockdiagram of VHF receiver

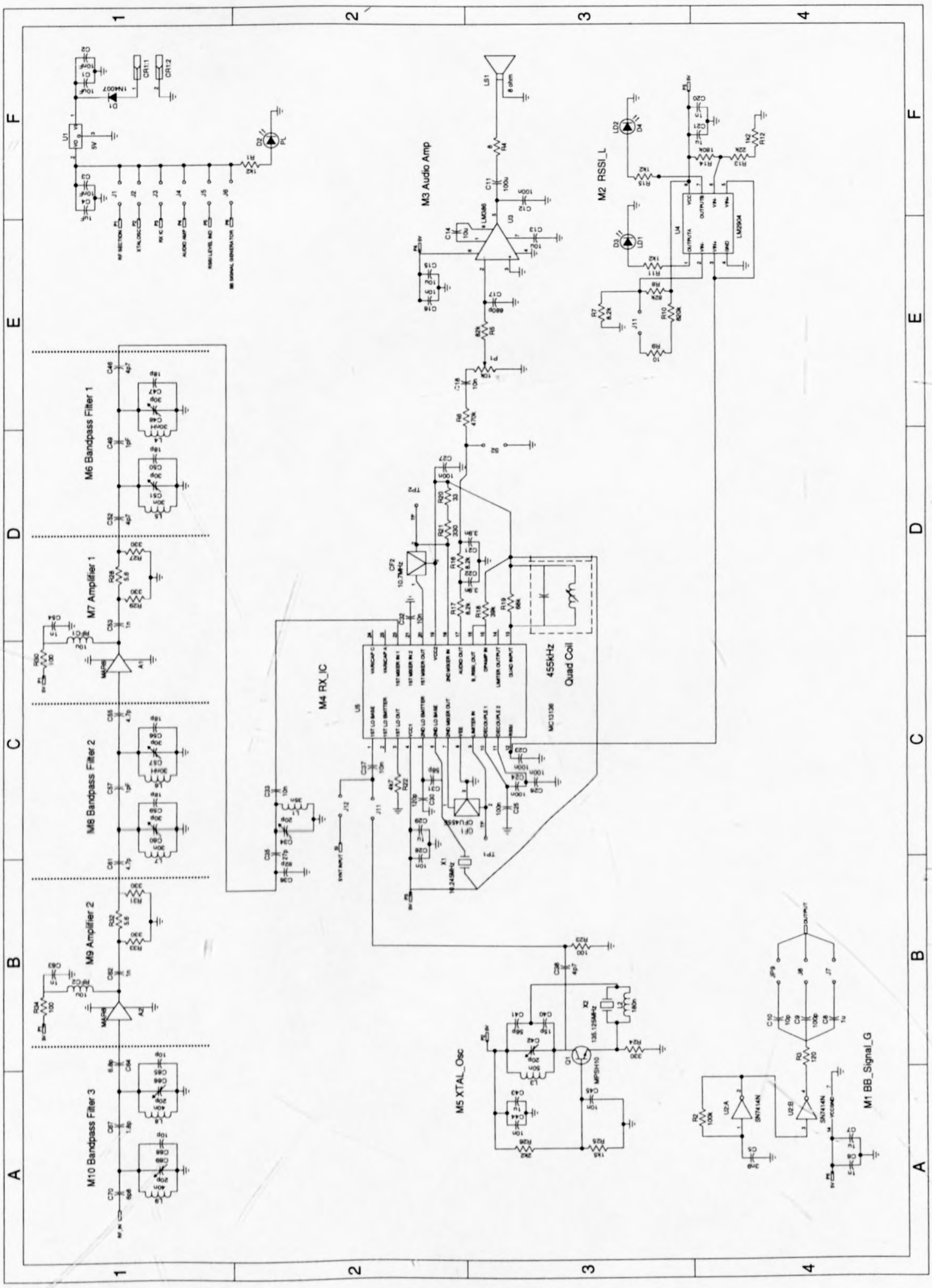

Figure B.2: Schematic of VHF receiver

Appendix C

Crystal Oscillator Appendix

Appendix C.1 gives the small signal AC model of a common base transistor and the Y-parameters of the MPSH10. The MATLAB code to determine the open loop gain is given in section C.2 . The calculated values, the computed responses and the measurements taken of the oscillators are listed separately in appendix C.3 to C.5 .

C.1 Small Signal Model and Y-Parameters of Common Base Transistor

In obtaining the small signal AC model the AC model of a common base transistor had to be determined. Figure C.1 shows a general Y parameter model for a common base transistor.

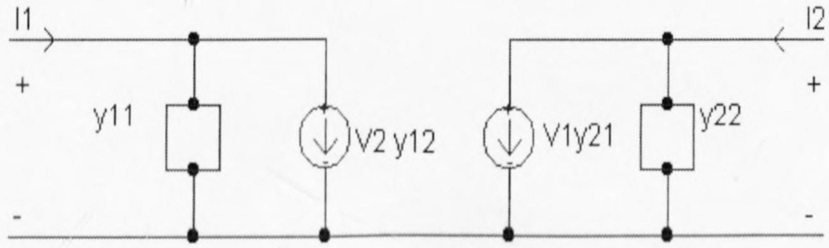

Figure C.1: General small signal model of y parameter model of a common base transitor

Equations C.1 and C.2 apply to the model in Figure C.1 [1].

$$I_1 = y_{11}V_1 + y_{12}V_2 \tag{C.1}$$

$$I_2 = y_{21}V_1 + y_{22}V_2 \tag{C.2}$$

Table C.1 lists the y parameters for the MPSH10 bipolar transistor which were read from the graphs in [17].

Table C.1: Y parameters of MPSH10 transistor

Frequency [Hz]	y11	y12	y21	y22
100e6	80e-3 - i*50e-3	0 - i*0.4e-3	-75e-3 + i*50e-3	0 + i*1e-3
135e6	77e-3 - i*50e-3	0 - i*0.45e-3	-63e-3 + i*52e-3	0 + i*1.25e-3
500e6	35e-3 - i*45e-3	0 - i*1.25e-3	-12e-3 + i*55e-3	0.5e-3 + i*5.5e-3
700e6	20e-3 - i*40e-3	0 - i*2.25e-3	-8e-3 + i*40e-3	1e-3 + i*7.2e-3
900e6	10e-3 - i*33e-3	0 - i*2.75e-3	10e-3 + i*33e-3	1.5e-3 + i*9e-3
1000e6	2.2e-3- i*30e-3	0 - i*3.2e-3	30e-3 + i*30e-3	2e-3 + i*9.8e-3

C.2 MATLAB Code used in Simulations

```
% This is an experimental program with G.W. Milne's RF Toolbox
% of MATLAB by F. Nel
close all;
clear;
f  = linspace(110e6,180e6,20000);  % Generates frequency
% Y parameters of MPSH10 transistor
Y11= [100e6,80e-3 - i*50e-3, 0 - i*0.4e-3 , -75e-3 + i*50e-3, 0 + i*1e-3
   135e6, 77e-3 - i*50e-3, 0  - i*0.45e-3 , -63e-3 + i*52e-3, 0  + i*1.25e-3
   500e6, 35e-3 - i*45e-3, 0  - i*1.25e-3 , -12e-3 + i*55e-3, 0.5e-3 + i*5.5e-3
   700e6, 20e-3 - i*40e-3, 0  - i*2.25e-3 , -8e-3  + i*40e-3, 1e-3 + i*7.2e-3
   900e6, 10e-3 - i*33e-3, 0  - i*2.75e-3 ,10e-3 + i*33e-3, 1.5e-3 + i*9e-3
   1000e6, 2.2e-3- i*30e-3, 0  - i*3.2e-3  ,30e-3 + i*30e-3, 2e-3 + i*9.8e-3];

ladder(f,0);                  % terminates short circuit
yparam(Y11);                  % transistor with common base Y parameters as input
rp(330);                      % RE in the oscillator circuit
c0 = 5.5e-12;                 %  stray capacitance c0
c1 = 5.5e-12/(250*25);        % fundamental mode cap C1
c3 = c1/9;                    % 3rd overtone mode cap C3
c5 = c1/25;                   % 5th overtone mode cap C5
c7 = c1/49;                   % 7th overtone mode cap C7
c9 = c1/81;                   % 9th overtone mode cap C9
L1 = 1/((2*pi*135.125000e6)\^2*c5);  %L1 inductor in 5th overtone mode branch
rs(23);                       % replacing crystal with 23 ohm resistor
cp(56e-12);                   % C3 in the oscillator circuit
cs(15e-12);                   % C2 in the oscillator circuit
rcsp(100,4.7e-12);            % C5 in series with RL
lp(50e-9);                    % L1
rp(5000);                     % Rp of L1 = Q
cp(10.5e-12);                 % C1
cp(1.5e-12);                  %  C22/y22 must be included here.
plot(f,ai\_db,'r');           %  Linear plot of voltage gain in dB
grid;
title('Phase and gain plot of oscillator');
xlabel('Frequency[MHz]','fontsize',18);
ylabel('Magnitude[dB] \& Phase[degrees]','fontsize',18);
hold on;
plot(f,ai\_deg,'b');          % linear plot (angle of voltage gain)
hold off;
```

C.3 Calculated Values

Table C.2 to Table C.4 list the values for $C2$, $C3$, $C5$ and RL for the ratios of $n = 11$, $7.5 < n < 9.5$ and $6 < n < 7$ such that the upper and lower frequency boundaries are $\leq$ 120 MHz and $\geq$ 150 MHz and the other boundaries set in section 3.9.1 are met. Table C.2 lists the values for $C2$, $C3$, $C5$ and RL for the ratio $5 < n < 4$ so that the lower and upper frequency boundaries as well as some of the boundaries set in section 3.9.1 are met.

Table C.2: The values for $C2$, $C3$, $C5$ and RL for $n = 11$

RL [Ω]	$C2$ [pF]	$C3$ [pF]	$C5$ [pF]	f_{lower} [MHz]	f_{upper} [MHz]
50	12	100	4.7	117.5	155.5
	12	120	3.9	118.5	157.5
	12	120	4.7	117	155
100	12	120	2.7	120.5	162.5
	12	120	3.3	119.5	160.5
	15	150	2.7	116	152
	15	150	3.3	115.25	150.5
150	15	150	2.7	116.25	152.5
500	15	150	1.5	118.25	157.5
	15	150	1.8	118.25	157.25
1000	15	150	1.2	119	159.5
	15	150	3.3	119.5	160.5

The values for $C5$ from 1.5 pF to 2.7 pF will all be valid for $RL = 1000$ Ω.

Table C.3: Values for $C2$, $C3$, $C5$ and RL for $7.5 \leq n \leq 9.5$

RL	$C2$ [pF]	$C3$ [pF]	$C5$ [pF]	f_{lower} [MHz]	f_{upper} [MHz]
50	12	82	3.9	119	159.25
	12	82	4.7	118	156.5
	12	100	3.9	118.75	158.5
	12	100	4.7	117.5	155.5
100	15	100	2.7	117	154.5
150	15	120	2.7	116.5	153.75
500	15	120	1.5	118.75	158.75
	15	120	1.8	118.75	158.5
1000	15	120	3.3	120	161.5
	18	150	3.3	115.5	151.25

The $C2 : C3$ ratios 18:150 and 15:120 with 1 pF $\leq C5 \leq$ 3.3 pF will all give a frequency range of 120 MHz to 150 MHz.

Table C.4: Values for $C2$,$C3$,$C5$ and RL for $6 < n < 7$

RL [Ω]	$C2$ [pF]	$C3$ [pF]	$C5$ [pF]	f_{lower} [MHz]	f_{upper} [MHz]
50	12	68	3.9	119.5	160.5
	12	68	4.7	118.25	157.5
	15	82	3.9	115.5	151.25
100	15	82	2.7	117.5	155.5
	15	82	3.3	117	153.75
150	15	82	2.7	117.75	156
500	15	82	1.5	120	161.5
	15	82	1.8	119.75	161
	18	100	1.5	115.75	151.75
	18	100	1.8	115.75	151.5
1000	18	100	1.2	116.5	153.5
	18	100	3.3	117	154

Values for $C5$ from 1 pF to 3.3 pF will be valid for $RL = 1000\ \Omega$, $C2 = 18$ pF and $C3 = 100$ pF .

Table C.5: Values for $C2$,$C3$,$C5$ and RL for $4 < n < 5$.

RL [Ω]	$C2$ [pF]	$C3$ [pF]	$C5$ [pF]	f_{lower} [MHz]	f_{upper} [MHz]
50	15	47	4.7	116.5	153
	15	56	4.7	115.5	151.5
100	15	56	3.3	118	156.75
	15	56	4.7	116.5	153.25
	18	56	3.3	115.25	150.5
150	15	56	2.7	119.5	159.5
	18	56	2.7	116.25	152.75
500	18	56	1.5	118.25	157.5
	18	56	1.8	118.25	157.5
1000	18	56	3.3	119.5	160.5

C.4 Computed Results

The computed responses for $n = 11$, $7.5 < n < 9.5$, $6 < n < 7$ and $4 < n < 5$ are listed in this section. Table C.6 to Table C.9 list the lower and upper frequencies for $Zfb = 23\ \Omega$, open loop gain $|A_L|$ for Zfb = crystal and the maximum value for $|A_L|$ when $Zfb = 7$ pF. The lower and upper frequencies are indicated with f_{lower} and f_{upper} separately in Table C.6 to C.9. Only some of the options for $C2$, $C3$, $C5$ and RL that were given in Appendix C.3 were simulated to determine what the results are for different values of Zfb as mentioned previously.

Table C.6: Computed responses for $n = 11$

R_L	$C5$ [pF]	$C2$ [pF] : $C3$ [pF]	f_{lower} [MHz] $Zfb = 23\ \Omega$	f_{upper} [MHz] $Zfb = 23\ \Omega$	$\|A_L\|$ [dB] Zfb = crystal	$\|A_L\|$ [dB] $Zfb = 7$ pF
50	3.9	12:120	118.3	157.25	8	- 3.3
	4.7	12:120	117.12	154.5	6	- 5.5
100	2.7	12:120	120.3	162.24	8.25	- 3.0
	2.7	15:150	115.8	151.7	8.3	- 2.8
150	2.7	15:150	116	152.2	6.63	- 5.0
500	1.8	15:150	118	157	6.4	- 5.8
1000	3.3	15:150	119	160	5.25	- 6.1

Table C.7: Computed results for $7 < n < 9$

R_L	$C5$ [pF]	$C2$ [pF] : $C3$ [pF]	f_{lower} [MHz] $Zfb = 23\ \Omega$	f_{upper} [MHz] $Zfb = 23\ \Omega$	$\|A_L\|$ [dB] Zfb = crystal	$\|A_L\|$ [dB] $Zfb = 7$pF
50	3.9	12:82	119.12	159.16	8.85	- 0.6
	3.9	12:100	118.65	158	8.5	- 2.0
100	2.7	15:120	116.3	152.8	8.9	- 1.2
	2.7	15:100	116.8	154	9.1	0.0
150	2.7	15:120	116.55	153.4	7.34	- 3.3
500	1.5	15:120	118.5	158.6	8.2	- 2.6
	1.8	15:120	118.52	158.4	7.15	- 4.1
1000	3.3	15:120	120	161.28	6.2	- 4.4
	3.3	18:150	115.3	150.8	6.14	- 5.1

Table C.8: Computed results for $6 < n < 7$

R_L [Ω]	$C5$ [pF]	$C2$ [pF] : $C3$ [pF]	f_{lower} [MHz] $Zfb = 23\ \Omega$	f_{upper} [MHz] $Zfb = 23\ \Omega$	$\lvert A_L \rvert$ [dB] Zfb = crystal	$\lvert A_L \rvert$ [dB] $Zfb = 7$ pF
50	4.7	12:68	18.5	157.5	7.5	- 1.6
	3.9	15:82	115.5	150.5	9.1	1.2
100	2.7	15:82	117.5	155.25	9.1	1.25
	3.3	15:82	116.7	153.5	8	- 0.8
150	2.7	15:82	117.7	156	8	- 0.7
500	1.8	15:82	120	161.2	7.91	- 1.5
	1.8	18:100	115.5	151.2	7.9	- 1.4
1000	1.8	18:100	116.4	153.3	8.1	- 0.8
	3.3	18:100	116.75	153.5	7.1	- 2.6

Table C.9: Computed results for $4 < n < 5$

R_L	$C5$ [pF]	$C2$ [pF] : $C3$ [pF]	f_{lower} [MHz] $Zfb = 23\ \Omega$	f_{upper} [MHz] $Zfb = 23\ \Omega$	$\lvert A_L \rvert$ [dB] Zfb = crystal	$\lvert A_L \rvert$ [dB] $Zfb = 7$ pF
50	4.7	15:47	116.5	153	6.6	2.4
	4.7	15:56	116	151.5	7.6	1.4
100	3.3	15:56	118	156	7.7	1.5
	4.7	15:56	117	154	5.4	- 2.9
	3.3	18:56	115.28	150	7.5	3.0
150	2.7	15:56	119.2	159.5	7.85	1.6
	2.7	18:56	116.25	152.4	7.65	3.0
500	1.5	18:56	118.25	157.25	8	3.8
	1.8	18:56	118.2	157.25	7.45	2.2
1000	3.3	18:56	119.5	160	7	1.75

C.5 Measured Results

Table C.10 to Table C.11 list the frequency range measurements taken for $n = 11$, $6 < n < 7$ and $4 < n < 5$.

Table C.10: Frequency range measurements for $n = 11$, $6 < n < 7$

n ratio	R_L [Ω]	$C2$ [pF]	$C3$ [pF]	$C5$ [pF]	f_{upper} [MHz]	f_{lower} [MHz]
11	47 (50)	12	120	4.7	109	157.5
	100	12	120	2.7	112	163.5
	1000	12	120	3.3	116.5	170
$6 < n < 7$	47 (50)	15	82	3.9	102	140
	100	15	82	3.3	103	144

Table C.11: Frequency range measurements for $4 < n < 5$

R_L [Ω]	$C2$ [pF]	$C3$ [pF]	$C5$ [pF]	f_{upper} [MHz]	f_{lower} [MHz]
47 (50)	15	56	4.7	108	151.5
100	15	56	4.7	109	154

Appendix D

Calculations to Evaluate Harmonics of Square Waves

The derivations for the exponential Fourier components for a square wave with a certain duty cycle are in section D.1 . The formulas which were used in the MATLAB file Harm4.m are in section D.2. The MATLAB program Harm4.m which is listed in section D.4 was used to determine the total power in a resistive load for a certain frequency band, to plot the PSD of the harmonics in a resistive load when there is only a resistive load or RC filter in series with the resistive load on the output of the signal generator. Section D.3 lists calculated values for a spectral line for Rout, Cout and a resistive load at 455 kHz, 10.7 MHz and 145.825 MHz .

D.1 Derivations

Equations D.9, D.10 and D.11 give the exponential Fourier components ($C_N = (a_n - jb_n)/2$) for square waves with a duty cycle of 70%, 75% and 80%. The derivations to determine the exponential Fourier components for a square wave with a duty cycle of 75% are outlined in equations D.1 to D.8. The derivative method was used in the derivations of Fourier components $C_N = (a_n - jb_n)/2$. Equations D.1 and D.5 were obtained from [3]. Figure D.1 shows the steps of deriving the second derivative of the square wave with a 75% duty cycle.

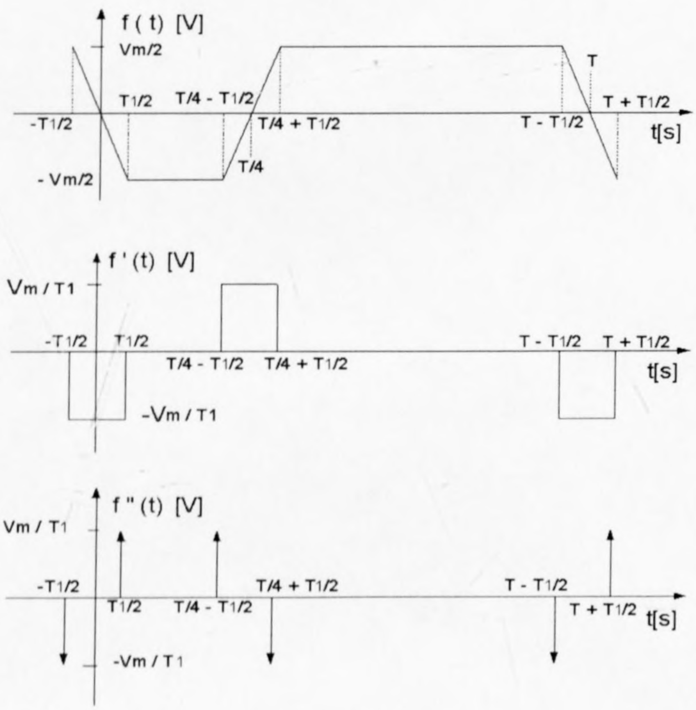

Figure D.1: Graphs of f(t), f'(t) and f"(t)

The derivation of a_n:

$$-(n\omega_0)^2 a_n = a_n^{''} = \frac{2}{T}\int_0^T f^{''}(t)\cos n\omega_0 t \tag{D.1}$$

$$-(n\omega_0)^2 a_n = \frac{2}{T}\frac{V_m}{T1}\left[\cos\frac{n2\pi}{T}\frac{T1}{2} + \cos\frac{n2\pi}{T}(\frac{T}{4} - \frac{T1}{2}) - \cos n\frac{2\pi}{T}(\frac{T}{4} + \frac{T1}{2}) - \cos n\frac{2\pi}{T}(T - T1/2)\right] \tag{D.2}$$

$$-(n\omega_0)^2 a_n = \frac{2}{T}\frac{V_m}{T1}\left[2\sin\frac{n\pi}{2}\sin n\pi\frac{T1}{T}\right] \tag{D.3}$$

$$a_n = -\frac{TV_m}{(n\pi)^2 T1}\sin\frac{n\pi}{2}\sin n\pi\frac{T1}{T} \tag{D.4}$$

The derivation of b_n:

$$-(n\omega_0)^2 b_n = b_n^{''} = \frac{2}{T}\int_0^T f^{''}(t)\sin n\omega_0 t \tag{D.5}$$

$$-(n\omega_0)^2 b_n = \frac{2}{T}\frac{V_m}{T1}\left[\sin\frac{n2\pi}{T}\frac{T1}{2} + \sin\frac{n2\pi}{T}(\frac{T}{4} - \frac{T1}{2}) - \sin n\frac{2\pi}{T}(\frac{T}{4} + \frac{T1}{2}) - \sin n\frac{2\pi}{T}(T - T1/2)\right] \tag{D.6}$$

$$-(n\omega_0)^2 b_n = \frac{2}{T}\frac{V_m}{T1}\left(1 - 2\cos\frac{n\pi}{2}\right)\sin n\pi\frac{T1}{T} \tag{D.7}$$

$$b_n = \frac{TV_m}{(n\pi)^2 T1}\left(\cos\frac{n\pi}{2} - 1\right)\sin n\pi\frac{T1}{T} \tag{D.8}$$

The Fourier coefficient for a duty cycle of 70%, 75% and 80% are as follows:

- 70% duty cycle:

$$C_n = \frac{Vm}{2(\pi n)^2}\left[\left(-\sin\frac{3\pi n}{5}\right) + j\left(1 - \cos\frac{3\pi n}{5}\right)\right]\sin n\pi\frac{T1}{T} \tag{D.9}$$

- 75% duty cycle:

$$C_n = \frac{Vm}{2(\pi n)^2}\left[\left(-\sin\frac{\pi n}{2}\right) + j\left(1 - \cos\frac{\pi n}{2}\right)\right]\sin n\pi\frac{T1}{T} \tag{D.10}$$

- 80% duty cycle:

$$C_n = \frac{Vm}{2(\pi n)^2}\left[\left(-\sin\frac{2\pi n}{5}\right) + j\left(1 - \cos\frac{2\pi n}{5}\right)\right]\sin n\pi\frac{T1}{T} \tag{D.11}$$

D.2 Formulas to Calculate the Power of a Harmonic in a Resistive Load

The power in a single harmonic in a resistive load is calculated with the formula in equation D.12.

$$P = \frac{2|G_n|^2}{R_{resistiveload}} \tag{D.12}$$

were $|G_n|^2 = |C_n|^2$ for only a resistive output at the broadband signal generator and $|G_n|^2 = |C_n|^2|H(n\omega_0)|^2$ when there is a RC highpass filter in series with the resistive output at the broadband signal generator. The MATLAB file Harm4.m determines the power of each harmonic in a resistive load.

D.3 Calculated Results

Tables D.1 to D.3 list the power of a single harmonic in a resistive load for an 1 kHz and 4 kHz square wave.

Table D.1: Calculated values of a single harmonic at 455 kHz ± 6 kHz

f_0	$Cout$ [pF]	$R_{resistive*load}$	Power [dBm]
1 kHz	10	1.5 kΩ	- 78
	100		- 60
4 kHz	10	1.5 kΩ	- 66.2
	100		- 47.5
1 kHz	10	50 Ω	- 93
	100		- 73
4 kHz	10	50 Ω	- 81
	100		- 61

Table D.2: Calculated values of a single harmonic at 10.7 MHz ± 6 kHz

f_0	$Cout$ [pF]	$R_{resistive*load}$	Power [dBm]
1 kHz	10	4 kΩ	- 83
	100		- 82
4 kHz	10	4 kΩ	- 71
	100		- 70.5
1 kHz	10	50 Ω	- 93
	100		- 76.7
4 kHz	10	50 Ω	- 81
	100		- 65

Table D.3: Calculated values of a single harmonic at 145.825 MHz ± 6 kHz

f_0	rise/fall time [ns]	$Cout$ [pF]	$R_{resistive*load}$	Power [dBm]
1 kHz	2.5	10	50 Ω	- 100
	5	10	50 Ω	- 108
	10	10	50 Ω	- 111.5
	12.5	10	50 Ω	- 116
1 kHz	2.5	100	50 Ω	- 99
	5		50 Ω	- 107
	10		50 Ω	- 110
	12.5		50 Ω	- 114.5
4 kHz	2.5	10	50 Ω	- 88.5
	5		50 Ω	- 96
	10		50 Ω	- 99.5
	12.5		50 Ω	- 107
	2.5	100	50 Ω	- 87
	5		50 Ω	- 94.5
	10		50 Ω	- 98
	12.5		50 Ω	- 105.5

D.4 Matlab File Harm4.m

```
%This file calculates the distribution as well as the
% power of a the harmonics in a resistive load for for a square wave.

close all;
clear;

Vm   = input('Enter the peak to peak voltage of the square wave: ');
T1   = input('Enter the rise/fall time: ');
fo   = input('Enter the fundamental frequency of the square wave: ');
C1   = input('Enter the value of Cout/C1: ')
R2   = input('Enter the resistive load: ');

klo  = input('Enter the first harmonic: ');
kup  = input('Enter the last harmonic: ');
incr = input('Enter the increments between the first and last harmonic: ');

f    = klo:incr:kup;      %increments to generate vector of "frequencies"
T    = 1/fo;
R1 = 120;                 %Rout

%If incr = 1 set to plot + and* otherwise plot envelope

%75 percent duty cycle without any rise/fall time
CN1 = ((Vm/(2*pi))./f).*((-1).*sin((pi/2).*f)
+ i.*(1 - 1.*cos((pi/2).*f)));
%75 percent duty cycle with rise/fall time included
CN2 = ((Vm/(2*pi))./f).*((-1).*sin((pi/2).*f)
+ i.*(1 - 1.*cos((pi/2).*f))).*sinc((T1/T).*f);

 %Transfer function for RC highpass filter
H1 = (R2./(R2 + R1 - j./(2*pi*C1*fo.*f)));
G1 = (abs(H1).^2).*(abs(CN1).^2).*(2/R2);
G2 = (abs(H1).^2).*(abs(CN2).^2).*(2/R2);
```

```
%PSD plot of the harmonics at the output of the signal generator
if incr == 1
 figure(1);
 plot(fo.*f,10*log10(abs(H1).^2),'r*');
 hold on;
 plot(fo.*f,10*log10(1000*G2),'b*');
 plot(fo.*f,10*log10((abs(CN2).^2).*(2000/R2)),'r+');
 grid;
 title('PSD plot (output of signal generator) & plot of RC highpass filter');
 label('PSD in 50 ohm [dBm]','fontsize',18);
 xlabel('Frequency[Hz]','fontsize',18);
 hold off;
else
 figure(1);
 plot(fo.*f,10*log10(abs(H1).^2),'r');
 hold on;
 plot(fo.*f,10*log10(1000*G2),'b');
 plot(fo.*f,10*log10((abs(CN2).^2).*(2000/R2)),'r');
 grid;
 title('PSD plot (output of signal generator) &  plot of RC highpass filter');
 ylabel('PSD in 50 ohm [dBm]','fontsize',18);
 xlabel('Frequency[Hz]','fontsize',18);
 hold off;
end

P1  = 0;
P2  = 0;
PG1 = 0;
PG2 = 0;

% Determine the total power in for the number
% of harmonics that have been plotted.
if incr == 1
   for n = 1:1:(kup - klo + 1),
   P1  = ((abs(CN1(:,n))).^2).*(2/R2) + P1;
   P2  = ((abs(CN2(:,n))).^2).*(2/R2) + P2;
   PG1 = G1(:,n) + PG1;
   PG2 = G2(:,n) + PG2;
   end
 P1_dBm  = 10*log10(1000*P1);
 P2_dBm  = 10*log10(1000*P2);
 PG1_dBm = 10*log10(1000*PG1);
 PG2_dBm = 10*log10(1000*PG2);
end
```

Appendix E

Alignment of the VHF Fixed Frequency Receiver

This appendix gives step by step instructions to align the VHF receiver. The observation that should be made if the section in the receiver operates or is correctly aligned will be mentioned. This manual does not include a PCB lay out to indicate where the components are. The receiver IC section is the only section in the receiver where the individual components that have to be soldered are mentioned. The reason is that the receiver IC section is tested section by section.

Refer to the schematics in appendix B when component numbers are given.

Assemble the Power Supply section first. The LED labeled PL will shine if there is an output voltage at the output of the 5 V regulator. The LED will not shine if the power supply is incorrectly connected at connector $CR1$.

The Broadband Signal Generator section is soldered/assembled next.

The Audio Speaker is the first component to be tested. Select the 1 uF capacitor ($C8$) on the output of the signal generator by selecting jumper $J7$. Then test if the audio speaker operates. An audio tone should be heard if the audio speaker is operating.

The Audio amplifier is assembled and tested next. Select the 100 pF capacitor ($C9$) at the output of the Broadband Signal Generator output with jumper $J8$. The volume control can also be tested. An audio tone should be heard if the audio amplifier is correctly assembled.

The RSSI Level Indicator is assembled next.

The following components should be assembled next:

- Receiver IC (MC13136 - U5)
- Sockets $S2$ and $S3$
- Audio lowpass filter ($C21$, $C22$, $R16$ and $R17$)
- Capacitors $C23$ to $C26$
- V_{CC2}, V_{CC1}, $C27$, $C28$, $C29$
- $R19$ (68 kΩ), 455 kHz Sub - Miniature IF Transformer
- Testpoint $TP1$
- Resistors $R18$ (36 Ω), $R20$ (33 Ω) and $R21$ (330 Ω)

Select the 100 pF capacitor ($C9$) at the output of the Broadband Signal Generator and test the limiter and detector by applying a signal at $TP1$. A distorted audio tone should be heard at the audio speaker if everything is operating. Nothing should be heard if no signal is applied at the $TP1$. The LED labeled $LD1$ of the RSSI level indicator should shine ($J11$ not selected) if the limiter section is operating.

The following components should be assembled:

- $CF1$ (455 kHz IF filter)
- Capacitors $C31$and $C31$
- Resistors $R20$ and $R21$
- Testpoint $TP2$

Select the 100 pF capacitor (C2) at the output of the Broadband signal generator and apply a signal at $TP2$ to test if the second mixer, limiter and detector sections operate. An audio tone should be heard at the speaker. The LED labeled $LD1$ of the RSSI level indicator should shine ($J11$ selected) if the second mixer section is operating.

Solder the following components:

- Oscillator section
- Receiver IC
  - CF2 10.7 MHz IF filter
  - Capacitors $C32$ to $C36$
  - Resitor $R22$ and inductor $L1$

Select the 10 pF capacitor ($C10$) to align the bandpass matching network, the crystal oscillator and 455 kHz Sub - Miniature IF Transformer. The jumper $J12$ must be selected to align the receiver IC and crystal oscillator. The steps to align the bandpass matching network, the crystal oscillator and the 455 kHz Sub - Miniature IF Transformer are as follows:

- Turn the LC discriminator so that the noise output at the speaker is loudest. Then turn it 45^o left or right.
- Turn the trimmer capacitor of the crystal oscillator a 1/4 of the trimming range.
- Turn the trimmer capacitor of the bandpass matching network through 360^o.
- When at any stage a distorted audio tone is heard at the speaker and it can be affirm that it is due to the Broadband Signal Generator, it means the oscillator is operating.
- If no audio tone is heard and the LED's labeled $LD1$ and $LD2$ do not shine repeat the previous routine till an audio tone is heard at the speaker.
- The matching network will be aligned best if the LED's of the RSSI level indicator labeled $LD1$ and $LD2$ shine.

Bandpass Filter 1 is assembled and aligned next. Select the 10 pF ($C10$) capacitor on the Broadband Signal Generator and apply a signal at the input of Bandpass Filter 1. Tune the trimmer capacitors $C51$ and $C48$ until an audio tone is heard over the speaker and the LED's labeled $LD1$ and $LD2$ shine. Remember to turn the sub - miniature IF transformer so that it is $\pm 45^o$ from the place where the noise output at the speaker is loudest. This is to ensure that an audio tone is heard.

Bandpass Filter 2 and Amplifier 1 are assembled and aligned next. Select the the 10 pF capacitor at the output of the Broadband Signal Generator and apply a signal at the input of Bandpass Filter 2. Tune the trimmer capacitors $C57$ and $C60$ until a distorted audio tone is heard and $LD1$ and $LD2$ of the RSSI level indicator shine.

Bandpass Filter 3 and Amplifier 2 are assembled and aligned next. Select the the 10 pF capacitor at the output of the Broadband Signal Generator and apply a signal at the input of Bandpass Filter 3. Tune the trimmer capacitors $C57$ and $C60$ until a distorted audio tone is heard and $LD1$ and $LD2$ of the RSSI level indicator shine.