9 VHF/UHF Receivers, Transmitters and Transceivers

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The purpose of this chapter is to give the reader an insight into what is available to the radio amateur on the VHF and UHF bands in the UK.

There is some theory about the choice of the equipment to use but the emphasis is on the practical aspects of choosing equipment and enhancing it with preamplifiers, and power amplifiers to give you a station that can be used to its maximum effect.

GETTING THE BEST OUT OF YOUR VHF/UHF STATION

The following paragraphs comprise an abridged version of an overview of the VHF/UHF bands by David Butler, G4ASR, published in the February 1999 edition of RadCom.

One of the great attractions of operating on the VHF/UHF bands is that there are so many different aspects of the hobby that can be utilised at these frequencies. Interested in voice communications? You can use the VHF bands for both local and international contacts. Perhaps your interest lies in digital communications. Well you can join the growing band of enthusiasts that use packet radio (AX25) to access mailboxes or the DX Cluster.

A further aspect of this technology is the automatic packet reporting system (APRS) that allows real-time tracking of mobile (or fixed) stations. Image communication such as slow scan television (SSTV) is also popular, especially now that most of the processing is achieved using a computer and sound card. And don't forget Morse! This 'digital' mode is still very much used on the VHF bands by the DX community. Once you get hooked on working DX you'll then discover exotic propagation modes such as trans-equatorial propagation (TEP), Sporadic-E (Sp-E or Es), Aurora and meteor scatter (MS). And it doesn't have to be two-way terrestrial contacts.

You can also make use of amateur satellites or even bounce your VHF/UHF signals off the moon (earth moon earth EME) to make world-wide contacts. You can operate from home, in the car or go out back-packing from the hill tops. Other activities include low power or high power, rag chews or contesting. The VHF/UHF bands really do have something for everyone.

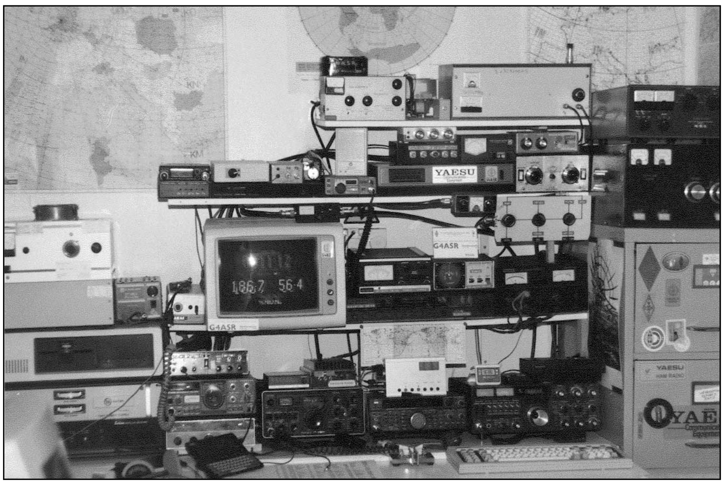


Fig 9.1: The VHF station of David Butler, G4ASR

Prime Mover

The one piece of equipment that determines exactly what facilities you can ultimately use on the VHF bands is the station transceiver. This will either be a single-mode or multi-mode base station, mobile unit or portable hand-held radio. Most single-mode transceivers available today are mobile units (often pressed into service for home use) and portable hand-helds. These are designed to operate exclusively on FM and are very popular, as they can be used for short-range telephony (either direct or via a repeater) and for data communications such as packet radio. FM transceivers can be obtained from amateur radio retailers, but that's not the only source of this type of equipment.

Commercial operators regularly upgrade their private mobile radio (PMR) equipment, and this can be obtained from traders who specialise in electronic surplus. It will get you operational very quickly and at a price that will suit most pockets. Indeed, for many fixed station applications, I would recommend that you use dedicated PMR equipment as it does possess many advantages. It is designed for use by a wide range of operators in varying environments. Because of this the equipment is normally of rugged construction. Drop it and it will probably keep working. The majority of PMR equipment has to be built to a high technical performance and reliability. Spectral purity of the transmitted signal is very good. Some amateur band allocations are very close to the commercial PMR bands. By looking around you should find equipment suitable for the 50MHz, 70MHz, 144MHz and 430MHz bands. Most equipment is relatively easy to modify and in some instances may not need any modification at all. However, before you hand over your money there are a few points to note. Is the equipment working on a frequency range close to an amateur band? What transmission mode does it use? Is it AM or FM? What is the channel spacing of the equipment? Is it 50kHz, 25kHz or 12.5kHz? The latter two are preferable, whereas the 50kHz channel spacing would indicate that the equipment is many years old and uses wide bandwidth filters which may not be suitable for use on the VHF bands today.

The VHF/UHF Bands

The three UK VHF_amateur bands are 50MHz, 70MHz and 144MHz, and the only UHF band is 430MHz. Profiles of these bands, their different propagation characteristics and the bandplans can be found in the *Amateur Radio Operating Manual*, by Don Field, G3XTT, and available from the RSGB.

Long Distance

As I've just mentioned, the use of FM equipment is for short-range communication links. If you want to broaden your horizons and contact stations much further away then you'll need to procure a multi-mode rig which in addition to FM includes CW and SSB transmission modes. Unfortunately you won't be able to find surplus PMR equipment that can be pressed into service as a multi-mode rig, so it really is a case of digging deep into your pockets and buying a suitable transceiver.

If you already possess a multi-mode HF transceiver. you may wish to consider the use of a transverter. A transverter is a transmitting converter, a receiving converter and a local oscillator source all combined into one unit. It connects to the antenna socket of an existing transceiver that provides the driving signal, typically at 28MHz. The transverter then mixes the IF drive from the transceiver with its own local oscillator to produce an output on the VHF/UHF band of your choice. On receive a similar process takes place, the VHF signals being down-converted to provide an output signal in the 28MHz band.

In practice, transverters are available for all VHF/UHF bands and for a variety of IF drive frequencies. Although the majority will be at 28MHz you'll also find models that will accept drive at 144MHz. So if you already have a transceiver on this VHF/UHF band you should have no problem finding a transverter that will allow you to operate on the 50MHz band. The advantage of using a transverter is that it allows all the functions and performance of the driving transceiver to be used on the VHF/UHF band of your choice; more on this topic later.

Optimisation

Now it's time to take a look at what a VHF/UHF station comprises and how you can make simple improvements. Fig 9.1 shows G4ASR's station. No matter what VHF/UHF band or transmission mode you wish to use, the basic system will always be the same. It's a transceiver feeding an antenna via a length of coaxial cable.

So why do some stations consistently perform better than others? One of the most important factors is the site on which the VHF/UHF station is located. Ideally, a hill-top location is best, but good results can be obtained in low lying areas that are clear of local obstructions. Results depend very much on the band used, obstructions having considerably less effect at 50MHz than at 144MHz or 430MHz.

We can't all live at 250m above sea level with a clear take-off, so you need to pay special attention to the most significant item in your station. That of course is the antenna; Fig 9.2 shows a typical VHF/UHF antenna array.

Convention dictates that FM operation, for both telephony and digital communications, an antenna with vertical polarisation is required. If you want to make local contacts then you'll probably need omni-directional coverage. For packet radio you will require a similar vertically polarised antenna, although you might consider using a small 4 or 5-element beam for a fixed link.

For serious VHF DX work, using CW or SSB, a horizontally polarised directional Yagi is recommended. There are many types of beam antennas available, some very good and some, well, not so good. But the difference between the poorest designs to that of the very best may only amount to 4dB or so.

The point here is that if you are only interested in working occasional DX when the band is open, what 'real' difference do a few decibels make when propagation conditions can vary by many tens of dB? So, unless you really want to eke out the very last vestige of antenna gain, the most important criterion is not ultimate gain but build quality. After all, a long boom antenna is no good if it folds in half during the winter gales. Similarly, the longer the antenna boom the sharper the directivity of the array becomes.

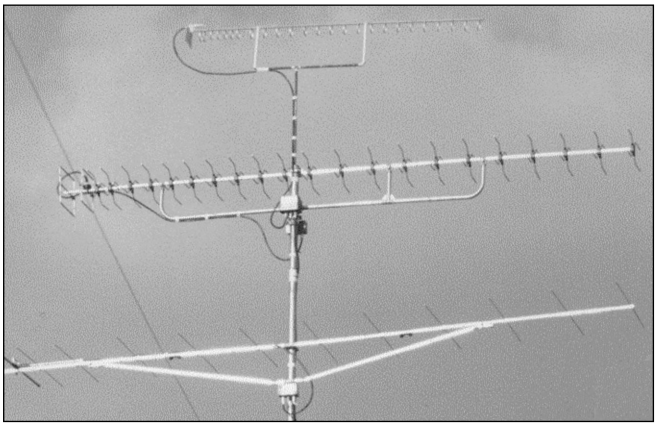


Fig 9.2: VHF/UHF antenna system of Richard Girling, G4FCD (circa 1991)

The possibility of missing stations away from the main antenna lobe becomes increasingly likely. So you might consider trading off some gain for an increase in beamwidth. Taking all these factors into account you might find that a pair of stacked 9-element Yagis (on the 144MHz band) will provide a more practical solution than using a single 18-element Yagi.

Much more information can be found in the chapter on VHF/UHF_antennas.

Siting and Cabling

The siting of an antenna is just as important as the type of antenna used. A ground-plane antenna located on a chimney top, clear of any obstructions, may give better results than a beam antenna located in a loft space. Unless you have restrictions imposed at your home, the best place for a VHF/UHF antenna is always outside in an uncluttered location. If possible, mount it on a suitable pole, elevating it above the roof and away from nearby television aerials.

The coaxial feeder connecting the antenna to the transceiver should have a low loss at the frequency in use and this is especially important on the VHF/UHF bands. A poor quality cable will lose valuable transmit and receive signal power, so be prepared to spend more money on the main feeder than on the antenna. It really will be an investment.

Finally, make sure that the connectors you use are of the highest quality. Although the use of N-type plugs and sockets is recommended, they are not essential, especially on the lower VHF bands.

Background Noise

Having paid attention to the antenna and feeder, it's now time to look at the receiver. The background sky noise arriving at the antenna effectively limits the maximum receiver sensitivity required for normal communications. On the lower VHF bands of 50MHz and 70MHz man-made noise often exceeds the background noise by 10dB or so. Consequently, receiver noise figures as high as 12dB and 10dB respectively are quite adequate for these bands. At 144MHz and 430MHz however, the sky noise is much less and a receiver noise figure of around 2.5dB will be quite adequate for most types of terrestrial communication. Unfortunately you probably won't find out the overall noise figure of your commercially made transceiver because it is rarely given. Normally the specification is given in terms of so many µV for a signal to noise ratio of so many dB. For example, one 144MHz transceiver quotes "better than 0.5µV for 11dB s/n", making the most favourable assumptions, this translates to a noise figure of 11dB.

Now you can see how little some manufacturers are really offering the VHF/UHF enthusiast. Much effort seems to be exerted in producing rigs with 100 memories, air-band receive facilities, computer control and displays that say "Hello", when what is really required is a VHF/UHF transceiver with a low noise figure, a dynamic range in excess of 100dB, switchable filters, IF shift, notch filtering, adjustable noise blankers and full CW break-in. All these features can be found on a modern HF radio, which brings us nicely back to the original suggestion of using a VHF/UHF transverter with an HF transceiver. You really do get the best system performance by adopting this technique.

Pre-amps

Another way of overcoming the basic lack of sensitivity is to use an external pre-amplifier and, if this is mounted at the antenna, it will also eliminate the effect of feeder loss in the receive direction. Unfortunately, the receive sensitivity is only improved if the pre-amplifier has sufficient gain, but this extra gain also decreases the strong-signal handling capability of the receiver. Therefore the use of a pre-amplifier may show overload effects on some signals that originally didn't cause any problems. Try to use a pre-amplifier that has adjustable gain, so that you can adjust it to suit your receiver. Typically, a gain of between 6 and 15dB will be sufficient for most needs.

Summary

The biggest improvements to your VHF/UHF station always come first. Changes to the antenna system, coaxial feeder, making the receiver more sensitive and increasing your transmit power will easily improve your system performance. After that it becomes a little bit more difficult. The rewards are still available but each improvement will be less significant.

D-STAR

We are at the start of the digital revolution in amateur radio so it is important to understand the new modes that are becoming available. One that has the support of a major amateur radio equipment supplier, Icom, is likely to gain a significant part of the market. The following section by Gavin Nesbitt, M1BXF, describes how D-Star works and how to use the mode.

Imagine a radio system that finds your friends for you; A system where you are able to call CQ on a handheld and making the contact exactly the same whether your friend is next door or on another continent; an amateur radio system that lets you send position reports in a similar way to analogue APRS and other messages without nasty noises on the audio channel.

Until recently, this sort of connectivity has been the stuff of dreams. But now, D-Star makes it all a reality. It transparently links mobiles, handhelds, home stations and repeaters world-wide using the Internet. In normal use it simply doesn't matter if your friend is on the local repeater or on holiday miles away, D-Star sorts it all out and lets you communicate effortlessly.

The great thing about D-Star is that you don't have to learn about the technical stuff to operate using it - unless you want to.

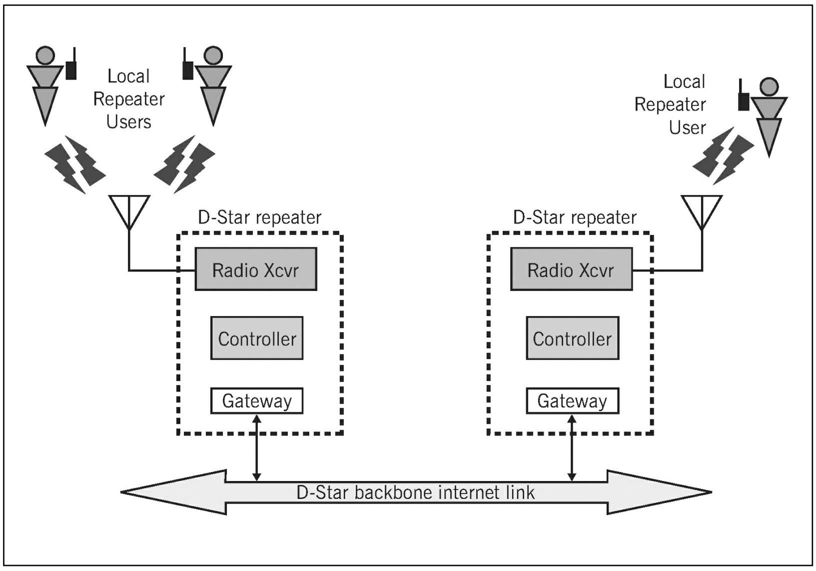


Fig 9.3: D-Star can connect individuals directly, via a local repeater or over large distances via the Internet

Overview

The D-Star system consists of individual radios, repeaters, servers known as gateways and Internet connections. **Fig 9.3** shows a diagram of the D-Star network. D-Star repeaters work like standard analogue voice repeaters, with some added bonuses. For a start, they can operate on several bands at once, with 2m and 70cm being most common.

So, what are the fundamentals of getting active on D-Star? First, you'll need a D-Star compatible radio (or DV-Dongle, more of which later). The only people currently making D-Star radios are Icom, although there is nothing to stop others climbing aboard the bandwagon because D-Star is an open standard so that anyone can make equipment.

You need to tell the D-Star radio who you are, which identifies you on the D-Star network. Next, you need to work out who you want to talk to. If you haven't got any D-Star equipped friends yet, just set it to *CQCQCQ* and someone will come back, just like any other mode.

Finally, you need to make the radio talk to your local repeater by entering its frequency and callsign.

So, the radio knows who you are, who you want to talk to, and the name of the repeater you want to work through. What are you waiting for? Call CQ!

The Basics

D-Star stands for Digital Smart Technology for Amateur Radio and was developed by the Japanese Amateur Radio League (JARL) back in 1999, in conjunction with Japanese universities and amateur radio suppliers. Their goal was to advance the hobby into the digital age by developing an open standard, using off-the-shelf parts, such as the AMBE 2020 voice encoder chip, which would allow anyone to design and produce digital radios or interfaces for radio amateurs. So D-Star was born.

The D-Star standard has two modes. The one causing the most excitement, and the main one used in the UK, is Digital Voice. DV mode allows simultaneous voice and serial data and easily fits in a 12.5kHz channel. The other mode is a high-speed digital data mode, DD, which gives a 128kHz half-duplex data connection. DD requires wide channels and is only used on 23cm.

Digital Voice uses a 4800 baud data stream. This is split into 3600 baud for voice, including 1200 baud of error correction, and 1200 baud for non-error-corrected data. When DV mode signals are marginal, the first noticeable effects are errors on the 1200 baud serial data stream. The voice holds up reasonably well as signals degrade further, eventually failing quickly as the signal quality falls below the error correction capability.

Think of D-Star as the transport mechanism, just like the meaning of FM or SSB. It's what you send over it, in DV or DD mode, which makes it interesting. A little like packet radio, packet by itself is useful but using packet for APRS or the DX cluster is what makes packet what it is today. D-Star uses GMSK (Gaussian Minimum Shift Keying) modulation to send the 4800 baud data stream over the air. As it is digital, the data stream includes headers such as the route the data should take and where it came from. There are four values that are used for routing and are easily programmed in all the radios available. These are MYCALL, URCALL (sometimes seen as YOURCALL), RPT1 and RPT2. In real life, these are used as follows:

The Icom IC-E92 D-Star handheld radio

Abridged from a review by Gavin Nesbitt, M1BXF published in the November 2008 edition of RadCom

Icom's IC-E92, shown in **Fig 9.4**, is the European version of the IC-92A dual-band FM/D-Star handheld. It was the first Icom radio to come with D-Star as part of the default feature set and not an additional option, a sign of Icom's commitment to the D-Star standard. The IC-E92 remedies many of the problems of its popular predecessor, the IC-E91, and adds more functionality. This brings it in line with some of the mobile D-Star radios available.

The European version of the IC-E92 receives from 495kHz through to 999.995MHz. This range is only available on band A, with band B limited to 118 to 174MHz and 350 to 470MHz. The radio has a large liquid crystal display that can show both bands at once when in dual receive mode, supporting V/V, U/U & V/U. Modes available are FM, FM-N, WFM, AM and DV (Digital Voice), DV only being available on band B. Transmit is limited to the 2m and 70cm amateur bands, 144-146MHz and 430-440MHz.

The transmitter has four power level options, high at 5W, med at 2.5W, low at 0.5W and super low at 0.1W. The radio still gets rather warm on the higher powers, which is not so bad when operating on a cold day. The manual actually contains more than one warning that the transceiver becomes hot and "may cause a burn". Stated operating life is 6 hours and there are two auto power save modes available.

The IC-E92 has a total of 1304 memories. This staggering number is made up of 800 normal memories on band A with an additional 400 on band B. Both bands A and B have 50 scan edge channels and two call channels each. The memory channels can be arranged into 26 scan bands. All memories including scan edges and memory banks can be alpha tagged with an 8-character identifier. Using the control software discussed later you can upload, save and manipulate the memories on a PC. Many memory files are available online to download.

The IC-E92 also has the regular features you would expect from modern FM handhelds: 1750Hz toneburst, 10 channel DTMF memories, multiple scan functions, built-in CTCSS/DTCS encoder and decoder and a simple bandscope.

On the DV side, the IC-E92 can report the GPS position of another DV station including supporting GPS A mode for easy D-PRS operation, includes a one touch reply button and has a built-in voice recorder with auto reply voice message. It can be fully controlled from a PC with the RS-92 software and serial cable (see below).

Fig 9.4: The Icom IC-E92

Construction

People who like rugged handheld radios won't be disappointed with the IPX7 submersible construction of the IC-E92, which allows the radio to be submerged to a depth of 1 metre for 30 minutes. This might not seem relevant as nobody can imagine using the radio underwater but it does give you confidence to use it in the rain.

The buttons on the front panel are well spaced out, and all are dual function with a momentary press for one and a long press for the other. All frequent functions are on these keys. The keypad also doubles up to allow direct frequency input or to enter a memory channel.

Also on the front are band switch (dual receive), power, band and menu (key lock) keys, features in brackets indicate a long press function. There is a dedicated menu button. Using the 2, 4, 6, 8 & 5 keys for up, left, right, down and enter navigation makes navigating the menus simple. A rotary control at the top is used for frequency and memory navigation in normal use. Below this control is a rotary volume control. There is a horizontal bar graph that appears on the display when you adjust the volume to see what the level is.

Between these controls and the SMS antenna connector is the external microphone connector which is also used for the PC interface. On the left side of the radio is the PTT, and below the squelch key. Holding the squelch key and turning the control dial adjusts the squelch from open, auto through levels 1 to 9.

On the right side is the DC jack that can be used to power the radio and charge the battery. It accepts voltages between 1V and 16V.

On-air and usability

All in all the radio feels good in the hand, it actually feels you could hammer a tent peg in with it while still holding a QSO. Using it with one hand is very easy because the volume and dial are in the correct place for your forefinger if held in your left hand and your thumb if held in your right hand. Using the menu from one-handed operation is also easy and the menu is good for ease of navigation.

The microphone feels sturdy in the hand and has a good strong clip on the back that can be rotated around 360 degrees with locking points every 45 degrees, it is also constructed to IPX7 rating including the connector. There is a button on the microphone to enable and disable the GPS functionality, the radio can display your latitude and longitude on the screen and, in DV mode, will send it over the slow data channel when you transmit. If you are talking to someone else on DV mode that is sending GPS data, you can display the location of the remote station including distance and direction from you on the screen too.

Using the local FM repeaters was no problem with the IC-E92 after setting all the memories including tones. Using the alpha tags on the memories helps to find them with the ability to monitor two repeaters at once.

PC software

Loading the RS-92 PC software is trouble free making it easy to talk to the radio as shown in **Fig 9.5**. The price of the software is a little high, mainly because it is supplied with the serial cable with the proprietary connector. It's a real shame because the cable is needed to take advantage of the slow speed data facility but the cable can only be bought with the RS-92 software at present. The RS-92 software does most things expected of rig control software. It displays exactly what is on the main radio display down to the signal meter and battery indicator, this is quite useful because the update rate is almost instantaneous making it useful and not a gimmick.

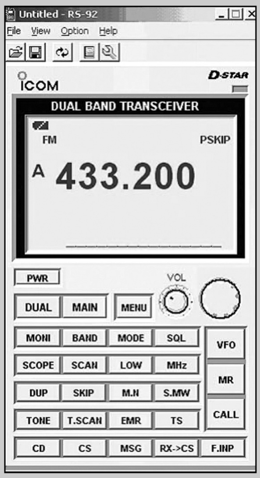


Fig 9.5: The Icom IC-E92 PC interface

The software has the ability to show a wealth of options and current status of the radio, most is mainly for D-Star like GPS RX position, Received call record and messages, all with history. Some options like Edit set mode make using the IC-E92 at home much more versatile than just using the radio on its own. The RS-92 software also provides the ability to edit, save and import memories, although using the keypad is not difficult. There are many saved memory files available to download saving lots of time when setting up the radio or swapping memories around if going on holiday. Using the software on air with D-Star I had all the received callsigns showing on the display while at the same time showing what GPS beacons are being sent and which messages are passing over the repeater, yes the RS-92 can show all this at once.

MYCALL: Station/Terminal identifier, this is usually your own callsign but can be any name you wish such as CAR 1, USER A, etc. There is no licence requirement for it to be your callsign and this entry is what is used to authenticate and track you on the D-Star network (more later) so using your callsign is favoured.

URCALL: The destination address. Can be another users callsign, a repeater or, if there is no intended recipient, *CQCQCQ* is used.

RPT1: The callsign of the repeater being accessed, this is used in the same way as CTCSS on analogue repeaters - meaning if you are in range of two D-Star repeaters only the one corresponding to the RPT1 value will trigger.

RPT2: This value is optional; it is only used if you wish to route traffic to a different place such as another local port on a different band, through the gateway if you are connecting to a remote repeater or finding the location of a user and the echo test server - more about this later.

All the D-Star radios have these values that can be updated and, in the case of Icom radios, each memory channel supports having unique URCALL, RPT1 and RPT2 saved alongside other memory data. This means you can save a channel with the correct values to route to a remote repeater or user, and name the channel appropriately.

Advantages

Being digital is what makes D-Star an interesting mode. It brings all manner of new and enhanced features to radio amateurs such as data and call routing, simultaneous voice and data traffic, web reporting of stats and a real-time last heard list of users on the network. But, it's when in range of a D-Star repeater on the gateway, ie connected to the Internet, that the system provides most of the advantages over analogue repeaters, being able to form a wide area network.

A key feature is that the D-Star network can track a user's last known location in the form of the last repeater and port the user was heard on.

This can be used like a cellular system and automatically route calls so, if someone wants to contact another station without knowing where that station is located, the gateway can do a lookup of where he was last heard and can route the call there automatically.

Other features of gateways are that they are able to connect to reflectors, GPS (DPRS) position reporting to the web and echo test, which is just a simple way to test your transmission. It's worth saying that the gateway has user access control and users must be registered to use its features, there is no access control on using the repeater for local contacts.

The main reason for registration is to add your callsign, a MYCALL value, to the gateway that assigns it an ID (an IP address) that is used by all the gateways for tracking and routing calls. Contact your local D-Star repeater keeper or group to be added to the gateway once you get a D-Star radio.

Repeaters

Currently only Icom have a D-Star repeater system available but there are open source repeaters in the pipeline. Icom's repeater controller can support four repeaters, simultaneously associated with three ports, simply known as Port A (same for DV or DD), B or C. The ports are harmonised across the network so the user knows which repeater/band they are connecting to. Normal aliases are Port A for 23cm DV and/or DD, B is 70cm DV and C is 2m DV. There are other ports such as G for the gateway (internet connection) or E for echo test. When routing, you always need to specify the port you are connecting to on a remote repeater and, in Icom's case, the port number always has to be in the 8th position in the URCALL, RPT1 and RPT2 fields. If you start the URCALL with a forward slash '/' this indicates the callsign is that of a repeater and not a user.

Let's look at a few examples of how routeing works. If I wanted to connect to another repeater from the 70cm port on GB7PI to, say, the 70cm port on GB7NM (north Manchester) I would set my values as:

MYCALL = M1BXF,
RPT1 = GB7PI B,
RPT2 = GB7PI G,
URCALL =/GB7NM B.

When I transmit with these settings I would be heard locally on the 70cm port of GB7PI and would also be heard, almost instantly, on the 70cm port on GB7NM. If I used this connection a lot, I could save these values in a memory channel and call it something like GB7NM.

If I wanted to call a user directly, I would simply change the URCALL from the above example to the callsign of the station I wish to contact like URCALL = G7LWT. What happens now is the gateway PC does a lookup of G7LWT's last known repeater/port and then routes my transmission to that repeater/port. It's that easy.

If I were just calling locally, I would just need to set RPT1 = GB7PI B and URCALL = CQCQCQ. However, with the advent of reflectors and DV Dongle users it is suggested that everyone put the local gateway into RPT2. This is because DV Dongles and reflectors connect to repeaters as a gateway and not a user. So,

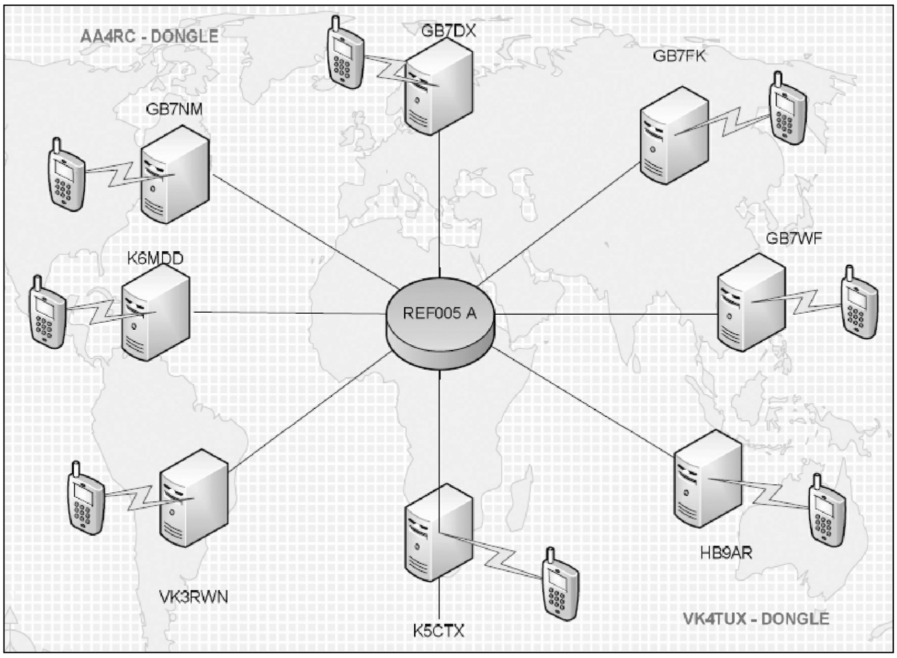


Fig 9.6: Reflectors link D-Star repeaters, RF stations and DV dongle users from around the world into a single meeting place

Fig 9.7: A collection of D-Star radios

for your data to be routed back to those connected via reflectors or DV Dongles, you must use the gateway, hence you need to set RPT2 = GB7PI G.

Reflectors

I keep mentioning reflectors. Users of *Echo Link*, *IRLP* and *eQSO* will be familiar with conference rooms, and reflectors are the D-Star equivalent. D-Star reflectors were developed by Robin, AA4RC, who has written many enhancement applications for the D-Star gateways. His *dPlus* is the code that reflectors use. Reflectors link D-Star repeaters, RF stations and DV Dongle users from around the world into a single meeting place as shown in **Fig 9.6**. There are reflectors located in many parts of the world. Two of these are in London: REF005 and REF006. Each has three ports, A, B and C, giving a total of six conference rooms that are used by different European countries or groups.

Connecting to reflectors is simple and only requires the name and port of the required reflector plus an L for 'link' or U for 'unlink' in the URCALL field, eg REF005AL to connect and REF005AU to disconnect. Once connected, an audible announcement is given and the users should then change their URCALL back to CQCQCQ so *dPlus* doesn't keep trying to connect to the reflector every time you transmit. Again these settings can be saved as a memory channel that can be recalled whenever needed.

DV Dongle

The DV Dongle is an AMBE2020 chip connected to a computer via a USB interface. *DV Tool*, the DV Dongle control software, is supported in Windows, Mac and Linux and connects to repeater gateways or reflectors to allow DV Dongle users to speak with D-Star radio users from their computer. As mentioned earlier, it does require remote RF users to have their RPT2 value configured correctly to the gateway port. The DV Dongle is not the same as an *Echo Link* user sitting at their PC as it connects to the gateway and not a user. In a comparative sense, it's like a link connecting to a repeater. At present, there is no radio used at the DV Dongle end but I believe there are plans to make DV Tool interface into a 9k6 port of an existing radio that will give the user the joy of being RF again.

Radios and Interfaces

D-Star is not Icom's own standard or technology, but they are the first and so far only company currently to make D-Star equipment commercially available. At the time of writing there are six different Icom radios available, four mobiles and two handhelds, some examples are shown in **Fig 9.7**. These comprise ID-1 for 23cm DD and DV mode, ID-800 for 2m and 70cm DV, IC-2200 for 2m DV, IC-2820 2m and 70cm DV mobiles and IC-E91 and IC-E92 2m and 70cm DV handhelds. All six radios are effectively normal FM radios with D-Star being an optional extra on some and fitted as standard in the other. I've had the chance to play with many of these radios and they are all feature rich even if just used on FM.

If buying a radio is not for you then there are many projects appearing in the form of adapters that connect to a 9k6 port in an existing FM radio and allow users to join in on D-Star. A word of caution: not all 9k6 radios are built the same and some have problems with transmitting GMSK. Have a look at [1] for project details. A final word if you are thinking of D-Star, there are many forums available where you can ask questions or just monitor if you prefer. The UK forum can be found at [2].

DESIGN THEORY

This section will explore some of the theory and practice of designing receiving and transmitting equipment for the 50, 70, 144 and 432MHz amateur bands.

Receivers

Standards for VHF/UHF receivers are strongly based on the performance expected from HF receivers, in particular the ability of the receiver to detect, without any deterioration in performance, a weak signal in the presence of one or more unwanted strong signals present at the same time.

Above 50MHz background noise is much lower, so with a good receiver it is possible to realise a performance superior in terms of sensitivity and signal-to-noise ratio. An HF signal of a few microvolts is often down in the noise but at VHF and UHF, communication between stations can be achieved with signal levels as low as a few nanovolts (1nV = 1V x 10^{-9}). On the HF bands a limit is imposed by both man-made and natural interference, beyond which any attempt to recover signals is fruitless. In VHF/UHF signal reception, there is no appreciable atmospheric noise with the exception of that caused by lightning discharges or from electrically charged rain drops. The limiting factor, when the receiver (and antenna) is in a good location, is extraterrestrial noise but the receiver can be designed to respond to signals only slightly above this level.

Definition of noise

Broadly, noise is unwanted signal of a more or less random nature within the pass-band of the receiver. It may be natural or man-made. Examples of natural noise are the radiation from the Sun or, as described earlier, that from electrical storms and charged rain drops. These can only be avoided by excluding the Sun or the electrical storms from the 'field of view' of the antenna.

Also, there is the inescapable noise generated in a resistor at any temperature above absolute zero, and shot noise produced in semiconductors, caused by the random generation and recombination of electron-hole pairs in their operations.

Examples of man-made noise are the radiation from switches and thermostats when they break current, and the radiation from computers caused by their processing pulses with very fast rise and fall times.

In the design of VHF/UHF receivers only the inescapable natural noise needs to be considered. Resistors introduce thermal noise, due to the random motion of charge carriers that produce random voltages and currents in the resistive element. There is unfortunately no resistor that will not produce these random products unless the receiver is operated at a temperature at absolute zero (0K). However, resistor noise generation can be minimised, particularly in the receiver front-end, by the correct choice of resistor. Metal film types are recommended. Thermal noise is also known as Johnson or white noise.

Shot noise in semiconductors is due to charge carriers of a particle-like nature having fluctuations at any one instance of time when direct current is flowing through the device. The random fluctuations cause random instantaneous current changes. Shot noise is also known as Schottky noise.

Noise factor and noise figure

The noise factor is the ratio of the input signal-to-noise ratio to the output signal-to-noise ratio. The noise figure is the noise factor expressed in decibels and is used as a figure of merit for VHF and UHF circuits:

$$f = \frac{\text{Input S/N}}{\text{Output S/N}}$$

$$\text{NF} = 10 \log_{10} f$$

It is measured as the noise power present at the receiver output assuming a conventional S/N ratio of 1 at the input. An ideal noiseless receiver does not produce any noise in any stage. Thus the equation becomes 1/1 or a noise factor of 1 or 0dB. The noise factor of a practical receiver that will generate noise in any stage, particularly the front-end, is the factor by which the receiver falls short of perfection.

Amateur communication receiver manufacturers usually rate the noise characteristics with respect to the signal input at the antenna socket. It is commonly expressed as: (signal + noise) / noise, or “signal-to-noise ratio”.

The sensitivity is usually expressed as the voltage in microvolts at the antenna terminal required for a (signal + noise) / noise ratio of 10dB. Sensitivity can also be specified as the minimum discernible signal or noise floor of the receiver.

An important point to remember in VHF/UHF receivers is that the optimum noise figure of an RF amplifier does not necessarily coincide with the highest maximum usable gain from that stage.

Figs 9.8 and 9.9 illustrate this feature. The transistor is capable of operation up to 2.0GHz. As Fig 9.8 shows, however, the gain falls with increasing frequency but the actual noise also increases with increasing frequency. This characteristic is also shown in Fig 9.9 but here the gain at noise figure and the actual noise figure are plotted for 500MHz and 1GHz against variations in collector current. Note that the maximum gain occurs with good input matching but minimum noise does not. For example, a GaAsFET preamplifier may have an input VSWR as high as 10:1 when tuned for the lowest noise figure.

The definition of noise figure and degradation of receiver performance due to noise implies that the front-end stages, namely the RF amplifier and mixer, must use active devices, either bipolar or field effect types, with a low inherent noise figure. The noise figure quoted for the transistor illustrated in Figs 9.8 and 9.9 applies only when the device is connected to a 50 ohm source. As can be seen the noise figure increases by almost two times between 1 and 2GHz. This figure can be reduced by mismatching to the source at 1GHz and above.

Modern design theory and practice now employ S-parameters (scatter parameters) to obtain maximum performance from an RF amplifier and mixer while maintaining the noise figure at low levels. However, S-parameters require an advanced knowledge of design which includes the use of the Smith chart and the availability of some sophisticated equipment such as a network analyser. Manufacturers’ data on RF devices includes tables of S-parameters. This method of design of amplifiers and mixers is outside the scope of this handbook.

Intermodulation

Intermodulation occurs when two or more signals combine to produce additional (spurious) signals that were not originally present at the receiver input and possibly causing interference to a weak wanted signal. The receiver front-end is handling many incoming signals of different strengths but only those signals passing through the selective (IF) filters will eventually be detected. The RF circuits can have a bandwidth of several megahertz but the selective IF filters reduce the bandwidth to that required for adequate resolution of signals, dependent on

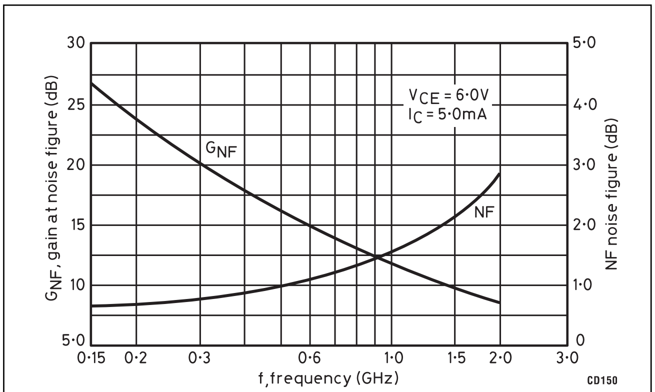


Fig 9.8: Showing gain at noise figure and noise figure versus frequency

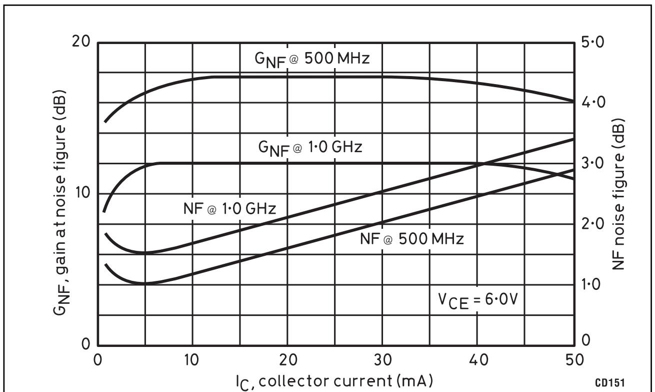


Fig 9.9: Gain at noise figure and noise figure versus collector current

the method of modulation of the wanted carrier. At low signal levels the front-end will have optimum linearity, ie there is no unwanted mixing between signals. However, as stated previously, very strong signals will cause the front-end to go into its non-linear region of operation and then these signals will mix together and produce new signals which can appear in the IF pass-band. Second-order intermodulation products (IPs) are caused by two signals mixing, viz f_1 and f_2, and generating new frequencies which appear as ($f_1 + f_2$) and ($f_1 - f_2$) and the second harmonics of each signal ($2f_1$ and $2f_2$), generating the second-order IPs. However, if f_1 and f_2 are close spaced, their second-order IPs will be well spaced and can be easily filtered out by the selective IF filters.

However, if the f_1 and f_2 signals are increased in strength then another set of IPs is generated. These are third-order intermodulation products, due to the fact that mixing occurs between three signals. The three signals can be independent but the same products can be generated by f_1 and f_2 by themselves. These frequencies f_1 and f_2 can add or subtract to produce the following third-order IPs:

Third harmonics:	(f_1+f_1+f_1) and (f_2+f_2+f_2)
Sum products:	(f_1+f_1+f_2) and (f_2+f_2+f_1)
Difference products:	(f_1+f_1-f_2) and (f_2+f_2-f_1)

It is clear that if f_1 and f_2 are equally spaced above and below the wanted frequency, interference will be severe. When $f_1 + f_2$ are close to the wanted frequency the third-order sum products will appear in the third harmonic area of this frequency and will be attenuated by the selective filters in the receiver. However, when the difference products containing a minus sign are close to f_1 and f_2 and are generated by the receiver, the filters, however selective, will not remove these spurious signals. These products could cause unwanted interference to a wanted weak signal.

When third-order intermodulation products are generated in the receiver, they will increase in level by 3dB for every 1dB increase in the levels of f_1 and f_2. Thus the appearance of intermodulation products above the receiver noise floor is quite usual. When further levels of f_1 and f_2 occur higher odd-order intermodulation products are generated, eg fifth, seventh etc, which can interfere with a weak wanted signal. These higher-order products will appear even more quickly than third-order products but require stronger f_1 and f_2 signals, eg fifth-order products will be generated five times as fast as f_1 and f_2 as the level of f_1 and f_2 is increased. Significant intermodulation products can only result from f_1 and f_2 when their strength is high. If either the f_1 or f_2 signal disappears, leaving only one signal, the intermodulation product will disappear. Optimising linearity in the receiver front-end will minimise generation of these unwanted intermodulation products and hence interference to wanted signals.

Gain compression

This occurs when a strong incoming signal appearing at the antenna socket causes one (or more) stages in the front-end to be driven into the non-linear region of its output characteristic. As an example, when an amplifier stage is operating in its linear region, an increase by 3dB in signal level at its input will cause by linear transfer, a 3dB increase in signal level at the output. However, a further increase in input signal level could cause non-linear transfer and limit the output level increase to 1dB. A very strong signal could drive the stage into extreme non-linearity, making the stage degenerative (gain less than 1) and desensitising the receiver. Background noise will decrease in level together with all other signals, including wanted weak signals. As with intermodulation, optimising front-end linearity will minimise receiver desensitisation by strong signals.

Reciprocal mixing

This phenomenon occurs when the receiver local oscillator produces excessive sideband noise on its carrier and a strong off-channel RF carrier mixes with this noise to produce the IF. Reciprocal mixing causes an increase in receiver noise level when a strong carrier appears, the opposite effect of gain compression. The receiver selectivity is not necessarily defined by front-end RF filters. The ‘cleanest’ local oscillators are LC (VFOs) and crystal-controlled oscillators. Some of the `noisiest’ oscillators are found in receivers employing a synthesised system, for example phase-locked-loop synthesised oscillators. Some earlier receivers suffered from reciprocal mixing effects, ie generation of an unwanted spurious signal in the IF pass-band, for example, noise on the voltage-controlled oscillator control line leading to FM noise sidebands. However, modern receivers now employ ‘quiet’ synthesised local oscillators that minimise reciprocal mixing.

Dynamic range

The main problems of front-end overload are gain compression, intermodulation and reciprocal mixing. Each phenomenon has its own characteristic and level at which strong unwanted signal(s) cause degradation in receiving wanted signals. Just one strong signal causes gain compression or reciprocal mixing, whereas two are required to cause intermodulation products. The gain of the front-end, ie the RF stage and the mixer, should be kept as low as possible, consistent with good sensitivity and signal/noise performance. Gain compression and intermodulation are caused when either or both stages are driven beyond their linear transfer range. The front-end should be designed so it cannot be overloaded by even the strongest amateur band signals. Intermodulation products will not be a problem if they are restricted to the level of the background noise and gain compression and reciprocal mixing effects are not a problem if they do not significantly change the system noise level.

The lowest end of the dynamic range will be designed for the lowest power audible signal and, conversely, the highest end will be designed for the unwanted signal of the highest power level, ie signals without any overload effects degrading the front-end performance. This principle is called spurious-free dynamic range but the range will change according to the differing power levels of unwanted signals.

Receiver Front-End Stages

Thus the requirements for front-end stage design in a VHF/UHF receiver are:

- Low noise figure
- Large dynamic range
- Power gain consistent with good sensitivity and signal-to-noise ratio

The noise figure and dynamic range requirements have already been described in detail. However, ‘power gain’ must be brought into the equation to complete the design philosophy. ‘Power gain’ needs some explanation because sheer power gain is not sufficient in itself or even desirable. In a multistage receiver with, say, eight stages of gain, input noise originating in the first stage, normally an RF amplifier, will be amplified by the eight gain stages, that in the second by seven and so on. If the effective noise voltages are denoted by V_1, V_2, V_3, ... V_8 and the stage gains by G_1, G_2, G_3,... G_8, the total noise present at the receiver detector will be:

$$V_1(G_1 G_2 G_3 .. G_8) + V_2(G_2 G_3 .. G_8) + V_3(G_3 G_4 .. G_8) \text{ and so on.}$$

If the voltage gain of the RF amplifier (G_1) is high, for example 20dB (10 times) or more, the important noise contribution is due to G_1. Provided the remaining gain stages are correctly designed and provide evenly distributed gain, the overall noise contribution from them will be very small.

Additional noise generated by any stage that is regenerative or is actually oscillating will degrade the overall receiver noise performance, and might actually cause receiver desensitisation and consequent poor weak-signal performance.

However, at the output of G_3 (eg the first IF stage) an amplified signal will be large compared to the noise contributed and G_4 ... G_8 should not degrade noise performance. The function of the RF stage is to provide just sufficient gain to overcome the noise contribution of the mixer stage. The mixer is by definition a non-linear device and normally contributes more noise than any other stage no matter how well designed. The RF stage therefore considerably improves the receiver signal-to-noise ratio, improving weak-signal performance. If the RF stage gain is too high then the problems of strong signals, ie intermodulation, gain compression and reciprocal mixing products, will, as described, degrade receiver performance.

A system of gain control on the RF amplifier may appear to be the answer. However, the use of AGC must be carefully considered, otherwise the weak-signal performance will be degraded if the onset of AGC is not delayed. Reducing the amplifier gain, using an AGC current for bipolar transistors or a voltage for field effect transistors, can degrade the RF stage linearity. The noise factor of a transistor amplifier is also dependent on emitter (source) current. A large variation from the manufacturer's data given for noise factor V I_E (I_S) will also degrade weak-signal performance. Circuit layout and correct shielding for VHF/UHF RF amplifiers is of paramount importance for stable operation. Instability and spurious oscillation can be produced via RF feedback through the amplifier transistor. Even the latest designs of transistor, both bipolar and field effect types, can give rise to these effects. Regeneration or actual oscillation can be prevented by neutralising the internal feedback, using an external circuit from output to input of the amplifier. This will feed back an equal amount of out-of-phase signal and, providing there is no other feedback path, the amplifier will be stable.

Circuit noise

Noise due to devices other than transistors is produced solely by the resistive component; inductive or capacitive reactances do not produce noise. Inductors of any form have negligible resistance at VHF and UHF but the leakage resistance of capacitors and insulators is important. It is imperative to choose high-quality capacitors such as silver mica, ceramic, polycarbonate, or polystyrene types with negligible leakage current and high-Q properties. Tantalum capacitors should be used where it is necessary to decouple at LF, as well as VHF and UHF. Attention should be given to the use of low-noise resistors. Carbon or metal film types (of adequate power dissipation) must be used. The old-style carbon composition resistors are very good noise generators. 25dB noise difference between film and composition resistors has been observed when a direct current was passed through the two types, both being the same value.

Circuit noise due to regeneration has already been discussed. Common causes of regeneration are:

(a) Insufficient decoupling of voltage supplies (at each stage) and particularly emitter (source) and collector (drain) circuits.

Suitable capacitor values and types are as follows:

10MHz	47nF (0.047µF) ceramic disc or plate
10-20MHz	22nF (0.022µF) ceramic disc or plate
20-30MHz	10nF (0.01µF} ceramic disc or plate
30-100MHz	4.7nF (0.0047µF) ceramic disc or plate
100-200MHz	2.2nF (0.0022µF) ceramic disc or plate
200-500MHz	1nF (0.001µF) ceramic disc or plate
500-1000MHz	220pF (0.00022µF) ceramic disc or plate

Leadless capacitors should be used above about 400MHz.

(b) Closely sited input and output circuits. Every attempt should be made to build the amplifier(s) in a straight line and where possible, to use shielded coils, particularly for 6m, 4m and 2m equipment.

(c) Insufficient or wrongly placed screening between input and output circuits. In transistor amplifiers where a screen is required, it should be mounted close fitting across the transistor with input and output circuits (normally base/gate and col-

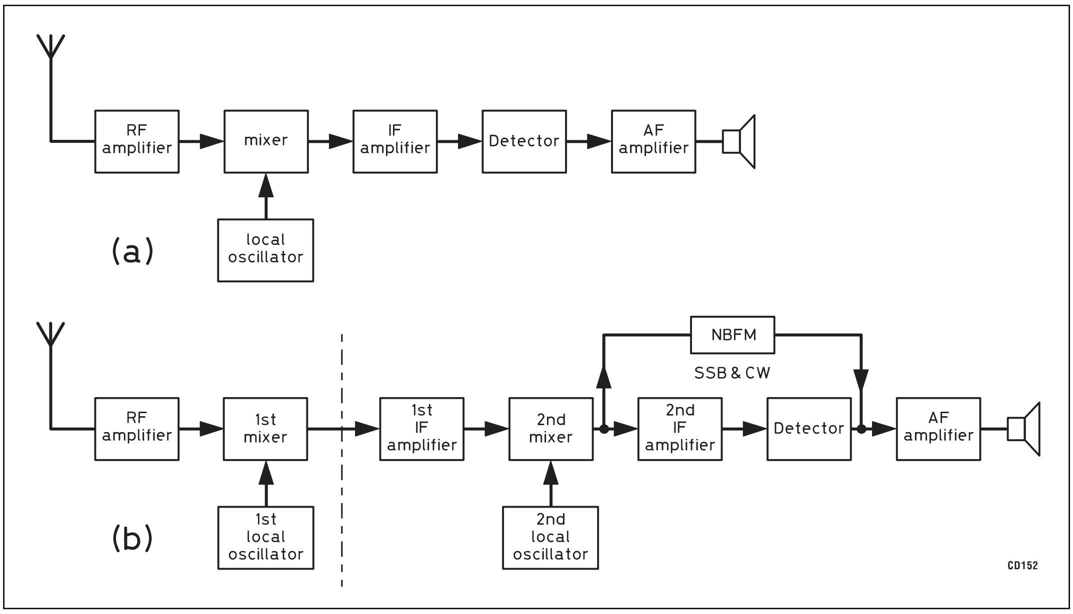


Fig 9.10: Two conventional superheterodyne configurations: (a) Self-contained single superhet with tuneable oscillator. (b) Self-contained double superhet or converter in front of a single superhet. Either or both oscillators may be tuneable

lector/drain) on opposite edges of the screen. The screen can conveniently be made of double-sided copper laminate and soldered to the main PCB.

(d) Circulating IF currents in the PCB due to multipoint grounding. Decoupling capacitors for each stage should be grounded at a single point very close to the emitter (source). It is preferable to use a double-sided PCB using one side as the ground plane.

Effect of bandwidth on noise

If the noise factor of a receiver is measured with a noise generator it is independent of receiver bandwidth. Generator noise has the same characteristics as circuit noise so, for instance, if the bandwidth doubled, the overall noise is doubled. For the reception of a signal of finite bandwidth however, the optimum signal-to-noise ratio is obtained when the bandwidth of the receiver is only just sufficient to accommodate the signal. Any further increase in bandwidth results in increased noise. The signal-to-noise ratio at the receiver detector therefore depends on the power per unit bandwidth of the transmitted signal.

As an example, a receiver may generate 0.25μV of noise for each 2.5kHz of bandwidth. Assuming an SSB transmitter radiates a sideband signal of 2.5kHz bandwidth and produces 2.5μV of signal at the receiver detector, the signal-to-noise ratio is therefore 10, provided the receiver overall bandwidth is 2.5kHz.

If the transmission bandwidth is reduced to 1.25kHz for a CW signal and the radiated power is unchanged, the receiver input will remain at 0.25μV but if the bandwidth is also reduced to 1.25kHz, the receiver detects only 0.125μV of noise and the signal-to-noise ratio will increase to 20.

Using receiver bandwidths that exceed transmission bandwidths is therefore undesirable when optimum signal-to-noise ratio is the prime factor. Transmitters with poor frequency stability will either require the receiver to be retuned or the use of wider bandwidth, resulting in a degraded signal-to-noise ratio. Fortunately, well designed transmitters with PLL synthesised oscillators or crystal-controlled oscillators are now employed in the majority of the VHF/UHF bands.

Choice of Receiver Configuration

Receivers using other than superheterodyne techniques are rare on VHF or UHF. Modern superheterodyne receivers may have one, two or three frequency changes before the final IF, each with its own oscillator which may be tuneable (by the receiver tuning control) or of fixed frequency. Receivers may have a variety of configurations; two are illustrated here in **Fig 9.10**. **Fig 9.10(a)** shows a conventional single superheterodyne for use on the HF bands. The local oscillator will be partially synthesised, ie use a pre-mixer driven by an HF crystal-controlled oscillator and a LF VFO to produce the local oscillator for the main mixer. **Fig 9.10(b)** shows a double superheterodyne. This can be a purpose-built receiver (or as illustrated in Fig 9.10a)) to which is added (to the left of the dotted line) a VHF/UHF converter. The first local oscillator is crystal controlled and tuning is accomplished using the HF receiver (second) oscillator.

The main disadvantage of this method of using an HF receiver as a 'tuneable IF' amplifier preceded by a converter is again the problem of overloading the first amplifier and second mixer with strong signals. This will result in intermodulation and reciprocal mixing products, if not gain compression, particularly if the converter gain is high, say, 20 to 30dB.

A superior arrangement is to build a tuneable IF amplifier containing all the refinements of a normal HF receiver, including an NBFM IF amplifier and detector, and restrict the tuning range to a few megahertz to cover the VHF/UHF ranges of the converters. The HF receiver gain in front of the second mixer must be low. The first IF amplifier can be omitted but the pre-mixer selectivity should be retained. The converter gain (from VHF/UHF to first IF) should also be low: 10 to 14dB. This will result in a VHF/UHF receiver with a very good noise factor and dynamic range, and demodulation of NBFM in addition to CW and SSB signals.

Choice of the first IF

In any superheterodyne receiver it is possible for two incoming frequencies to mix with the local oscillator to give the IF; these are the desired signal and the image frequency. A few figures should make the position clear. It will be assumed that the receiver is to cover the 144 to 146MHz band, and that the first IF is to be 4 to 6MHz. The crystal oscillator frequency must differ from that of the signal by this range of frequencies as the band is tuned and could therefore be 144 - 4MHz = 140MHz or, alternatively, 144 + 4MHz. However, the choice of 144 + 4MHz would invert the sideband being received. If the signal transmitted were USB, the tuneable IF would need to be set to LSB to resolve the signal correctly. As this is inconvenient the best choice is 144 - 4MHz. Assuming that the lower of the two crystal frequencies is used, a signal on 136MHz would also produce a difference of 4MHz and unless the RF and mixer stages are selective enough to discriminate against such a signal, it will be heard along with the desired signal on 144MHz. From the foregoing, it will be appreciated that the image frequency is always removed from the signal frequency by twice the IF and is on the same side as the local oscillator.

It should be noted that even if no actual signal is present at the image frequency, there will be some contributed noise which will be added to that already present on the desired signal. It is usual to set the RF and mixer tuned circuits to the centre of the band in use so that on the 144MHz band they should be at least 2MHz wide in order to respond to signals anywhere in the band. This bandwidth only represents approximately 1.4% of the mid-band frequency and it is not surprising that appreciable response will be obtained over the image frequency range of 134 to 136MHz unless additional RF filtering is employed. Naturally the higher the first IF, the greater the separation between desired and image frequencies. However, an IF as low as 4 to 6MHz is feasible, provided some attempt is made to restrict the bandwidth of the converter by, for example, employing two inductively-coupled tuned circuits between the RF and mixer stages, thus providing a band-pass effect.

The choice of the first IF is also conditioned by other factors. Firstly, it is desirable that no harmonic of the oscillator in the main receiver should fall in the VHF band in use and secondly, there should be no breakthrough from stations operating on the frequency or band of frequencies selected for the first IF.

Many HF receiver oscillators produce quite strong harmonics in the VHF bands and, although these are high-order harmonics and are therefore tuned through quickly, they can be distracting when searching for signals in the band in question. The problem only exists when the converter oscillator is crystal controlled, as freedom from harmonic interference is then required over a band equal in width to the VHF band to be covered. This also applies of course to IF breakthrough.

As it is practically impossible to find a band some hundreds of kilohertz wide which is unoccupied by at least some strong signals, it is necessary to take steps to ensure that the main receiver does not respond to them when an antenna is not connected. Frequencies in the range 20 to 30MHz are often chosen, since fewer strong signals are normally found there than on the lower frequencies but this state of affairs may well be reversed during periods of high sunspot activity.

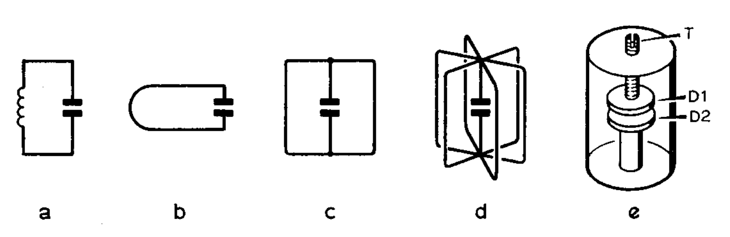


Fig 9.11: Progressive development of tuned circuits from a coil to a cavity as the frequency is increased

With the greatly increased use of general-coverage receivers covering 100kHz to 30MHz, the best part of the spectrum for 6m, 4m and 2m is from 28 to 30MHz. Full coverage of the 70cm band will require the receiver to be tuned from 10 to 30MHz. IF breakthrough is minimised and frequency calibration is simple.

Tuned Circuits

Tuning is readily achieved at HF by lumped circuits, ie those in which the inductor and capacitor are substantially discrete components. At VHF the two components are never wholly separate, the capacitance between the turns of the inductor often being a significant part of the total circuit capacitance. The self inductance of the plates of the capacitor is similarly important. Often the capacitance required is equal to, or less than, the necessary minimum capacitance associated with the wiring and active devices, in which case no physical component identifiable as 'the capacitor' is present and the circuit is said to be tuned by the 'stray' circuit capacitance.

As the required frequency of a tuned circuit increases, obviously the physical sizes of the inductor and capacitor become smaller until they can no longer be manipulated with conventional tools. For amateur purposes the limits of physical coils and capacitors occur in the lower UHF hands: lumped circuits are often used in the 432MHz band but are rare in the 1.3GHz band.

Distributed circuits

Fig 9.11 illustrates how progressively lower inductances are used to tune a fixed capacitor to higher frequencies. In Fig 9.11(b) the 'coil' is reduced to a single hairpin loop, this configuration being commonly used at 432MHz. Two loops can be connected to the same capacitor as in Fig 9.11(c). This halves the inductance and can be very convenient for filters.

Fig 9.11(d) represents a multiplication of this structure and in Fig 9.11(e) there are in effect an infinite number of loops in parallel, ie a cylinder closed at both ends with a central rod in series with the capacitor. If the diameter of the structure is greater than its height it is termed a rhumbatron, otherwise it is a coaxial cavity. The simple hairpin, shown at Fig 9.11(b), is a very convenient form of construction: it can be made of wide strip rather than wire and is especially suitable for push-pull circuits. It may be tuned by parallel capacitance at the open end, or by a series capacitance at the closed end.

In a modification of the hairpin loop, the loop can be produced from good-quality double-sided printed circuit board and such an arrangement is known as microstripline. The loop is formed on one side; the ground-plane side of the PCB, through the dielectric, makes the stripline. When the PCB is very thin, the result is called microstrip and that is used in many commercial receivers.

Bandpass circuits

Tuning of antenna and RF circuits to maintain selectivity in the front-end of a VHF/UHF receiver cannot be undertaken with normal ganged tuning capacitors, not only due to the effects of strong capacitance coupling between circuits, which become prominent in these frequencies, but also due to the difficulty of procuring small-swing (say 20pF) multi-ganged capacitors. The varicap diode can be used to replace mechanical capacitors, but can degrade the receiver performance when strong signals are present by rectifying these signals and introducing intermodulation products into the mixer, thus seriously reducing the receiver's dynamic range. Fortunately modern construction techniques have enabled coil manufacturers to introduce a band-pass circuit in a very small screened unit, namely the helical filter.

The helical filter in simple terms is a coil within a shield. However, a more accurate description is a shielded, resonant section of helically wound transmission line, having relatively high characteristic impedance. The electrical length is approximately 94% of an axial quarter-wavelength. One lead of the winding is connected to the shield; the other end is open-circuit. The Q of the resonator is dictated by the size of the shield, which can be round or square. Q is made higher by silver plating the shield. Resonance can be adjusted over a small range by opening or closing the turns of the helix. The adjustment is limited over the small frequency range to prevent degradation of Q. Modern miniature helical resonators can be obtained in a shield only 5mm square, but the minimum resonant frequency is normally 350MHz. Maximum F_r can be 1.5GHz. Large-size resonators (10mm square) will resonate down to 130MHz.

The band-pass filter is obtained by combining two to four resonators in one unit with slots cut in each resonator screen of defined shape to couple the resonators. This forms a high-selectivity tuned circuit with minimum in-band insertion loss and maximum out-of-band attenuation.

Helical filters can be cascaded to increase out-of-band attenuation. As an example, a quadruple filter with a centre frequency of 435MHz might have a 3dB bandwidth of 11MHz and 25dB attenuation at plus or minus 15MHz; with a ripple factor of 2dB and insertion loss of 4dB (see Table 9.1). Fig 9.12 shows pictures of the helical filters tested. The test circuit is shown in Fig 9.13, and Figs 9.14 and 9.15 show the test results.

Tuning is accomplished by brass screws in the top of the screen, one for each helix. The nominal input/output impedance

Parameter	HRW (231MT-10001A)	HRQ (232MT-1001A)
Centre frequency (MHz)	435	435
Bandwidth at 3dB (MHz)	12 min	11
Attenuation (dB)	20 min at ±30MHz	25 min at ±15MHz
Max ripple (dB)	1.5	2
Max insertion loss (dB)	2.5	4
Impedance (Ω)	50	50

Table 9.1: Electrical characteristics of Toko HRW and HRQ helical resonators

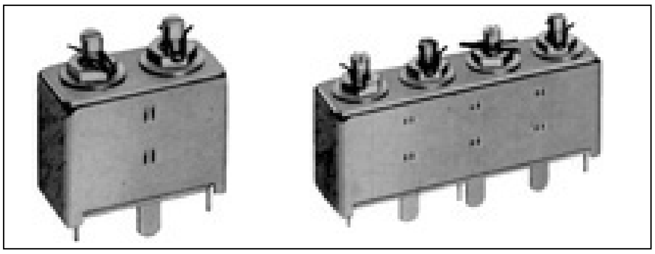


Fig 9.12: (left) Toko resonators. (left) HRW and (right) HRQ

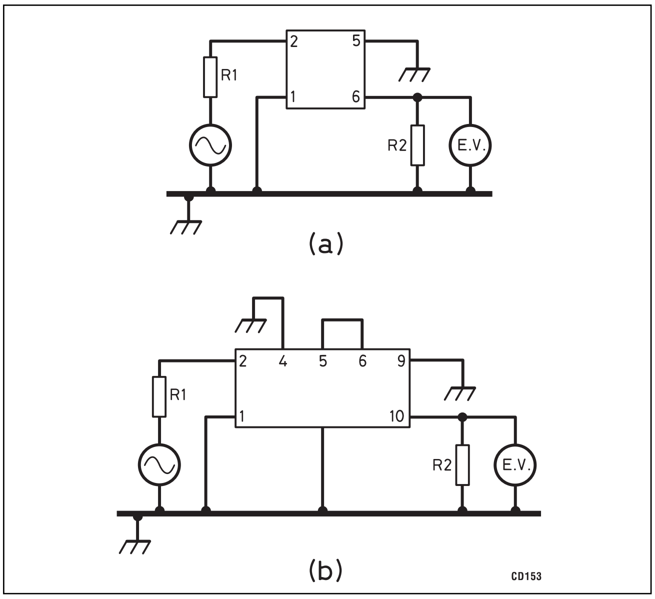


Fig 9.13: Test circuit for (a) Toko HRW and (b) Toko HRQ filters. The case lugs must be grounded. R1 = R2 = 50Ω

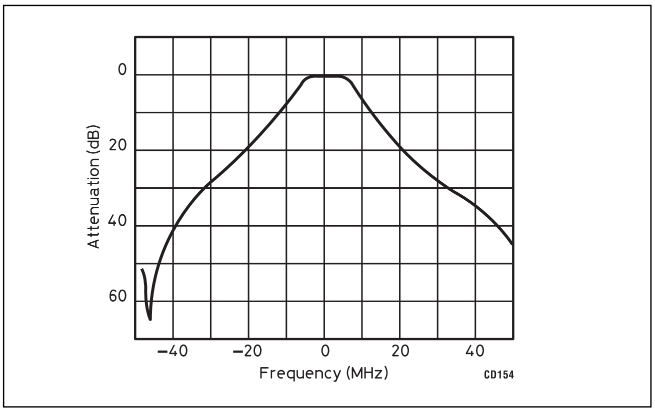


Fig 9.14: The Toko HRW frequency response

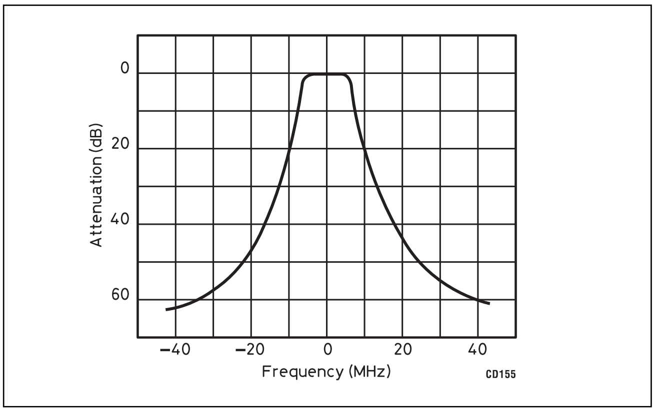


Fig 9.15: The Toko HRQ frequency response

is 50 ohms formed by placing a tap on the helix. This impedance is ideal for matching to antennas and to RF amplifiers and mixers designed using S-parameters. Thus the helical filter replaces conventional tuned circuits in the receiver front-end, resulting in a considerable improvement in selectivity. One note of caution; the screening can lugs must be soldered perfectly to the PCB, otherwise the out-of-band attenuation characteristics of the filters will be degraded. These filters can be used for 2m, 70cm and 23cm receiver front-ends.

Preamplifiers

As available devices improve and new circuit designs are published, it will become apparent that a receiver that may have been considered a first-rate design when built is no longer as good as may be desired. Specifically, a receiver using early types of transistor may not be as sensitive as required, although the local oscillator may perform satisfactorily. The sensitivity of such a receiver can be improved without radical redesign by means of an additional separate RF amplifier, usually referred to as a preamplifier. Such an amplifier should have the lowest possible noise figure and just sufficient gain to ensure that the overall performance is satisfactory.

Fig 9.16 shows the improvement to be expected from a preamplifier given its gain and the noise factor of the preamplifier alone and the main receiver.

An example will suffice to show the application. An existing 145MHz receiver has a measured noise figure of 6dB and is connected to its antenna via a feeder with 3dB loss. It is desired to fit a preamplifier at the masthead; what is the performance required of the preamplifier?

Suppose a BF180 transistor is available: this has a specified maximum noise figure of 2.5dB at 200MHz and will be slightly better than this at 145MHz. The main receiver and feeder can be treated as having an overall noise figure of 3 + 6 = 9dB. From Fig 9.16, if the preamplifier has a gain of 10dB, the overall noise figure will be better than 4.1dB. Increasing the gain of the preamplifier to 15dB will only reduce the overall noise figure to 3.6dB and may lead to difficulty due to the effect of varying temperatures on critical adjustments. The addition of so much gain in front of an existing receiver is also very likely to give rise to intermodulation products from strong local signals. If it is desired to operate under such conditions, it is essential that provision is made for disconnecting the preamplifier when a local station is transmitting.

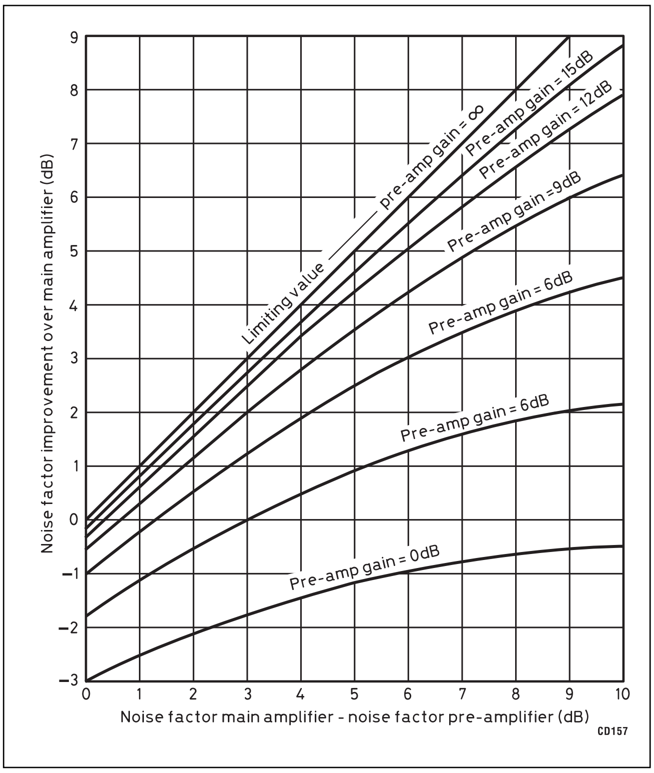


Fig 9.16: Receiver noise figures

RF Amplifier Design

The advent of the field effect transistor (FET) eased the design of VHF RF amplifiers in two ways. The relatively high input resistance at the gate permits reasonably high-Q tuned circuits providing protection against strong out-of-band signals such as from broadcast or vehicle mobile stations. Also the drain current is exactly proportional to the square of the gate voltage; this form of non-linearity gives rise to harmonics (and the FET is a very efficient frequency doubler) but a very low level of intermodulation.

The use of a square-law RF stage is not, however, as straightforward as it first appears. Second-order products are still present, such as the sum of two strong signals from transmitters outside the band, typically Band II broadcasts, mixing to generate the unwanted signals on the 144MHz band. It follows, of course, that if two FET amplifiers are operated in cascade a band-pass filter is required between them to ensure that the distortion products generated in the first stage are not passed to the next stage where they will be re-mixed with the wanted signal.

A development of the FET was the metal oxide semiconductor FET (MOSFET), the gate is insulated by a very thin layer of silica. The gate therefore draws no current and a high input resistance is possible, limited only by the losses in the gate capacitance. These devices may be damaged by static charges and must be protected against antenna pick-up during electrical storms and also by RF from the transmitter feeding through an antenna changeover relay with an excessively high contact capacitance. MOSFETs now have protective diodes incorporated in the device which limit the input voltage to a safe level and these devices are thus much more rugged.

In the dual-gate MOSFET the drain current is controlled by two gates, resulting in various useful circuit improvements. If it is desired to apply gain control to a bipolar transistor stage the control voltage is applied to the same electrode as the signal but the result can be a reduction in signal handling capacity, showing up as intermodulation and/or blocking. The dual-gate FET avoids this problem and automatic or manual gain control can be applied to gate 2 without reducing the signal-handling capability at gate 1. Fig 9.17 shows how such an RF stage for a 2m converter is arranged. When a strong local station causes intermodulation at the mixer or an early stage of the main receiver, the RF gain can be reduced until interference-free reception is again possible. Note that reducing the gain by this method increases the noise figure. For a gain reduction of more than a few dB the noise figure will begin to degrade.

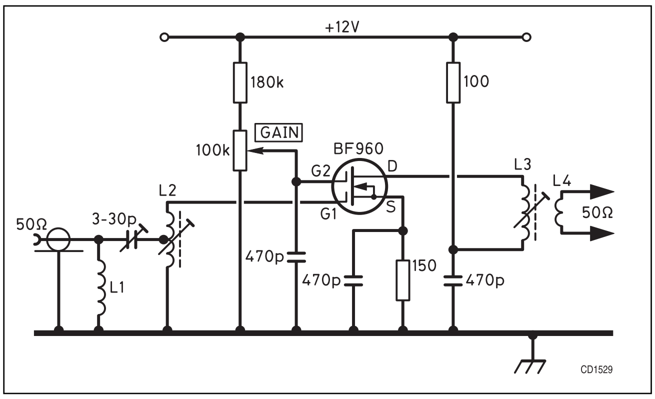


Fig 9.17: Dual-gate MOSFET RF amplifier for 2m with gain control. L1: 3 turns 1.0mm enamelled wire, 6.0mm ID, 8.0mm long. L2: 2 turns 1.0mm tinned copper wire on 8.0mm former 10mm long, tapped at 1½ turns and tuned with dust core. L3: 6 turns 1.0mm enamelled wire on 8mm former tuned with dust core and coupled to L4. L4: 2 turns 1.0mm enamelled wire on same former, close spaced to capacitor end of L3

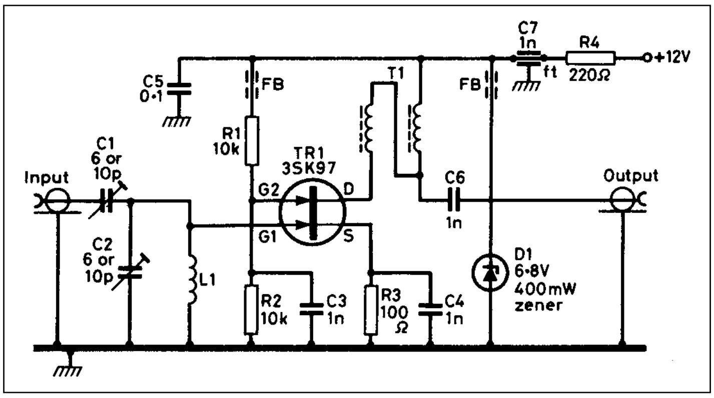


Fig 9.18: GaAsFET preamplifier for 144 or 432MHz

The GaAsFET is similar to the MOSFET but is based on gallium arsenide rather than silicon. Gallium arsenide has larger electron mobility than silicon and therefore has a better performance at UHF. GaAsFETs were originally designed for use in television receiver tuners and are entirely suitable for 144MHz and 432MHz preamplifiers or as a replacement for an existing RF amplifier. These devices are available from major semiconductor manufacturers and include the 3SK97, 3SK112, C3000 and the CF739 (Siemens).

In a correctly designed amplifier the GaAsFET will provide excellent performance at these frequencies. The quoted noise figure for the 3SK97 is 1dB at 900MHz. Silicon diodes are mounted in the GaAsFET chip to protect the gates against ESD (electrostatic discharge) breakdown. The noise figure obtainable is related to the manufacturer's data sheet. Biasing is the same as for silicon MOSFETs, ensuring, however, that the gates are never positively biased with respect to the channel.

The circuit for a GaAsFET preamplifier [3] that incorporates a self-biasing arrangement is shown in Fig 9.18. Construction of L1 and T1 for 144MHz and 432MHz is given below.

L1 144MHz: 6 turns 2.0mm tinned copper wire 6mm ID 13mm long
432MHz: copper line 15mm wide 57mm long spaced 4mm above ground plane

T1 144MHz: 12 turns bifilar wound 0.5mm enam copper wire.
432MHz: 5 turns 0.5mm enam copper wire centre tapped as a 4:1 transformer on an Amidon T-20-12 toroid Core

Double-sided copper laminate is used for mounting most of the components with a vertical screen of tinplate forming a screen between input and output circuits. Care must be taken against static when mounting the GaAsFET in the circuit. After careful checks for constructional errors the current should be checked before alignment. This should be between 25-30mA.

Power gain alignment will be in the order of 26dB at 144MHz and 23dB at 432MHz. Any tendency towards instability can be cured by fitting a ferrite head over the drain lead close to the FET.

An attenuator must be used between the preamplifier and input to an existing receiver or converter to prevent degradation of the strong-signal performance.

Mixers

The most common types of mixers in use today are designed around diodes, bipolar and field effect transistors; integrated circuits are now available for use in mixer applications.

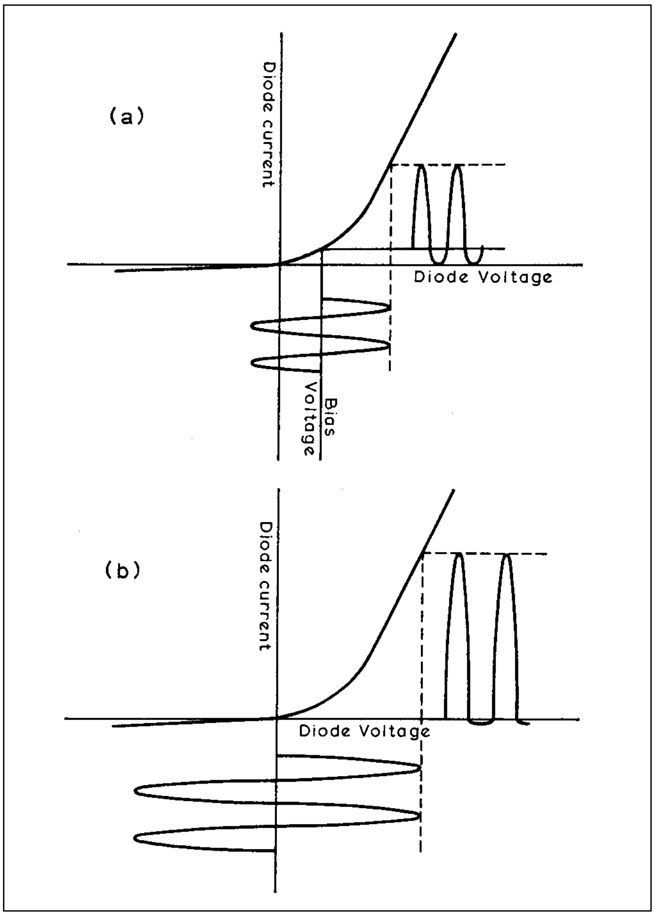


Fig 9.19: Working conditions of diodes mixers. In (a) forward bias is required to provide an optimum point. In (b) the local oscillator is higher and no bias is required

Diode mixers

A diode mixer operates non-linearly, either around the bottom bend of its characteristic curve where the current through the diode is proportional to the square of the applied voltage - see **Fig 9.19(a)**, or by the switching action between forward and reverse conduction as shown in **Fig 9.19(b)**. For optimum working conditions in (a), it is usually necessary to apply DC forward bias to the diode of typically 100 to 200mV; this will vary with the type of diode.

The switching diode mixer (Fig 9.18(b)) is used where a high overload level (strong signals) is required. Signal levels approaching one tenth of the local oscillator power can be handled successfully, and the oscillator level is limited only by the power handling capacity of the diode. The noise generated in the mixer rises with increasing diode current, however, and this sets the limit on the usable overload level if maximum sensitivity is required. The LO power can be adjusted to select a compromise between sensitivity and overload capacity.

Diode mixers display high intercept points and almost all of them are balanced. Conversion loss is an inherent characteristic of diode mixers. In practice this is usually between 3 and 6dB. It is therefore essential that the stage following the mixer, eg an IF amplifier, has the lowest possible noise figure.

Balanced-diode mixers

Noise component transfer from the LO and the mixer to the post-mixer stage can be reduced to a low level by using a balanced two-diode design employing modern low-noise diodes.

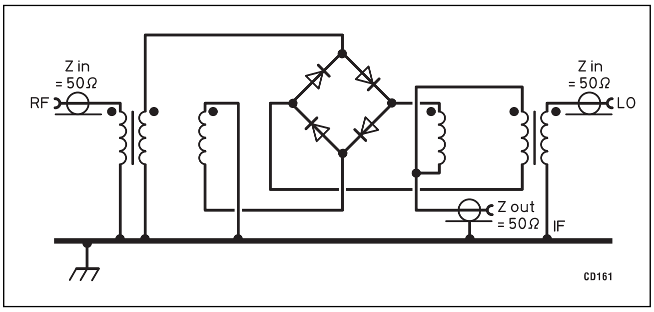


Fig 9.20: Typical diode ring mixer circuit (*ARRL Handbook*)

However with modern diodes it is not usually necessary to provide adjustment of balance at the LO frequency for best noise performance.

Conversion loss is minimised with as little injection as 1mW (0dBm), but IMD (intermodulation) product rejection is improved by increasing the LO drive level to +7 to +10dBm (5mW to 10mW).

Diodes suitable for VHF and UHF receiver mixers are the hot-carrier, high-speed silicon switching types and Schottky barrier diodes. The latter are the only types used in mixers designed specifically to have a high overload capability. A Schottky barrier diode such as the HP 2800 is probably the best choice for mixers operating at these frequencies. The diode ring mixer as shown in **Fig 9.20** is capable of very high performance at VHF and UHF at all the usual signal levels. The transformer cores are normally a low-permeability ferrite with a μ value of typically 125. Toroids are not the best type to use but multi-hole ferrite beads are ideal and make transformer construction simple. For optimum balance the diodes themselves must be dynamically balanced, and for this reason it is easier to purchase a ready-built diode ring mixer, which is usually built into a screened and potted assembly.

Bipolar transistor mixers

A bipolar transistor can be used as a mixer at VHF and will provide a reasonable performance. From the circuit shown in **Fig 9.21** it will be seen that RF input is made to the base and LO injection is made to the emitter. The injection is made at low impedance - the LO coupling capacitor has a low reactance. Bipolar transistor mixers require a fairly low value of LO input power. This reduces the output power requirement of the LO. A typical level is -10dBm. Conversion gain is moderate and correct choice of the transistor will minimise the noise figure.

However, the intrinsic exponential forward-transfer characteristic of the bipolar transistor severely limits the large-signal handling capabilities. Blocking and IMD products become evident even with moderate signal input levels to the mixer. The gain of the preceding RF stage (if used) must be kept low, eg 6dB, to minimise these unwanted products.

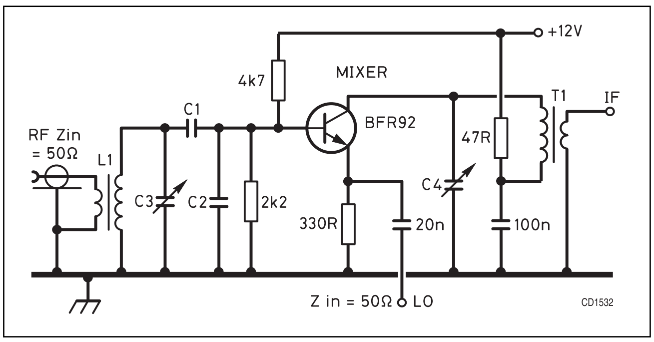


Fig 9.21: Bipolar mixer circuit (*ARRL Handbook*)

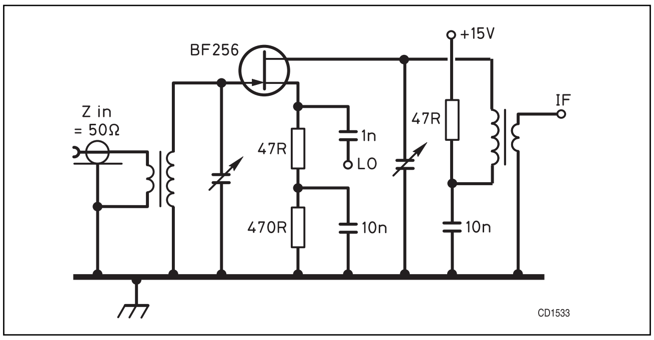


Fig 9.22: A JFET mixer (*ARRL Handbook*)

Junction field effect transistor mixers

Junction FETs can provide very good performance as mixers at VHF. A typical JFET mixer circuit is shown in **Fig 9.22**. The input impedance is high but the conversion gain is only 25% of the gain of the same device used as a VHF amplifier. Bias is critical, and is normally chosen so the gate-source voltage is 50% of the pinch-off voltage of the device. The LO voltage is injected into the source, the level being chosen to avoid the pinch-off region, but not sufficient to cause the source-gate diode to be driven into conduction. Normally the LO peak-to-peak amplitude should be kept a little below the pinch-off voltage.

The JFET has moderate output impedance, typically 10kΩ. This eases impedance matching to the following IF filter, which is usually performed via a step-down transformer as shown in Fig 9.22. The JFET does not have an ideal square-law input characteristic due to the effect of bulk resistance associated with the source. However, the generation of unwanted IMD products under strong signal conditions is much lower than the bipolar mixer, although the noise figure is similar to that of the latter.

Dual-gate MOSFET mixers

MOSFET mixers can provide a superior performance compared with both bipolar and junction field transistors. They have excellent characteristics including a low noise figure, almost perfect square-law forward transfer, together with high input and output impedances. Conversion gain can be high and at the same time there is very low generation of IMD products at the IF output under large signal input levels. Overall performance is extremely good at both VHF and UHF.

Fig 9.23 shows the dual-gate MOSFET in a mixer circuit. The signal input is applied to gate 1 as for an RF amplifier, while the LO is applied to gate 2, the IF output at the drain is controlled by the input levels. Optimum conversion gain is obtained with about 5V peak-to-peak of LO. This mixer has the advantage of having inherent isolation between the signal and LO gate inputs. Next to the ring diode mixer the dual-gate MOSFET mixer has the highest 'overload' level, at the same time giving conversion gain instead of loss.

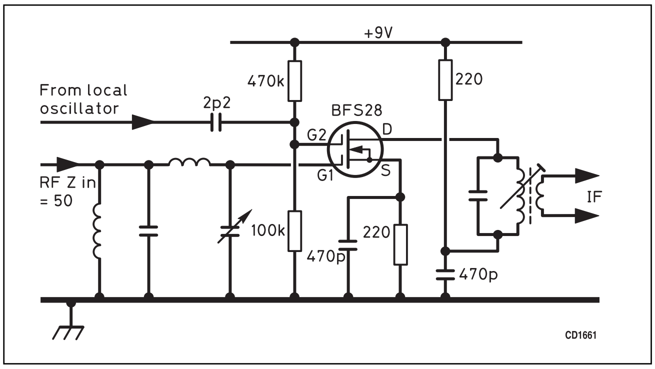


Fig 9.23: A dual-gate MOSFET mixer

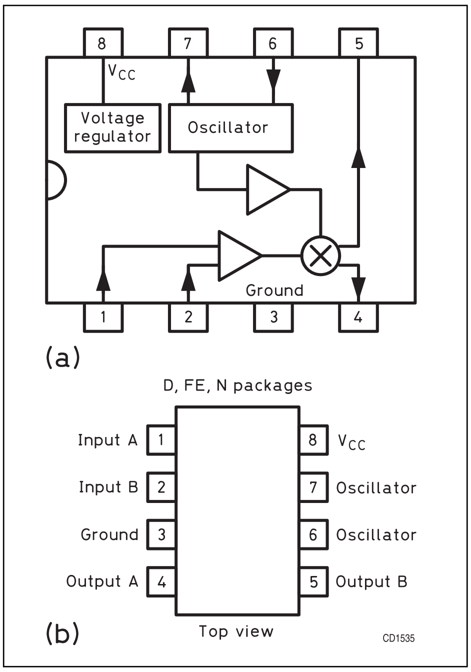


Fig 9.24: NE612 double-balanced mixer and oscillator. (a) Block diagram. (b) Pin configuration (Signetics)

Balanced mixers, whether they use bipolar transistors, junction FETs or MOSFETs, will give a much improved performance where low noise and the largest dynamic range must be achieved at the same time as maximum suppression of unwanted mixer outputs (typically IMD products and LO feedthrough) either to the antenna or to the IF amplifier.

Integrated circuit mixers

Using IC types can obviate the problems of matching transistors and components in balanced mixers. These are available in the form of a monolithic bipolar double-balanced mixer intended for use at VHF and UHF. External circuit layout is simplified, bias adjustment is eliminated and results are more predictable.

The NE612 is an example of this type of mixer. The block diagram is shown in **Fig 9.24** and the equivalent circuit in **Fig 9.25**. Input signal frequencies can be as high as 500MHz. The mixer is a Gilbert cell multiplier configuration which can provide 14dB or more conversion gain. The Gilbert cell is a differential amplifier that drives a balanced switching cell. The differential input stage provides gain and determines the noise figure and signal handling performance. The mixer noise figure at 50MHz is typically <6dB.

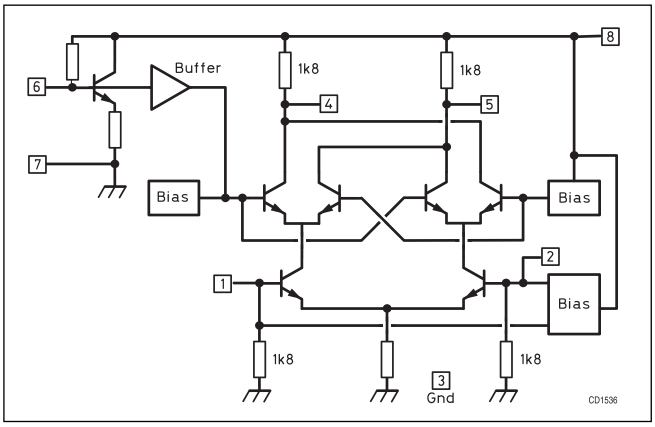


Fig 9.25: Equivalent of NE612 double-balanced mixer and oscillator (Signetics)

The NE612 contains a local oscillator. This can be configured as a CCO or a VFO. The oscillator can also be reconfigured as a buffer amplifier for an external oscillator (the latter is to be preferred). The low power consumption, typically 2.4mA at Vcc = 6V, makes this type of device well suited for use in battery operated receivers.

IF Filters

The performance of a purpose-built double-conversion VHF/UHF receiver with a 'tuneable' first local oscillator can be enhanced by using a crystal filter between the first mixer and IF amplifier. Commercial crystal filters are now readily available, not only for the well-known IFs of 10.7 and 21.4MHz, but also for 45 and 75MHz. Until recently, crystal filters above about 25MHz were only available with third-overtone-mode crystal elements. Now 45 and 75MHz filters are available with fundamental crystal elements. Recommended filters for VHF/UHF bands are:

6m and 4m	10.7MHz
2m	21.4MHz
70cm	45.0MHz

Modern crystal filters are available between 2-poles and 10-poles, the higher the number of poles the greater the attenuation to out of band signals.

Crystal filters need to be correctly matched to the circuitry to attain the specified response. Some types of filters have built-in matching transformers and others do not. Generally types with built-in matching networks have low impedance, typically about 470Ω, whereas types without matching networks have higher impedance, typically 3k3Ω. Often the filter will require a reactive impedance and this is shown by, for example, "3k3Ω//2pF".

The value of the terminating impedance quoted by the manufacturer in the data sheet needs to be achieved for the correct pass-band response, skirt attenuation and minimum insertion loss. Note that the "terminating impedance" is the value the circuitry needs to present to the filter terminals and is the conjugate of the filter impedance seen when you look into the filter with a Vector Network Analyser or other test equipment.

The preferred matching mechanism to the first mixer is the 'constant-R' network which ensures constant filter terminating impedance.

Bandwidth

Crystal filters must be correctly chosen for the type of modulation to be detected by the receiver. The -3dB bandwidth for SSB filters can be as little as 2.1kHz (500Hz for CW receivers). The -60dB bandwidth can be 4.2kHz, giving a shape factor of 2:1.

However, filters for NBFM receivers must have a -3dB bandwidth of ±3.75kHz for 12.5kHz channel spacing and ±7.5kHz for 25kHz channel spacing for linear 'detection' of NBFM transmitters using maximum frequency deviation of the modulator.

Pass-band ripple must be minimal (not more than 1dB) to avoid phase modulation distortion. The -60dB bandwidth for a typical six-pole NBFM crystal filter can be 16-20kHz for 12.5kHz channel spacing and 32-40kHz for 25kHz channel spacing, giving a shape factor of around 2.5:1 All crystal filters have some insertion loss and it is usual to build a post-filter IF amplifier in the receiver to negate this loss. **Fig 9.26** illustrates the difference in SSB and NBFM IF filter characteristics.

RECEPTION OF FM SIGNALS

There are two principal features in receivers designed to receive FM signals, namely limiting rather than linear amplifiers precede the detector and the latter is designed to convert IF variations into AF signals of varying amplitude, dependent on the degree of frequency variation in the transmitter carrier.

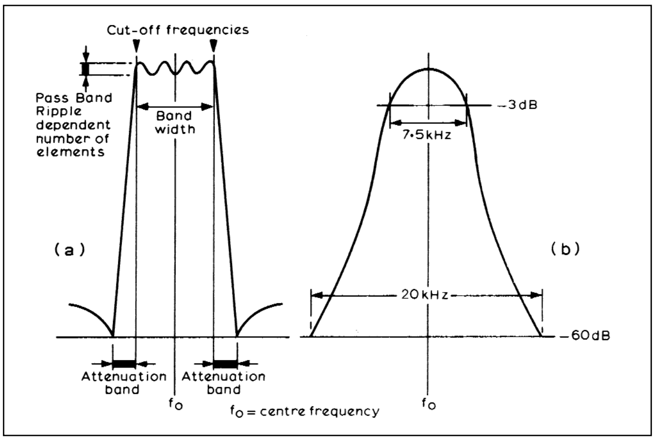


Fig 9.26: Block filter characteristics (a) SSB, (b) NBFM

The FM Receiver

The block diagrams of an FM receiver and AM/SSB receiver are shown in **Figs 9.27 and 9.28**. The principal difference between the receivers are the IF filter bandwidths (see above) and the IF amplifier gains required before the detector.

It is necessary to provide sufficient gain between the antenna and detector of an FM receiver to ensure receiver quieting; ie optimum signal-to-noise ratio with the weakest signal. Usually this is less than 0.35μV PD or -116dBm (into 50 ohms).

Thus it is necessary to use the double superheterodyne principle to achieve the required voltage gain, usually greater than 1 million or 120dB, whilst ensuring optimum stability independent of the input frequency. Other receiver stages, particularly the RF amplifier, mixer, oscillator and audio stages, can be identical to those employed in AM/SSB/CW receivers.

In a multimode receiver designed for reception and detection of all principal methods of modulation, the difference in signal-to-noise ratio and effect of interference is very noticeable between FM and AM/SSB signals. The limiter and detector (discriminator) for FM signals reduce interference effects, usually impulse noise, to a very low level, thus achieving a high signal-to-noise ratio. However, it is necessary to align the detector correctly and in use tune the receiver accurately to achieve noise suppression.

An unusual effect peculiar only to FM receivers, and known as capture effect, occurs when a strong signal appears exactly on the frequency to which the receiver is tuned. If this strong signal has a carrier amplitude more than two to three times that of the wanted signal, the strong signal will be detected. This effect can be a problem in mobile operation, particularly in a geographical area between two repeater outputs on the same frequency.

Weak-signal reception in AM and FM receivers can be degraded by a much stronger carrier on or near the frequency of the weak carrier.

Selectivity

As already stated in the previous section on filters, it is essential to choose the IF filters designed for NBFM reception. A narrow-bandwidth filter will introduce unwanted harmonic distortion. Too wide a bandwidth will degrade adjacent channel selectivity. However, transmitters exceeding the recommended limit on frequency deviation, will aggravate distortion effects and may cause 'squelch-blocking' where a squelch momentarily closes on peaks of the deviation. With modern NBFM transmitters this

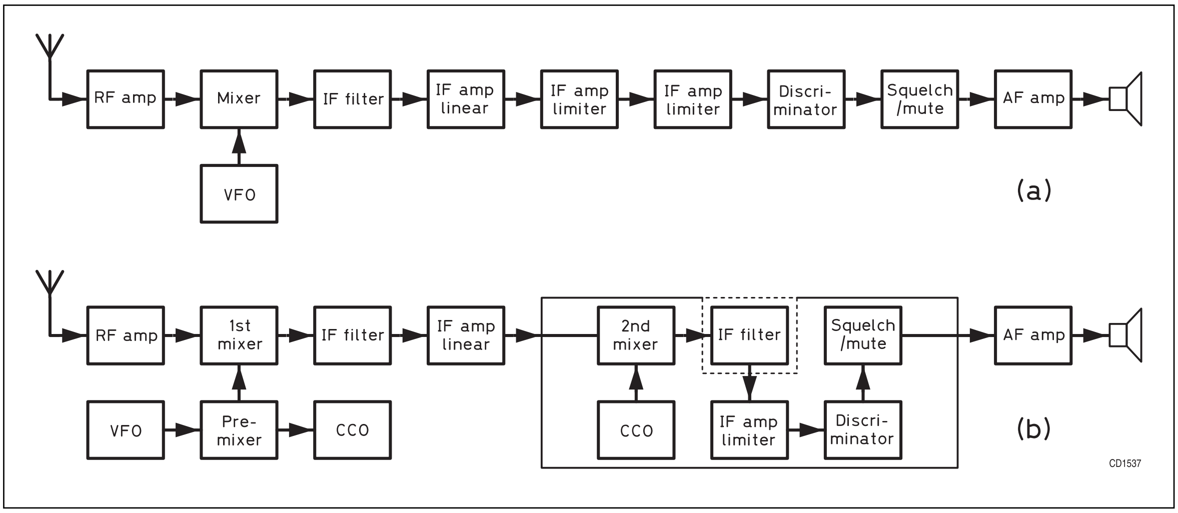


Fig 9.27: Comparison of the essential stages of receivers for (a) NBFM with discrete circuits and (b) NBFM with integrated circuits

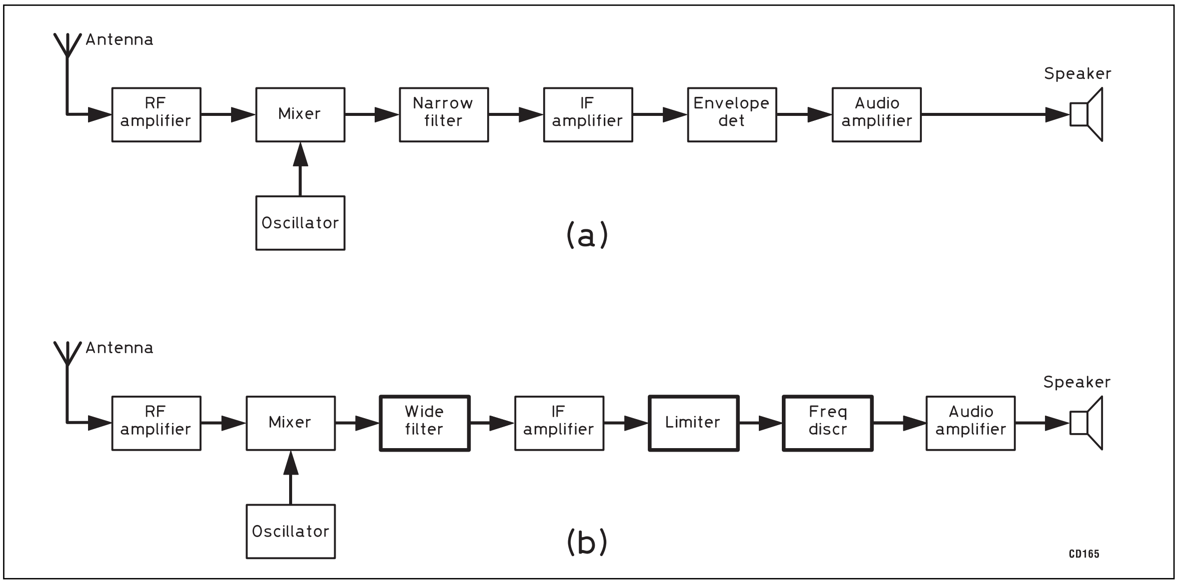


Fig 9.28: Block diagrams of (a) an AM and (b) an FM receiver. Dark borders outline the sections that are different from the AM set (*ARRL Handbook*)

is less of a problem. Poor adjacent channel selectivity can cause receiver desensitisation, particularly when strong signals are present on either or both adjacent channels.

A transmitter with poor adjacent channel power rejection can also degrade weak-signal reception. These effects again cause receiver desensitisation but the transmitter modulation will not appear on the wanted signal (cross-modulation cannot in theory occur in FM systems).

Limiters

Limiting IF amplifiers are specifically designed to introduce gain compression into their forward-transfer characteristics. If an amplifier is driven into limiting, its output signal level remains unchanged as the input signal level is varied. This effectively removes any sudden amplitude change, which is important as it is necessary to remove any impulse noise and AM on the carrier prior to an FM detector.

Fig 9.29 shows the difference between a linear and a limiting IF amplifier output waveform. The clipping action removes the AM component. The overall amplifier gain must be high enough to ensure the limiting stages are limiting even with weak signals or with large changes of signal level IF to the receiver input. With an IF input of typically 5.0μV (equivalent to 0.25μV RF input to a receiver front-end with a conversion gain of 26dB) to the IF amplifier input, a minimum of three stages are required to raise the level of the signal for limiting action to commence. As the IF carrier level increases above 5.0μV the limiting action starts.

Now the signal-to-noise ratio improves until at a certain level the noise disappears. This is known as the receiver quieting characteristic referred to earlier in this section, usually the input for 20dB signal-to-noise ratio.

Discrete limiting amplifier(s) preceded by linear amplifier(s) with interstage IF transformers are still to be found in some early designs but are not now employed in modern NBFM receivers. Examples are shown in **Figs 9.30 and 9.31**. Linear amplifiers precede the limiting amplifiers. The base bias on the final IF amplifier in Fig 9.30 is varied to set the required limiting input level. In Fig 9.31 the base bias on the two final amplifiers is varied. This

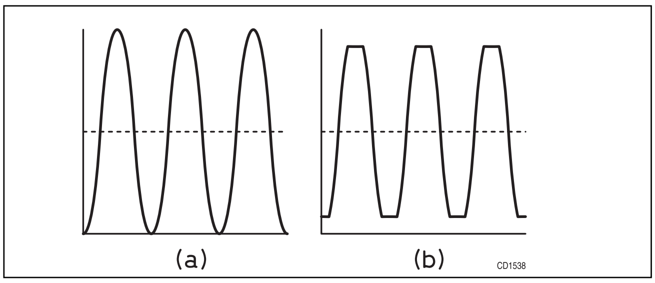


Fig 9.29: (a) Linear amplifier output waveform. (b) Limiting IF amplifier 'clipped' output waveform

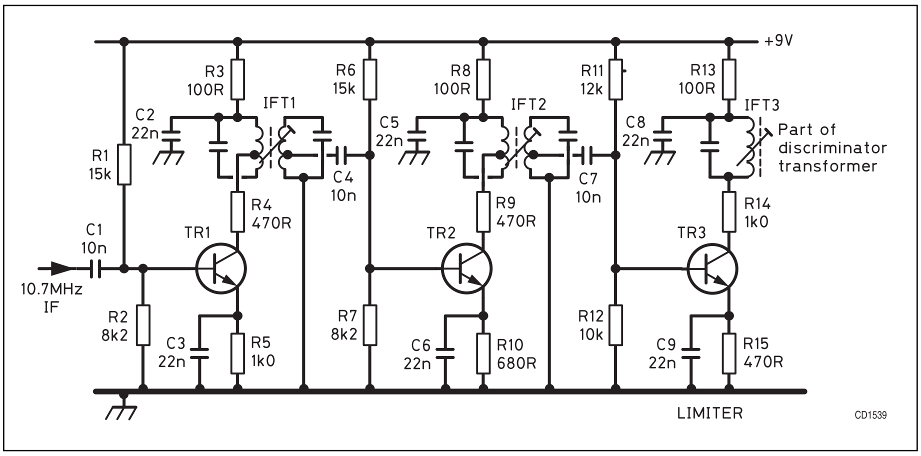


Fig 9.30: A three-stage amplifier and limiter. Transistors are BF194 or similar

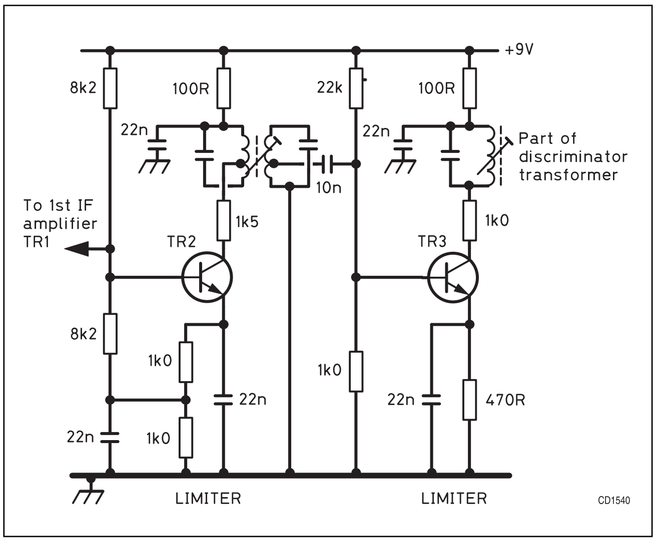


Fig 9.31: A two-stage limiter developed from the circuit of Fig 9.30

sets the limiting knee characteristic of the transistors to a point at which, for an increasing input, there will be no further increase in collector current. The amplifiers saturate, giving the required limiting and consequent noise level reduction for good receiver quieting. The circuit in Fig 9.31 gives an improved limiting performance compared to the circuit in Fig 9.30.

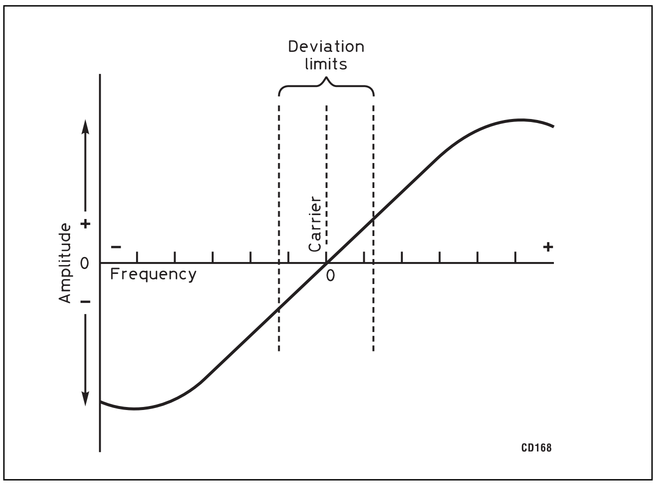


Fig 9.32: The characteristic of an FM discriminator

Some FM receiver manufacturers incorporated an IC limiting amplifier containing two or more stages, as they gave superior limiting action compared to discrete designs such as the MC1590 and the CA3028A. These ICs became obsolete with the introduction of devices containing six or more differential DC coupled IF amplifiers with improved and consistent limiting characteristics, operating at a relatively low IF, typically 455kHz. This system removed the requirement for IF transformers between each stage, but caused layout problems. However, it considerably simplified alignment. These ICs are called NBFM IF subsystems, and will be described after the section on FM demodulators.

FM demodulators

The FM detector, or more correctly the FM discriminator, was evolved to be able to respond only to changes in frequency as received from an FM transmitter and not to amplitude variations as received from an AM transmitter when the carriers are modulated. The degree of frequency change (frequency deviation) corresponds to the amplitude of the audio signal applied to the transmitter modulator. For example, if the transmitted carrier is deviated by ±2.5kHz by the modulator, the received carrier will be changing in frequency by ±2.5kHz.

This frequency 'swing' is detected by an FM discriminator. **Fig 9.32** shows clearly the 'S' curve characteristic of the discriminator. Provided the swing is in the linear (straight) portion of the curve, ie about ±6kHz, very little distortion will be present in the recovered audio signal.

Maximum frequency deviation in amateur FM equipment is limited to ±5kHz compared to ±75kHz for Band 2 broadcast transmitters and receivers. This explains why FM voice and data communications systems are normally known as narrow-band FM systems.

A practical discriminator, known as the Foster-Seeley discriminator, after its inventors, is shown in **Fig 9.33**. T1 is the discriminator transformer. Voltage, due to the IF carrier, is developed across the primary of T1. Primary-to-secondary inductive coupling induces a current in the secondary 90° out of phase with the primary current. The IF carrier is also coupled to the secondary centre tap via a coupling capacitor.

The voltages on the secondary are combined in such a way that these lead and lag the primary voltage by equal amounts (degrees) when an unmodulated carrier is present. The resultant rectified voltages are of equal and opposite polarity.

When the received carrier is deviated the phase is changed between primary and secondary, resulting in an increased output

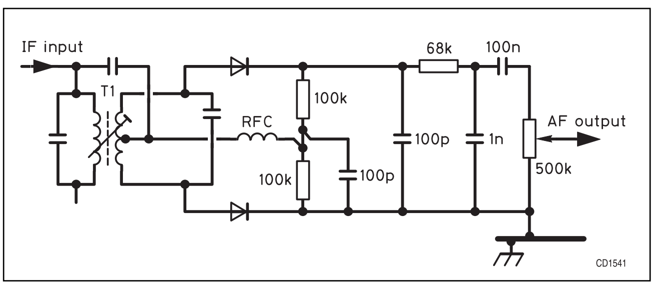


Fig 9.33: The Foster-Seeley discriminator

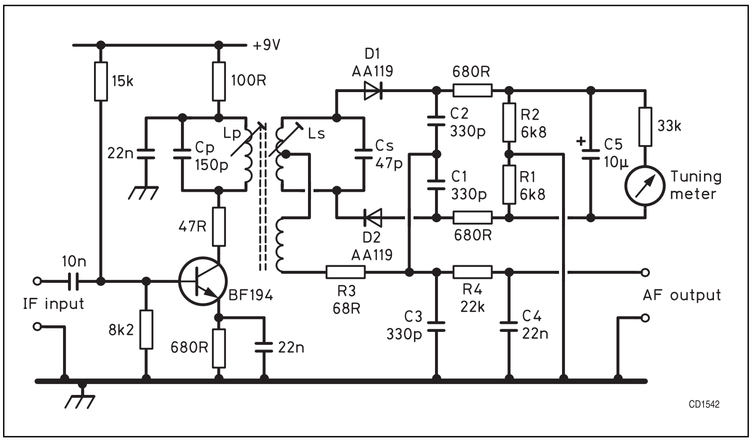


Fig 3.34: A practical ratio detector circuit

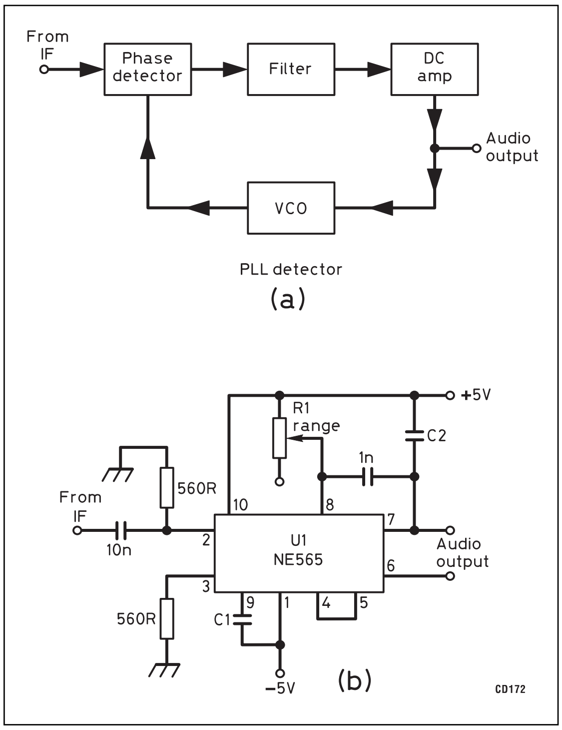


Fig 9.35: (a) Block diagram of a PLL demodulator; (b) Complete PLL circuit

level on one side and a decreased output on the other side. These voltage-level differences, after rectification by D1 and D2, represent recovered audio. This discriminator responds to carrier amplitude variations (AM) unless driven hard by the limiting IF amplifier stages and, for this reason, it fell from favour.

Ratio detectors

This type of FM detector was developed from the frequency discriminator and became very popular in broadcast FM receivers - it has also been widely used in amateur receivers. The ratio detector is far less susceptible to carrier amplitude variations, hence less limiting is required before it. The detector is shown in **Fig 9.34**. T1 is a ratio detector transformer comprising of a primary, centre-tapped secondary and a tertiary winding, tightly coupled to the primary.

As its name implies, it functions by dividing the rectified DC voltages from D1 and D2 appearing across R1 and R2 into a ratio equal to the amplitude ratio present on each side of transformer secondary (L_S). The DC voltage sum appears across the electrolytic capacitor C5. This has sufficient capacitance to maintain this voltage at a constant level during fluctuations in carrier levels, caused for example by AM or noise signals being present on the carrier. Hence the ratio detector has its own inherent limiting characteristic. With a detector that responds only to ratios, the strength of the IF carrier can vary considerably without causing the output level to change. Therefore only FM can be detected and not AM. The carrier level should not fall below a level to cause D1 and D2 to become partially or fully non-conductive.

When the carrier is deviated by FM the audio signal is recovered from the tertiary winding L1, The IF carrier is filtered out by R3 and C3. The diode load resistors are lower in value than for the discriminator. It will be noted the diodes are connected in series rather than series-opposing as in the discriminator. This makes the ratio detector 6dB less sensitive. The diodes such as the gold-bonded AA 119 should preferably be matched dynamically.

Phase-lock-loop detectors (PLL)

With the advent of the single-chip PLL it has been possible to design a reliable NBFM detector without tuned circuits and, therefore, without the necessity for alignment.

An example of an IC PLL detector is shown in **Fig 9.35**. The block diagram is shown in **Fig 9.35(a)** and the circuit including external components in **Fig 9.35(b)**.

Referring to Fig 9.35(a), the VCO oscillates close to the carrier frequency, in most cases 455kHz. The phase detector produces an error voltage when the VCO frequency and the carrier frequency are not identical. This error voltage is a DC voltage that is amplified after filtering, and 'corrects' the VCO frequency. When the carrier is deviated by audio modulation the frequency change is sensed by the phase detector and the resultant error voltage corrects the VCO frequency, causing it to remain locked to the carrier frequency, The system bandwidth is controlled by the loop filter. As the error voltage corresponds exactly to the frequency deviation, the PLL circuit functions as a precise FM detector.

It has a high sensitivity, requiring typically a 1mV carrier level for the PLL circuit to function correctly. Referring to Fig 9.35(b), R1 and C1 set the VCO frequency close to the carrier frequency. C2 controls the loop filter bandwidth that in turn controls the PLL capture range. The capture range is the maximum deviation from the carrier frequency to which the loop will gain and maintain lock on the carrier frequency.

Because the VCO runs at the same frequency as the IF and is pulled into frequency and phase lock with the signal to be demodulated, any signal leakage from the VCO into the earlier IF stages will cause the system to self limit or 'jam' the VCO signal. This is often detected as a low amplitude 'whistle' or beat note which varies in frequency when the input signal is absent. For this reason the PLL Detector has largely fallen out of favour and today the preferred type is the Quadrature Detector.

Quadrature detectors

This is also known as the quadrature discriminator (sometimes as the coincidence detector). The symbolic circuit is shown in **Fig 9.36**. It is found in virtually all NBFM subsystem ICs. Alignment is simple, there being only one coil to align for maximum audio output.

The IF input is fed to the detector, both directly and via a 90° (quadrature) phase shift. The quadrature input is fed to the detector using an appropriate value capacitor (C2) and a phase-shift network C1 and L which resonates at the IF centre frequency. The detector itself is an analogue multiplier. The phase-shifted signal is multiplied by the IF signal deviation from centre frequency (when modulated with an audio signal). The direct signal is multiplied with the phase-shifted signal to recover the

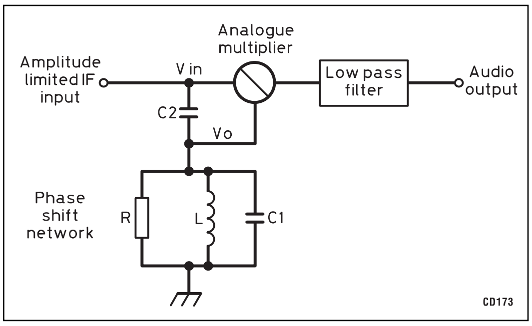


Fig 9.36: Symbolic circuit of quadrature detector

audio signal, via the low-pass filter from the multiplex output spectrum.

For small frequency deviations (as in NBFM systems) the phase shift, controlled by the quadrature network, is sufficiently linear to give acceptable audio quality, ie with very little distortion.

The working Q of L in the network can be controlled by shunting it with a resistor (R). The lower the resistor value, the better the linearity, as this will increase the peak deviation capability of the detector. However, the audio recovery level is reduced.

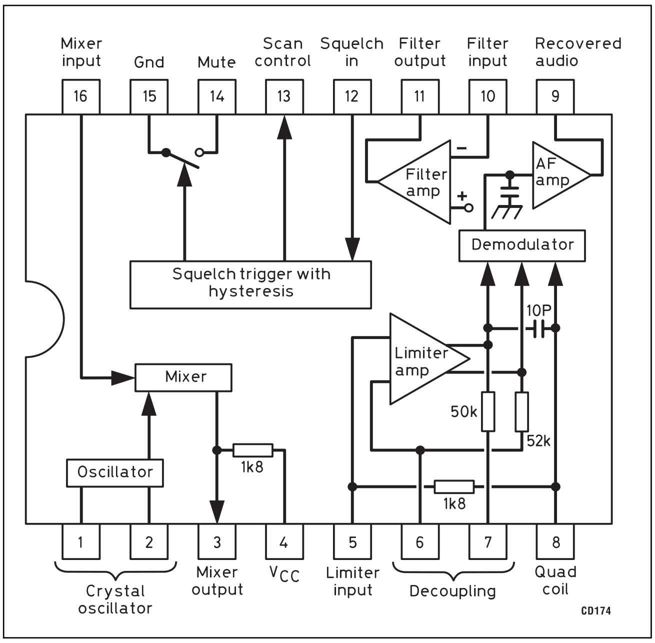


Fig 9.37: Internal block diagram of the MC3361 NBFM IC

De-emphasis

It is normal practice to insert a de-emphasis network, usually a resistor and capacitor combination, in the post-detector section of the receiver, irrespective of the type of detector employed, to attenuate noise and audio frequencies above 3kHz by 6dB/octave.

NBFM IF subsystems

These are used in modern dual-conversion receivers. The IC integrates the limiting amplifiers with a second mixer and oscillator for converting to a low IF, typically 455kHz, a quadrature detector, an active filter for driving a squelch circuit and a post-detector AF preamplifier in a single monolithic block. This results in considerable space saving, particularly for hand-held and mobile equipment. Power consumption is low and setting up and alignment is simple. Early examples of this type of IC are the Motorola MC3357 and the MC3359; these ICs can be found in many professional and amateur NBFM receivers.

More recently improved performance versions have appeared, eg the MC3361 and MC3362. The MC3361 is shown in block diagram form in **Fig 9.37** and in a practical circuit in **Fig 9.38**.

The mixer is balanced to reduce spurious radiation. It converts the first IF input signal to the second IF of 455kHz. After passing through an external band-pass filter, normally a multi-pole ceramic type, the IF signal is fed to the five-stage limiting amplifier and then to the quadrature detector where the audio signal is recovered. The 10pF on-chip capacitor produces the 90° phase shift between one output port of the final limiting amplifier and one input port of the quadrature detector. The quadrature coil is shunted by parallel resistor R2. This controls the linearity of the detector swing from centre frequency, hence the harmonic distortion in the recovered AF output which is buffered by the internal amplifier. R3 and C7 form the de-emphasis network. R4, R5 and C9 apply attenuated output to the filter amplifier. C10 and R6 peak the filter response to approximately l0kHz, ie above the normal audio pass-band of 300Hz to 3kHz.

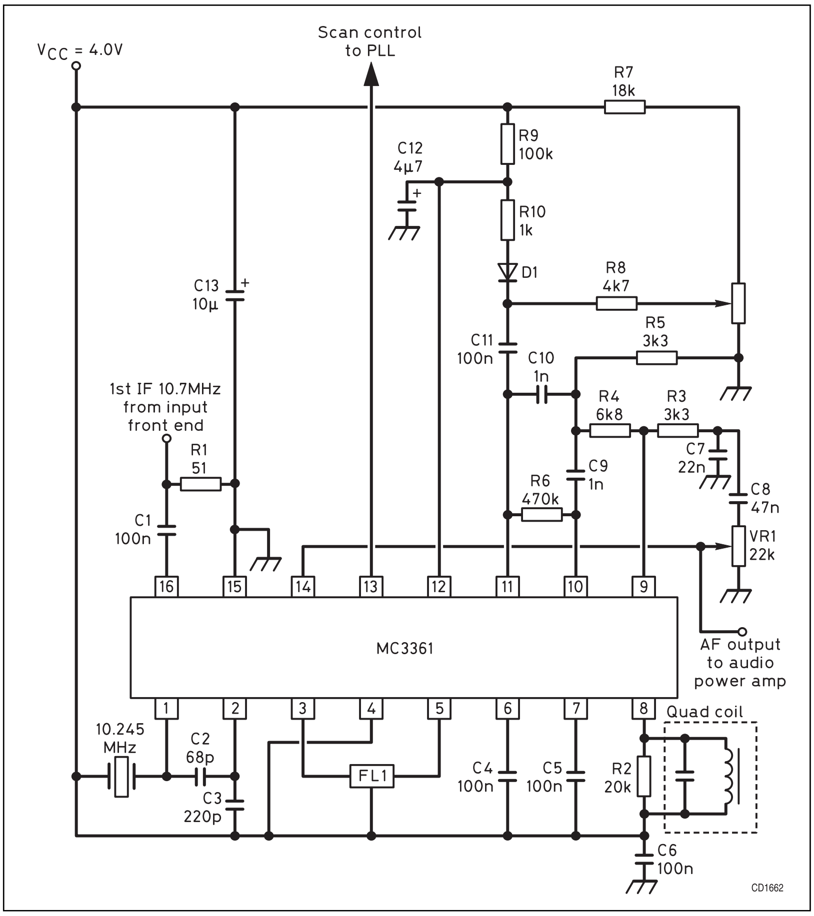
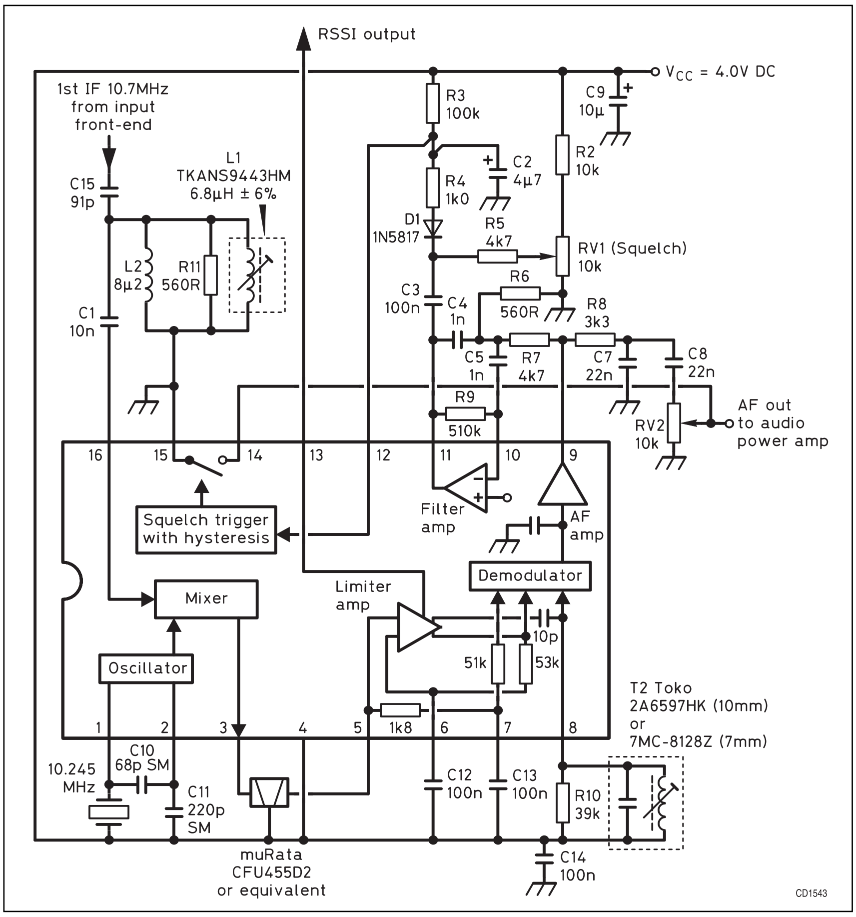


Fig 9.38: Practical circuit using the MC3361 (Motorola)

In the absence of a carrier, only the noise signal is amplified this is detected by D1, which conducts, causing the DC voltage at the junction of' R9/R10 to fall. This level change is applied to the squelch switch input. The internal switch that stops noise

Fig 9.39: Typical application for the MC3371 at 10.7MHz with a parallel LC discriminator (Motorola)



output appearing across RV1 grounds the mute output pin. R10 and C12 filter the DC output from D1 anode. R7 and R8, with RV2, the squelch control, set the initial biasing of D1 and hence the squelch threshold level. When a carrier is detected and audio signals are recovered, D1 will not conduct and the squelch mute switch will remain open, removing the muting of the audio signal across RVI. The scan control at pin 13 can be used in conjunction with a digital PLL tuning system for 'locking' onto channels where the receiver has scanning facilities.

Fig 9.38 shows the external components required to complete the circuit of a practical NBFM receiver IF system. An IF block filter, as described earlier, must be inserted between the 1st mixer output and the IC input (2nd mixer) to provide adequate adjacent selectivity. The oscillator is an internally biased Colpitts type. The circuit shows a 10.245MHz fundamental-mode crystal oscillator for converting the 1st IF of 10.7MHz into the 2nd IF of 455kHz. However, a 20.945MHz crystal can be used for a 21.4MHz IF input and a 44.454MHz third overtone crystal for a 45MHz IF input. C1 and C2 form the crystal load capacitance.

For 10.7MHz to 455kHz mixers the alternative crystal is 11.155MHz and this will be used if the harmonics of the 10.245MHz crystal fall closer than 100kHz to the input signal. For example, 102.45MHz as an input signal would be blocked by the harmonics from the 10.245MHz crystal. Changing the second conversion crystal obviates this problem. The other IFs have similar alternative crystals, 21.855MHz and 45.455MHz are commonly available from crystal manufacturers as standard items.

The doubly balanced mixer has a high input impedance of about 3kΩ. This characteristic enables crystal filters to be matched easily to the mixer input. Similarly the output impedance is fixed by the internal 1.8kΩ resistor for correct matching to the 455kHz ceramic bandpass filter (FL1). The filter -3dB bandwidth should be 7.5kHz for 12.5kHz and 15kHz for 25kHz channel spacing respectively. Ultimate adjacent-channel selectivity is a function of the filter stop-band attenuation. A six-pole filter will provide sufficient attenuation. Measured at ±100kHz the attenuation should be 40dB minimum: this is adequate for NBFM receivers. The filter is matched at the IF input by an internal 1.8kΩ resistor.

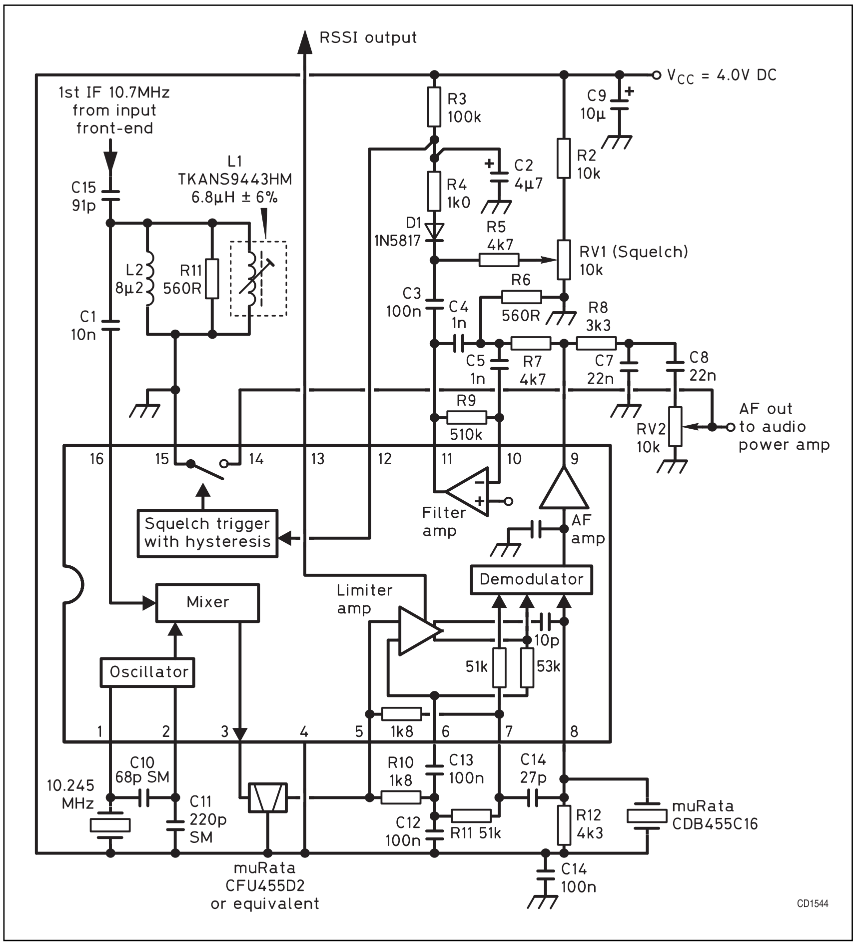


Fig 9.40: Typical application for the MC3372 at 10.7MHz with a ceramic discriminator (Motorola)

The Motorola MC3371 and MC3372 represent the two low-power NBFM sub-system ICs. They are basically very similar to their predecessors, the MC3361 and MC3362. The principal difference between the MC3371 an MC3372 is in the limiter and quadrature circuits. The MC3371 has internal components connecting the final limiter to the quadrature detector for use with parallel LC discriminator. In the MC3372 these components are omitted and must be added externally The MC3372 can be used with a ceramic discriminator. Application circuits are shown in **Figs 9.39 and 9.40**.

Both ICs have a meter output at pin 13 which indicates the strength of the IF level and the output current is proportional to the logarithm of the IF input signal. A maximum of 60µA is available to drive an S-meter and to detect the presence of an IF carrier. This feature is known as a received signal strength indicator (RSSI) or S-meter. Pin 13 is resistively terminated (to ground) to provide a DC voltage proportional to the IF signal level. The resistor value is estimated by V_{cc} - 1.0V/60µA, so for V_{cc} = 4.0V a 50kΩ resistor will provide a maximum swing of 3.0V.

PREAMPLIFIERS

Modern semiconductors have made it possible for amateurs to build high quality preamplifiers to improve the performance of their existing transceivers or transverters for the VHF and UHF bands. The examples included in this chapter have been designed by amateurs to meet the specific needs of operation on the VHF and UHF bands. The three designs for the lower bands are from Dragoslav Dodricic, YU1AW [4] and show that bipolar transistors rather than FET transistors still have their place. State of the art commercial preamplifiers are reviewed by Sam Jewell, G4DDK to show what can be obtained off the shelf. The design for 23cm comes from Sam Jewell, G4DDK and has a novel two stage design where the second stage can be bypassed to improve performance under local strong signal conditions.

Preamplifiers for 6m, 4m and 2m

A new type of low noise preamplifier is described here, which is recommended for its exceptional noise and inter-modulation characteristics not only for normal DX operation, but also for operation under difficult conditions when there are a significant number of powerful local stations, for example during competitions. The amplifiers are designed to have low noise, unconditional stability and exceptional linearity, thanks to the use of special ultra-linear, bipolar, low noise transistors designed for TV signals amplifiers. Since they are widely used, they are readily available at low cost. Construction is extremely simple with a small number of components, very simple adjustments and a high repeatability. This has been achieved using extensive computer non-linear and statistical optimisation. Designs are available for all amateur bands from 6m to 23cm [4]; watch that web site for latest updates to these designs.

For three decades, MOSFET or GaAsFET transistors have been used almost exclusively in preamplifier designs. The reason for this is their superior noise performance and amplification. What we inevitably encounter when using GaAsFETs is a stability problem due to their conditional stability on VHF and UHF frequencies [5 - 8]. However, with an increasing number of stations using greater output powers, especially during competitions, the majority of these low noise preamplifiers that are successfully used for DX, MS or EME activity, become overloaded. This is manifested by a large number of inter-modulation products that contaminate the band, this is attributed to other stations and especially those that use powerful amplifiers.

The problem, of course, could be in non-linear power amplifiers due to excess input power causing saturation. This generates a high level of inter-modulation products. However, in practice it is more frequently due to the receiver's excessively high amplification and insufficient linearity, ie its input stage is overloaded causing it to generate products that look as though they really exist on the band.

In order to understand how to cure this problem, it is necessary to know how, where and under what conditions it occurs. It turns out that the source of this problem is very high amplification, a feature that the majority of amateurs praise the most and should be praised the least or even avoided. Technically it is much more difficult to achieve the two other important properties of an amplifier: low noise factor and strong signal performance, ie linearity. These are the most important properties for an amplifier to those for whom decibels are not just numbers that cover ignorance.

How do we determine which amplifier is of good quality? In order to resolve this dilemma a measure of an amplifier's quality has been introduced which encompasses all of an amplifier's three characteristics: noise factor, amplification and the output level of a signal for a determined level of non-linear distortions. This measure of quality is called the dynamic range of an amplifier and represents a range in which the level of a signal on an amplifier's input can be changed, while the output signal degradation stays within defined limits. The lower limit of this range is determined by the minimum allowable signal/noise ratio of the output signal and it is directly determined by the amplifier's noise factor, and the upper limit is the allowable level of non-linear distortion.

The lower limit of a dynamic range is the level of the input signal that gives a previously determined minimal signal/noise ratio (S/N) at the output. If the lower limit value is a S/N = 0 (incoming signal and following noise are equal) and if the upper limit of this range is limited by the maximum output signal voltage at which the amplifier, due to non-linear distortion, generates products equal to the level of noise on the output of the amplifier, then this is the so-called SFDR (spurious free dynamic range) or a dynamic range free from distortion, ie products of intermodulation distortion or IMD.

Since the third order inter-modulation distortion (IMD3) is dependant on the cube of the input signal, that is with each increase or decrease of the input signal by 1dB, the third order inter-modulation products increase or decrease by 3dB. It is therefore possible to calculate the maximum output level for different values of relations between products and the signal that is being used, or the value of IMD3 products, at different output signal levels. Using an attenuator enables us to also check whether an amplifier is overloaded, ie to recognise whether an audible signal on our receiver really exists on the band or whether it is simply the 'imagination' of our overloaded receiver. This enables us to dispose of overload and IMD3.

Since the level of products rise faster than the basic signal, by increasing the input signal we reach the point at which third order inter-modulation products, IMD3, reach the level of a useful signal at the output and that point is known as IP3 (Intercept Point). When the IP3 value is quoted it is necessary to state if it is referenced to the input or output of the amplifier. These values naturally differ by the value of the amplifier's amplification. Occasionally, it is stated as the TOI (third order intercept). This point is often taken as a measure of an amplifier's linearity and is highly convenient when comparing different amplifiers. Knowing the value of an amplifier's IP3 enables us to precisely calculate the value of IMD3 products at some arbitrarily chosen output or input signal level.

If excessive amplification is used, for example in a multi stage amplifier, a danger exists where antenna noise and the noise of the first amplifier are amplified to such an extent that they exceed the limit of linear operation of the last transistor, at which point the amplifier is saturated with the noise itself without any signal.

The conclusion is clear: An amplifier is worth as much as its dynamic range value, rather than how great its amplification is!

Therefore, if we want to construct an amplifier with the maximum amount of SFDR we have to fulfil the following conditions:

- the noise factor is as low as possible
- the IP3 is as high as possible
- it has acceptable amplification

On the one hand, amplification should be as large as possible, to prevent second degree influence on noise factor, and on the other hand it should be as small as possible so that the IP3 input is as high as possible, ie so that the amplifier should withstand the highest possible input signals without distortion. Compromise is essential and it usually ranges between 13-20dB amplification, depending on which parameter is more important for us.

If we want a low noise amplifier with a high dynamic range, then the choice of a corresponding transistor is extremely important. It is necessary to choose the type of transistor that besides low noise and sufficient amplification on the given frequency fulfils the condition of good linearity, that is high IP3 along with unconditional stability. Hitherto, MOSFET and GaAsFET transistors did not fulfil this condition in a satisfactory manner. Specially built transistors for ultra linear working, primarily for CATV do fulfil these criteria. For that reason, Siemens BFP196 bipolar transistors in SMD packaging were chosen. The Philips transistor BFG540/X corresponds closely to the Siemens device, it requires only slightly different base bias resistors. This Philips transistor should be used on 1296MHz because it gives several decibels greater amplification. It should be stressed that BFG540 without /X could be used, but the layout of pins is different, - it is not pin-to-pin compatible with the BFP196 - therefore the printed circuit board has to be changed, which is not recommended.

Since we are talking about a broadband transistor whose Znf and S11 values are relatively close to 50 ohms, the input circuit has been chosen to optimally match the transistor with regards to noise, while at the same time it provides some selectivity at the input. By varying the circuit values a compromise is found which provides the highest selectivity with minimal degradation of the noise factor. On lower bands where the noise factor is not as important, the compromise was in favour of selectivity which is more important than noise on these bands. The operating point of the transistor was also chosen as a compromise between minimal noise and maximum IP3. The output circuit is relatively broadband and it is implemented using a printed inductor to reduce coupling with the input and to provide high repeatability. In order to maintain optimal output matching that gives minimal IMD, any matching by trimmer capacitors or by variable inductances is forbidden on the output. In order to achieve unconditional stability, minimal IMD, optimal amplification and minimal noise, negative feedback is applied which cannot be changed arbitrarily.

The printed circuit board is made with the dimensions shown in the relevant figure (**Fig 9.41** for 6m, **Fig 9.42** for 4m and **Fig 9.43** for 2m - all located in the Appendix B). Double sided board, type G10 or FR4 is suitable. The bottom copper surface is an unetched ground plane.

SMD components are the 1206 type and the ground connections are made using through plated holes or with wire links through the holes soldered on both sides. The parallel resistor and capacitor in the base bias circuit are soldered on top of each other and not next to each other. The transistor collector is connected to the wider track.

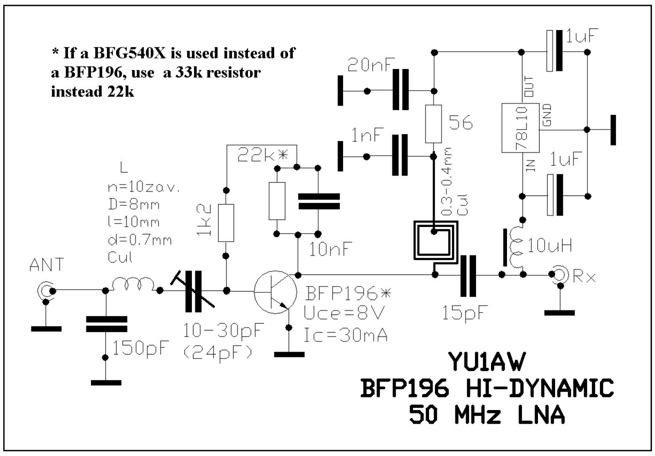


Fig 9.44: Circuit diagram for the 6m low noise amplifier

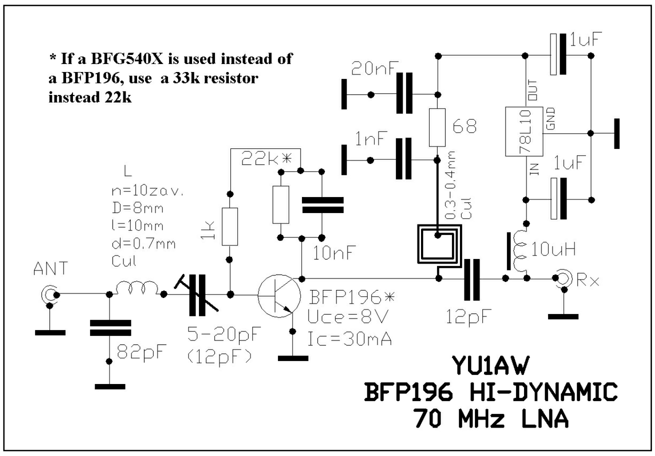


Fig 9.45: Circuit diagram for the 4m low noise preamplifier

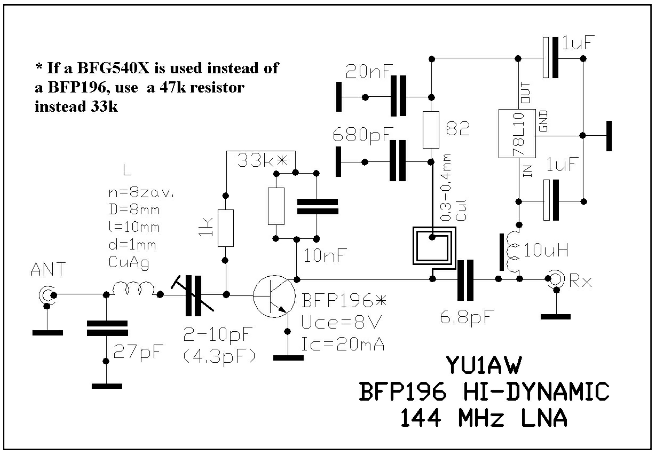


Fig 9.46: Circuit diagram for the 2m low noise amplifier

The trimmer used is either of the air or PTFE foil type, although a ceramic one can also be used if it has a suitable capacity range. It is especially important for the higher band amplifiers that the trimmer capacitor has a low enough minimum capacity.

The coil is wound, as shown in the relevant circuit diagram (**Fig 9.44** for 6m, **Fig 9.45** for 4m and **Fig 9.46** for 2m), with silver plated copper wire, thickness ‘d’ and ‘n’ turns with a body diameter ‘D’. The coil is to be expanded to length ‘l’. When the coil is fitted it has to be positioned so that the bottom is approximately 3mm above the printed circuit board.

The box for the amplifier is made so that the printed board is the bottom side of the box, as can be seen in Fig 9.47. The easiest way to do this is to solder 25-30mm wide copper or brass strip, 0.3mm thick, around the edges of the printed board. Connections are mounted onto the box created, and a lid is made out of the same kind of sheet metal.

Once everything is carefully soldered, check for any possible mistakes such as short circuits, then connect the DC supply voltage and measure the collector current and voltage. If everything is correct and properly connected and the transistor is functioning properly, the values should be close to the ones given in the circuit diagram. If the differences are within 10%, everything is OK. If the differences are greater, check the supply voltage and then reduce the value of the base bias resistor, which is in parallel with the capacitor. Make it lower to raise the collector current and vice-versa. Do not change the value of the other resistor in the base bias circuit. If the collector voltage is not correct at the correct value of collector current, adjust the value of the resistor in the supply line. Such corrections are extremely rare and are necessary only if the particular transistor used has different characteristics from the common characteristics for that transistor type.

When both collector voltage and current are within the expected range, connect an antenna to the amplifier input and the receiver to the output and adjust the trimmer capacitor for maximum received signal using a weak station. This completes the final adjustment; the performance will be very close to the predicted values. With higher band amplifiers, especially 23cm, there is a slight difference in matching for maximum amplification and for minimum noise. The amplifier should be set to maximum amplification and then adjusted to a slighter lower frequency, ie slightly raise the trimmer capacity, until the amplification falls by 1-2dB.

Any further changes or modifications except the ones stated above are absolutely not recommended, because the amplifier is optimised so that it immediately reaches the required characteristics. Any modification would prevent that and would produce much worse results than the expected.

The amplifier should always be mounted as close to the antenna as possible and connected to the antenna with the shortest possible cable, using coaxial relays to switch the anrtenna from receive to transmit. Its power supply should be fed through the coaxial cable that connects it to the receiver, using the adapter shown in Fig 9.48.

As expected, very good noise characteristics have been proved in practical use, which mainly satisfy every requirement for serious DX work. Only for EME work at 432 and 1296MHz you might try using lower noise value amplifiers, ie the GaAsFET amplifiers [5 - 8], but in all other cases the amplifier satisfies even the most rigorous noise requirements. These amplifiers have shown exceptional linearity with IP3 values far exceeding +30dBm on all bands, except at 1296MHz where it is 3-4dB lower.

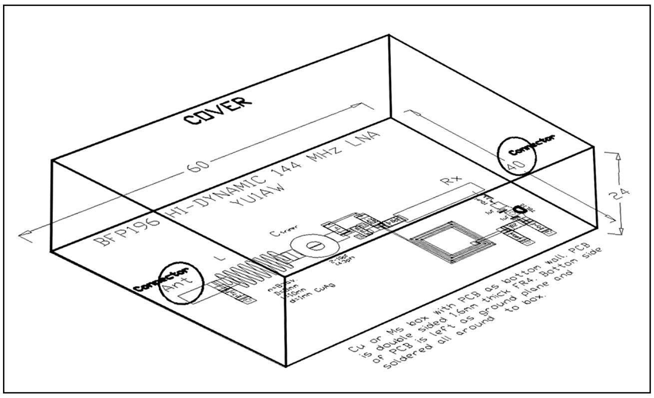


Fig 9.47: Mechanical layout of the low-noise amplifier

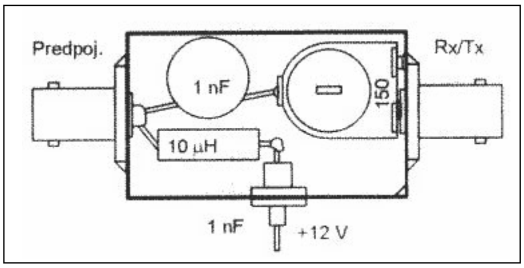


Fig 9.48: Power feed for the pre-amplifiers

As a comparison, Figs 9.49 and 9.50 show the performance of this amplifier and a common amplifier which uses an MGF1302 GaAsFET. Both amplifier inputs have three signals of 7.1mV (-30dBm) to simulate three strong stations on the band. The graphs show what would be heard with an ideal receiver without its own IMD. With a real receiver, because of its possible IMD, things would look even worse! Before you accuse someone of band wasting check with an input attenuator whether your receiver might actually be creating IMD due to a strong input signal!

The amplifier using the BFP196 is superior to the one using the MGF1302. The difference in the IMD products appearing on the output of an ideal receiver was over 30dB! Of course, in both cases the amplifiers had approximately the same amplification.

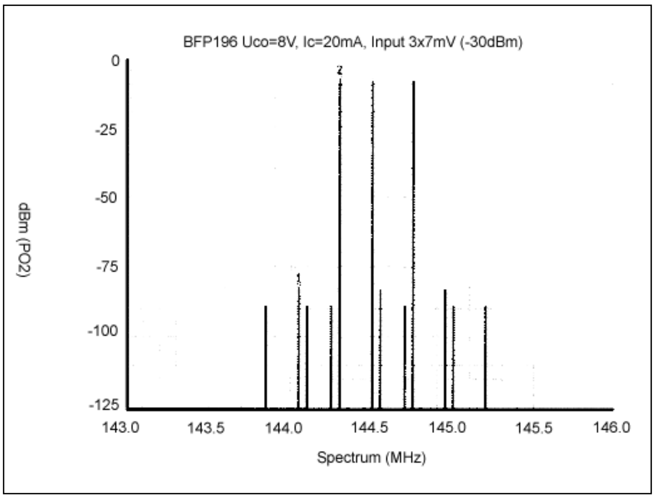


Fig 9.49: IMD products from a BFP196 preamplifier

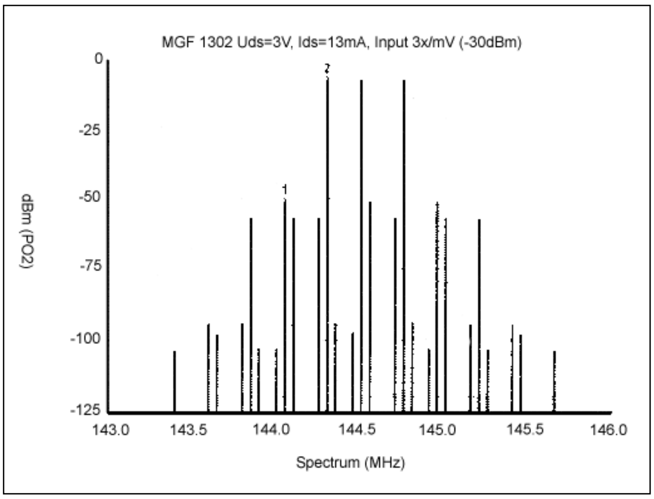


Fig 9.50: IMD products from a MGF1302 preamplifier

The component layouts for the amplifiers are shown in Figs 9.51 - 9.53 (all in Appendix B) and the predicted performance is shown in Figs 9.54 - 9.61. The values shown have been simulated on a computer. Also, in real life the values have been proven on a sufficient number of built and measured amplifiers that they do not differ more than usual for this type of construction. Strict adherence to the guidelines given here will produce amplifiers with performance very close to those shown.

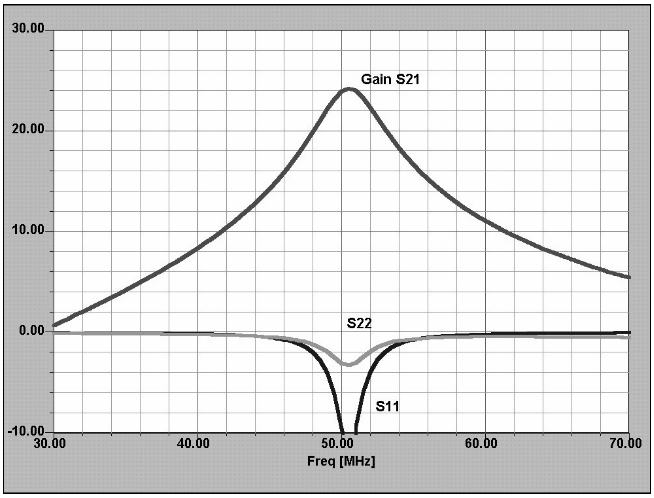


Fig 9.54: Amplification, input and output adjustment for the 6m low noise preamplifier

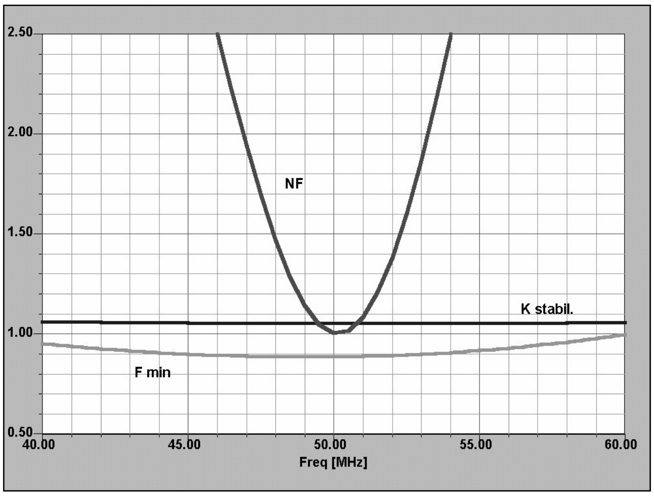


Fig 9.55: Noise figure, minimum noise and stability of the 6m low noise preamplifier

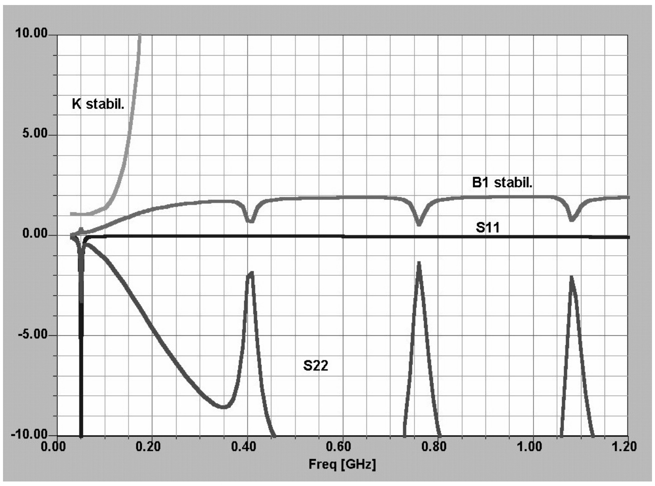


Fig 9.56: Stability factor and adjustment for the 6m low noise preamplifier

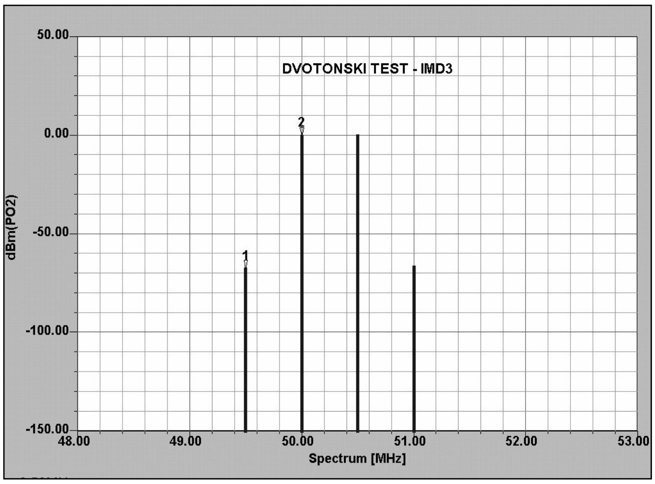


Fig 9.57: Two tone test and IMD products for the 6m low noise preamplifier

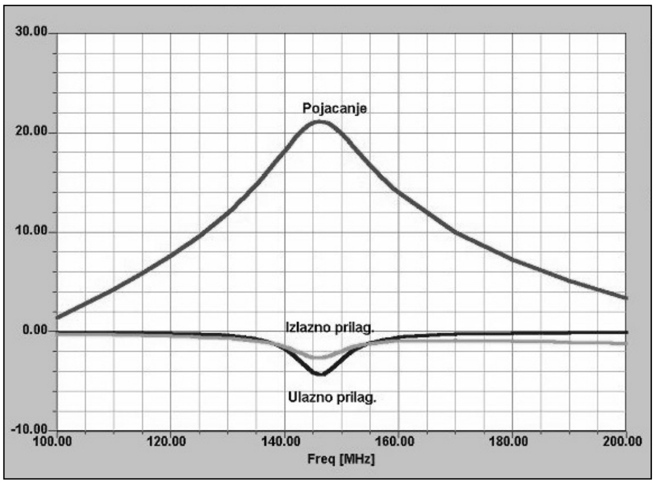


Fig 9.58: Amplification, input and output adjustment for the 2m low noise amplifier

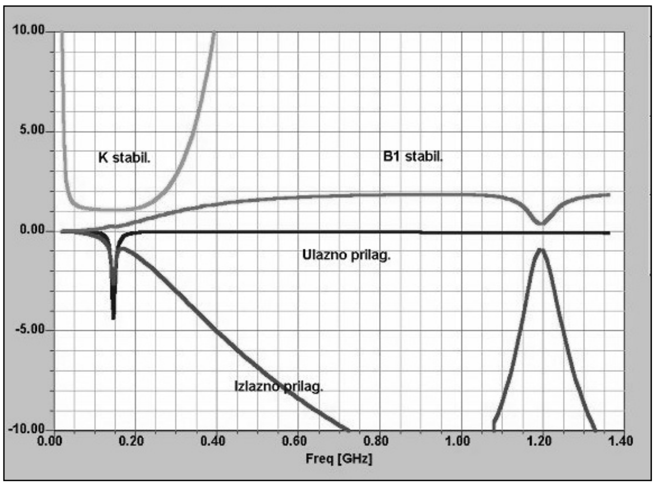


Fig 9.59: Noise figure, minimum noise and stability of the 2m low noise preamplifier

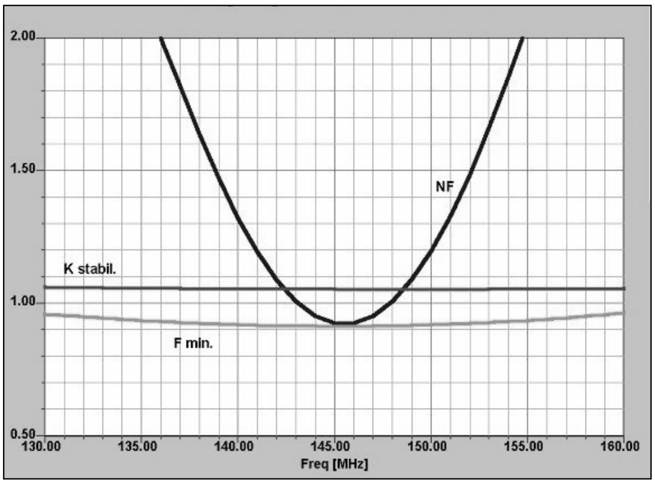


Fig 9.60: Stability factor and adjustment for the 2m low noise preamplifier

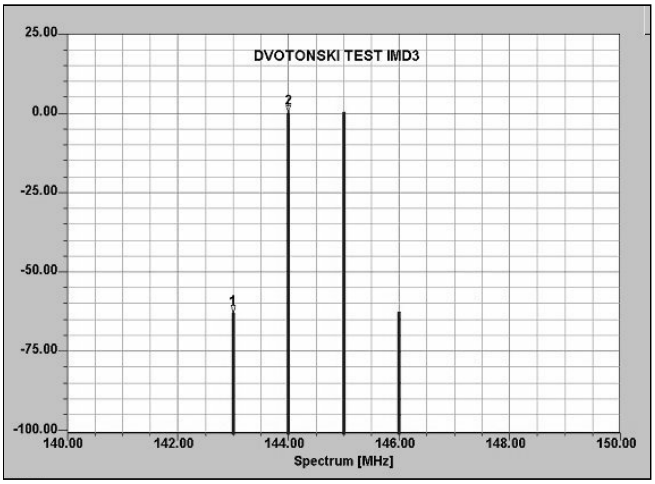


Fig 9.61: Two tone test and IMD products for the 2m low noise preamplifier

The final results achieved with these amplifiers in real life conditions largely depend on the IMD characteristics of the receiver used. If it has weaker characteristics, then the results may even be worse in respect to IMD because when signals, amplified in the preamplifier, reach the input of a bad receiver they cause overload and IMD and the results are poor. That is why the minimum necessary amplification is recommended between this amplifier and the first mixer in the receiver or the transverter in order to preserve as much of the dynamic range of the whole receiving system as possible.

If IMD is apparent in the receiver, put a variable attenuator between the amplifier and the receiver, define the lowest attenuation at which it disappears, replace it with a fix attenuation of the same value and work with that in circuit. This method is highly efficient because the IMD products are attenuated three times faster than the wanted signal, so that it is possible to weaken the products to the level where they are not heard whilst the useful signal has very little attenuation!

Don't be afraid that you will not hear the desired signal, there is too much amplification as soon as IMD appears - feel free to lower it!

A miniature 100-500 ohm trimmer potentiometer connected to the receiver input can be used in place of a variable attenuator. It can be built into the amplifier supply adaptor box as shown in Fig 9.48. This represents a very practical and rather elegant solution at least on of the lower bands. You can also use a variable 20dB attenuator used in CATV.

SSB 2m, 70cm and 23cm Preamplifiers

Introduction

Serious VHF/UHF weak signal operators use low noise amplifiers (LNA) to improve the sensitivity of their receiver systems. External noise, whether from earth or galactic in origin, tends to be at a very low level above a few hundred megahertz. This usually means that the limit to receiving weak signals is set by the internal noise of the receiver and by the loss of the signal in any antenna to receiver feeder.

Using an LNA immediately after the antenna, but before the feeder and the main receiver can often make a marked improvement in sensitivity. This is known as a preamplifier. In general, the lower the noise figure of the preamplifier, and the higher its gain, the better results will be. However, too much gain can result in the receiver system suffering from strong, unwanted, signal problems.

Low noise figure, adjustable gain, preamplifiers are a good solution. This way the optimum gain can be set for any particular situation, so as to minimise strong signal effects.

During transmit the preamplifier needs to be switched away from the antenna in order to protect its sensitive input stage from high transmit power. When a single feeder system is used the preamplifier needs to be bypassed so that a low loss path exists between the transmitter output and the antenna. In the SSB Electronics SP series preamplifiers this is done with the use of two coaxial changeover relays.

It is not advisable to allow transmit power to arrive at the relay contacts before they have closed. This greatly increases their power handling capacity. The relay contacts that do the path switching should ideally be allowed time to settle before the next stage of switching commences. The device that controls the timing sequence of the relays and stage switching is known as a sequencer.

SSB Electronics manufacture a range of high quality, weatherproof, preamplifiers for the VHF and UHF bands from 50MHz to 2.3GHz. Not only do these preamplifiers have low noise figures, but they also have user adjustable (preset) gain. They are also configured to use a single coaxial feeder for both transmit and receive and to be both powered and switched over the same cable using a sequencer, in what is known as the failsafe mode. Power can be supplied to the masthead preamplifier via an S0239 socket if preferred. **Fig 9.62** shows the SP2000 preamplifier. Externally the SP7000 and SP23 look identical.

In this review the results of RF measurements on the SSB Electronics SP2000 (144MHz band), SP7000 (432MHz band) and SP23 (1.3GHz band) preamplifiers, together with their complementary DCW2004B and DCW2004 SHF sequencers are presented.

Fig 9.62: SP2000 preamplifier. The SP7000 and SP23 look similar and only the label distinguishes the units externally

SP2000 and SP7000 preamplifiers

These two preamplifiers use very similar circuits, consisting of a Mitsubishi MGF1302 Gallium Arsenide Field Effect Transistor (GaAsFET) low noise first stage followed by an Agilent (Avago) MSA1104 Silicon Microwave Integrated Circuit (MMIC) second stage. Band pass filtering is incorporated in both amplifiers. In the SP2000 preamplifier the band pass filter consists of a simple top-coupled tuned filter pair whilst in the SP7000 preamplifier the filter is a commercial two-stage helical block filter. Low noise input matching to the GaAsFET is by means of a capacitor and inductor/capacitor tuned "L" match.

Microwave GaAsFETs, such as the MGF1302, can have very low inherent noise figures at VHF, but the noise matching input circuit must itself have low loss if it is not to dominate the overall noise figure, since it adds directly to the device noise figure. This requires very high Q parts in the matching circuit. SSB Electronics appear to have used components of an acceptable quality in these two amplifiers.

The amplifiers are gain adjustable with a variable resistive attenuator, accessible through the lid of the tin plate box that is housed within the plastic weatherproof outside housing.

Both the SP2000 and the SP7000 use type N connectors for the antenna and transceiver connections. A connector provides for external DC power, if this is preferred over powering via the coaxial feeder cable.

Two methods of switching the preamplifier out of circuit, for transmit, are incorporated. It can be either RF detection switching (RF VOX) when RF is detected on the cable from the transceiver/transverter, or removal of the DC powering (via transceiver PTT) on the coaxial cable. The sequencer allows use of higher transmit powers due to the carefully controlled relay timing. The manufacturer claims that the RF VOX method results in a significant reduction in power handling capability.

SP23 preamplifier

The SP23 1.3GHz band preamplifier also uses an MGF1302 GaAs FET front end, but with a revised low noise, air spaced matching circuit. A commercial helical filter is used to shape the pass band. The second stage uses an ERA55 GaAsFET MMIC, providing enough additional amplification to raise the overall gain to around 20dB. No user adjustable gain control is provided.

The SP23 uses the same connector arrangement as the VHF preamplifiers.

DCW2004B and DCW2004B SHF sequencer

Some multiband transceivers, such as the FT847 and TS2000, can provide the necessary DC powering and switching for the SP2000 and SP7000. The alternative is to use an SSB Electronics DC2004 sequencer. The DCW2004B is specified for 50MHz, 144MHz and 432MHz and not only provides the required DC supply to the preamplifier, via a suitable power injector (bias tee), it also provides suitable time delays to allow sequenced switching of the preamplifier, power amplifier (if used), and transceiver/transverter. Independent selection of preamplifier on or off and power amplifier on or off is provided for by front panel push (momentary) switches with LED indication of mode. **Fig 9.63** shows the front panel of the DCW2004 and **Fig 9.64** shows the DCW2004B SHF rear panel connectors.

Fig 9.63: DCW2004B sequencer, Note the preamplifier and power amplifier enable switches and indicator LEDs

Fig 9.64: Rear panel view of the DCW2004B SHF sequencer showing the N connectors and 9 pin D connector used for powering and interconnections to the transceiver and power amplifier (if used). The non-SHF version looks similar

The DCW2004B SHF is specified for use with the SP23 (1.3GHz) and SP13 (2.3GHz) preamplifiers. The difference in the two types of sequencer appears to be in the frequency range of the bias tee used for DC power injection.

Measurements

Accurate noise figure measurement is notoriously difficult for a number of reasons including external interference, noise source impedance changes, LNA input match and poor equipment calibration. These preamplifiers have been measured using a Hewlett Packard (HP) HP8970A and HP346A (5dB ENR) noise head.

This equipment has been used extensively to measure literally hundreds of low noise preamplifiers and transverters at a number of UK VHF and microwave events. Cross-checking between measurements performed on a number of the same LNAs, using different measurement equipment over a period of several years provides confidence that the noise figure results presented here are as accurate as is likely to be achieved using similar equipment.

The third order input intercept (IIP3) measurements were made using an Agilent E4432 vector signal generator providing multi-tone output (50kHz, 2-tone separation) signal drive and an Agilent E4405 spectrum analyser for the measurements, whilst the input return loss was measured using a HP8754 vector network analyser and HP8502A reflection test set.

More about noise figure and input intercept can be found in [9] with reference to using APPCAD to calculate system performance.

The estimated uncertainty for noise figure is 0.1dB based on measurement of 'golden' (ie known noise figure) LNAs. Measurement uncertainty is -0.5dB for gain, return loss and IIP3. All cable and sequencer losses have been taken into account in the overall preamplifier gain results.

Results

The measured results for the three preamplifiers are shown in **Table 9.2**, together with the manufacturers claimed figures.

Measurements on the sequencers gave the following results:-

DCW2004B Insertion loss (144MHz and 432MHz) = <0.1dB
DCW2004B SHF Insertion Loss= <0.1dB

Claimed Noise figure and gain (dB)	Noise figure and gain (dN) (High gain)	Noise figure and gain (dB) (Low gain)	3rd order input intercept (dBm)	Input return loss (dB)	3/20dB bandwidth (MHz)	Preamplifier off insertion loss (dB)
SP2000 (144MHz)						
144MHz 0.8/20	0.95/20.5	1.23/11.5	-8	<1	6/19	0.11
145MHz	0.9/21.6	1.14/12.5				
146MHz	0.88/21.8	1.11/12.8				
SP7000 (430MHz)						
430	1.5/19.4	1.92/9.6	-3	3.1	14/37	0.23
432 0.9/20	1.33/21.4	1.66/11.4				
434	1.31/21.4	1.6/12.0				
436	1.29/21.4	1.58/11.8				
438	1.3/21.2	1.58/11.8				
440	1.33/21.3	1.6/11.0				
SP23 (1200MHz)						
1240	2.34/14.53					
1260	1.28/19.65					
1280	1.05/21.0					
1300 0.9/20	1.2/18.9	0.5	7.2	65/195	0.41	
1320	2.09/13.5					

Table 9.2: Results of measurements on the SP2000 (144MHz), SP7000 (430MHz) and the SP23 (1.3GHz) preamplifiers

Conclusions

The SSB Electronics preamplifiers provide a convenient and cost effective solution to the problem of what to use to improve your VHF/UHF receive system.

Within the accuracy limits of the measurement system, the SP2000 noise figure and gain measured very close to the manufacturers typical figures, whilst the SP23 noise figure was a little higher than claimed by the manufacturer. To put this in context, an increase of 0.3dB will probably not be noticed in terrestrial operation. The SP7000 noise figure was a little disappointing but again, putting the measured noise figure in context, would give a big improvement over most transceivers in common use. The 3rd order input intercept measured a very respectable -3dBm.

The SP23 measured IIP3 of +0.5dBm is remarkably good and should allow this preamplifier to be used in some strong signal situations. A good preamplifier cannot compensate for an indifferent transverter or transceiver, so do not expect miracles.

English language data sheets are available on the Diode Communications [10] web site.

POWER AMPLIFIERS

There is a simple decision to be made when you are thinking of making a power amplifier: should it use valves or should it use semiconductors?

There is no doubt that you can get more power for less money with valves, the down side is that high power valve amplifiers require very high voltages, 1- 2kV, that can be very dangerous if you don't take the correct precautions. There are many designs for valve power amplifiers for all of the VHF and UHF bands, in this chapter two more novel designs are included. The single 4CX250B 6m power amplifier was designed by Geoffrey Brown, GJ4ICD [11] some time ago, if you visit his web site you will find other 6m power amplifier designs. The 70cm valve power amplifier is from *VHF Communications Magazine* 4/1998 [12]. Semiconductor amplifiers fall into two types, those that use hybrid modules and those that use discrete semiconductor devices. Hybrid modules are easier to use because they require very few external components to get them working. Unfortunately the most common modules used by radio amateurs, from Mitsubishi, have been discontinued. They are still available form some suppliers and on the second hand market and there are other modules available that can be used. The designs for transverters shown later in this chapter have optional amplifiers using hybrid modules. The 400W amplifier for 2m shows how to design and construct a reliable semiconductor power amplifier.

If you don't want to embark on a big constructional project you can get more power on the VHF and UHF bands by modifying a commercial amplifier. Modification of the ex PMR A200 amplifier is shown as an example of this possibility.

6m Amplifier

This 6m amplifier can be built in under a day and will provide 250+ watts out for about a couple of watts in. The circuit diagram is shown in **Fig 9.65** and the parts list is shown in **Table 9.3.**

You will need a diecast box, fan, SK600 or SK610 surplus socket for the valve, a 4CX250B valve, a couple of tuning capacitors,

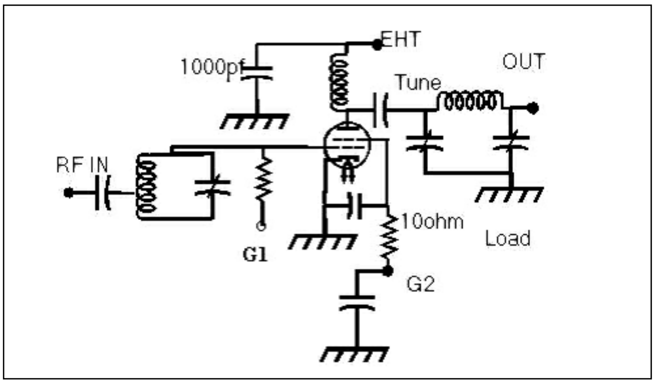


Fig 9.65: Circuit diagram of the 6 metre power amplifier

Anode Section:	
Anode Tuning Capacitor	Jackson C804 25pf (wide-spaced)
Loading Capacitor	Jackson C804 150pf
Anode Isolating Capacitor	1000pf 20kV 'door knob'
Anode Coil	Connects between Tune and Loading capacitors, 5t 12SWG, 1.375in dia
EHT RFC	36t 22SWG enamelled wire on 5/8in dia PTFE rod
Grid Section:	
Grid Tuning	Jackson C804 50pf
Grid Tuning Coil	6t 14SWG 1/2 in diameter
C1 input capacitor	Connects between input connector and Grid Tuning Coil, 1000pf mica

Table 9.3: Parts list for 6m power amplifier

some PTFE sheet to make the anode cooling chimney, and a multiple connector for the various voltages, plus a high voltage connector such as a PET 100 or TNC for the EHT of 1kV to 1.5kV.

Remember that any valve type amplifier has to operate with high voltages in order to work. The typical voltages applied to the 4CX250B series of valves is as follows:

- EHT (anode) is between 1 and 2kV. For this design 1.8kV at 250mA is ideal
- Screen grid + which MUST be regulated by zeners or stabilising valves, this should be 300 volts with a current capacity of 50mA
- A bias supply for the G1, this should be variable to -75 volts
- A heater voltage is also required which is 6.3 volts at a couple of amps
- A relay voltage is also required for the bias circuit and the antenna changeover relays

This amplifier was built in a die-cast box. The box measures 9 x 5 x 5 inches and a plate is fitted across the right hand end about 2.5 inches in. This plate (made of aluminium) has the SK600/SK610 socket fitted on it but offset towards the back (see **Fig 9.66**). The tuning capacitor and loading capacitor are fitted towards the front of the box along with the EHT choke, isolating capacitor and tuning coil. Cooling is via the anode compartment, so no manufacturer's chimney is used. A chimney made of PTFE sheet bonded together with silicon rubber is fabricated to fit onto the 4CX250B and run to the left hand end of the box. A small printed circuit board is used to support the

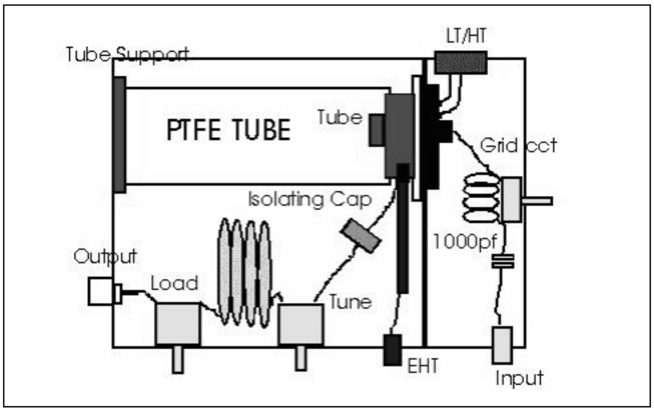


Fig 9.66: Component layout for the 6m power amplifier

Fig 9.67: The 6m power amplifier

chimney, it has a hole cut in it for the exhaust and a brass shim soldered to support the chimney. The blower is fitted to the lid of the box.

Anode and loading capacitors are fitted in the front panel of the box, as can be seen in Fig 9.65 and **Fig 9.67**. A band of brass, or an electrolytic capacitor mounting clamp is utilised to connect the isolating capacitor and the EHT choke onto the 4CX250B. RF out is fitted on the right hand end of the box. The grid circuit is straightforward with the tuned circuit being fitted behind the SK600/610 socket. A power connector is fitted to the rear wall of the grid compartment.

4m Amplifier

The Pye (later Philips) A200 was designed as a boot mounting linear amplifier to give more output power for their range of mobile radios. They are still available on the second hand market but many radio amateurs do not realise the potential of these units to add a useful amount of extra power to a 4m station, they can also be used on 2m and 6m. The A200 is built to last in a heavy weatherproof case with automatic RF sensing for transmit/receive switching, so it is unlikely that you will buy one that does not work.

Fig 9.68: The A200 amplifier. (Picture supplied by the Pye Museum [24])

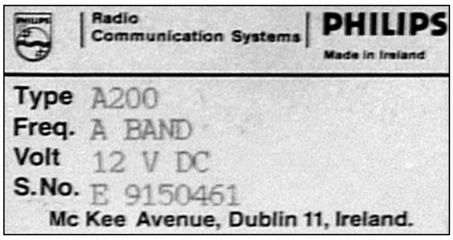


Fig 9.69: The identification plate on the A200 amplifier (Picture supplied by the Pye Museum [39])

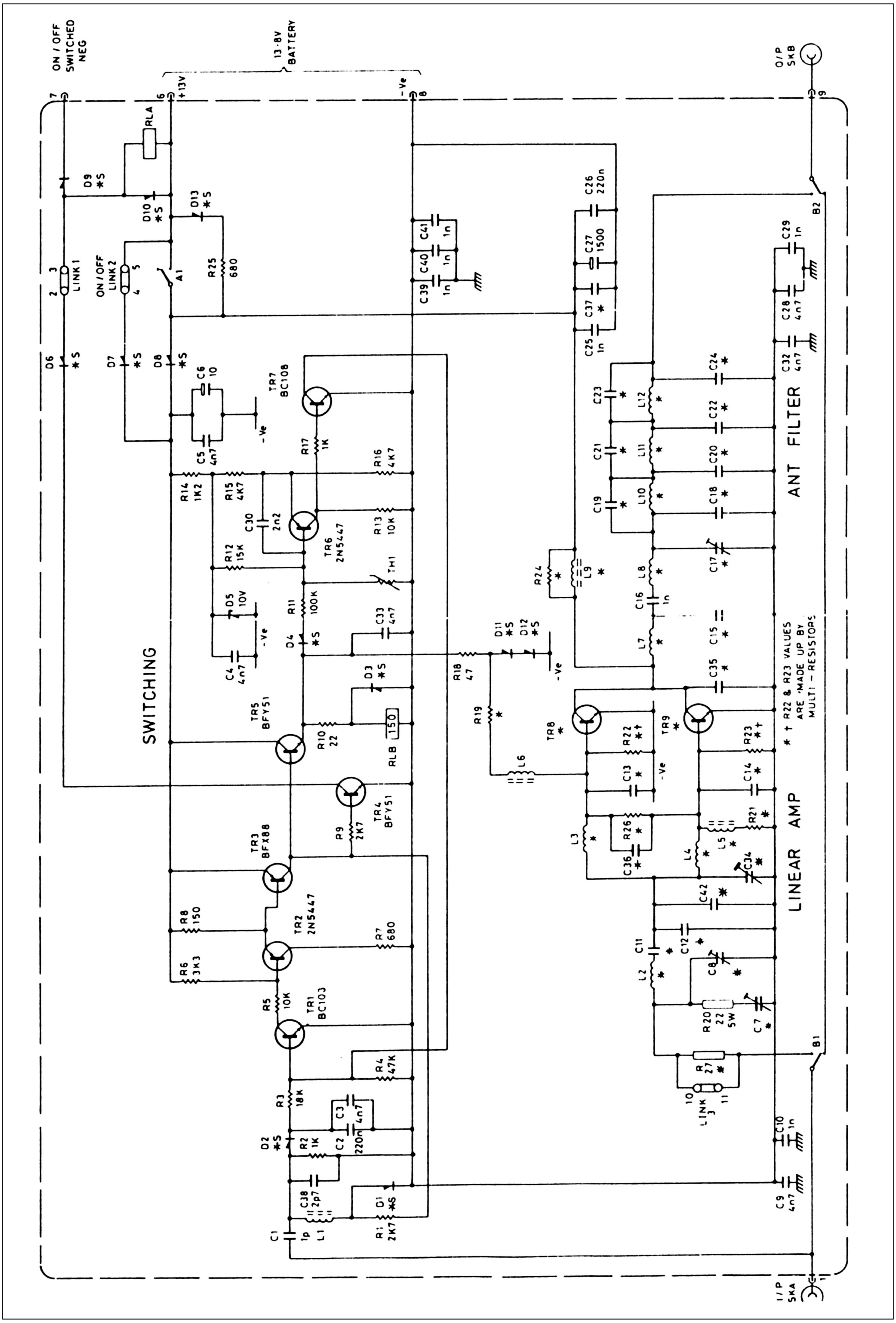


Fig 9.70: Circuit of the A200 amplifier. Reproduced with the permission of Pye Telecommunications

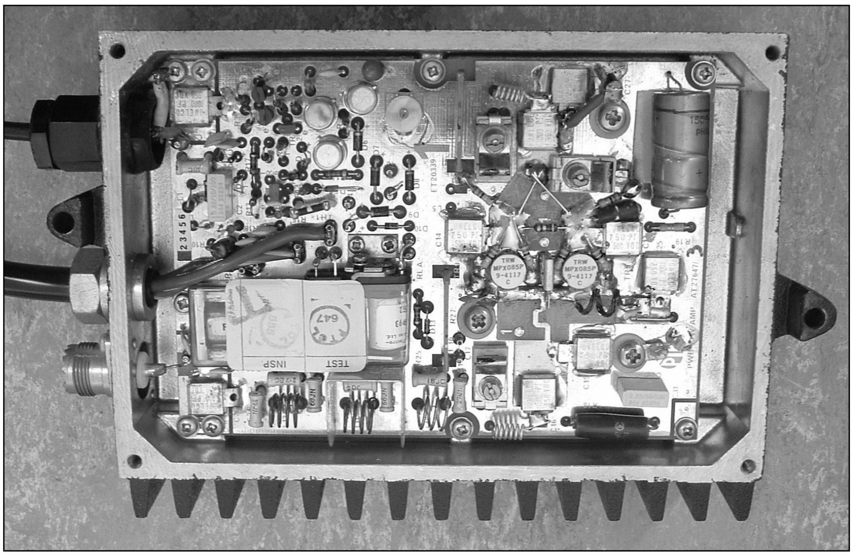


Fig 9.71: Internal view of the A200 amplifier. (Picture supplied by the Pye Museum [39])

It is easy to spot the A200 by the chunky black case shown in Fig 9.68. There are three connections at one end, these are RF input, RF output and a thick DC power lead. The DC power lead is actually heavy duty mains cable with brown being the positive 13.8V supply, blue is negative and green/yellow is for switching. Do not connect this lead to a mains supply, this is a sure fire way to destroy your new acquisition. Also be careful not to confuse a VR200 24V to 12V converter for an A200, it has a similar case but two DC cables coming out of the side.

Two types of A200 were manufactured; early models had a TNC connector for the RF input and an N-Type RF output socket. Later models had a flying lead for the RF input and an SO239 RF output socket. Both models are very similar inside.

To decide if the unit is suitable for 4m, look at the identification plate on the side, see Fig 9.69. They are marked "Cat No. A200" but the aligned frequency is often blank. Fortunately the 'Code' should be marked, something like "01 E0", this will tell you the frequency range:

E0:	68 - 88MHz
M1:	105 - 108MHz
B0:	132 - 156MHz
A0:	148 - 174MHz

The E0 model is suitable for 4m and the A0 or B0 models will work on 2m. The E0 model can be modified to work on 6m, this is described in [13].

Fig 9.70 shows the circuit diagram of the A200. A pair of MPX085P or BLW60 transistors are used in the output stage with bias derived using a wire wound resistor and two forward biased diodes. Printed circuit inductors are used for input and output circuits, tuned with compression trimmers. There is a three-stage low pass filter in the output. The RF sensing circuit switches the amplifier into circuit if DC power is applied to the A200.

The amplifier is well protected including a thermal cut out to shut down the unit if the output transistors are overheating. Fig 9.71 shows an internal picture of the amplifier and Fig 9.72 shows the component layout.

As an initial check, ensure that links between 2 and 3 plus 4 and 5 are fitted. This will ensure that the RF sensing is enabled. This switches power to the amplifier via relay A and the RF path through the amplifiers via relay B when RF is sensed on the input. If you want to use direct switching, remove these two links and switch the green/yellow wire to 0V to enable the amplifier.

The amplifier requires about 10 watts of drive to produce 60 - 70W output and will

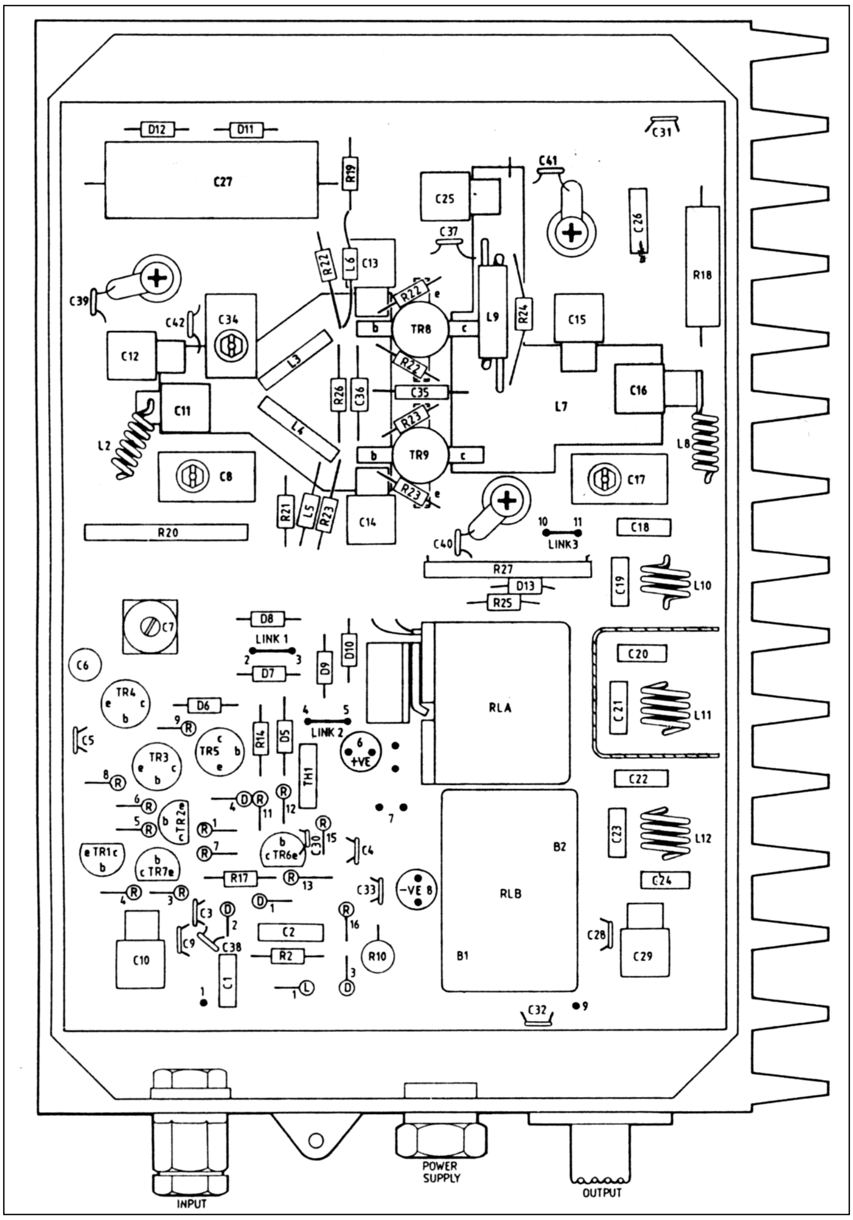


Fig 9.72: Component layout of the A200 amplifier. Reproduced with the permission of Pye Telecomm

draw 10 - 15A from a 13.8V supply. To align for 4m the following steps should be used:

- Set C7 to minimum, this reduces the input drive to the amplifier
- With 2 - 15W input power, check that the relays operate
- Tune C8 and C17 to achieve maximum output power. It may be necessary to repeat adjustment of these two capacitors to achieve optimum tuning.
- If an SWR bridge is available insert it between your transmitter and the A200 and tune C8 for minimum SWR. This should coincide with maximum power output.

Because the amplifier is linear, it will operate on AM, FM or SSB.

For AM operation C7 should be set for a maximum output of 25W with no modulation to prevent over driving the amplifier.

For SSB operation either direct switching should be used or the 'hang time' of the RF sensing circuit should be increased to prevent chatter. Fitting a 0.68µF across C2 and C3 will give a 'hang time' of approximately 0.75 seconds. It is also necessary to increase the sensitivity of the circuit by fitting a 4.7pF capacitor in parallel with C1. C7 should be adjusted to reduce the maximum power by about 10% from the maximum, to prevent overdriving the amplifier. This will still mean that you get 45-50W PEP output with the third order IMD products at least 28db down.

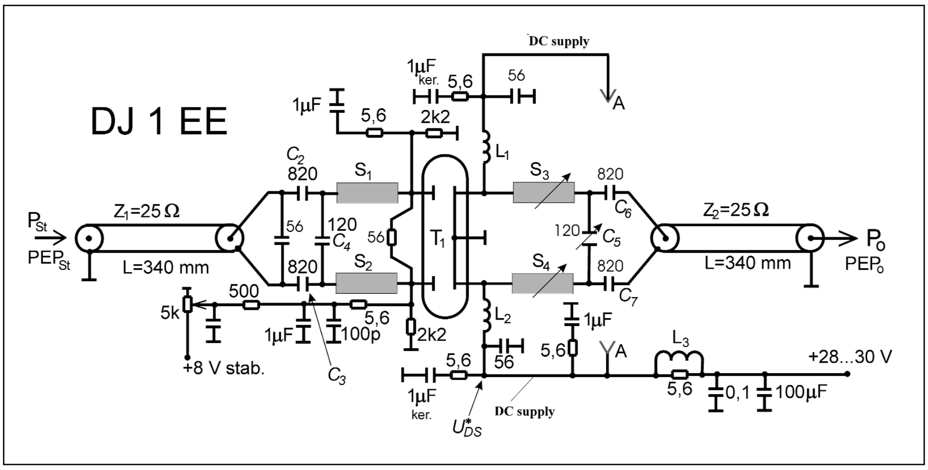


Fig 9.73: Circuit diagram of the 400 watt 2m power amplifier

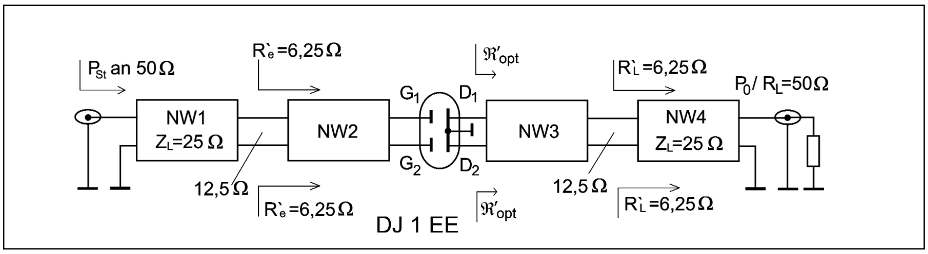


Fig 9.74: Block diagram of the 400 watt 2m power amplifier

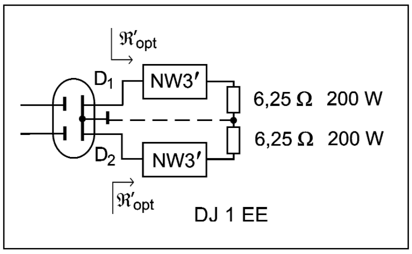


Fig 9.75: Network NW3

400W Power Amplifier for 2m

Because the 2m band is enjoying an increase in popularity a power amplifier with the maximum output of 400 - 450W with a supply voltage of 28 – 32V is described below. The active device was chosen as a proven 'VHF workhorse' from the manufacturer SEMELAB. It is very robust; the load VSWR may vary up to 20:1 thus giving more scope for output network optimisation without destroying the semiconductor. Full data for the transistor can be found in the datasheet [14]. The circuit is not complicated yet has good characteristics.

The circuit

The critical part of a transistor power amplifier circuit (**Fig 9.73**) is the output network. A mismatch at the output represents a danger to the semiconductor. Therefore the matching circuits shown in **Fig 9.74** will be described in more detail.

A narrow band solution would be sufficient for the 144MHz to 146MHz frequency range; it is nevertheless advisable to use a wider band solution in order to allow for adjustment tolerances and alignment sensitivity.

The input and output matching circuits NW1 and NW4 consists of quarter-wave matching transformers (25 ohm coax cable), they also provide a 50 ohm asymmetric to 12.5 ohm symmetric transformation. The networks NW2 and NW3 are very simple L/C circuits with inductances made from individual striplines.

Design of the circuit for network NW3

The quarter-wave matching transformer NW4 is made from 25 ohm coaxial cable that transforms the asymmetric 50W load RL to 2 x R´L = 2 x 6.25W = 12.5W. Flexible or semirigid cable should be used with an outside diameter not less than 3mm because of the 400W output power that it must handle)

To make the computation a bit clearer, **Fig 9.75** shows that the real load is divided from 12.5 ohms to 2 x 6.25 ohms. The network NW3 must be designed to match the optimal load resistance R'_{opt} at half of the total output of 200W with the internal values of the transistor shown in **Fig 9.76**. The value R´opt is shown in the SEMELAB data sheet. The internal load resistance R'_{DS} can be determined from the effective RF Drain voltage V'_{RF} at 200W. The effective RF voltage is close to the supply less 'the bottoming voltage' V_K (assumed to be 3V):

$$R'_{DS} = \frac{(V - V_K)^2}{2 \cdot P_o} = \frac{(28-3)^2}{400} = 1.56\Omega$$

In parallel with this is an output capacitance C_{ob} = 190pF (see data sheet). The effective capacity is:

$$C'_{ob} = 1.3 \cdot C_{ob} = 1.3 \cdot 190 = 250\,pF$$

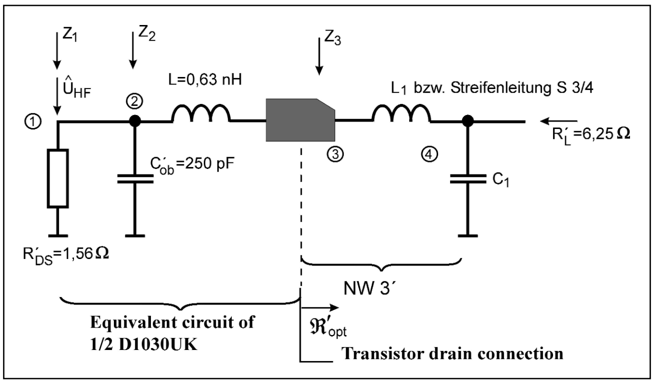


Fig 9.76: The equivalent circuit of half D1030UK

The inductance of the drain connection is taken from the SEMELAB data and is 0.63nH.

Calculating the value of L1 in Fig 9.76

L1 is made from a small low impedance stripline. The transformation procedure can be seen by using a Smith chart. The Smith chart program by Fritz Dellsperger [15] was used for this process and found to be extraordinarily helpful for this task. The Smith chart in **Fig 9.77** is standardised for the terminal resistance R’2 = 6.25 ohms. The internal load resistance R'_{DS} = 1.56 ohms and the goal value R'_L = 6.25 ohms can be seen. The chart is simplified to make it easier to understand. Point 1 (Z1) on the real axle represents R'_{DS} = 1.56 ohms. C'_{ob} = 250pF is point 2 on the conductance circle. The bond inductance is a series inductor L = 0.63nH and is represented as Z2 = 1.4 + j0.5 ohms giving point 3. This is the impedance at the drain connector lug; Z3 = 1.4 + j0.1 ohms. This ‘transistor connection resist-

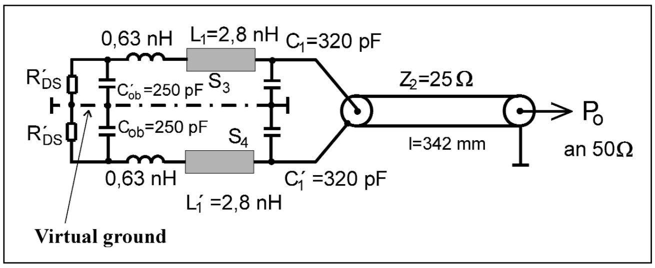


Fig 9.78: Circuit showing the two halves of the push-pull amplifier

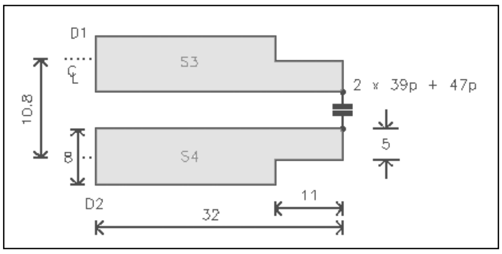


Fig 9.79: Dimensions of striplines S3 and S4 after optimisation

ance’ is now transformed with an inductance, formed by striplines, S3 and S4, on the circle around the centre of the diagram to point 4 giving Z4 = 1.4 + j2.6 ohms. This is the intersection with the conductance circle corresponding to the target resistance Z5 = 6.25 + j0 ohms.

The stripline width W was selected as 8mm and the substrate thickness, H, used was 0.83mm. The characteristic impedance Z_0 is given by:

$$\frac{W}{H} = \frac{8mm}{0.83mm}$$

This is for a dielectric constant ε_R = 1. In reality the characteristic impedance of the striplines S3 and S4 are 16.5 ohms because the substrate has a dielectric constant ε_R = 3.3. The inductance of this stripline is 2.8nH and is ‘quasi-stable’ the length of the line is l = 30 mm. The inductance of this very short line is independent of ε_R!

If a 320pF capacitor is connected from point 4 to ground then a real resistance of 6.25 ohms is achieved (point 5 on the Smith chart). This matches one half of the power transistor to 6.25 ohms. For the push-pull circuit twice this value is used, see **Fig 9.78**. Because the ‘virtual ground’ does not exist, the two single capacitors in Fig 9.78 are combined into a single capacitor (2 x 320pF in series = 160pF).

If the output networks are used with the calculated values the output power will be approximately 250 to 300W. The transistor equivalent circuit used do not exactly correspond to the actual values partly because of differences in the mounting of the transistor. In order to achieve the maximum output power of 400W with good efficiency (~70%), changes in the striplines S3 and S4 as well as the capacitor C5 are required. This is shown in Fig 9.73 by the variable arrow on these components. This tuning was carried out while watching the efficiency and the values shown in **Fig 9.79** were the result.

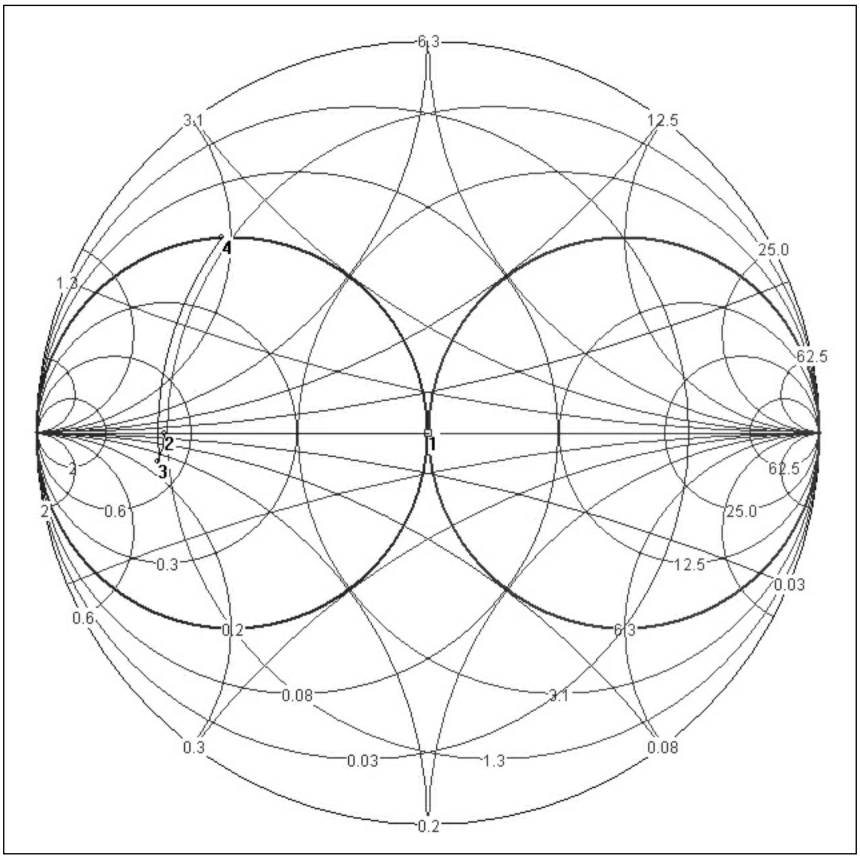


Fig 9.77: Smith chart showing the transformation to 6.25 ohms

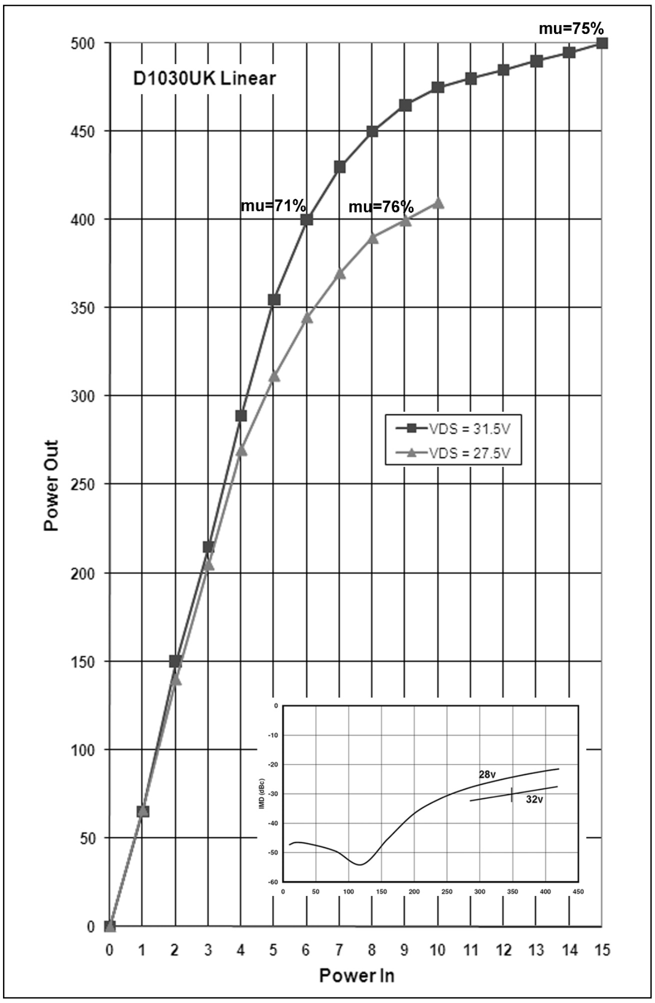


Fig 9.80: Output characteristics of the D1030UK

The characteristics of the output stage with a supply voltage V_{DS} of 28V and 32V are shown in **Fig 9.80**.

Measured performance

The performance of the amplifier was measured with the equipment shown in **Fig 9.81**. The results for a supply voltage of 28V are shown in **Table 9.4** and for a 32V supply voltage in **Table 9.5**.

CW:	P_o	400W
	P_{ST}	8.9W
	Efficiency	75%
	V'_{DS}	58V
Harmonics:	2 x f_0	-49dB
	3 x f_0	-44dB
SSB power:	PEP_0	300W
	PEP_{ST}	4.7W
	FMD_3	-29 to 31dB
FMD_5	-35dB	
Efficiency	49%	

Table 9.4: Measured parameters at 28V (V_{DS} = 28V; I_{DQ} = 1.7A; f = 145MHz)

P_o/W	P_{ST}/W	I_D/A	G/dB	Effic. %
50	0.7	6.8	18.7	24
100	1.35	9.2	18.9	
200	2.7	12.7	18.7	
300	4.3	15.9	18.5	
350	5.0	17.0	18.5	
400	5.9	18.1	18.3	71
450	8.2	19.6	17.4	74
SSB Power	PEP_0	350W		
	IM_3	-30dB		

Maximum drain voltage: (DC + RF) ≅ 65V for P_o = 480W

Table 9.5: Measured parameters at 32V. V_{DS} = 32V; I_D ≅ 1.8A; f=145MHz

Input matching

The input impedance of a half D1030UK has the typical values shown in **Fig 9.82**.

R'_{GS} is found, like R'_{DS}, in the data sheet [14]. R'_{GS} can be transformed to 6.25 +j0 ohms using a simple L/C circuit. The inductor is again made from a stripline. The transformation to 50 ohms is achieved with quarter-wave matching transformers

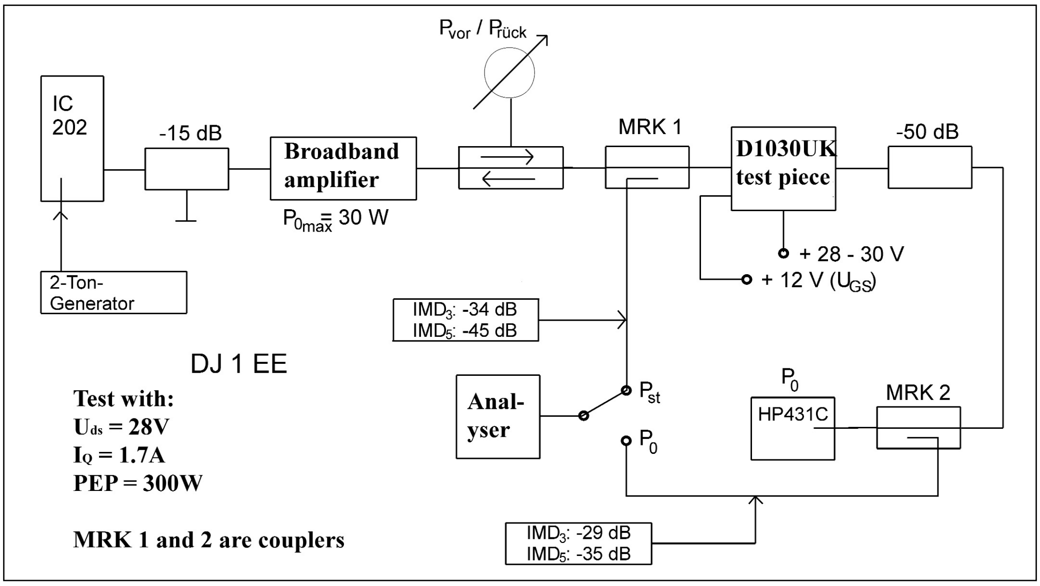


Fig 9.81: Block diagram of the performance measuring equipment

Fig 9.82: The input equivalent circuit

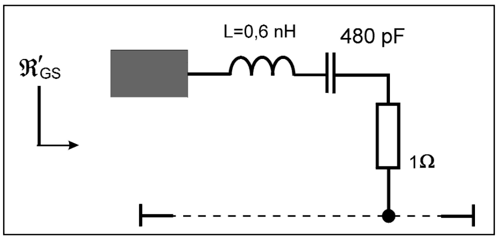
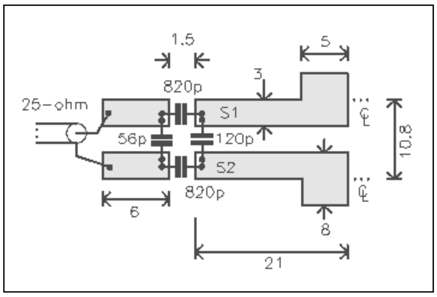


Fig 9.83: Dimensions of striplines S1 and S2 after optimisation



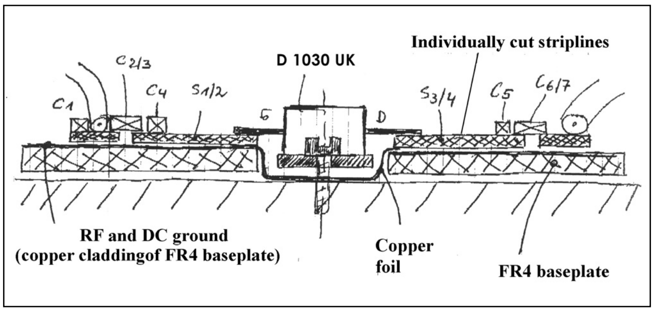


Fig 9.84: Rough sketch of the mechanical construction of the 400 watt 2m power amplifier

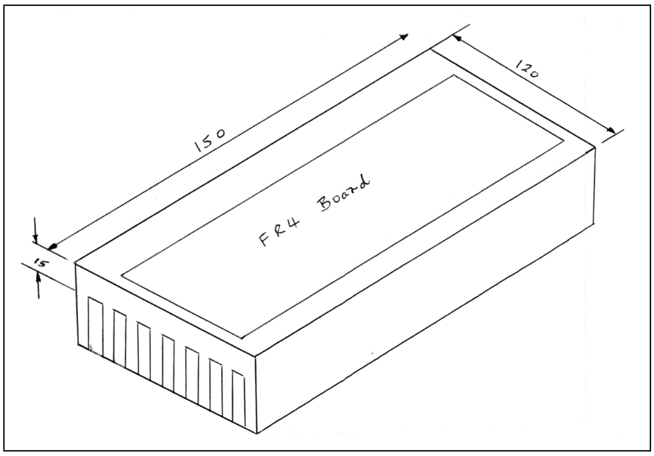


Fig 9.85: The heatsink baseboard

made from 25 ohm coax cable; thin cable is sufficient here because the power is low. From the values calculated using a Smith chart, the input matching from 5:1 to 2:1 is achieved. The striplines and capacitors must be optimised as shown in **Fig 9.83**. Unlike the output the input can be trimmed to give a return of zero.

Mechanical construction

Fig 9.84 shows the construction of the amplifier. The 50 x 150mm baseboard is fitted to the large 150 x 120 x 80mm heatsink as shown in **Fig 9.85**. The baseboard is 1.6mm thick copper clad FR4 material; the upper surface is the RF and DC ground. In order that fitting the power transistor does not interrupt this ground, a very thin copper foil (~ 0.1mm) is fitted into the cutout for the transistor. The striplines are individually cut from 0.83mm thickness material and soldered to the baseboard ground surface. The DC wiring can take place as desired.

Good heat transfer between the foil, transistor and heatsink is extremely important. Spread thermal compound on the individual surfaces very thinly because only the pores in the metal are to be filled! The heatsink should be well cooled with the aim of a maximum flange temperature of 60°C with a PEP output of 300W.

The RF circuits were computed with the aid of a computer program. Special attention was given to optimise the half-wave anode line so that with the given impedance the lowest possible loaded Q was obtained. Through a careful selection of the parameters, a unloaded Q of 39 was achieved which, for this application, is the lowest possible value. This ensures that the anode tank circuit and the amplifier work with the maximum efficiency.

The variable capacitors, shown in Fig 9.86, should have the following calculated values for optimum operation:

70cm Amplifier

The 70cm linear power amplifier described here can be built in little more than a weekend. It delivers the power required for satellite working, small antennas and short cables or larger antennas and longer cables. Three readily available 2C39 disc-seal triodes are used in parallel delivering 300W output for an input drive of 15W.

As can be seen in **Fig 9.86**, the three triodes operate in a grounded grid circuit with the cathodes being driven in parallel. The amplifier requires only two supply voltages for reliable operation: the anode and the filament voltages. The anode voltage may be between 1.3 and 1.5kV and the filament between 5.8 and 6.0VAC (at 3A).

With 1.3kV on the anode, the anode current can be driven up to 400mA giving an RF output power of some 300W for 15W drive power. It is quite possible that, if good tubes are used, the output power will be even more but they should not be overdriven. A good axial air-blower should be used for the anode cooling.

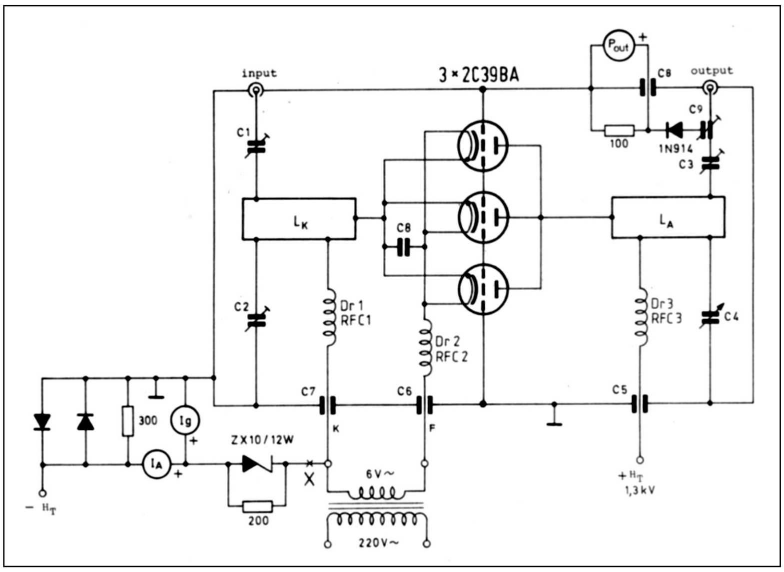


Fig 9.86: Circuit diagram of the 300 watt 12dB gain, 2C39A power amplifier for the 70cm band

C1	4.3pF
C2	5.3pF
C3	1.4pF
C4	5.3pF

The dimensions for the cathode (L_K) and the anode (L_A) line resonators as well as the coupling and tuning plates (C1 to C4) are shown in **Fig 9.87**.

The construction is very simple as can be seen from **Figs 9.88 to 9.90**. A few special parts of the circuit should be explained.

In order that the anode resonator (L_A) can be properly connected to the valves, the latter should be modified in the following manner:

- The cooling-fins are taken off and tapped, 4mm. The strip line can then be held tightly between the cooling-fin body and the tube's anode. A 25mm ceramic pillar supports the other end of the strip line
- Strip contact fingers are used to make the grid-ring contact. The cathode contact, on the other hand, can be fashioned from 10mm outer diameter copper tube of 0.5mm wall thickness. This tube is 12mm long and slit longitudinally down to the middle. The slotted half is then press-fitted over the cathode contact and the other end soldered to the cathode strip line L_K. The remote end of the strip line is secured to a PTFE or ceramic pillar.

The strip lines for the anode and cathode resonators are cut from 1 to 1.5mm stock and silvered, if at all possible.

The amplifier is built into a housing made from 1mm thick brass plate, see Figs 9.88, 9.89 and 9.90. The sides are soldered together.

Tuning capacitors C2 and C4 are made from 0.5mm thick brass plate (Fig 9.87) and hinged and rotated using nylon fishing line. A piece of insulating material - PTFE or polystyrene - is positioned as a stop to prevent direct contact with the opposite electrode. A couple of thick knots tied in the fishing line serve the same purpose.

These tubes require a lot of cooling air if they are to work reliably over a long period. The air blast must also be powerful in order to achieve sufficient cooling over all the surface of the cooling fins. The forced air comes in from above via C4 and cools both the anode resonator and the anode itself and is then vented out of the anode area. It is recommended that a couple of not too small holes be provided in the screening wall between anode and cathode enclosures (Fig 9.88) in order to allow a weak flow of air from the mainstream to flow over the cathode resonator and cathode.

The HT supply as with the drive power is connected to the amplifier by BNC panel sockets. An N socket is used for the RF output.

The valve heaters are connected in parallel. Between the inner heater contact and the cathode lead of every tube, a 1nF disc ceramic (C8) is fitted using the shortest possible connections. The RF chokes (RFCs) are wound using a 6 to 8mm shaft with 0.8 to 1mm diameter copper wire. They are 6 to 7 turn coils, supported from their soldered ends.

Tuning the amplifier is very straightforward, simply tune for maximum output power. This may be accomplished with the aid of a UHF SWR meter or by using the detector circuit shown in Fig 9.86. The coupling (C9) to the detector is adjusted by varying the

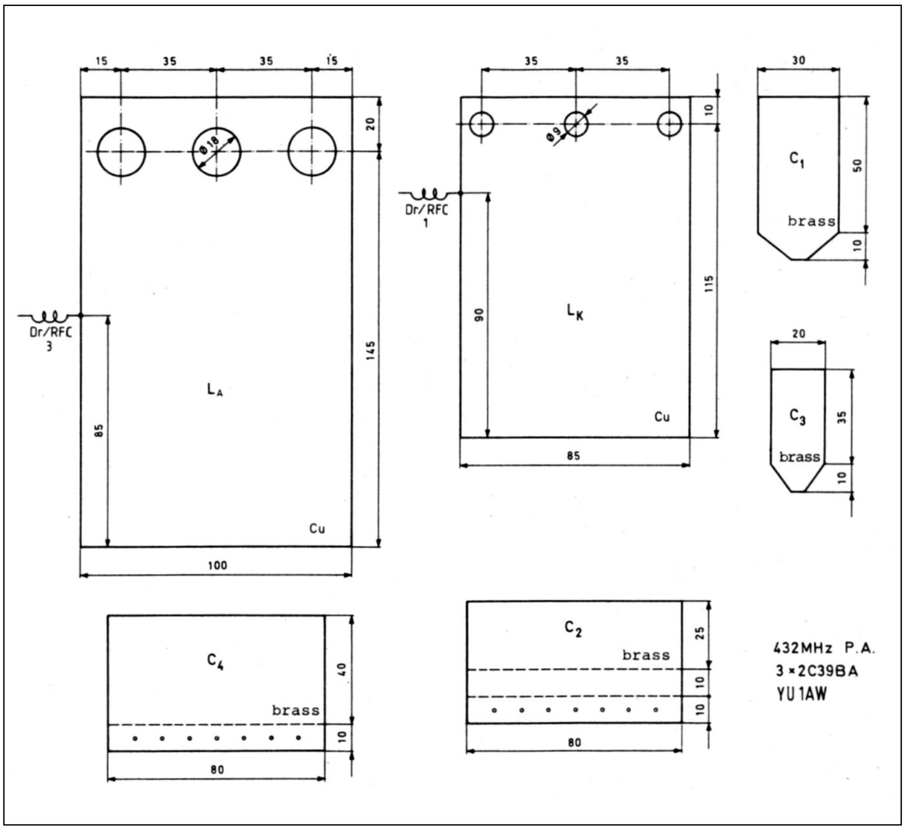


Fig 9.87: Dimensions of the housing parts for the 2C39A power amplifier for the 70cm band

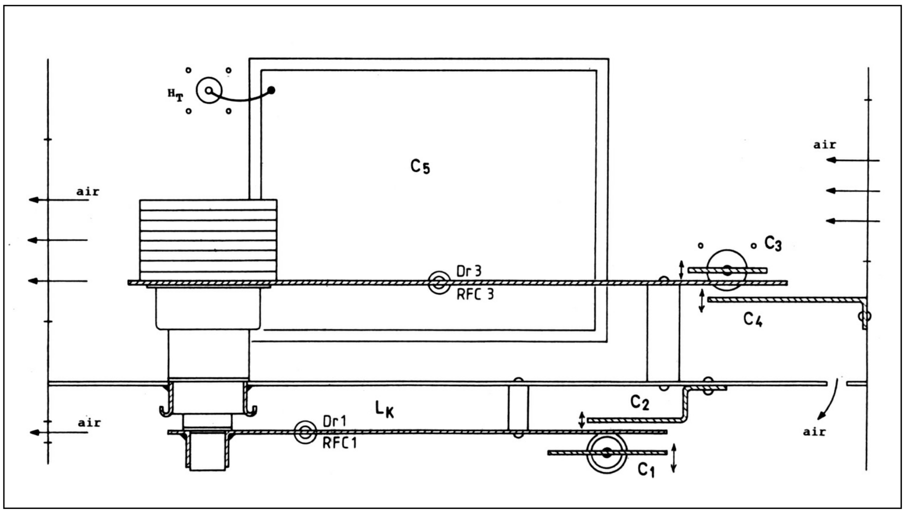


Fig 9.88: Side view of the 2C39A power amplifier for the 70cm band

distance of the silicon diode to the N socket centre pin. The first tuning attempt should take place with very low input drive power and then gradually increase it to maximum.

When the amplifier is in tune the following conditions should exist:

Anode voltage	1300V
Grid voltage	-10 to -12V
Filament voltage	5.8 to 5.9V
Filament current	3A
Quiescent anode current	120mA (40mA per valve)
Maximum anode current	400mA 130mA per valve)
Maximum grid current	100mA (32mA per valve)
Output power	280 to 300W
Power dissipation	210W (70W per valve)
DC Input power	520W
Efficiency	60%
Gain	13dB

It has been found that the anode voltage can remain on during transmit breaks and receive periods. If noise interference can be heard in the receiver, a 10kΩ resistor can be included in the circuit at the point marked X. This resistor must, of course, be short-circuited during transmit. Any type of available relay will do this job.

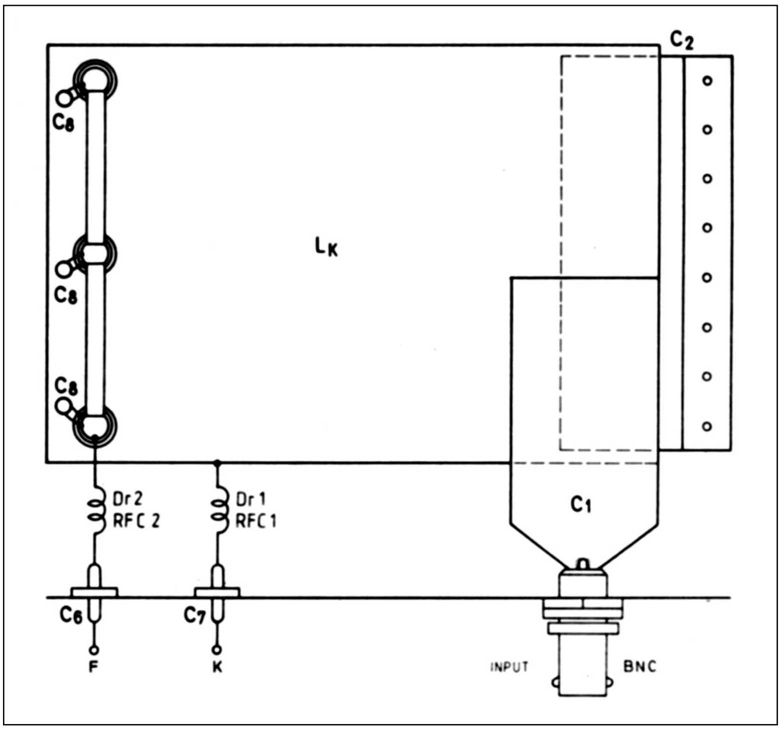


Fig 9.89: Bottom view of the 2C39A amplifier for the 70cm band

RECEIVERS

With the large number of commercial receivers available for the VHF and UHF bands, not many amateurs build their own. This design from *RadCom* by Andy Talbot, G4JNT, shows a new type of receiver [16].

The converter was designed with the primary aim of using it for the IF stage on microwave transverters. A linear receiver was needed with no AGC, but with a calibrated gain control to make accurate relative measurements of microwave beacons using a PC soundcard-based system for the actual level and signal-to-noise ratio measurements. A straightforward gain calibration could then be used to convert these into absolute readings, making this a useful piece of test equipment for propagation studies.

There is nothing inherently narrowband in the design - filtering limits the RF bandwidth to around 8MHz to eliminate strong signals from broadcast and PMR and the audio bandwidth is kept to about 20kHz, wide enough for the normal maximum soundcard sampling rate of 44100Hz. Any subsequent audio filtering for listening purposes is performed by the software or in separate audio processing circuitry.

The circuit diagram is shown in **Fig 9.91**. In the RF path two MMICs, a MAR-6 and a MAR-3, amplify the RF; there is a two stage band-pass filter between them with 10MHz bandwidth. The output feeds into two SRA-1 type DBMs via a resistive splitter, with the quadrature local oscillator (LO) signal generated using a MiniCircuits PSCQ-2-160 90° power splitter. This device guarantees less than 3° phase error over 100 to 160MHz; as 144MHz is near the middle of the range, we can expect better performance here.

The local oscillator is an AD9851 DDS, currently clocked at 100MHz, generating 16 to 16.67MHz followed by a x9 RF multiplier. The DDS source is not described here, but the module in a basic form is described in reference [17]. The active stages in the multiplier consist of MAR-6 MMICs configured as a pair of cascaded tuned x3 stages with a final MAR-6 as amplifier/limiter, this combination forming probably the simplest tuned RF multiplier possible! There are a couple of CW spurii generated by the DDS, but once you know where they are they can be ignored. All filtering is designed to allow the LO to tune over 144 to 150MHz to cover more than the normal 2MHz narrowband segments on

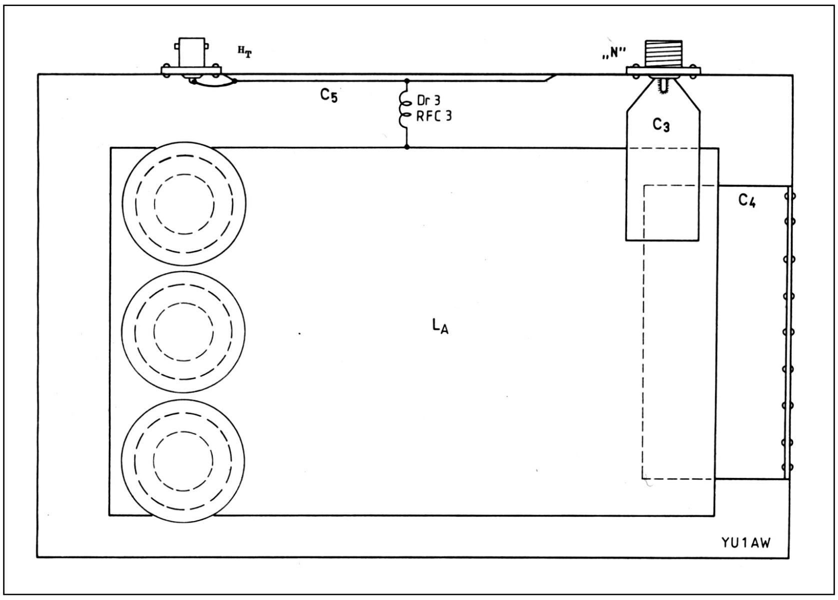


Fig 9.90: View of the anode enclosure of the 2C39A power amplifier for the 70cm band

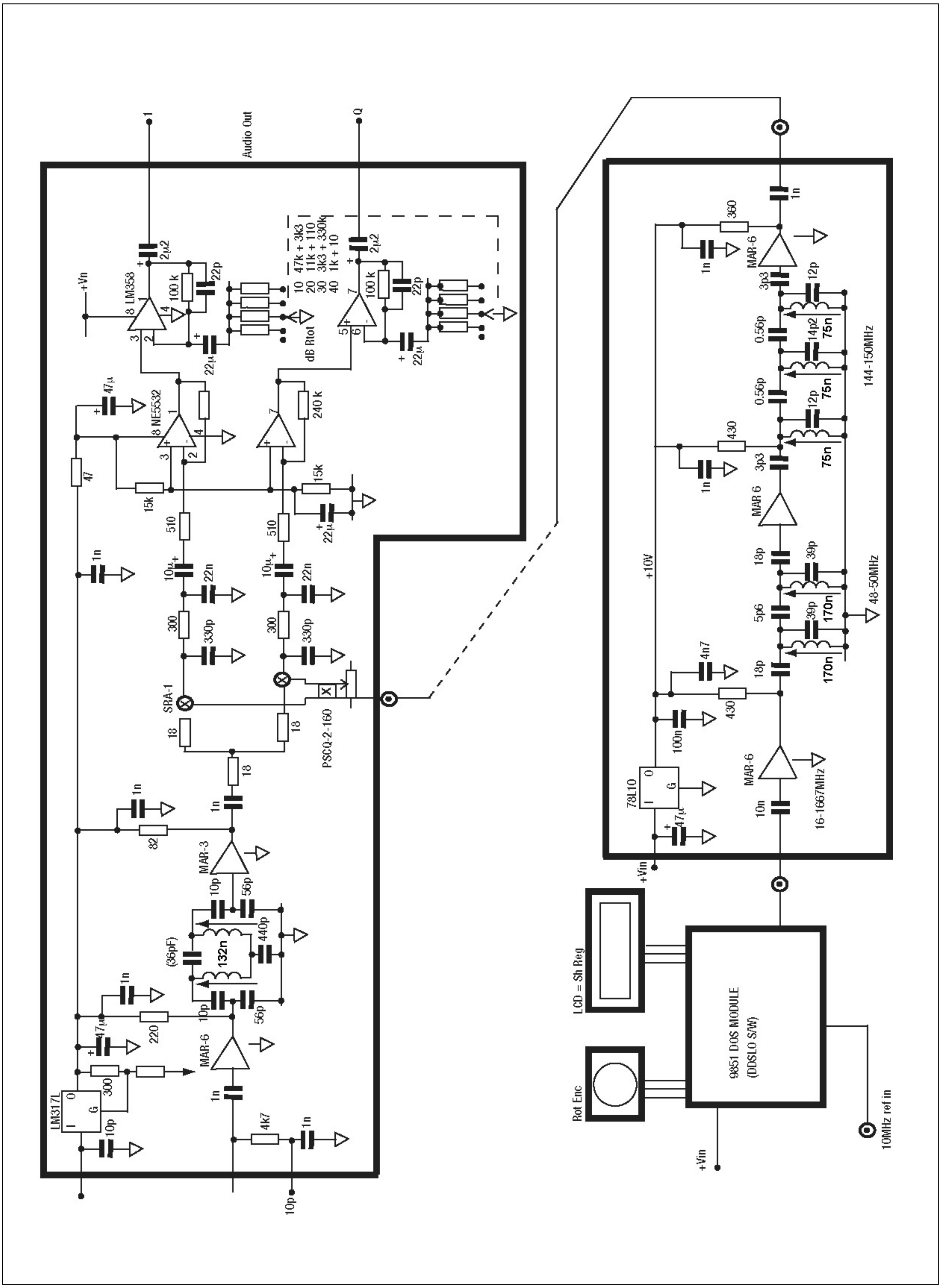


Fig 9.91: Circuit diagram of the 144MHz direct conversion receiver

the microwave bands, and allow for odd LO frequencies. The multiplier output level is +l0dBm drive to the quadrature hybrid.

By using the internal x6 option in the AD9851 DDS chip the LO could be driven from a 10MHz frequency reference, producing a clock of 60MHz, but this has not been tried.

The mixer outputs drive a pair of identical NE5532 op-amps with a voltage gain approaching 300 (the exact value is a bit uncertain due to the internal impedance of the mixer IF port). No clever matching is used, just the mixer feeding the inverting input, giving 800 ohm input resistance at audio, and low-pass filtering to get rid of RF leakage. The I/Q outputs feed another pair of op-amps with precisely switchable gain from 0 to 40dB in 10dB steps. Audio bandwidth is not especially tailored, but rolls off gently from about 20kHz to allow for 44100Hz sampling rate in a soundcard.

The total system gain and dynamic range is based on 16-bit digitisation, and is sufficient at maximum (+40dB) to place its own thermal noise at least 10dB above the quantisation noise pedestal. Strong signals and extra RF gain in transverters are catered for by backing off the audio gain. For signals too strong even for this (80db S/N in 20kHz) an external (calibrated) RF attenuator can be added.

No attempt was made to put this on a proper PCB. The converter and audio stages were built birds-nest style on a piece of un-etched copper clad PCB as can be seen in the photograph. Plenty of decoupling and short direct wires ensure stable performance. As there is a lot of gain - particularly at audio - the whole unit was built into a tinplate box for screening

Using parallel and series 1% resistors for the switchable gain stage, no special trimming or adjustment was necessary, the traces looked well enough matched on an oscilloscope and, as the aim was only 20 - 25dB sideband rejection to make opposite sideband noise insignificant, tweaking was not necessary. 3° phase error will give 25dB rejection, assuming the amplitude is correct, which is about equivalent to 5% amplitude imbalance. All power rails are regulated and well-filtered for operation from a portable 12V supply.

The LO multiplier was made by cutting a 50 ohm microstrip line into a double-sided PCB. To make a 50 ohm line quickly without etching, score two lines 2.8mm apart through the copper on the top face of the PCB for the full width; use a Stanley knife or similar, making sure you penetrate the copper fully. A 2.8mm width on normal 1.6mm-thick fibreglass PBC gives about 50 ohm characteristic impedance. Then, score two more lines about 1mm from each of these.

Using a hot soldering iron, use this to soften the adhesive and with a pair of tweezers, lift up and remove the two 1mm wide strips, which will give a single 50 ohm line surrounded by a copper ground-plane. Drill a number of 0.8 to 1mm holes through the top ground plane to the underside and fit wire links to give a solid RF ground structure. Wire links are best fitted close to where grounding and decoupling components are connected.

Cut the 50 ohm line into segments with gaps for the MMICs, DC blocking capacitors and filters. Other connections around the filters are made up bird's-nest style. When completed and aligned, coils can be held in place with glue (a hot glue gun is a useful accessory to have around).

For the stand alone unit for use as a receiver in the field, a simple quadrature network and loudspeaker amplifier can be added to make a complete receiver. A high/low pass pair of all-pass networks will give 15dB sideband rejection over 400Hz to 2kHz, which is good enough for listening to beacon signals on hill tops. Alternatively, look at [18] for phasing-type SSB networks to give an improved SSB performance.

A meter driven from the audio level via a precision rectifier circuit can be added to allow quite precise signal strength measurements to be made in conjunction with the calibrated attenuator. Alternatively, take at look at the Software-Defined Radio software [19] from I2PHD, for another solution

The DDS module, described in [17], has new PIC software, along with a rotary encoder and LCD display to give a user friendly interface. For anyone who has the original DDS board, G4JNT can supply PIC software for this modification. However, the AD9850 and AD9851 chips are in short supply now - they have been replaced in most cases by larger, faster, new devices in a different package. G4JNT has also developed a rotary encoder / display for the AD9852 DDS which gives a better route for a local oscillator as it can generate up to 100MHz. He can be reached at [20].

Alternatively, emulate the venerable IC-202 transceiver and build a VCXO to supply the signal to the multiplier. Or use a VFO/mixer, or a synthesiser - the choice is yours!

TRANSCEIVERS

Building a transceiver may seem to be a thankless task but seasoned contester André Jamet, F9HX has some different views as he explains below:

A 144MHz Transceiver – for SHF

Typical equipment for SHF operation

For operating on the 5.7,10, 24 and 47GHz bands and beyond a transverter is usually used to reduce the signal to be received or transmitted to lower frequencies. The 144MHz band is in frequent use as an intermediate frequency up to 10GHz, but the 432MHz and 1,296MHz bands are also used for higher frequencies.

We therefore need a VHF or UHF transceiver with the characteristics required to work in combination with the transverter – ie one that can generate SSB and CW, and also has certain accessories which are very useful for SHF use.

One 144MHz transceiver which is very widely used for this application is the famous IC-202. In spite of its faults:

- Imprecise frequency display
- An S-meter which is just as imprecise
- No receive selectivity adjustment (to adjust the pass band in order to improve the signal-to-noise ratio)
- No transmit power control
- No pip generator (to make it easier to get into contact)
- Very frequently poor health

Also bear in mind the great age of those in service and the amount of travelling they have had to endure.

So among SHF enthusiasts a wish has often been expressed to replace this old companion with a more modern transceiver that performs better. Unfortunately, tests carried out using modern transceivers fitted with a very large number of accessories have not always given the expected results. If the various faults mentioned above have disappeared, a new one has seen the light of day. The spectrum purity of their local oscillator is not up to that of the older equipment! This is a hindrance to the reception of weak signals [21, 22], when high amplitude signals are received, and to the generation of a 'narrow' transmission. Modern transceivers use PL's and, above all, DDSs, and their spectrum purity close to the carrier frequency (and also at a distance, in spite of numerous filters) does not attain that of a simple crystal oscillator, even when pulled in frequency in a VXO, as used in the IC-202.

All this is perhaps slightly exaggerated, but the 10GHz specialists (and not only in France) have a lot of trouble in replacing

Fig 9.92: The completed 144MHz transceiver

their IC-202s, and several have reconditioned them to give them a new lease of life, adding on the new equipment required.

What if we replaced our IC-202s?

F9HX has written several articles for the French SHF magazine [23]. The original idea was to create a 144MHz transceiver that would have precisely the characteristics required, without any unnecessary accessories.

The transceiver principle implemented

It would have been simple to retain the IC-202 structure, ie

- A simple intermediate-frequency conversion receiver operating around 10MHz, comprising a quartz filter to obtain the desired selectivity followed by a product detector for the demodulation of the CW and SSB.
- A transmitter using the same quartz filter to reject the unwanted sideband and generate SSB.

This solution was adopted by F1BUU, and has been described in several articles [24]. But we can also generate and demodulate SSB using the phasing method [25], ie using phase converters to cancel the unwanted sideband. (For more on methods of SSB generation, see the chapter on HF Transmitters in this Handbook)

Reception is based on a single frequency change. But, since the local oscillator is on the same frequency as the signal received, the intermediate frequency is directly in the audio range. This is referred to as being at 'zero intermediate frequency', since if the modulation signal is at zero frequency, the intermediate frequency is as well (and not at 10MHz).

In English publications, the expression 'direct conversion' refers simultaneously to the single frequency change and to the zero intermediate frequency [26, 27], whereas in France some assume that direct conversion corresponds to the single frequency change, without the intermediate frequency being at zero.

The resulting transceiver is shown in **Fig 9.92** and the block diagram is shown in **Fig 9.93**.

In the receiver, the antenna is matched to a low noise FET by a simple LC circuit. The output is fed to an MMIC through a band-pass filter. This feeds a Mini Circuits double balanced mixer to demodulate the signal into I and Q audio signals. These very low level audio signals (in the order of a microvolt for VHF reception in nanovolts) are amplified by two identical channels of amplifiers fitted with automatic gain controls. They also have active low-pass and high-pass filters in order to limit the pass band received. When the level is sufficient, the two square-wave signals are phase-converted in what are known as Hilbert filters, in such a way that, when they are subsequently added together, the signals from the wanted sideband are added and those from the other are cancelled out. An elliptic 8th order filter using a capacitor switching IC gives an adjustable bandwidth from one to three

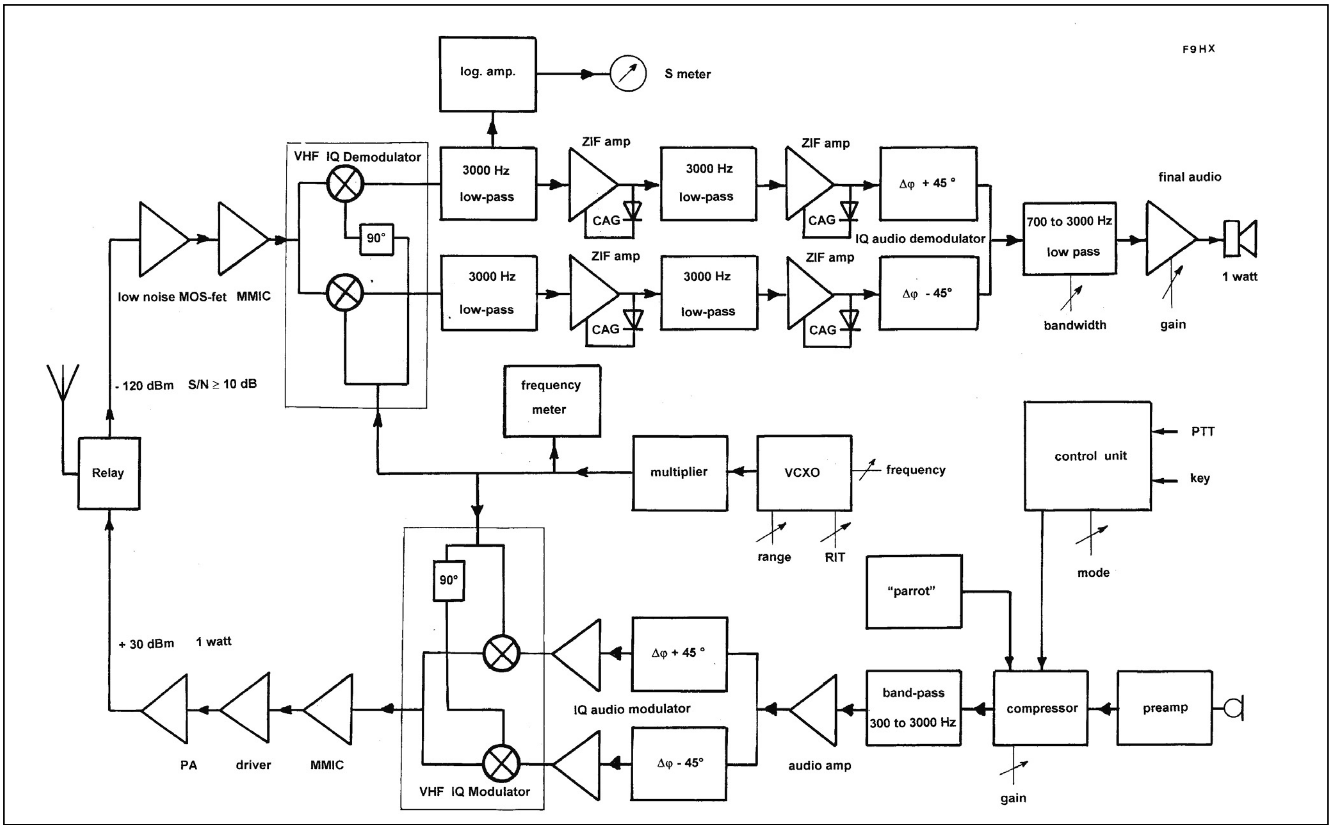


Fig 9.93: Block diagram of the 144MHz transceiver

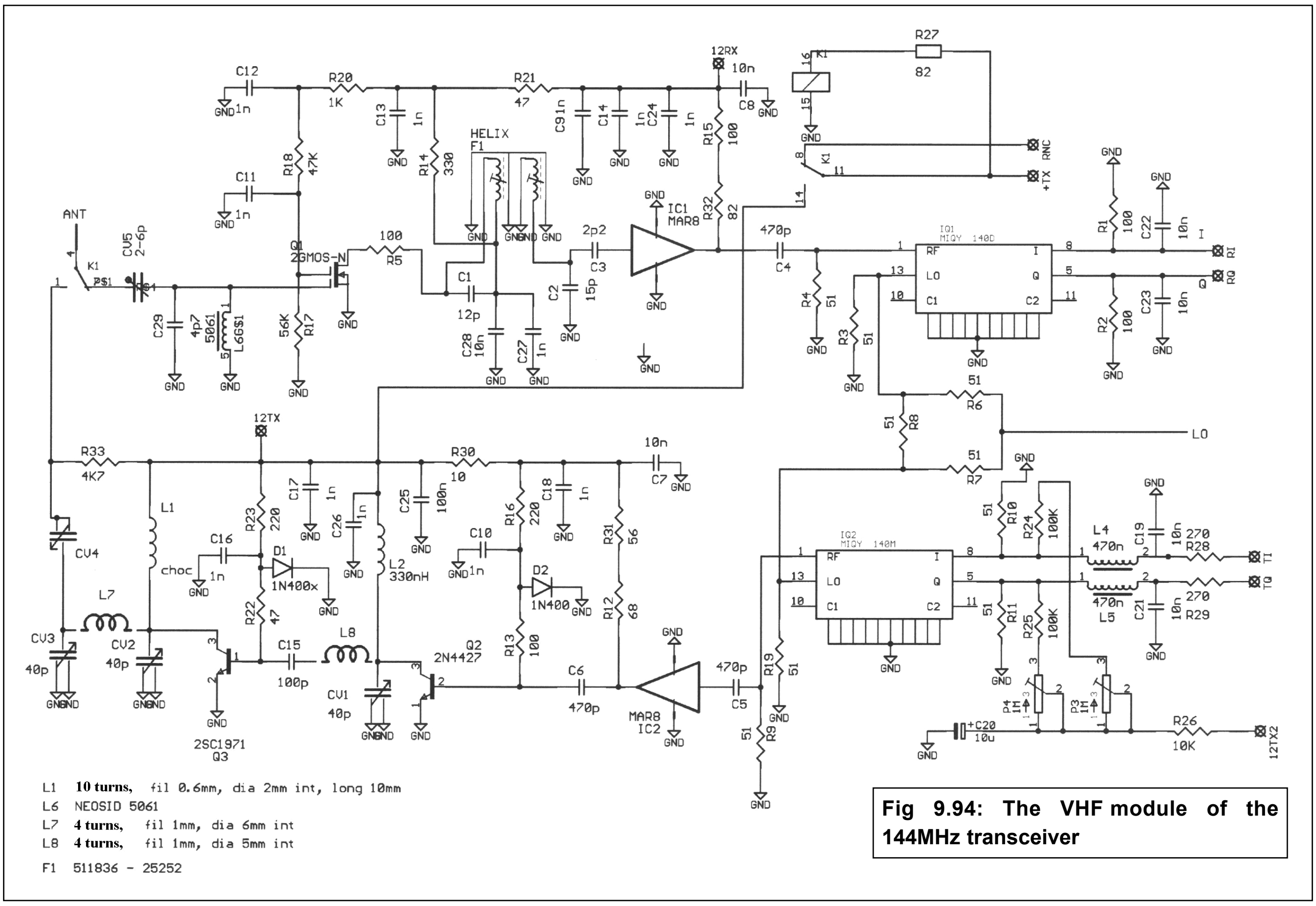


Fig 9.94: The VHF module of the 144MHz transceiver

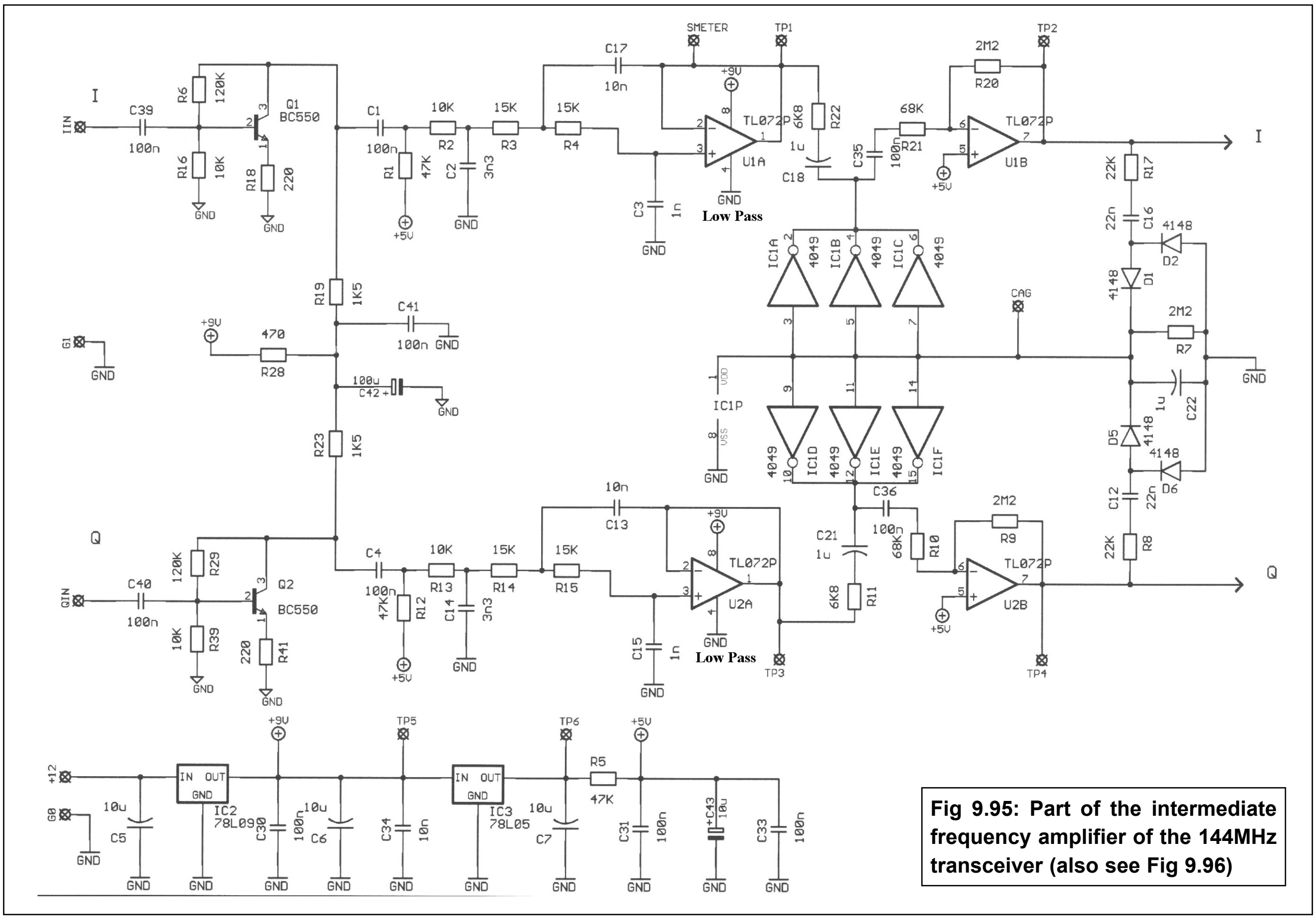


Fig 9.95: Part of the intermediate frequency amplifier of the 144MHz transceiver (also see Fig 9.96)

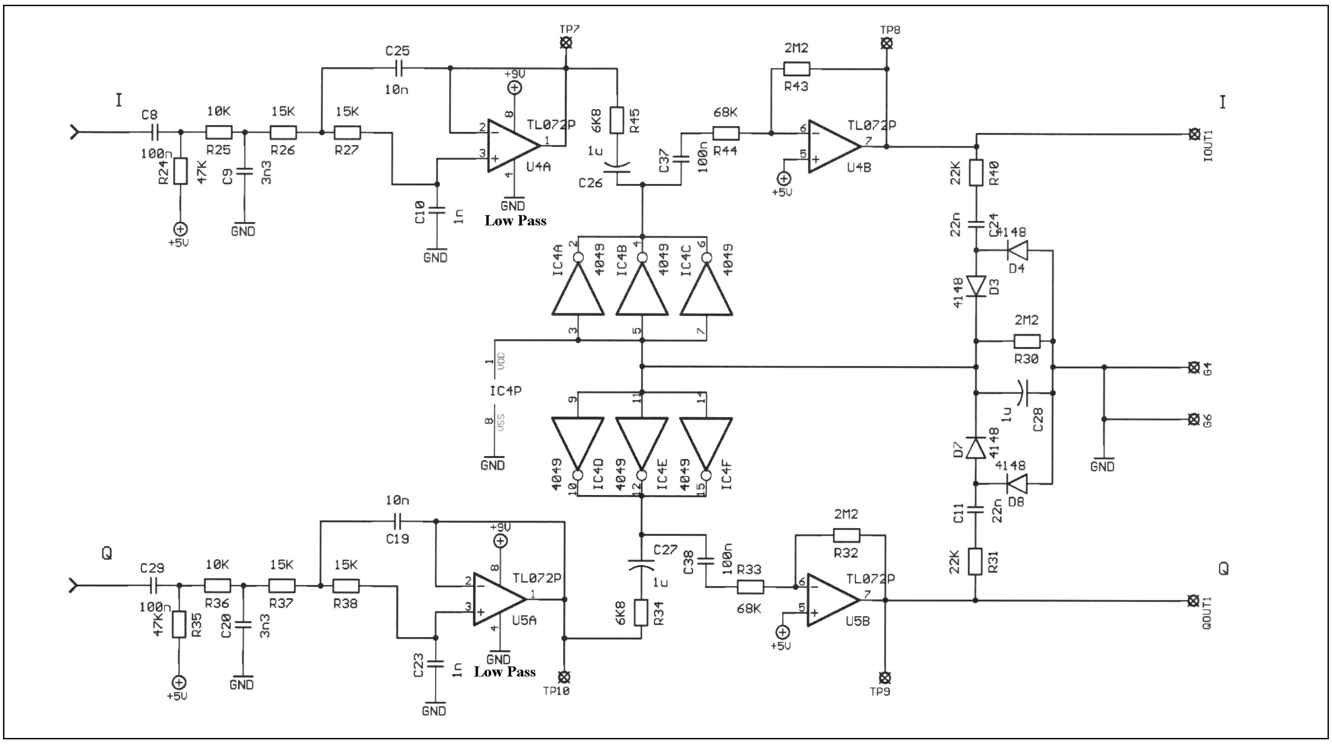


Fig 9.96: Part of the intermediate frequency amplifier of the 144MHz transceiver (see also Fig 9.95)

kilohertz. A one watt audio amplifier ensures a loud signal from the speaker. The demodulator is fed by the LO which comprises four VCXOs switched from the front panel and multiplied to give VHF reception. A logarithmic high dynamic range IC is used for the S-meter.

In the transmitter, modulation is obtained from a microphone or a one kilohertz signal for CW, tune or dots. This signal is initially amplified, then rigorously filtered to allow only the band needed for SSB to pass.

Two Hilbert filters produce two square wave signals to feed the double balanced modulator that produces a VHF SSB signal that only needs to be amplified up to the desired power. A voice record and playback IC stores a twenty second message for calling CQ.

For those interested in the theory, articles have been published in the amateur press [28, 29] explaining mathematically the functioning of this method and also the Weaver method, which is a refinement of it. Some intermediate-frequency direct conversion UHF and even SHF Weaver transceivers, up to 10GHz, are described in [30]. These are models of application of modern techniques. A direct conversion zero IF decametric receiver is described in [31].

Review of various functions of transceiver

To study the behaviour of the transceiver, modules were created to handle one or more related functions, each on a printed circuit. This also proved to be useful for the final design, and the idea of a single printed circuit was set aside for the final assembly.

Starting from the antenna in **Fig 9.94**, we first find a 50Ω relay, which handles the transmit receive switching for the VHF section.

Receive section

VHF module (Fig 9.94): The VHF signal is amplified by a low noise selective stage fitted with a robust BF 998 dual gate FET transistor, with a performance level at least equal to that of the CF 300, which is well known for its voltage fragility. A filter limits the pass band to the limits of the 2m band and feeds an untuned amplifier fitted with an MMIC. The amplified VHF signal feeds a Mini Circuits quadrature demodulator, which also receives the signal from the local oscillator described below. The output signals from the demodulator are two square wave audio signals referred to as I and Q.

Intermediate frequency amplifier (Figs 9.95 and 9.96): The printed circuit comprises two identical amplification channels with a low noise transistor at the input of each of them, followed by a low pass filter and a variable gain amplifier acting as an automatic gain control. Next is another low pass stage and another variable gain stage. The outputs from this module are thus always two square wave audio signals, but amplified, calibrated for the pass band and amplitude compressed.

Audio demodulator (Figs 9.97 and 9.98): On another printed circuit, there are two channels with different phase conversion. These are the Hilbert circuits that bring the signals from the desired sideband into phase and those from the other sideband into opposition. A passive circuit combines the two channels to obtain only the desired sideband. A first order active high pass filter and an eighth order elliptical low pass filter actively limit the pass band and play the major role in defining the transceiver band. A knob on the front panel can control the low pass filter. This adjusts the cut off frequency from 700 to 3,000Hz to cover the SSB and CW requirements.

Finally, a power amplifier stage feeds the internal loudspeaker and/or a headset.

Transmit section:

As shown in **Fig 9.99**, the signal from the microphone, which can be ceramic, electret or magnetic, is amplified by a stage followed by an adjustable compressor. Then high pass and low pass filters, as efficient as those used in the receiver, limit the pass band to 300 to 3,000Hz. An input is provided for the signal from the 'parrot' and 800Hz generator for CQ calls, CW and the generation of pips to assist when aligning parabolic antennas.

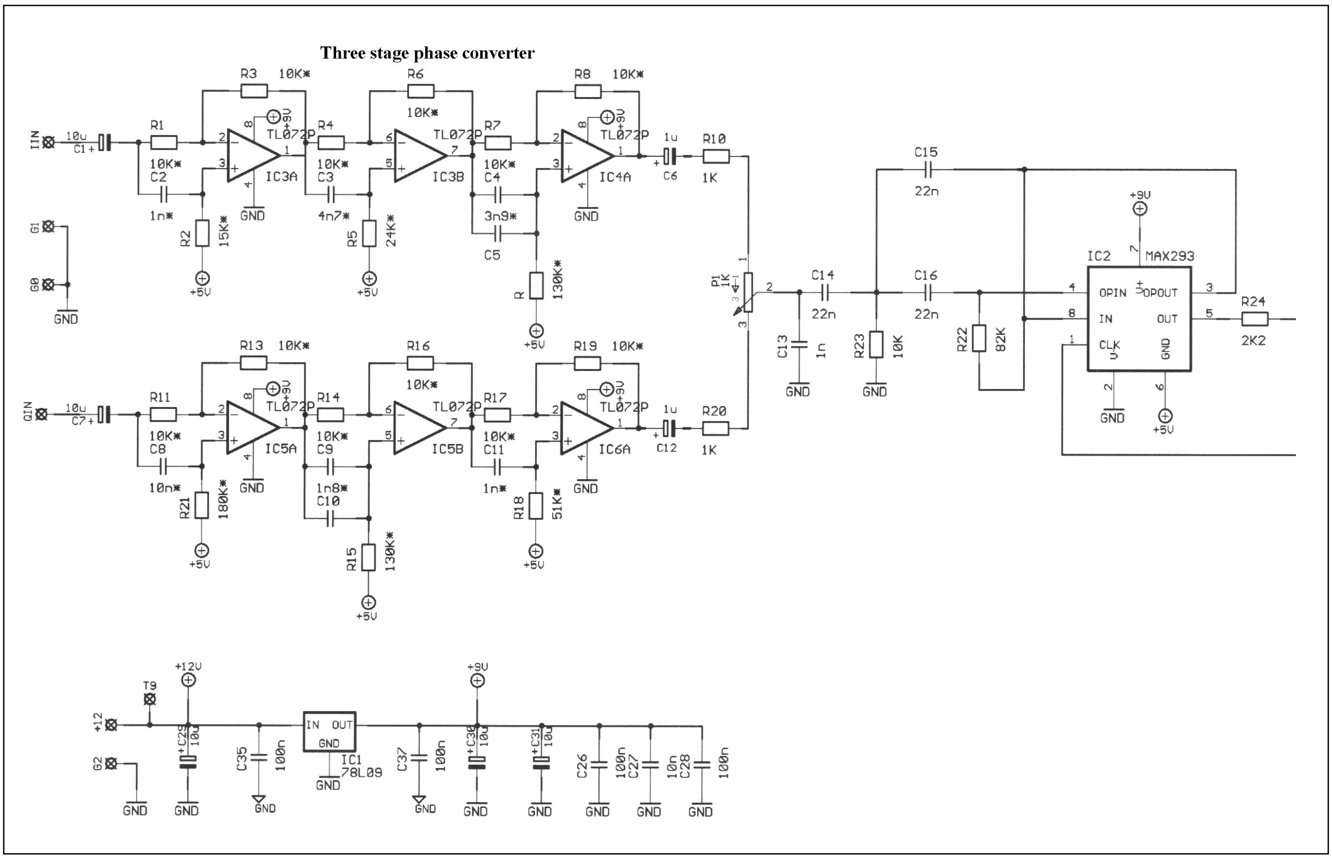


Fig 9.97: The audio demodulator of the 144MHz transceiver

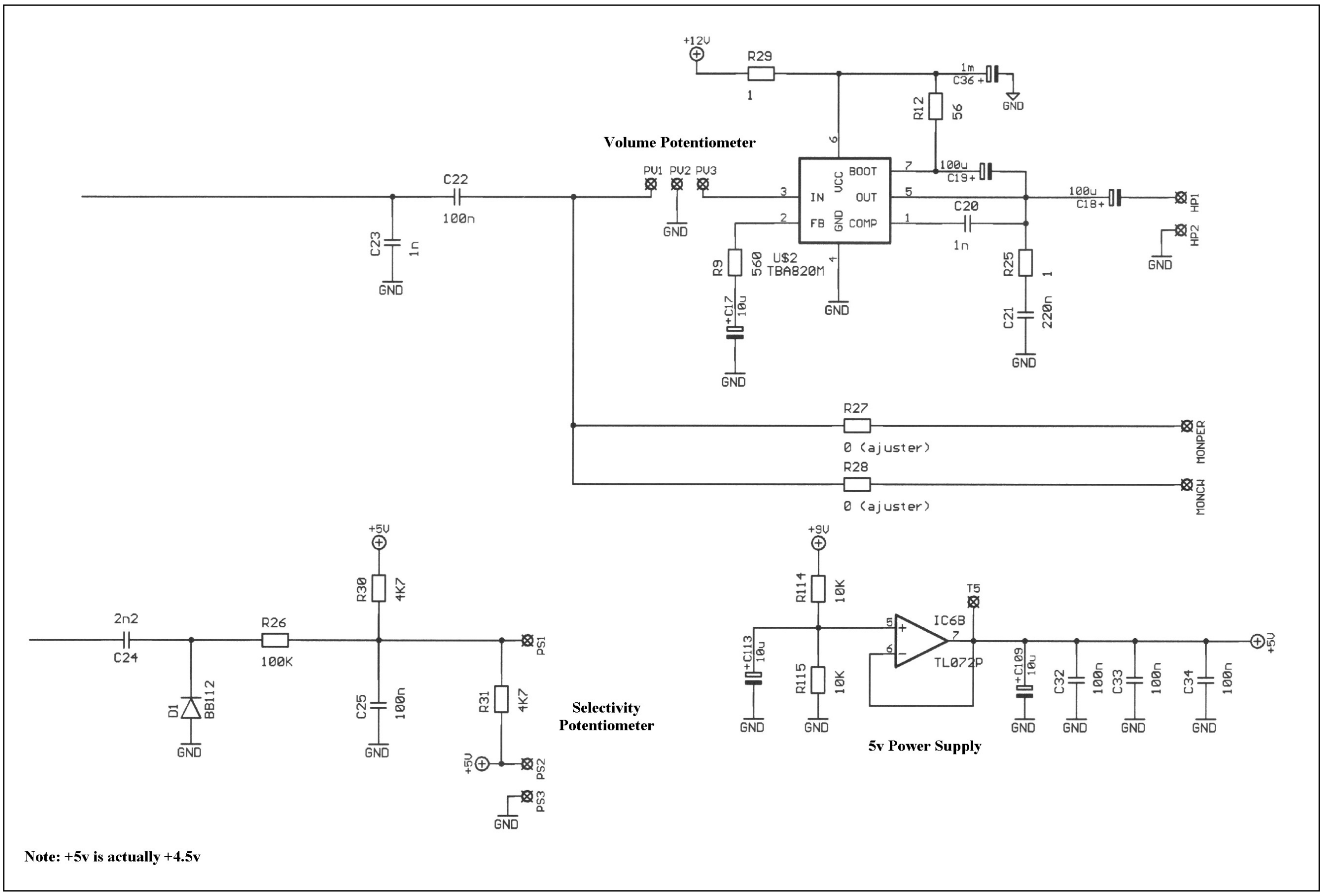


Fig 9.98: The audio amplifier of the 144MHz transceiver

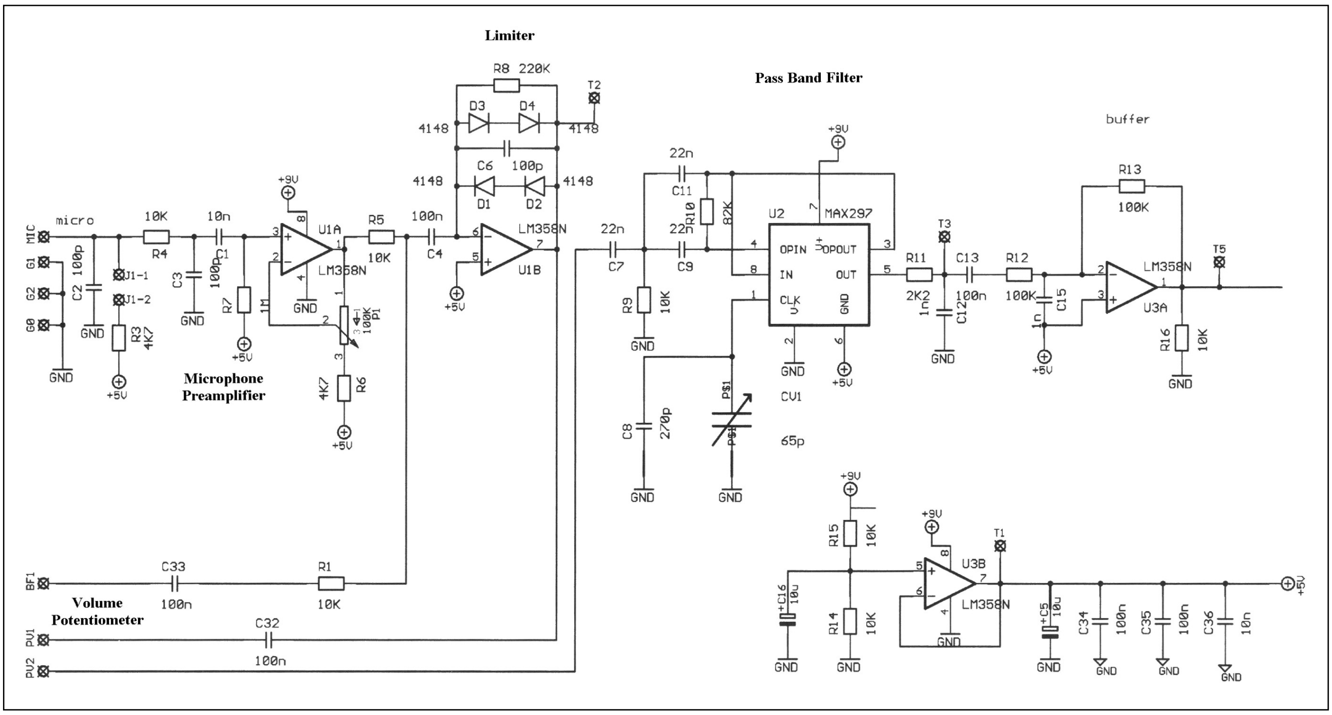


Fig 9.99: The transmit audio input of the 144MHz transceiver

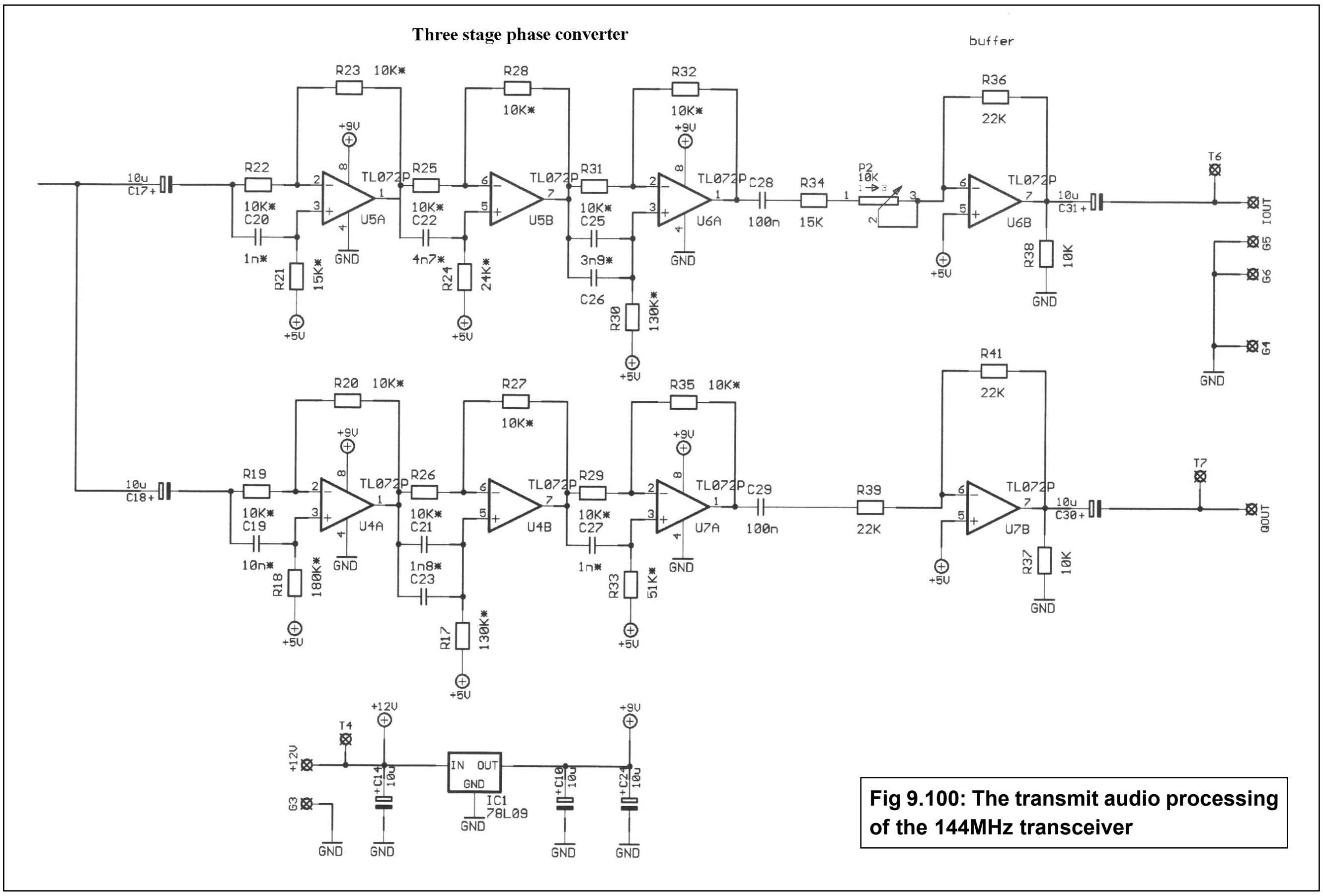


Fig 9.100: The transmit audio processing of the 144MHz transceiver

The signal is then fed to two channels, each including Hilbert phase converters to generate square wave audio signals (Fig 9.100).

On the same printed circuit as the receiver section, we find the transmit section (Fig 9.94). It receives square wave audio signals and feeds a Mini Circuits modulator, which is also fed by the signal from the local oscillator. The local oscillator is on a separate module, and is divided into two outputs by a 3dB resistive divider to feed the receive demodulators and transmit modulator. The output from the modulator is amplified by an MMIC, followed by two temperature stabilised class AB stages, each having a diode thermally linked to its casing. The output power can be adjusted

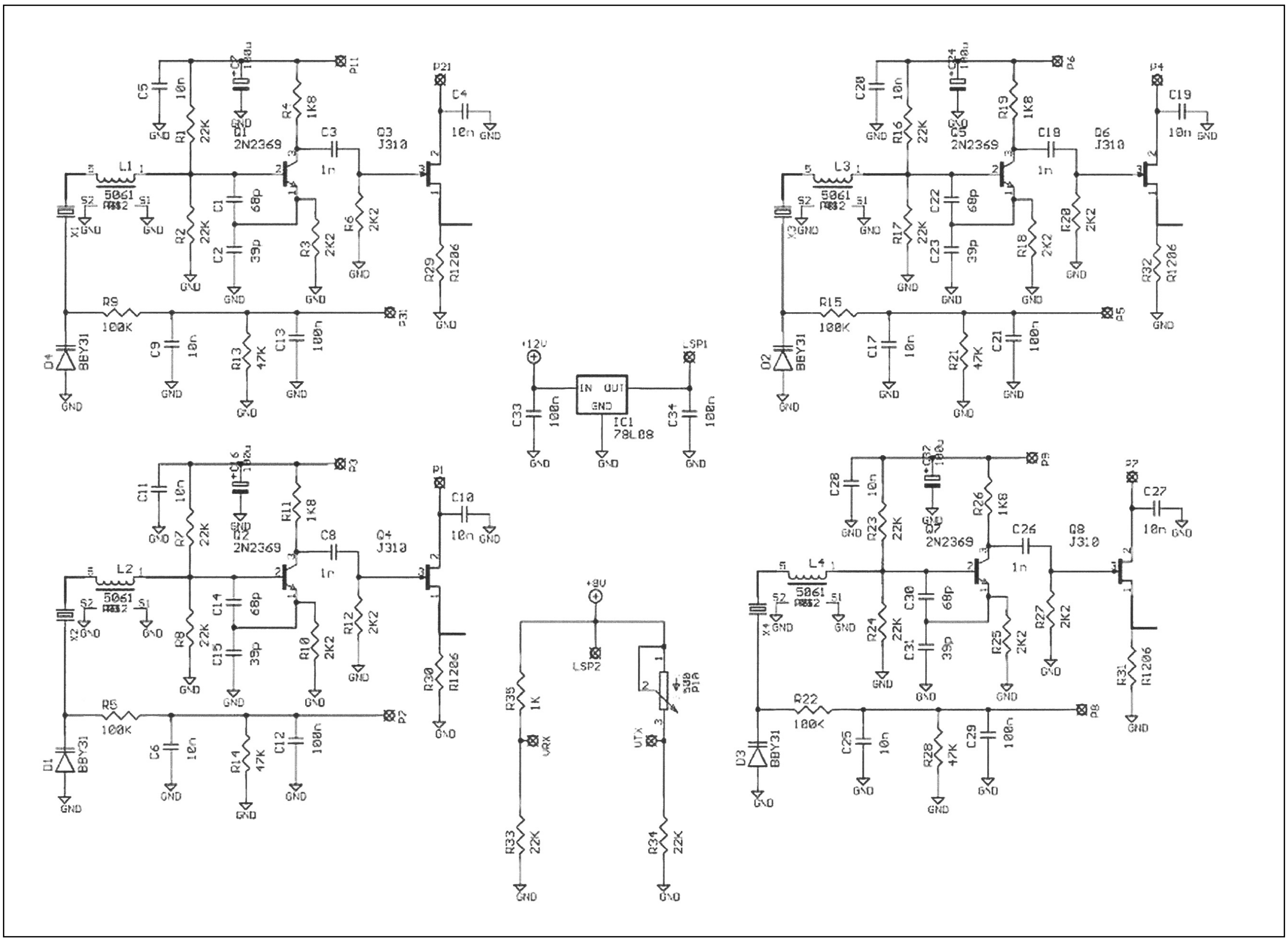


Fig 9.101: The local oscillator of the 144MHz transceiver

using a knob located on the rear face of the transceiver by controlling the level of I and Q signals feeding the modulator.

Local oscillator

This consists of a 24MHz VXO, the frequency is adjusted by means of a varicap diode, and a ten turn potentiometer (**Fig 9.101**). An RIT can be used for reception using a potentiometer with a notch at the central position, thus a click can be felt when the knob is rotated. A switch makes it possible to select one of four oscillators to cover four ranges of at least 200kHz within the 144-146MHz band. In contrast to the VXO of the IC-202, the switching is not effected via VHF, but through the DC feed of the selected oscillator. This avoids interference from other capacities, which would reduce the range covered by the varicap diode. The crystals used on the equipment and the ranges covered are:

- 24.038MHz crystal:- 144 to 144,200MHz
- 24.071MHz crystal:- 144.271 to 144.400MHz
- 24.133 MHz crystal:- 144.600 to 144.800MHz
- 24.172 MHz crystal:- 144.800 to 145.000MHz

The oscillator is followed by the multiplier stages and an amplifier stage, to provide the level required in the 144MHz band (**Fig 9.102**). In the same module there is a divider (x 10) supplying the signal for the frequency meter.

Auxiliary circuits

The following auxiliary functions are on a single printed circuit (**Figs 9.104 and 9.105**):

A DC voltage regulator, with reverse polarity protection, limits the voltage applied to various modules to 12V. The other modules include second regulation if necessary, for example for the VXO.

The PIC-based circuit controls the selection of the type of transmission and its generation: CW, SSB, pips, message, tune. It also controls the switching from transmit to receive, with a 'K' at the end of the message. Signals in CW, tune and pips are at approximately 800Hz.

An S-meter, using a logarithmic amplifier, receives one of the audio signals, taken from the output of the first IF amplifier stage, before the automatic gain control. This allows a linear deviation, in decibels, from the signal received (scale 100dB).

Frequency meter

The 24MHz signal generated by the local oscillator is divided by ten using an ECL divider to feed the frequency meter module. This is made up of a gate, two counters (16 bit counting), a PIC and a two line by 16 character back lit display (Fig 9.105 and **Fig 9.106**).

The PIC and its 20MHz crystal control the frequency meter, generating the gate opening time (0.1 seconds) and all the signals required for LCD display. The PIC code is optimised in order to measure frequencies in the 2m band, with a refresh time in the order of 120ms. The PIC clock frequency can be adjusted using a capacitor for an accurate display.

The analogue/digital converter function of the PIC is used to produce an S-meter display as a bar graph with a length proportional to the logarithm of the signal received.

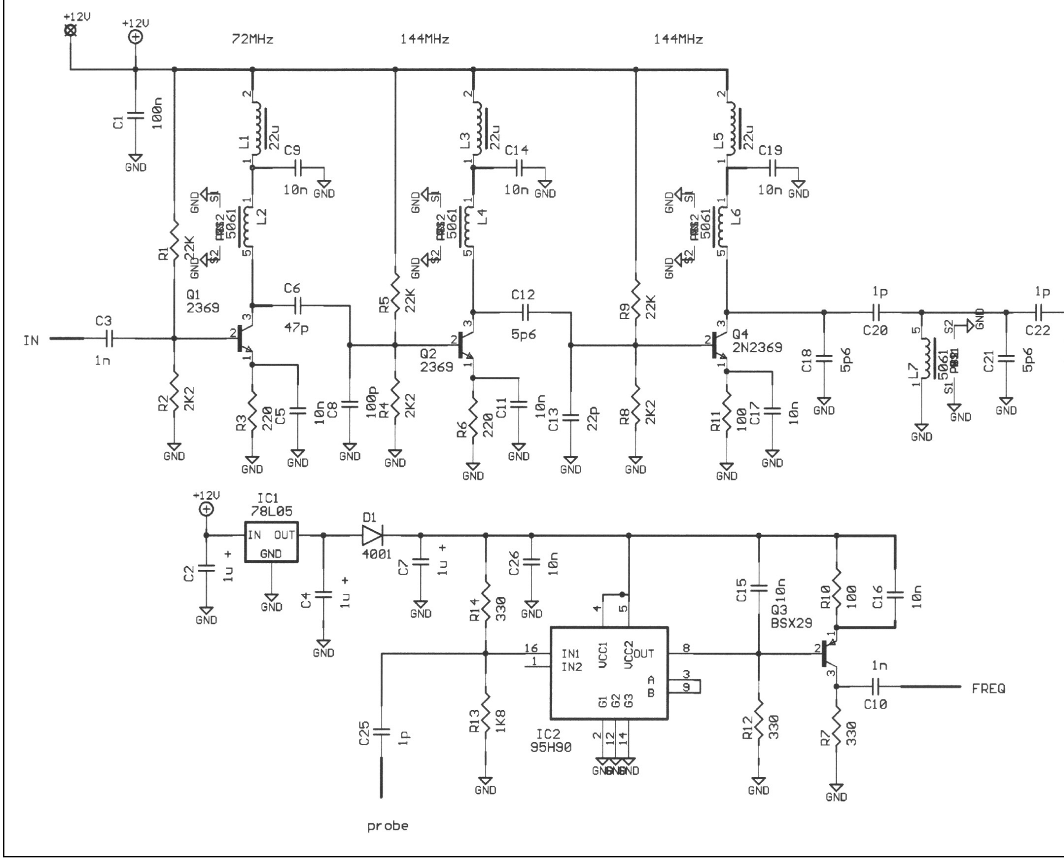


Fig 9.102: The local oscillator multiplier of the 144MHz transceiver

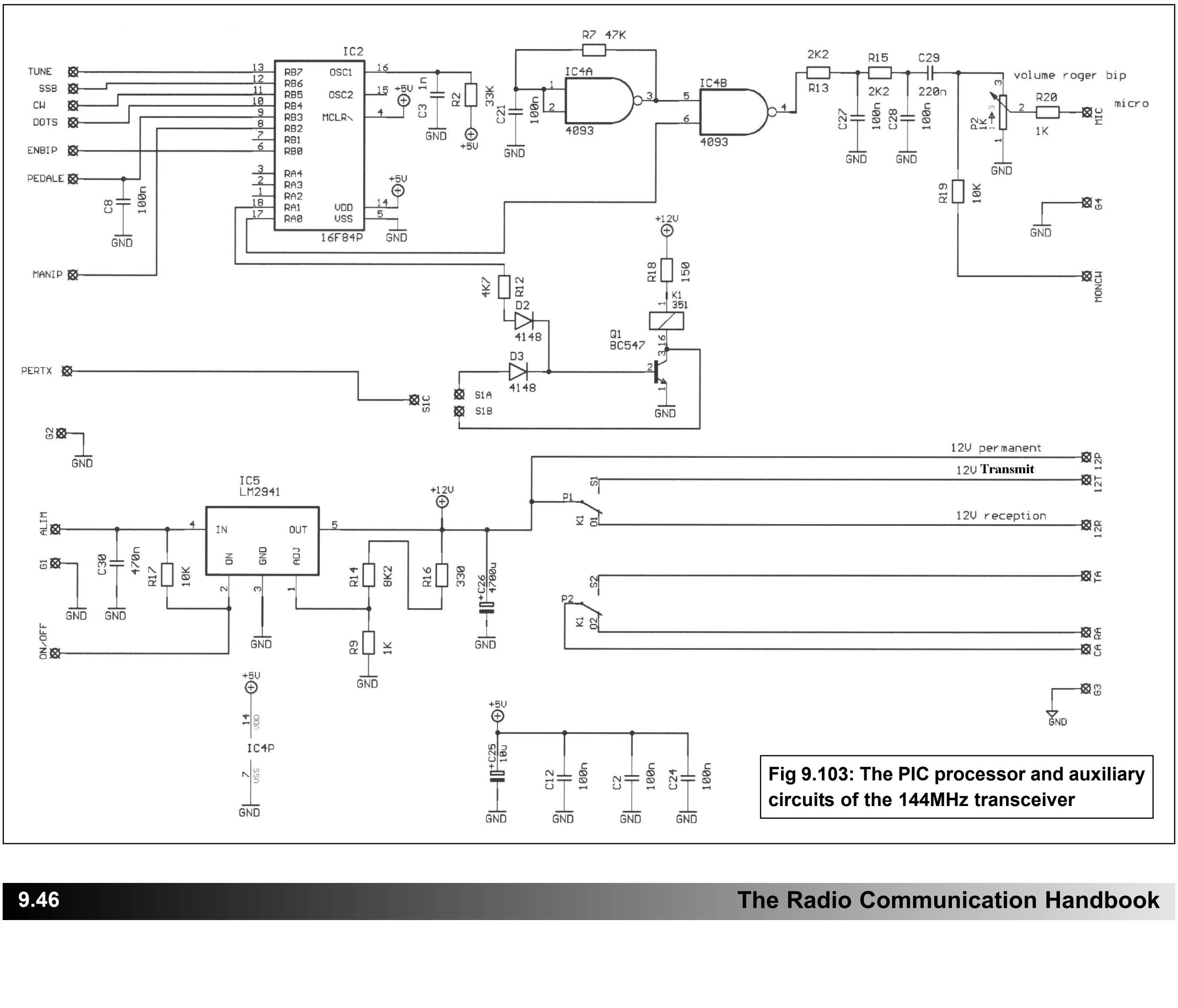


Fig 9.103: The PIC processor and auxiliary circuits of the 144MHz transceiver

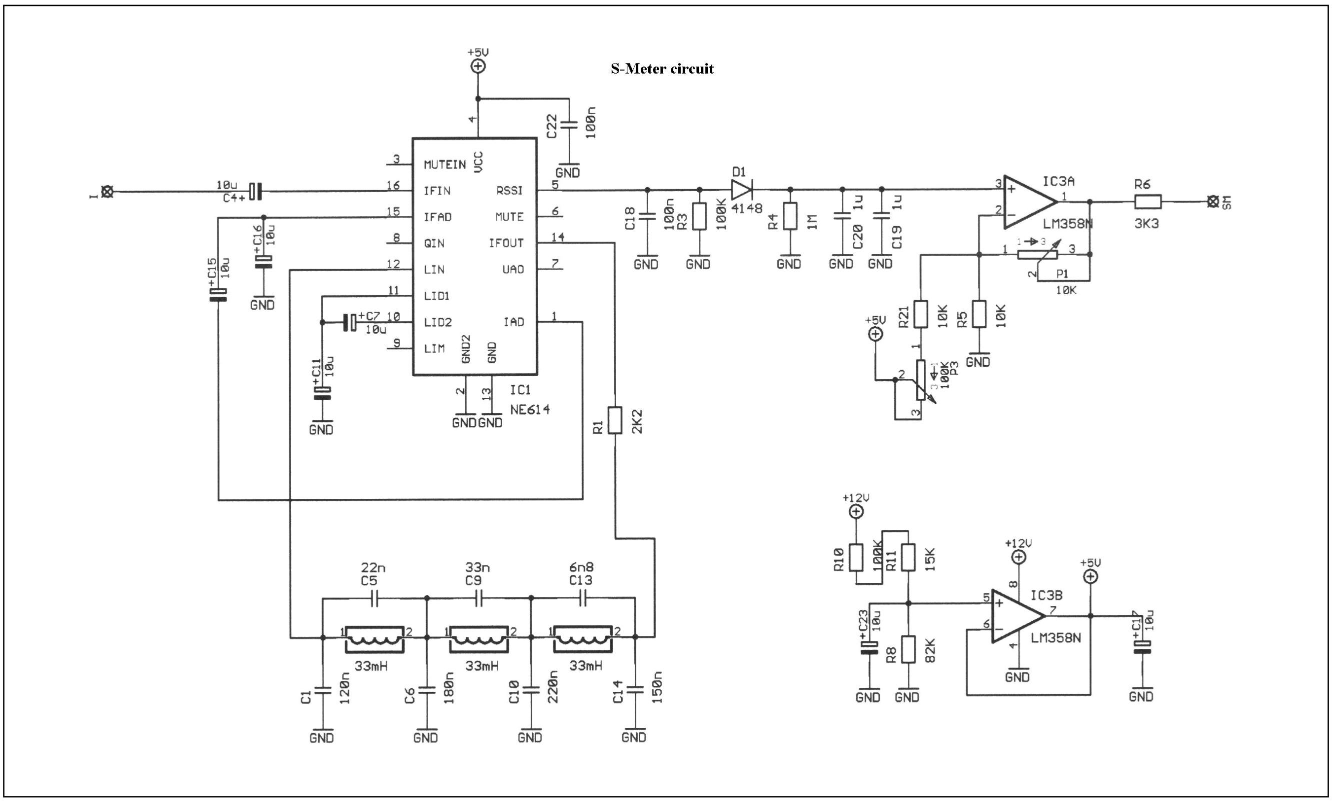


Fig 9.104: The S meter of the 144MHz transceiver

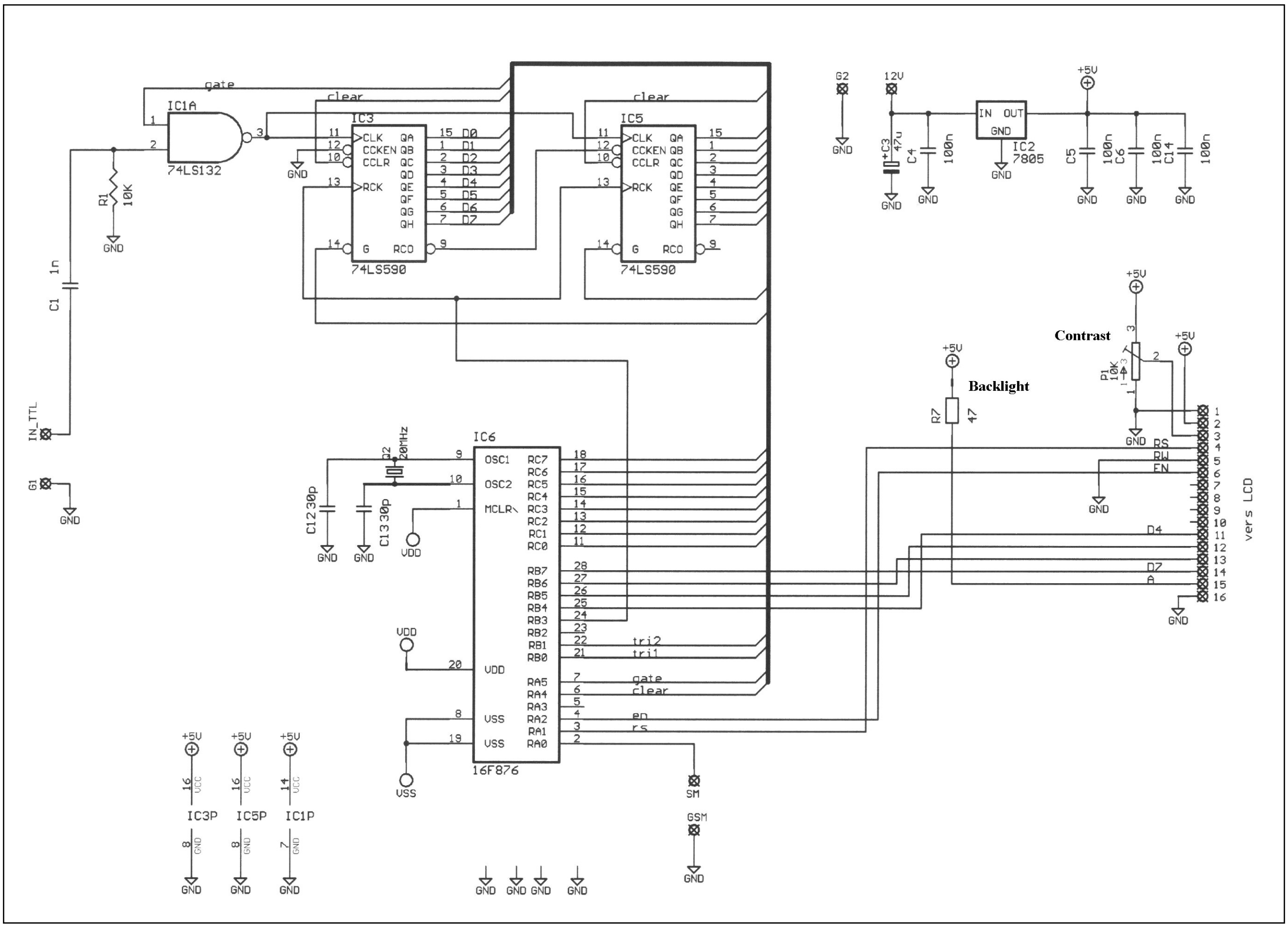


Fig 9.105: The frequency counter of the 144MHz transceiver

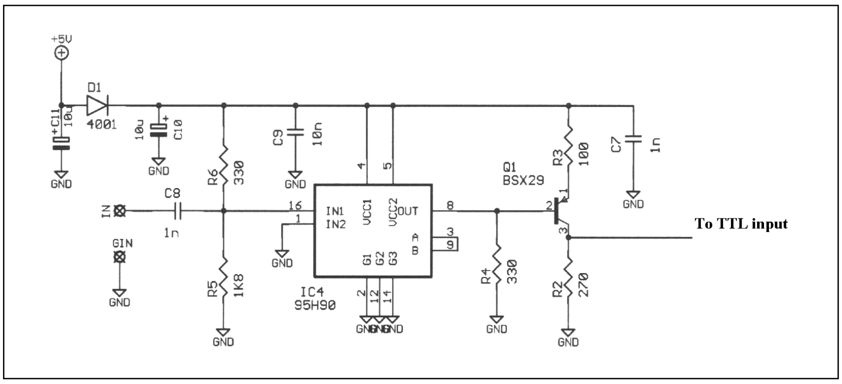


Fig 9.106: The divider for feeding the frequency counter

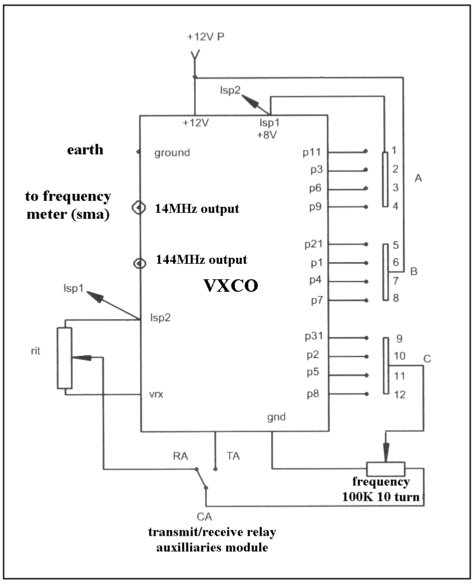


Fig 9.108: Local oscillator interconnections

Parrot

A recorder repeater makes it possible to modulate the transmitter, using a single or repeated message (Fig 9.107). The recording uses an electret microphone, mounted at the back of the transceiver, with knobs for the various operations necessary.

Switching operations

Two switches make it possible to select the range of frequencies received and the functioning mode. Figs 9.108 to 9.110 show how they are connected up to the various modules.

Assembly

The transceiver is housed in a metal box with only the controls strictly necessary for SHF operating on its front panel:

- Selection of range covered at 144MHz
- Frequency control
- RIT
- Receive audio volume
- Receive pass band
- Transmission type selection: message, tune, SSB, CW, pips
- Frequency and S-meter display
- Green indicator light: reception
- Flashing red indicator light: transmission
- Microphone socket

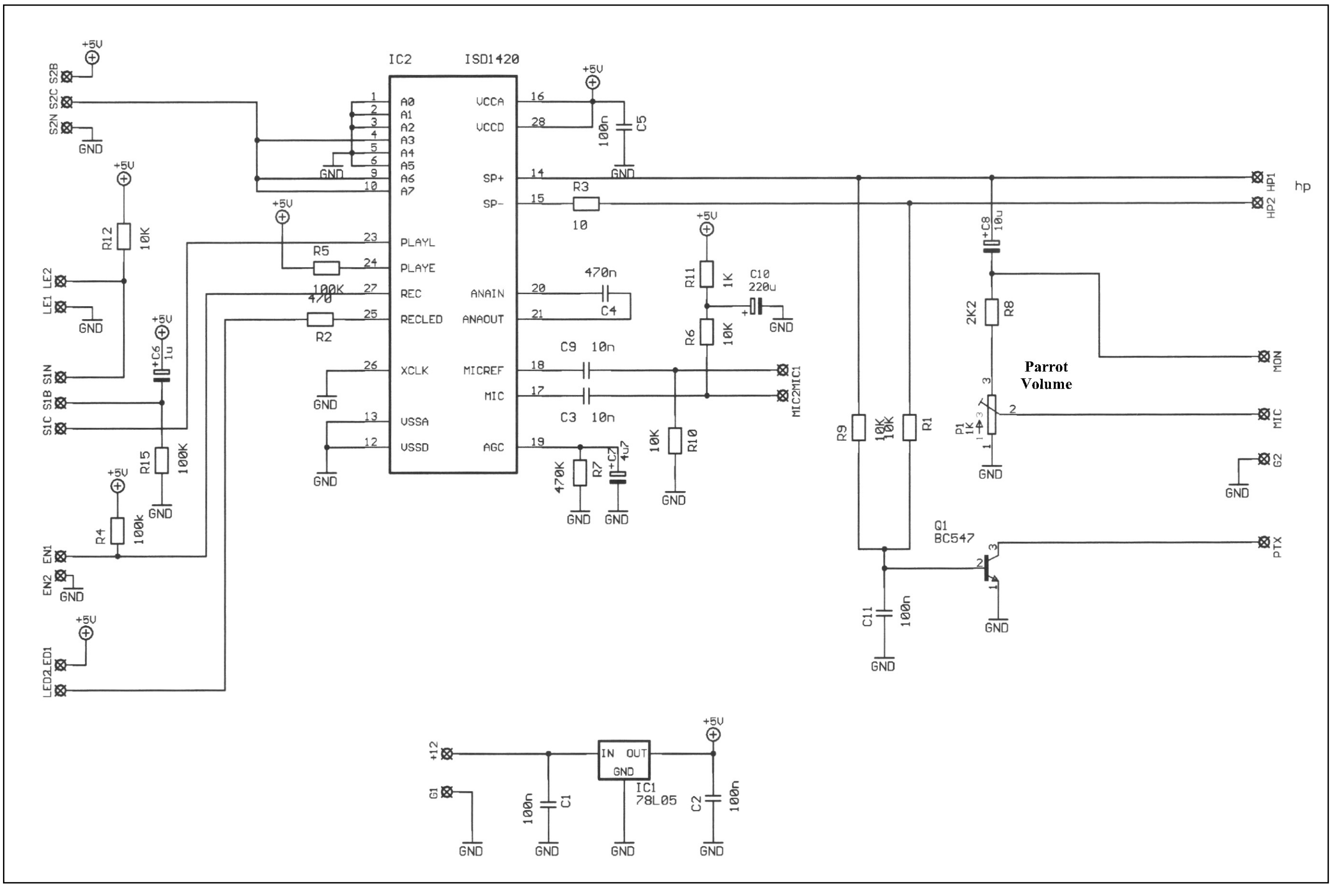


Fig 9.107: The parrot of the 144MHz transceiver

144MHz Transceiver Specification

Dimensions:	292 x 230 x 103mm, weight 1.8 kg.
Power supply:	12 to 15 volts for full transmit power, reduced power operation from 10 to 12 Volts
Transmit power at output:	1 Watt (+ 30dBm) in tune, CW, peak SSB positions
Sensitivity:	0.16µV (-120dBm) with S/N = 10dB for 3kHz pass band for a signal modulated at 1,500Hz
Receive AGC:	For 100dB variation in the received signal, the output varies by only 30dB
Pass band:	
Transmit:	300-3,000Hz, with –50dB drop at 4,000Hz (**Fig 9.111**)
Receive:	Adjustable on front panel from 700 to 3,000Hz, with –50dB drop at 4,000Hz for setting at 3,000Hz
Audio power:	1 Watt into 8 ohms

The other knobs and jacks are at the rear, since they do not have to be operated during a contact:

- 12 Volt input
- Fuse
- Connector for 144MHz input and output
- Parrot microphone
- Indicator light and operating knobs for same
- Jack for loudspeaker or headset, without internal loudspeaker cut-off
- Jack as above, but with cut-off (can also be wired up for insertion of a DSP)
- Jack for key
- Transmit power adjust knob

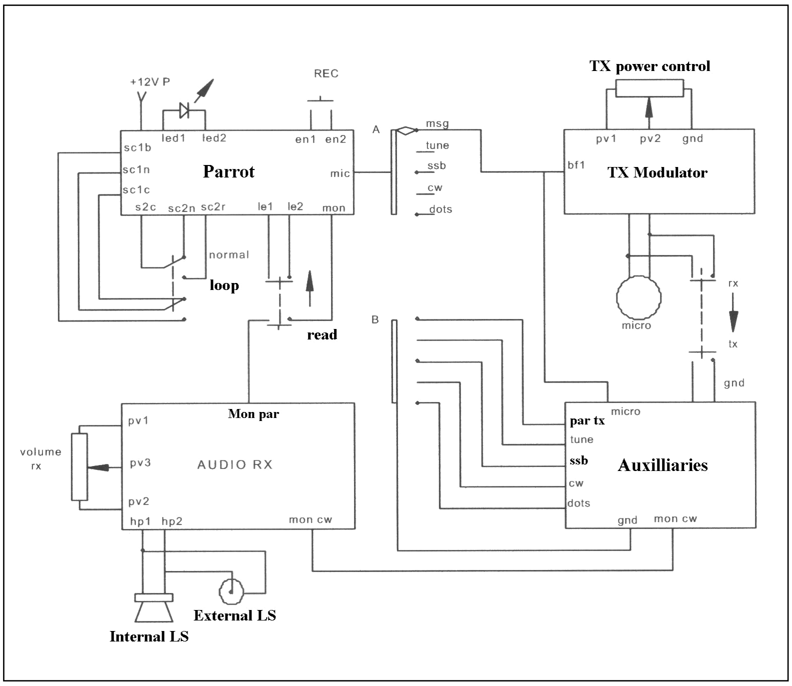


Fig 9.109: Main interconnections for the 144MHz transceiver

All the modules are mounted on printed circuit boards (single sided or double sided, depending on requirements). The finished prototype is shown in **Figs 9.112 and 9.113**.

The RF and local oscillator modules are housed in tinned sheet metal enclosures, which measure 74 x 111 x 50mm. (reduced to 40mm.) and 74 x 111 x 30mm. respectively. 1nF feed through capacitors are used for all the connections unless they are high frequency links. The antenna is connected via a TNC connector, providing for a very reliable contact, as with an N type, but taking up less space. Professional standard BNCs can also be used, but the mass market models should not be used, since they generate crackling due to earth contact resistance variations. The lower frequency connections and the frequency meter use SMA connectors. They are assembled using CMS components, except the RF power circuits. The interconnection is in the form of a star, using normal wire for DC circuits, shielded wires for the audio circuits, and small diameter coaxial for RF.

Using the transceiver without an SHF transverter

As with all receivers based on the principle of direct conversion to zero intermediate frequency there is a phenomenon that can be a real nuisance. This is the reception of interference from very strong signals, AM or SSB even if they are located outside the band. This is due to their demodulation by the demodulator's diodes, as soon as their detection threshold is reached. In a receiver with an intermediate frequency of about 10MHz, this detection produces an audio signal that is removed by the intermediate frequency amplifier. By contrast in this design the signal is amplified. We hear powerful stations from the 144MHz band, and even outside it, if they are strong enough to get through the filters located on the VHF module. This does not cause problems when the transceiver is used with an SHF transverter, but using it on 144MHz from a good location can quickly lead to problems due to the reception of interference. Trials now in progress appear to show that a solution requiring only some modest additions to the intermediate frequency/AGC circuit would lead to an appreciable reduction in this fault.

Another problem has been noted with radiation from the local oscillator. In a traditional receiver, this is not in the band of the frequencies received, but offset by the the intermediate frequency, if it emits radiation it is not noticed. By contrast, our transmitter emits radiation at the reception frequency, and can be heard by another receiver if it is nearby. This can be a possible source of conflict with neighbours!

Some direct conversion receiver makers have been worried by microphony. By avoiding any ceramic capacitors in the low level audio circuits, we can exclude the microphonic behaviour of these components.

Tests and adjustments

Each printed circuit board should be carefully inspected, preferably under an illuminating magnifying glass, to detect any poor soldering or any bridges between connections.

Each module should be individually tested before being fitted into the housing – initially, without putting the integrated circuits in place on their sockets. The modules run at 12 volts, apart from auxiliary equipment circuits, which operate at 13 volts. We can check whether the voltages are correct at various significant points on the circuit. Then, with the integrated circuits in place, we can proceed to the functioning tests.

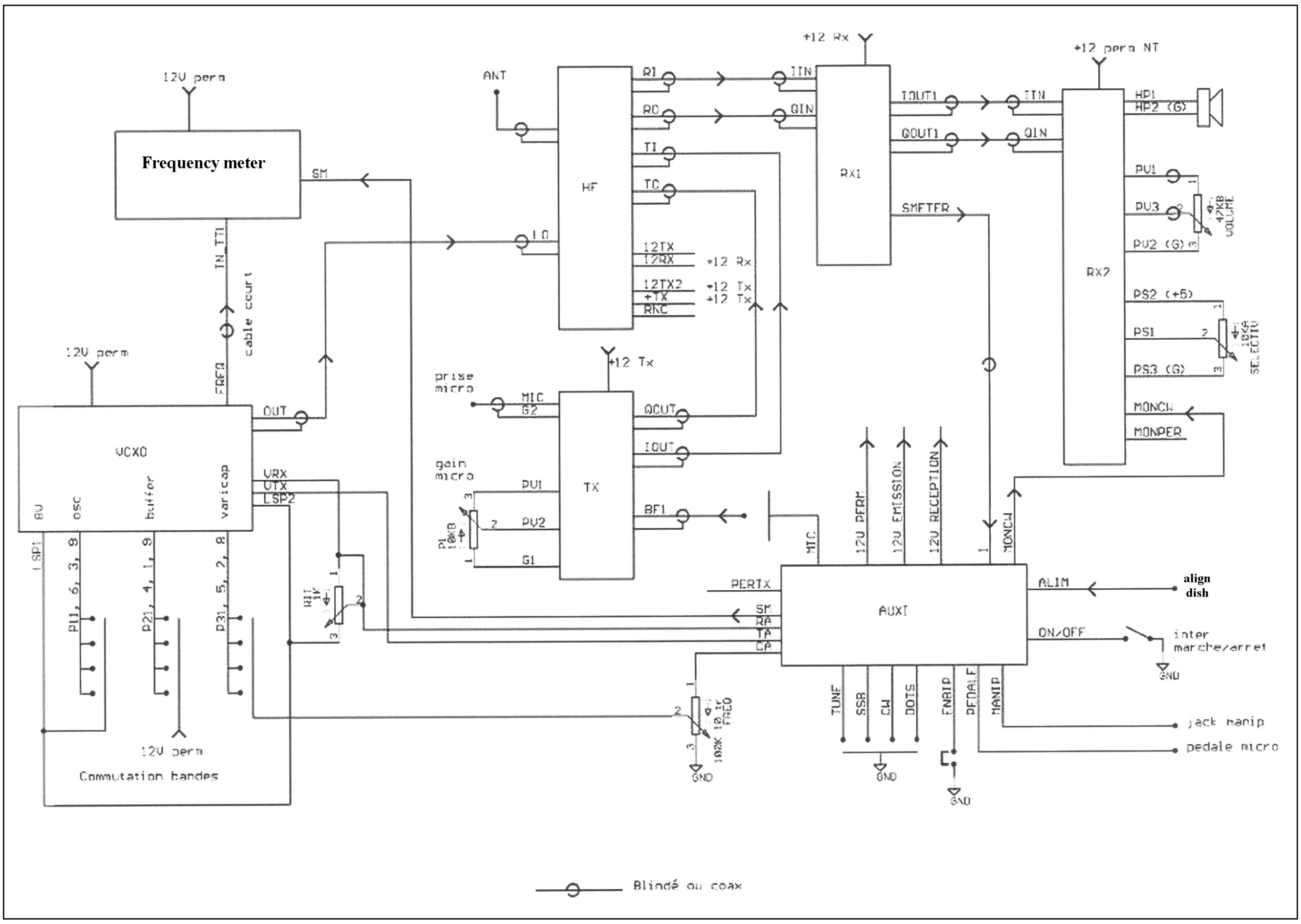


Fig 9.110: Details of all interconnections for the 144MHz transceiver

The VXO should be tested and adjusted for each of the crystals to obtain the band coverage required. The multiplier section should be set for maximum output at 145MHz. The intermediate frequency module is tested using an audio generator by putting the two inputs in parallel.

The two outputs should supply the same amplified and limited signal, thanks to the AGC. The second receiver module can be tested using an audio generator, preferably followed by a buffer stage supplying two square wave signals. The transmit modulator should be driven by an audio generator, to check that the outputs are square wave. The compressor should be adjusted as desired.

The VHF module should be tested in receive mode, with the local oscillator connected. Applying a 145MHz signal to the TNC socket will make it possible to adjust the input stage by measuring the output voltages I and Q, the input VHF signal being offset by approximately 1.5kHz in relation to the local oscillator frequency. It is possible to adjust the three adjustments, antenna and coupling, to obtain a flat pass band of between 144 and 146MHz, or to favour the most frequently used range, for example, from 144 to 144.200MHz for use with a 10GHz transverter. The phase difference of outputs I and Q can be checked by creating a lissajous figure with a two channel oscilloscope, which should show a circle.

During transmit, with the local oscillator connected, two square wave I and Qs, produced as referred to for the receive section will make it possible to generate a 145MHz signal. This can be tuned for a maximum using the adjustable capacitors.

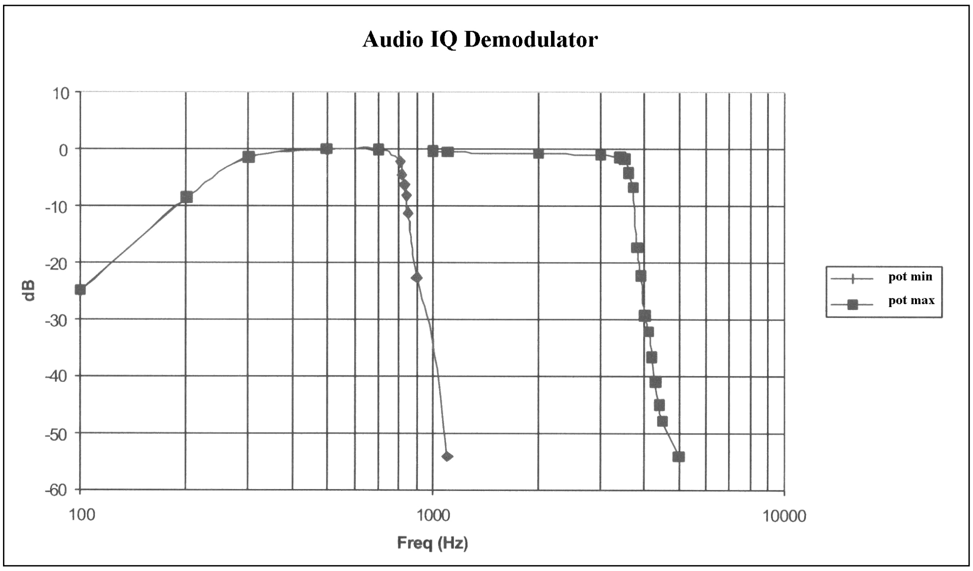
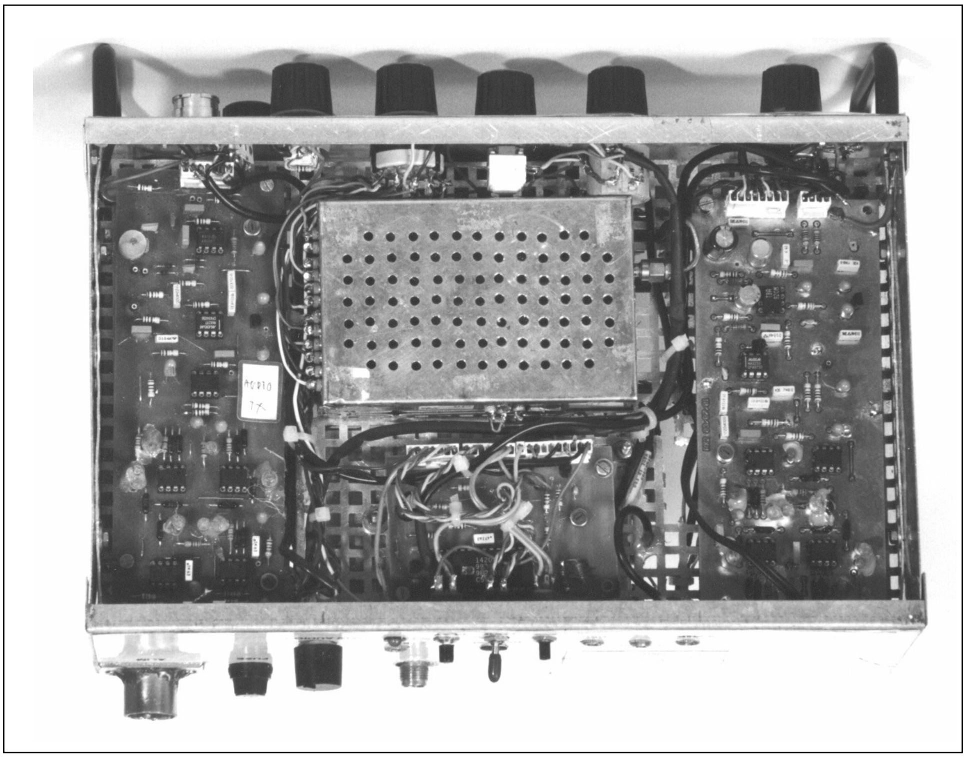
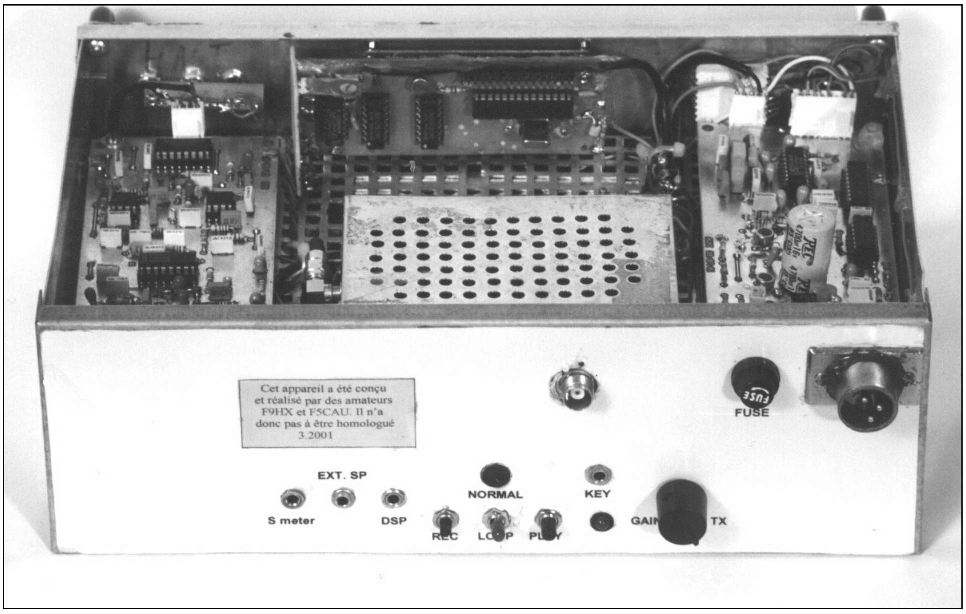


Fig 9.111: Frequency response of the audio filter of the 144MHz transceiver





(Top) Fig 9.112: Underside of the completed 144MHz transceiver

(Bottom) Fig 9.113: The top side of the completed 144MHz transceiver

When all the modules are functioning correctly, it is time to install them into the housing, to check that the assembly is operating properly, and to fine tune the settings.

Conclusion

This long project took two years to complete After several months of use on activity days, some very satisfactory results were achieved at 10GHz with numerous contacts at distances of close to 500 kilometres, whilst using a very modest parabola with a diameter of 48 centimetres.

F5CAU's contribution

F5CAU carried out the work of designing printed circuits to professional standards, and also designed the frequency meter, the 800Hz generator and the PIC for the auxiliary modules. In addition, as the assembly instructions are too extensive to be published in full here, he has posted it on his Internet site [32], where the printed circuit drawings and some other useful documents can be download free of charge.

TRANSVERTERS

If you already have an HF band transceiver, one of the easiest ways of getting onto the VHF and UHF bands is to use a transverter. This takes the output of your transceiver, usually the 28 - 30MHz band, converts it to the chosen VHF or UHF band and converts received signals on the VHF or UHF band so that they are received on your transceiver. The transceiver output and input is commonly called a tuneable IF. The advantage of this approach is that all of the facilities of your HF transceiver are available on the VHF or UHF band.

This section contains transverter designs for all of the VHF and UHF bands. The first set of designs cover 6m, 2m and 70cm using similar circuits. The 4m transverter design was used as a club project by the Andover Radio Amateurs Club [33].

2m Transverter

In 1990, Wilhem Schüerings, DK4TJ, and Wolfgang Schneider, DJ8ES, presented a paper at the 35th VHF Congress in Weinheim on a universal transverter concept [34] and [35]. The following design is the resulting 28/144MHz transverter [36]. It should be possible for the transverter to directly feed a standard power amplifier, the design of a suitable amplifier using a hybrid amplifier module is shown.

Transverters for the 2m band are always of interest, in an attempt to match the current state of art in amateur radio technology this transverter was developed using modern components to convert the 144 - 146MHz range into the 10m band.

Concepts such as high-level signal strength and oscillator signal spectral purity have taken on increasing significance. It is also important that the equipment can be reproduced easily. The transverter described below represents a circuit that corresponds to today’s requirements.

Fig 9.114 shows the circuit of the 28/144MHz transverter. The local oscillator uses a tried and tested U310 crystal oscillator at 116MHz [note: the U310 has been superseded by the J310].

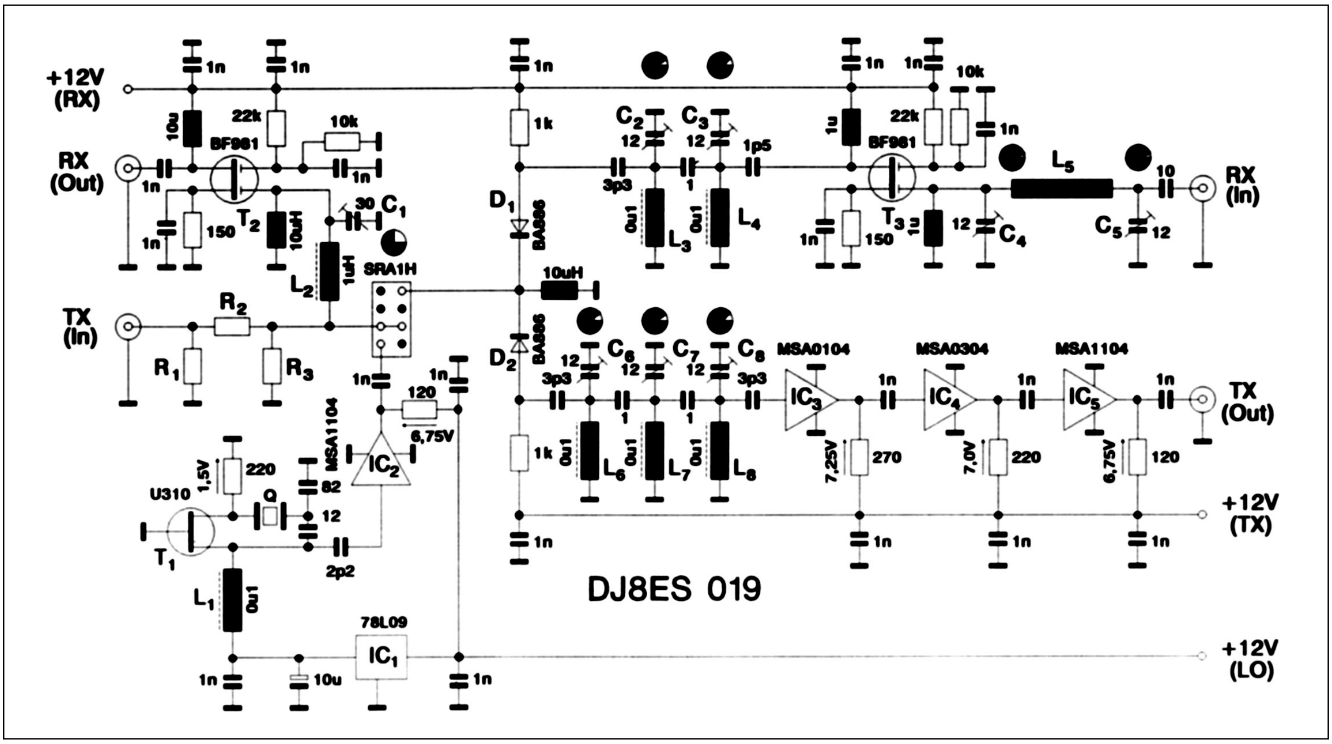


Fig 9.114: Circuit diagram of the 2m transverter. Note: the U310 has been superseded by the J310 and the BF981 by the BF992

This signal is amplified by the next stage, an MSA1104 MMIC, giving an output level of 50mW. The SRA1H high-level ring mixer requires a local oscillator level of +17dBm (50mW) and can be used at up to 500MHz.

A pi attenuator, consisting of R1 to R3 is used to the control the output from the driving transmitter (IF). For a 'clean' signal (intermodulation products <50 dB), the ring mixer must be driven by a maximum of 1mW (0dBm) at the IF port. Table 9.6 shows the resistance values needed for the attenuator for various IF input power levels. The attenuator uses standard value resistors. The attenuator also acts as a 50 ohm termination for the ring mixer.

The converted receiver signal is fed from the mixer by L2 and C1 to a high impedance amplifier using a BF992 low-noise transistor to give the required intermediate frequency amplification.

The 2m received signal is fed to the gate of the BF992 RF amplifier through a pi filter from the 50 ohm aerial input. The RF amplifier is followed by a two-pole filter. The received signal is switched to the ring mixer by the +12V receiver supply voltage through the PIN diode, D1 (BA886).

In transmit mode, diode D2 is activated. The 2m signal from the ring mixer first passes through a three-pole filter. The signal is then amplified by three MMIC amplifiers (IC3, IC4, IC5). The combination of MSA0104, MSA0304 and MSA1104 guarantees an output level of 50mW (+17dBm). The transverter can be used with any power amplifier but additional harmonic filtering is recommended.

The 28/144MHz transverter is assembled on a double sided PCB measuring 54mm x 108mm. The board can be mounted in standard tinplate housing of 55.5mm x 111mm x 30mm, Fig 9.115 shows pictures of the completed transverter. The board

Pin	dB	R1 in	R2 in	R3 in
1mW	0	–	0	51
2mW	3	300	18	300
5mW	7	120	47	120
10mW	10	100	68	100
20mW	13	82	100	82
50mW	17	68	180	68
100mW	20	62	240	62

Table 9.6: Resistance values for attenuator used in 6m, 2m, and 70cm transverters

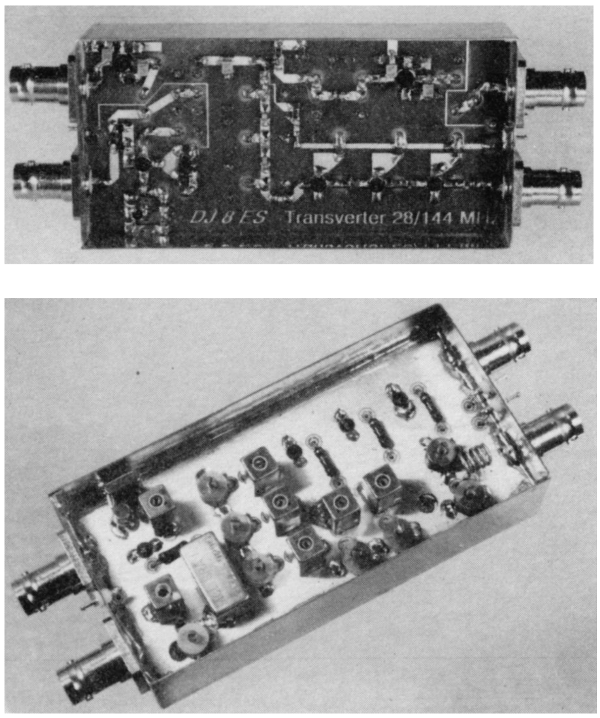


Fig 9.115: The completed 2m transverter

should either be made as a through-plated PCB, or copper rivets used to make the earth connections for the coils and ring mixer. The parts list is shown in **Table 9.7** and component layouts for both sides of the PCB are shown in **Figs 9.116 and 9.117** (both in Appendix B). Suitable holes for the crystal, trimmers and Neosid coils, etc are drilled on the earth side of the boards (fully coated side) using a 2.5mm drill. Holes that are not used for earth through connections should be countersunk using an 8mm drill. Suitable slots are to be sawn out in the printed circuit board for the BNC connectors. When the board has been soldered to the sides of the housing, the actual assembly can be undertaken. The boards are fitted into the housing so that the connector pins of the RF connectors are level with the surface of the PCB (cut off projecting Teflon collars with a knife first). When the mechanically large components (filter coils, trimmers, crystal and ring mixer) have been fitted it must be possible to fit the housing cover without any obstruction.

When the equipment is used for the first time, the following test equipment should be available: multimeter, frequency counter, diode probe, wattmeter and received signal (eg beacon). First the crystal oscillator is tuned by adjusting L1, the power consumption should be approximately 65mA. In transmit modes the only adjustment required is to tune the three-pole filter, the power consumption should be approximately 130mA. In receive mode the input filter should be tuned using a weak signal eg a beacon, then the 28MHz filter, C1 / L2, should be tuned.

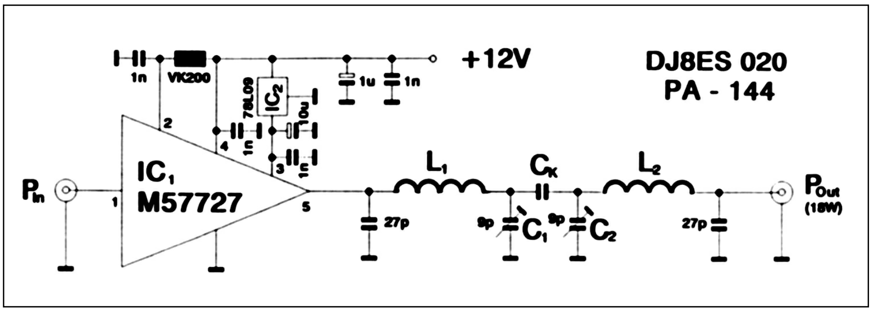


Fig 9.118: Circuit of the 2m power amplifier to be used with the 2m transverter

IC1	TA78L09F SMD voltage regulator
IC2, IC5	MSA1104 (Avago Semiconductor)
IC3	MSA0104 (Avago Semiconductor)
IC4	MSA0304 (Avago Semiconductor)
T1	J310 (TO-92) (Vishay Siliconix)
T2, T3	BF992 (SMD) (Vishay Siliconix)
D1, D2	BA886 PIN diode (SMD)
L1, L3, L4 , L6, L7, L8	Neosid BV5061 0.1μH blue/brown coil
L2	Neosid BV5048 1μH yellow/grey coil
L5	4.5 turns, 1mm gold plated copper wire
C1	30pF foil trimmer (red), 7.5mm grid (Valvo)
C2, C3	12pF foil trimmer (yellow), 7.5mm grid (Valvo)
R1, R2, R3	Attenuator, see Table
Q	116MHz crystal, HC18U or HC25U
1x	SRA1H high-level ring mixer
2 x	Carbon film: 120Ω, 0.5 W
1 x	Carbon film: 220Ω, 0.5 W
1 x	Carbon film: 270Ω , 0.5 W
4 x	BNC flanged socket (UG-290 A/U)
3x	Teflon bushing
1 x	Tinplate housing: 55.5mm x 111mm x 30mm
9x	Copper rivets (1.5mm dia.)

All other components in SMD format

Ceramic capacitors	Resistors
3 x 1pF	1 x 150Ω
1 x 1.5pF	2 x 220Ω
1 x 2.2pF	2 x 1kΩ
4 x 3.3pF	2x 10kΩ
1 x 10pF	2 x 22kΩ
1 x 12pF	**Inductors**
1 x 82pF	2 x 1μH choke
17 x 1nF	3 x 10μH choke
1 x 10μF / 20V Tantalum	

Table 9.7: Parts list for 2m transverter

A power amplifier can be added to the 2m transverter to increase the output power to 20W. The amplifier uses a Mitsubishi hybrid module, these are still available from some suppliers and can be found in surplus equipment.

Fig 9.118 shows the relatively simple circuit for the 144MHz power amplifier. The core of the circuit is a Mitsubishi M57727 hybrid module (IC1). This module operates at a working voltage of 12 volts. With exactly 27dB amplification, the transverter signal is raised to an output voltage of 20W. The output power to input power ratio is shown in **Fig 9.119**. The current consumption of the module is directly proportional to this ratio.

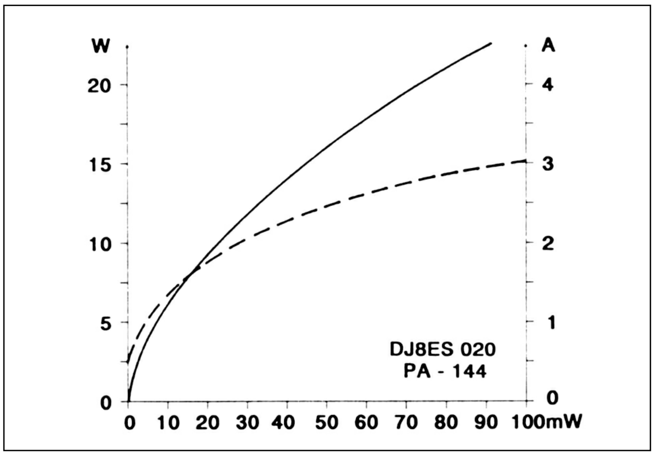


Fig 9.119: Power transfer characteristics of the Mitsubishi M57727 hybrid module

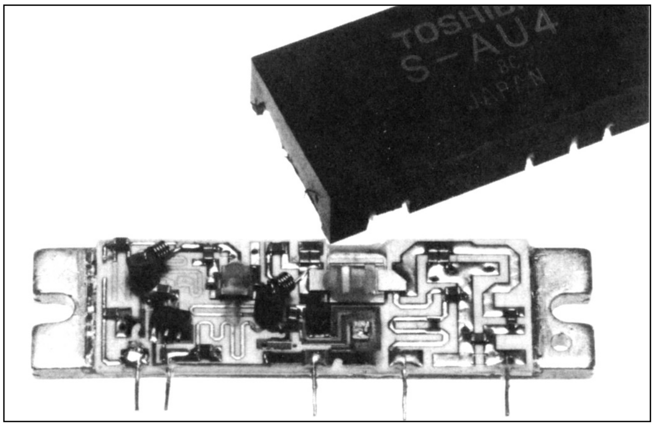


Fig 9.120: An internal view of a hybrid module. This is the Toshiba S-AU4 which is a 70cm amplifier

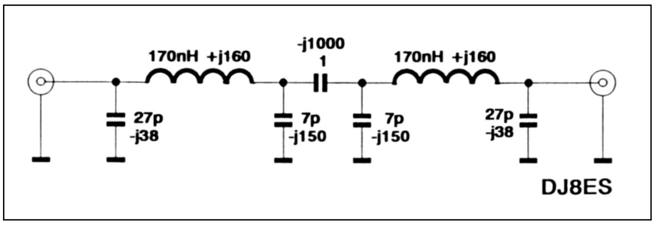


Fig 9.121: Circuit of the filter used in the two metre power amplifier

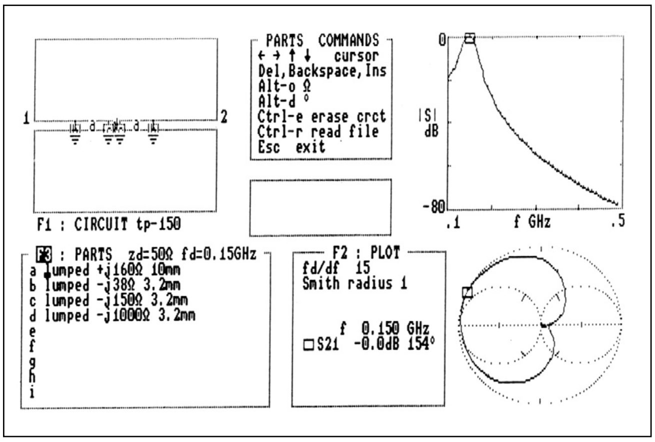


Fig 9.122: PUFF CAD software output used to design the two metre filter

Such amplifier modules are constructed using thick film technology. This module is designed for the 144 - 148MHz frequency range, and the amplification is achieved in two stages. Fig 9.120 shows what a typical module looks like from the inside. The 50 ohm input and output matching circuits are clearly visible.

A filter (**Fig 9.121**) on the output provides the harmonic suppression required. Amazing suppression is obtained using only two pi filters wired together. **Fig 9.122** shows the output of the PUFF CAD design software used to design the filter.

IC1	M57727 (Mitsubishi)
IC2	TA78L09F voltage regulator (SMD)
L1, L2, C_K	see text
C1, C2	9pF trimmer with soldering lug
1 x VK200 VHF broad-band choke	
1 x 1nF feed-through capacitor, solderable	
2 x BNC flanged bush (UG-290 A/U)	
1 x Tinplate housing 55.5 x 111 x 30mm	
All other components in SMD format:	
1 x lμF/20V tantalum	
1 x 10μF/20V tantalum	
2 x 27pF, ATC chip	
3 x 1nF, ceramic capacitor	

Table 9.8: Parts list for 2m hybrid amplifier

The amplifier is assembled on a double-sided printed circuit board measuring 54mm x 108mm. The board can fit into a standard tinplate housing (55.5mm x 111 mm x 30mm), the parts list is shown in **Table 9.8**. A suitable sized hole is sawn out for the hybrid module. Fastening holes are drilled along the edge as shown in **Fig 9.123** (in Appendix B). Good earth connections are essential for the circuit to operate correctly. The through contacts required are made by the M3 screws that secure the assembly to the heat sink. The BNC connectors are placed at suitable points on the side wall of the housing. Also positioned in the side wall is the feed-through capacitor for the power supply. The components are not inserted until the board has been soldered to the sides of the housing. The board should be fitted at the edges of the sides, so that the amplifier module will lie flat on the heat sink.

The two coils (L1, L2) and the coupling capacitor, CK, are hand made. The coils are 8.5 turns of silvered copper wire with a diameter of 1mm. The wire is wound around a 6mm mandrel (eg a 6mm drill shank) and soldered on with a 1mm clearance from the PCB. The coupling capacitor, C_K, is made from a 1cm long piece of coaxial cable (RG174), the length is chosen to give the required 1pF capacitance. A standard chip capacitor cannot be

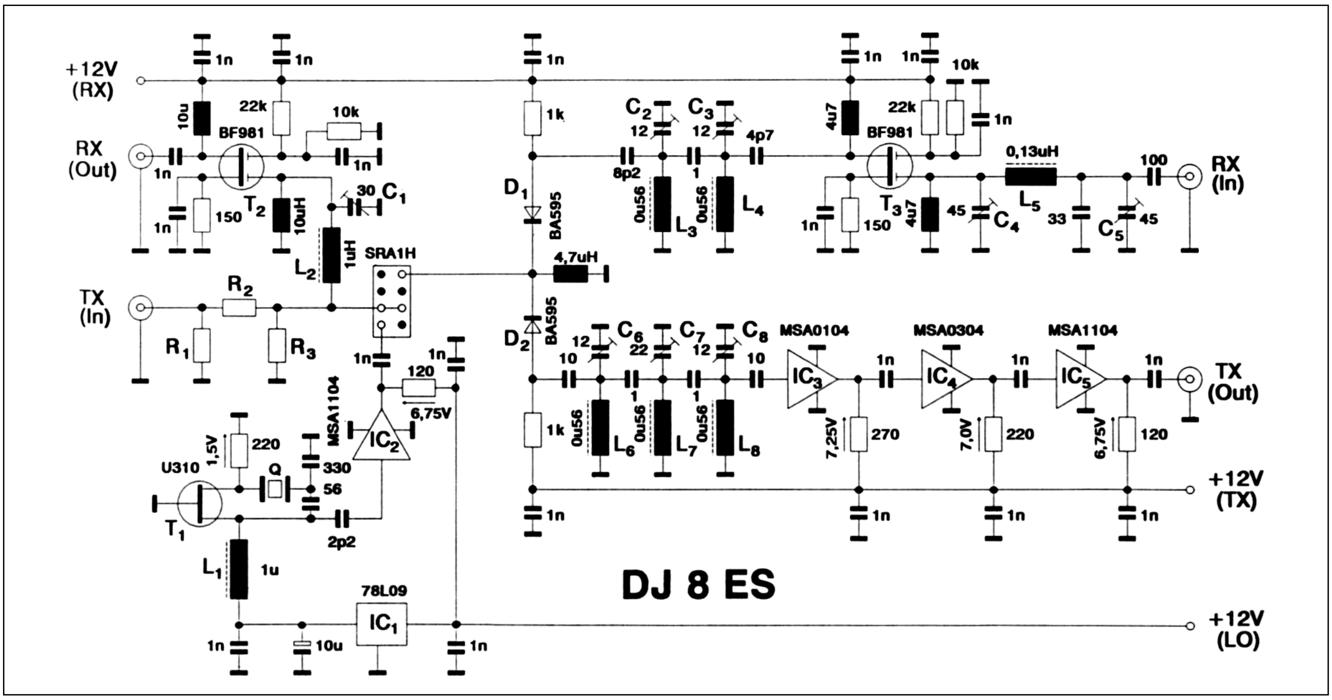


Fig 9.124: Circuit diagram of the 6m transverter. Note: the U310 has been superseded by the J310 and the BF981 by the BF992

IC1	TA78L diameter 9F voltage regulator (SMD)
IC2, IC5	MSA1104 (Avago Semiconductor)
IC3	MSA0104 (Avago Semiconductor)
IC4	MSA0304 (Avago Semiconductor)
T1	J310 (TO-92) (Vishay Siliconix)
T2, T3	BF992 (Siemens)
DI, D2	BA595 PIN diode (SMD)
L1, L2	BV5048 Neosid coil,1 µH, yellow/grey
L3, L4	BV5036 Neosid coil, 0.58µH, orange/blue
L5	BV5063 Neosid coil, 0.58µH, blue-orange
L6, L7, L8	BV5063 Neosid coil, L8 0.58µH, orange/blue
C1	30pF foil trimmer (red) 7.5 mm grid (Valvo)
C2, C3	l2pF foil trimmer (yellow) 7.5 mm grid (Valvo)
C4, C5	45pF foil trimmer (violet) 7.5 mm grid (Valvo)
C6, C8	12pF foil trimmer (yellow) 7.5 mm grid (Valvo)
C7	22pF foil trimmer (green) 7.5 mm grid (Valvo)
Q	22MHz crystal, HC18U or HC25U
1 x SRA1H ring mixer	
2 x 120Ω / 0.5W Carbon film	

1 x 220Ω / 0.5W Carbon film
1 x 270Ω / 0.5 W Carbon layer
4 x BNC flanged connector (UG-290 A/U)
3 x Teflon bushing
1 x Tinplate housing 55.5 x 111 x 30mm
9 x 1.5mm dia. Copper rivets

All other components are SMD format:

Ceramic capacitors	Resistors etc
3 x 1pF	1 x 150Ω
1 x 2.2pF	2 x 220Ω
1 x 4.7pF	2 x 1kΩ
1 x 8.2pF	2 x 10kΩ
2 x 10nF	2 x 22kΩ
1x 33pF	1x 10µF / 20V Tantalum
1 x 56pF	3x Choke, 4.7µH
1 x 330pF	2x Choke, 10µH
17 x 1nF	

Table 9.9: Parts list for the 6m transverter

used here, due to the relatively high power level. A thin copper plate is soldered between the two pi filters for screening, see Fig 9.123 (in Appendix B). Finally, the remaining components are added. The module is screwed directly onto the heatsink using two M4 screws after applying heat conducting paste.

A power meter and a multimeter are required for putting the equipment into operation. The quiescent current should be approximately 400mA, which rises to around 2.5A under full drive with an input power of 60mW. This gives an output power of the order of 18W.

Only the low-pass filter (C1, C2) requires tuning in the hybrid amplifier. The trimmers are normally screwed about half way in when the unit is correctly tuned. In order protect the hybrid module, carry out this tuning procedure with only a low drive power level (max 10mW).

6m Transverter

A transverter for the 6m band can be produced based on the 28/144MHz transverter described above [37]. All that is required is modification of the oscillator and the filter.

Fig 9.124 shows the complete circuit for the 28/50MHz transverter. The circuit can be assembled using the printed circuit board used for the 2m transverter. The pi filter at the input of the receiver needs to be altered; **Fig 9.125** (in Appendix B) shows details of the modification. All the coils and some of the capacitors have different values for the lower frequency range, **Table 9.9** show the parts list. To make it easier to produce the 6m version of the transverter, the layout of the printed circuit board with the appropriate components for the 50MHz version is illustrated in **Fig 9.126 and Fig 9.127** (both in Appendix B).

To increase the power output an M57735 hybrid module is used in a separate amplifier stage for the 6m band, the circuit diagram is shown in **Fig 9.128**. The M57735 module was developed for use around 50MHz and is still available from some suppliers or in surplus radio equipment. About 10W can be expected at the output of the PA from the 50mW output from the transverter.

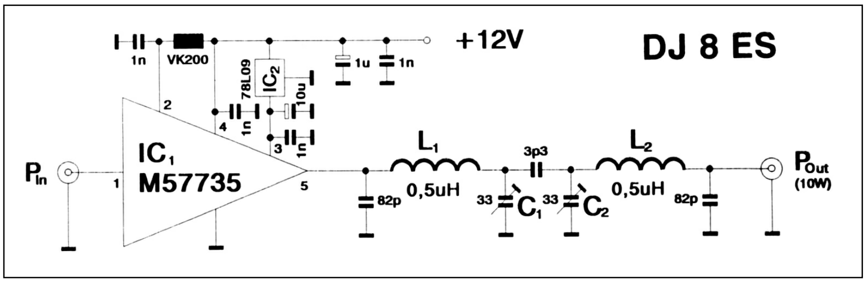


Fig 9.128: Circuit of the 6m power amplifier to be used with the 6m converter

IC1	M57735 (Mitsubishi)
IC2	TA78L09F voltage regulator (SMD)
L1, L2	0.5µH air-core coil
C1, C2	33pF trimmer with soldering lugs
1 x VK200 VHF wide-band choke	
1 x 1nF feed through capacitor, solderable	
2 x BNC flanged bush (UG-290 A/U)	
1x Tinplate housing 55.5 x 111 x 30mm	
1x 1µF / 20V Tantalum	
1x 10µF / 20V Tantalum	
1x 3.3pF, ATC chip	
2x 82pF, ATC chip	
3x 1nF, ceramic capacitor	

Table 9.10: Parts list for 6m hybrid amplifier

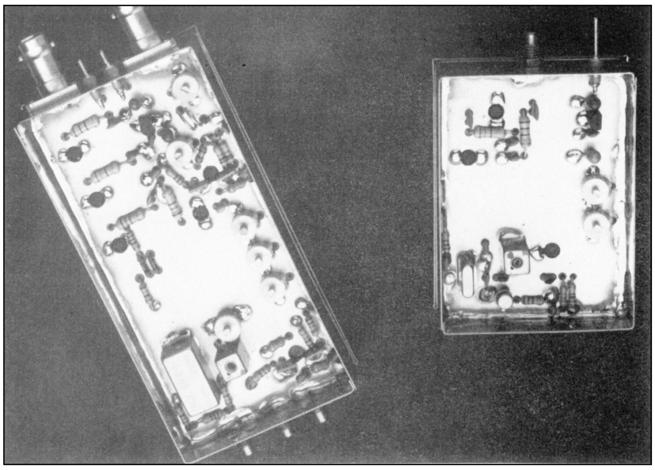


Fig 9.129: The completed local oscillator and transverter units of the 70cm transverter

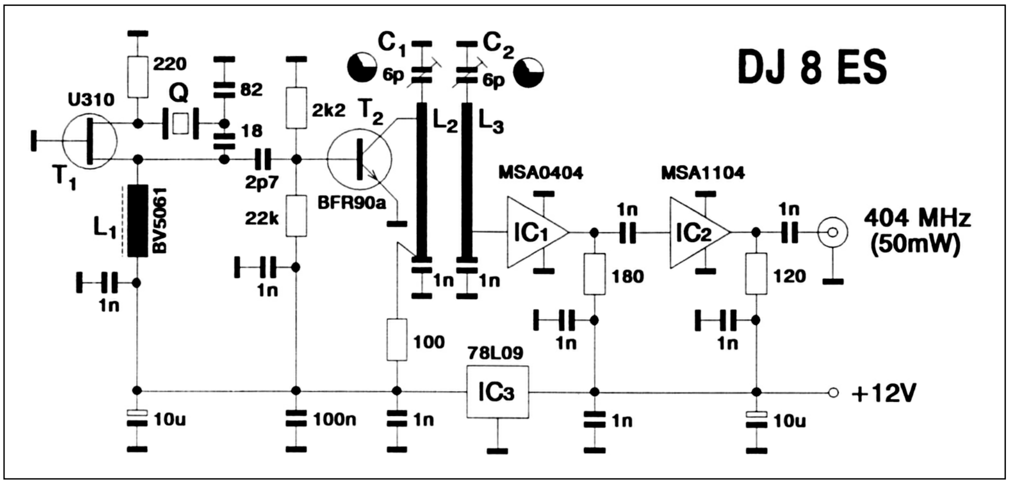
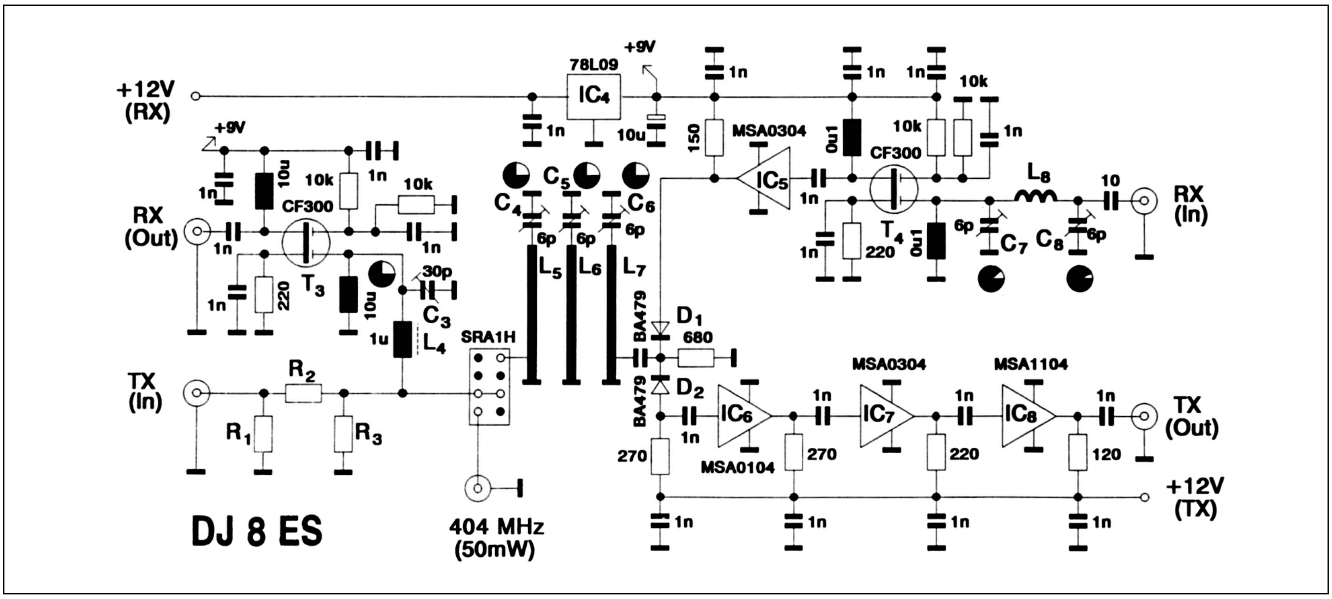


Fig 9.130: Circuit of the local oscillator used in the 70cm transverter

(below) Fig 9.131: Circuit diagram of the 70cm transverter. Note that the CF300 has been replaced by a BF994 and the U310 by a J310 (see parts list)



The low-pass filter provides the harmonic filtration required. Only components of appropriate quality (eg air-core coils and air-spaced trimmers) should be used here. The 50MHz amplifier can be assembled on the printed circuit board used for the 2m version using the same construction techniques. The parts list is shown in Table 9.10.

70cm Transverter

The following design for a 28/432 MHz transverter [38] is similar to the 28/144MHz transverter described above. It uses two boards; the oscillator and the transverter Fig 9.129 shows a picture of the completed units. It should be possible for the transverter to directly feed a standard power amplifier.

Using wide-band amplifier ICs and a ring mixer makes the circuit very flexible, by just changing the filters and the crystal oscillator, the tuning range can be changed to suit the requirements.

Fig 9.130 shows the circuit of the local oscillator, it uses a U310 as the crystal oscillator at 101MHz. The 404MHz required for the local oscillator is produced using a quadrupler. A printed circuit 2-pole filter provides the necessary filtering. Two wide-band integrated amplifiers, MSA0404 (IC1) and MSA1104 (IC2) supply the desired output of 50mW. The correct level of amplification is important, only the amplification, which is actually necessary, should be used. Any excess increases the spurious outputs.

Fig 9.131 shows the circuit diagram of the transverter. The SRA1H ring mixer used in the transmit/receive converter is suitable for use up to 500MHz, and requires a local oscillator level of 50mW. The mixer is controlled using an attenuator, which should provide an intermediate frequency (IF) level of no more than 1mW at the ring mixer. The attenuator must be designed on the basis of the IF output available.

Table 9.6 shows the resistor values required for the attenuator in relation to the IF power level. All the values are based on the standard values from the E12 to E24 ranges. The attenuator also serves as a wide-band 50-ohm termination for the ring mixer (SRA1H). The received signal is matched at high impedance to the CF300 (T3) using L4 and C3. This low-noise transistor stage provides the necessary intermediate frequency amplification.

The 70cm received signal is passed to the gate of the BF994 (T4) through a pi filter (aerial impedance 50 ohms) that is followed by an MSA0304 amplifier. When the receive +12V power supply is connected, the PIN diode D1 (BA479) is biased on and the signal passed through.

The printed circuit 3-pole 70cm filter is used for both receive and transmit. The transmit signal initially passes through the filter and diode D2 which is biased on. The subsequent amplifier uses integrated wide-band amplifiers (IC6, IC7, IC8). The combination of MSA0104, MSA0304 and MSA1104 provides an output of 50mW (+17dBm) with 40dB of amplification.

In practical operation, such transverters are used with the same driving unit; an additional filter for harmonics and spurious transmissions is recommended. Fig 9.132 shows a possible

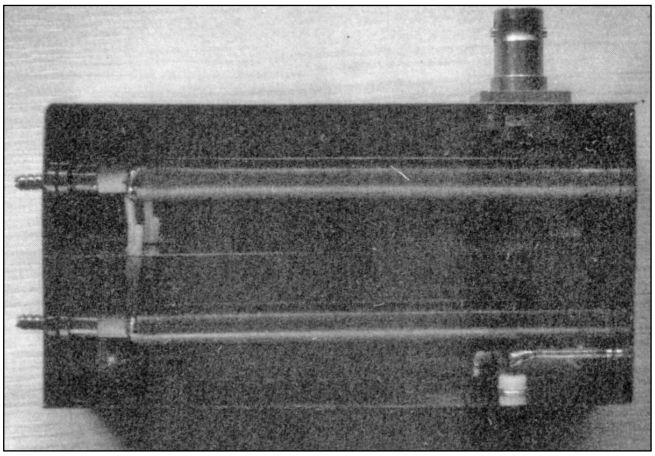


Fig 9.132: 70cm bandpass filter that can be used with the 70cm transverter

two-pole bandpass filter. It can be assembled as an air core construction using a standard tinplate housing measuring 55.5 x 111 x 30mm.

The 28/432 MHz transverter is divided into two independent assemblies: the local oscillator and the transmit/receive converter. The double sided printed circuit boards measure 54mm x 72mm for the local oscillator and a 54mm x 108mm for the transmit/receive converter. The parts list for the local oscillator and transverter are shown in Table 9.11 and 9.12 respectively. The component layout for the local oscillator is shown in Figs 9.133 and 9.134, and for the transverter in Figs 9.135 and 9.136 (all four of these are in Appendix B). The PCB are mounted in standard tinplate housings, suitable holes are drilled for the stripline transistors and the wide-band amplifiers; these components are mounted level with the surface of the board. The holes for the crystal, trimmers and Neosid coils, etc are drilled on the earth side of the boards (fully coated side) using a 2.5mm drill. Suitable slots are to be sawn out in the printed circuit board for the SMC or BNC connectors. The same applies to the 1nF capacitors at the source connection of the amplifier transistors, T3 and T4. When the individual boards have been soldered to the sides of the housing, the actual assembly can be undertaken. The boards are fitted into the housing so that the connector pins of the RF connectors are level with the surface of the PCB (cut off projecting Teflon collars with a knife first). When the mechanically large components (filter coils, trimmers, crystal and ring mixer) have been fitted it must be possible to fit the housing cover without any obstruction.

When the equipment is used for the first time, the following test equipment should be available: Multimeter, Frequency counter, Wattmeter and Received signal (eg beacon). The assemblies switched on one after another.

Firstly, the crystal oscillator is set to its operating frequency of 101MHz by adjusting coil, L1. The onset of oscillation results in a slight increase in the collector current of T2 (monitoring voltage drop across 100 ohm resistor). A frequency counter is loosely coupled and the oscillator frequency measured.

The two-pole filter after the quadrupler T2 (BFR90a) filters out the 404MHz frequency required. To adjust this, the two trimmers, C1 and C2 are adjusted one at a time for maximum output. The local oscillator should supply an output of at least 50mW. The current consumption for an operating voltage of +12V is about 120mA.

The transmit branch of the transmit/receive converter is put into operation first. Only the three-pole filter (C4, C5, C6) has to

IC1	MSA0404 (Avago Semiconductor)
IC2	MSA1104 (Avago Semiconductor)
IC3	78L09 voltage regulator
T1	J310 (TO-92) (Vishay Siliconix)
T2	BFR90a (Valvo)
L1	Neosid BV5061 0.1µH blue/brown coil
L2, L3	λ/4 stripline, etched
C1, C2	6pF foil trimmer (grey), 7.5mm grid (Valvo)
Q	101MHz crystal, HC18U or HC25U
1 x	Carbon film: 180Ω, 0.5 W
1 x	Carbon film: 120Ω, 0.5 W
1 x	SMC or BNC flanged socket (UG-290 A/U)
1 x	Teflon bushing
1 x	Tinplate housing: 55.5mm x 74mm x 30mm
2 x	1nF trapezoid capacitor
2 x	l0µF 20V tantalum capacitor
Ceramic Capacitors (2.5mm grid)	*Resistors (1/8W, l0mm)*
1 x 2.7pF	1 x 100
1 x 18nF	1 x 220
1 x 82pF	1 x 2.2k
6 x 1nF	1 x22k
1 x 100nF	
SMD Capacitor (model 1206 or 0805)	
2 x 1nF	

Table 9.11: Parts list for 70cm transverter local oscillator

IC4	78L09 voltage regulator
IC6	MSA0104 (Avago Semiconductor)
IC5, IC7	MSA0304 (Avago Semiconductor)
IC8	MSA1104 (Avago Semiconductor)
T3, T4	BF994 (Vishay Siliconix)
DI, D2	PIN diode BA479
L4	BV5048 Neosid coil, 1µH, yellow/grey
L5, L6, L7	λ/4 stripline, etched
L8	1.5 turns, 1mm CuAg wire
1x	SRA1H high-level ring mixer
C3	30pF foil trimmer (red) 7.5mm grid (Valvo)
C4, C5, C6	6pF foil trimmer (grey) 7.5mm grid (Valvo)
C7, C8	6pF foil trimmer (grey) 7.5mm grid (Valvo)
R1, R2, R3	Attenuator, see Table
1 x Carbon film: 120Ω , 0.5W	
1 x Carbon film: 150Ω, 0.5W	
1 x Carbon film: 220Ω, 0.5W	
l x Carbon film: 270Ω, 5W	
5 x SMC sockets (some of which may be BNC flanged: UG-290 A/U) (see photo of specimen assembly)	
2 x Teflon bushing	
1 x Tinplate housing 55.5 x 111 x 30mm	
4 x 1nF trapezoid capacitor	
2 x 0.1 µH choke, l0mm grid, axial	
2 x 10µH choke, l0mm grid, axial	
1 x 10µF 20V tantalum	
Resistors (1/8W, 10mm)	*Ceramic Capacitors (2.5mm grid)*
2 x 220Ω	1 x 10pF
1 x 270Ω	12 x 1nF
1 x 680Ω	SMD Capacitor (model 1206 or 0805)
4 x 10kΩ	6 x1n

Table 9.12: Parts list for 70cm transverter

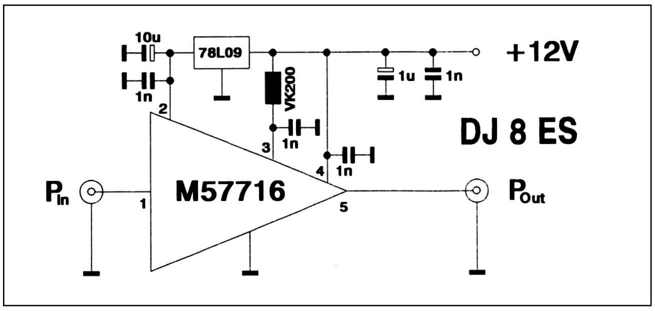


Fig 9.137: Circuit diagram of the 70cm power amplifier to be used with the 70cm transverter

be adjusted. A current of approximately 130mA should be measured for an operating voltage of +12V. This is an indication that the amplifier stages are operating satisfactorily. If the input attenuator is selected as described in Table 9.6, an output greater than 50mW can be expected. Possible spurious outputs (oscillator, image frequency, etc.) are suppressed by better than 50dB.

The receiver can be calibrated using a strong received signal (eg a beacon). Because the same filter is used as in the transmit branch, the beacon signal should be audible immediately. Another filter is used at the intermediate frequency (28MHz) after the mixer. The trimmer C3 should be adjusted to give maximum signal output. Optimising the signal-to-noise ratio using the pi filter, C7, C8 and L8 completes the calibration. The current consumption of the receive converter is very low, only 50mA. The noise figure is approximately 2dB with a conversion gain of approximately 30dB.

The designer uses the transverter described in association with an external preamplifier and power amplifier. Modern hybrid modules are just the thing for power amplifier stages. The output signal can be increased from 50mW to 10 - 20W in one go using such components. **Fig 9.137** shows the circuit for such a module using a Mitsubishi M55716. A 2C39 valve PA can be fully driven using this 10W output.

25kHz, the two spacings coincide on a number of frequencies including all the calling frequencies (70.45 FM, 70.26 All Mode and 70.2MHz SSB) and two of the most commonly used simplex FM working frequencies 70.35 and 70.40MHz.

The design of this transverter was not intended to push the frontiers of technology, but to provide a simple, repeatable design based on readily available components many of which could be found in the 'junk box' thus keeping down costs. The circuit diagram for the transverter is shown in **Figs 9.138 and 9.139**.

The receive converter uses Dual Gate MOSFET RF and mixer stages which provide good gain, noise performance and stability. L1 is resonated by the series combination of C1 and C2 to provide the first stage of input filtering, while the ratio C2:C1 provides matching from the 50 ohm input to the higher impedance of Q1 Gate 1. The RF and mixer stages are band-pass coupled (L2 and L3) to improve rejection of unwanted, out of band signals.

The local oscillator uses a third overtone crystal and is shared between the receive and transmit mixers. The choice of crystal frequency depends on the drive source to be used. For 28MHz the crystal frequency is 42MHz, for a CB 27/81 driver a frequency of 42.49875MHz is required while for the newer PR 27/94 CB Rigs a frequency of 43.0850MHz is required. The zener diode ZD1 is an option which may be required to improve stability if the transverter is to be used for SSB or CW operation (Initial tests indicate that it is not necessary so long as a regulated supply is used).

The transmit mixer uses a proprietary doubly balanced mixer to provide additional suppression of the oscillator and driver frequencies and their products. There are a number of different mixer units that will operate satisfactorily, but if substituting a different device beware of its pin connections as two different pin layouts are in common use and only the right one will work! Also, some types are not as shown in **Fig 9.140** but have pins 2, 5, 6 and 7 bonded directly to the case. This is not a problem. Note that pin 1 is identified by a different coloured bead that may be lighter or darker than its neighbours.

4m Transverter

This transverter was designed as a club project for the Andover Radio Amateur Club [33]. Various schemes such as transverting from two meters were considered but after some discussion, it was decided to transvert from 27MHz or optionally 28MHz. Using 27MHz as the drive source was particularly attractive for a number of reasons. Many members own or could cheaply obtain a 27MHz CB rig which would mean that the completed transverter could be permanently connected to a dedicated drive source without 'tying-up' and restricting the use of equipment used regularly on other bands. This would make it more practical to establish a club frequency that could be monitored whenever one is in the shack. It was felt this would help to build up band activity. Although CB radios have 10kHz channel spacing whilst the 4m band uses

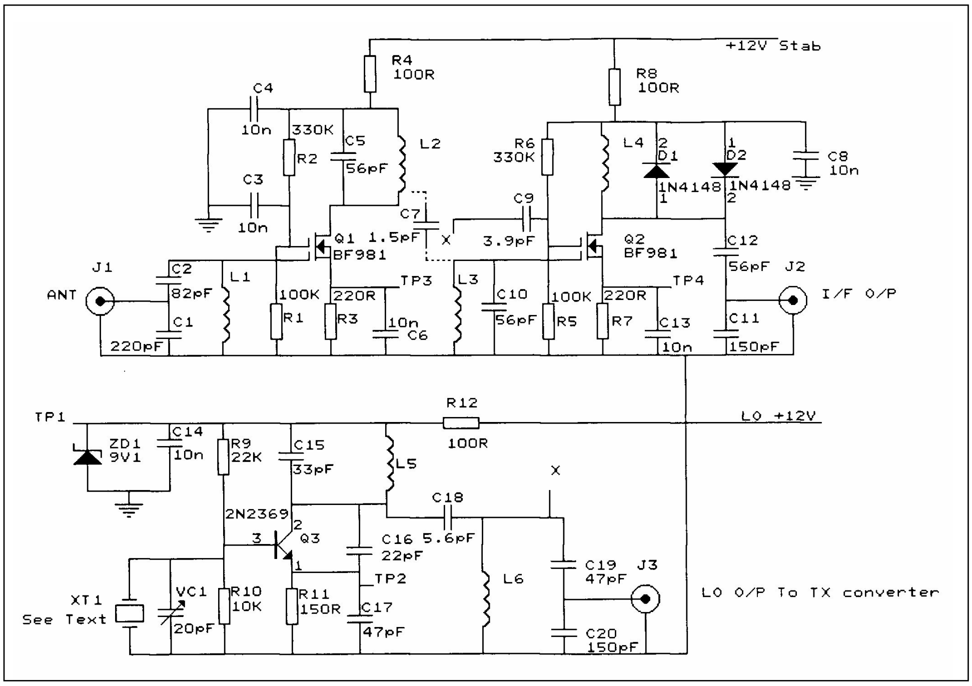


Fig 9.138: Receive converter circuit diagram for the 4m transverter. ***[Note: BF981 and 2N2369 are now obsolete devices. A suitable DG mosfet would be the BF992 in SMD and the 2N2369 can be replaced with the MPSH-10 in TO-92 with a slightly different pin out.]***

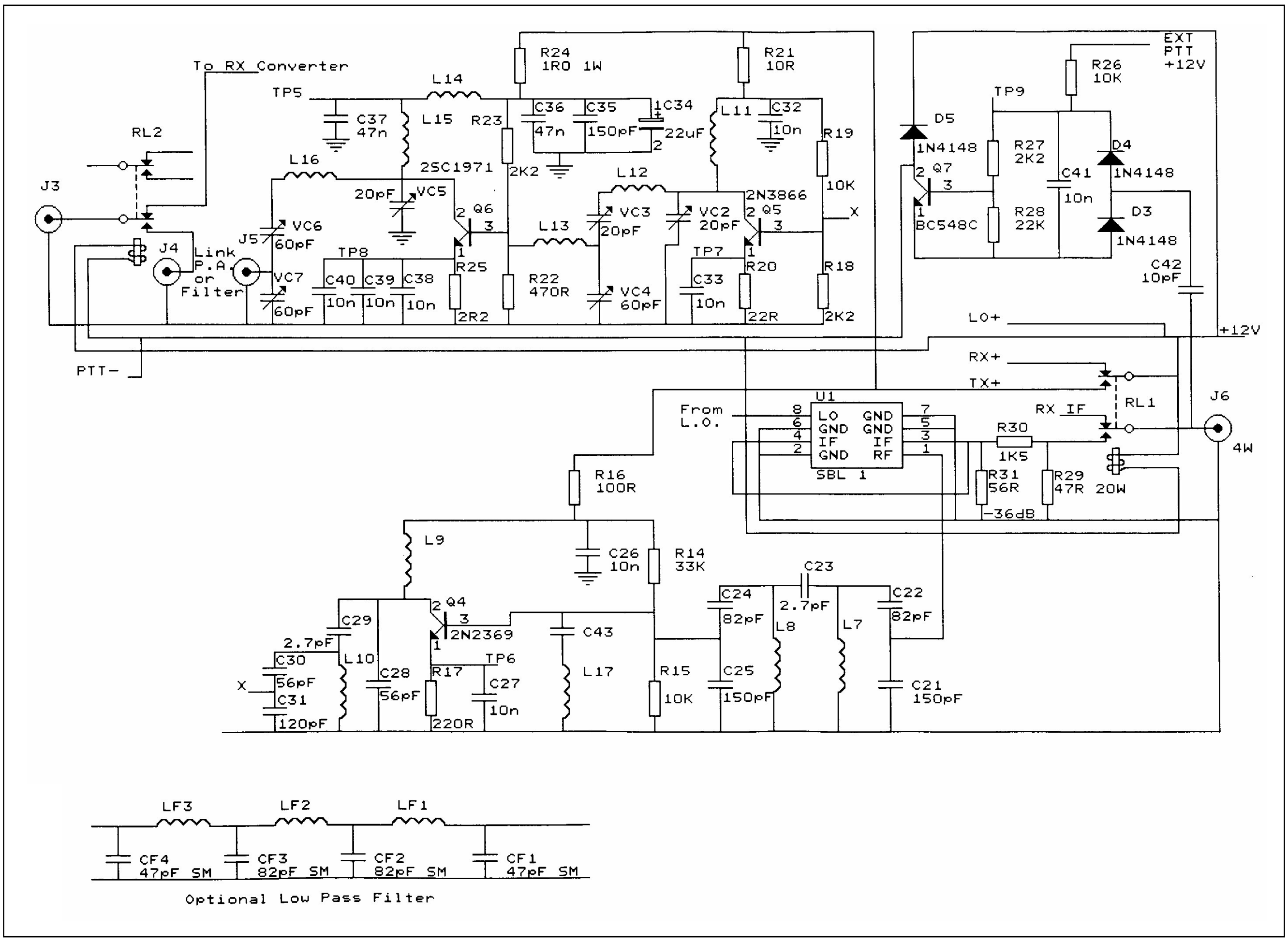


Fig 9.139: Transmitter converter circuit diagram for the 4m transverter ***[Note: 2N2369 is now obsolete but it can be replaced with the MPSH-10 in TO-92 with a slightly different pin out. The 2N3866 is now only made in the SMD version (SO-8 package)]***

These doubly balanced mixers require about +7dbm of local oscillator injection and less than 0dBm (1mW) of Drive (IF) for optimum IMD performance. The output of the oscillator is loosely coupled to L6 providing a band-pass arrangement to reduce unwanted harmonics. C19 and C20 in series resonate L6 while their ratios provide an impedance transformation to match the oscillator output to the mixer (U1) at 50 ohms.

The input from the driving 'rig' is first coupled through C42 to a voltage doubling detector D3/D4 which acts as an RF sensor driving Q7 to operate the change over relays RL1 and RL2. Note that the choice of device for Q7 was based largely on the need for a high current gain (H_{fe}), the specified BC548C having an H_{fe} min. of 420. Once a sufficient level of RF is present at the input, Q7 turns ON and relays RL1 (and RL2) close routing the RF through an attenuator, R29, R30 and R31 to reduce its level to about 0dbm (1mW). Please note that RF switching is generally acceptable for FM but can be problematic for SSB operation so direct switching is recommended for this mode. There are two options for external switching: a positive voltage on Transmit applied to the EXT PTT + input or a contact closure to ground applied to EXT PTT LO. Note that the latter should be from voltage free contacts or possibly through a series diode.

Assuming that a standard CB rig is being used with an output of 4 watts, the attenuation required is 36dB. The values shown for these resistors produce about 36dB of attenuation. If you wish to use a different input level, you will need to change these values, Table 9.13 shows some common values. Bear in mind that R29 will dissipate the bulk of the power output from the driving rig and should be rated accordingly. This design uses a TO220 style non-inductive power resistor for R29, and this is bolted to the front panel and hence chassis of the transverter to dissipate the heat. Although rated at 20W, a maximum drive

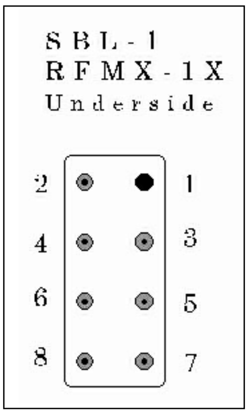


Fig 9.140: Pin connections for the mixer U1 used in the 4m transverter

Input Level	Required Attenuation	R30
1W	30dB	820
4W	36dB	1.5K
10W	40dB	2.7K

Table 9.13: Attenuator values for 4m transverter

C3, 4, 6, 8, 13,		R14	33k
14 ,26, 27,		R1, 5	100k
32, 33, 38,	0.1µF Multi	R2, 6	330k
36, 37, 39,	layer	Q3, 4	2N2369
40, 41, 44	ceramic	Q1, 2	BF981
C7	1pF/1.5pF	Q5	2N3866
C23, 29	2.7pF	Q6	2SC1971
C9	3.3/3.9pF	Q7	BC548C
C18	5.6pF	U1	DB Mixer SBL-1
C42	10pF	D1, 2, 3, 4, 5	1N4148
C16	22pF	ZD1	Zener 9V1
C15	33pF	RL1, 2	DPCO Relays
C17, 19	47pF	L1, 2, 3, 7, 8, 9, 10	Toko style
C5, 10, 12, 28, 30	56pF		MC119 3.5T
C2, 22, 24	82pF	L4	Toko style
C31	120pF		MC119 15.5T
C11, 20, 21, 25, 35	150pF		without can
C1	220pF	L5, 6	Toko style
C34	22 F 25V		MC119 9.5T
VC1, 2,3,5	Variable 22pF		without can
VC4,6,7	Var 5-60pF	10mm screening cans	
R24,25	1R0 2W	L11, 12, 13,	
R21	10R	14, 15, 16	Airspaced coils
R20	22R	XT1	Crystal, see text
R29	47R 20W	Enclosure AB10	Maplin LF11M
R31	56R	Heatsink	Farnell 170-071
R4, 8, 12, 16	100R	Sockets BNC	
R11	150R	Misc. Nuts bolts wire	
R3, 7, 17	220R	M3 x 6mm PH Screw	
R22	470R	M3 x 10mm Spacer	
R30	1k5	M3 x 10mm Screw	
R18, 23, 27	2k2	Nuts + washer + spw	
R10, 15, 19, 26	10k	Feet	
R9, 28	22k		

Table 9.14: Parts list for 4m transverter

level of 10 watts should not be exceeded and if you intend to have long overs using FM, you should keep to a maximum input level of 4 watts or the case will get mightily hot!

The mixer is band pass coupled (L7/L8) to a pre driver stage Q4 which is in turn band-pass coupled (L9/L10) to the driver Q5. Note that space has been provided for a trap (L17/C43) from the base of Q4 to ground. This trap which is expected to be necessary only if a 144 or 145MHz IF is used should be resonated at the LO frequency (74/75MHz). With a 28MHz IF and no trap, the local oscillator output level was measured at about 47dB below full output power which was felt to be acceptable. As in the receive part of the transverter, split capacitance is used (C21/C22, C24/C25 and C30/C31) to effect impedance matching.

The Driver and PA devices were chosen on the basis of price and availability. The PA is characterised for FM (Class C) operation (5W at 150MHz) but it was felt that there was a good chance that it could be made to operate satisfactorily in Class B and that would allow the transverter to be used for SSB (and AM) if required.

The first prototype appeared to provide about 4W and this power level was considered entirely adequate for the local activity that this project was designed to stimulate.

Subsequent improvements to the layout yielded the full 5W. For those wanting higher power, the layout provides for connection of an external power amplifier within the RF switching provided. Please note, however, that relays with a higher contact current rating may be needed to perform the DC switching of such a PA as their absolute maximum limit is 2A DC and in practice this should not be closely approached.

Space on the PCB has been left free for additional output filtering which it was thought may be necessary to reduce the harmonic output from the -35dB level observed on the prototype. Although not absolutely necessary at the 5 watt level, it would nonetheless represent good practice and tests suggest that this filter reduces harmonics to about 60dB down on the wanted signal at full power. Tests on a sample filter suggest a loss of about 0.85dB reducing the output from 5W to 4W. In real terms this is insignificant. If a PA is to be fitted a filter is essential and should be fitted to its output.

PCB assembly commences with the smallest components first as placement of the larger components makes it difficult to reach and inspect the smaller ones. Start with the pins, whose positions are marked on the PCB component layout overlay as circles, the PCB is shown in **Figs 9.141 and 9.142** in Appendix B. Push them firmly into place and this is most easily done before any other components have been fitted. The ridges on the pins hold them in place until soldering which need not begin until almost all the components are fitted. Leave the fitting of the following components until much later: MOSFETS Q1 and Q2. The power resistor R29 is not fitted until the unit is assembled in its box, the parts list for the transverter is shown in **Table 9.14**.

When fitting components, pre-form all leads so that the component will sit as close to the PCB as possible except for the self supporting coils which should sit 2mm above the board. Some capacitors such as C3, 7, 8, 9 must be pre-formed to a slightly wider pitch (than 0.1in) as conductors pass between their pins and more clearance is required. Don't cut the leads until you have fitted the component and bent its leads over to about 45 degrees from the board. Leave 1.5-2mm protruding. This will hold them in place until soldering which should not start until all the components except the MOSFETS, Inductors and R29 have been fitted. Note that the PCB layout shows VC5 as a large capacitor, in fact a small (green) one should be fitted in this position. The circuit diagram and parts list are correct. Note that Q5, the driver transistor should be mounted on a TO5 transipad.

After fitting the small components, carry out a careful inspection to confirm that everything is correctly placed before soldering. Carefully inspect your soldering to ensure that all the joints are good. Remove all surplus flux using a PCB cleaning solvent and a stiff brush (such as a half-inch paint brush with its bristles cut short) before inspection if possible as surplus flux has been known to mask a bad joint. Modern fluxes are hygroscopic and will absorb moisture from the atmosphere and cause corrosion if not removed.

When all the small components have been fitted it is time to fit the pre-wound inductors. These should be fitted one at a time

L1,2,3,7,8,9,10	Cirkit	35-11934	
L4	Cirkit	35-13415	
L5,6	Cirkit	35-13492	
L11	5t	22SWG	6mm dia close wound
L12	6t	22SWG	6mm dia - ditto -
L13	1t	22SWG	6mm dia - ditto -
L14	7t	22SWG	6mm dia - ditto -
L15	7t	22SWG	6mm dia - ditto -
L16	6t	22SWG	6mm dia - ditto -
L17	2m IF version only.		
LF1,2,3	6t	22SWG	6mm dia close wound

Table 9.15: Coil winding data for 4m transverter

and one pin 'tack soldered' merely to hold them in place. Note that L17 and C43 are only required if a 2m IF is used.

Next, the self supporting coils should be wound using a mandrel such as a drill bit with the correct diameter (6mm). All coils are specified for close (no) spacing between turns, details for winding the coils are shown in **Table 9.15**. While each coil is still on its mandrel, carefully scrape off the insulation enamel at the ends, then remove it from the mandrel and pre-tin the ends. Next, pre-form the leads to fit the PCB in the appointed space observing the orientation as shown on the component layout. Fit the coils so that they lie 2mm above the PCB (a matchstick makes a useful spacing tool) and bend the leads over at the rear of the PCB by about 45 degrees to hold them in place as you did with the other components. Be sure to observe the correct orientation of the axis of the self supporting coils. The sense (clockwise/anticlockwise) in which they are wound doesn't matter. When all inductors are fitted, inspect the PCB again for correct placement then solder them. With the pre-wound coils in particular, solder the unsoldered pin first to avoid them dropping out of the PCB!

Next fit the wire link LK1 and, if you have opted to fit the low pass filter, place a link from the transmitter output to the filter input which is nearby. If you are not fitting the filter, run a miniature co-axial link from the PA stage to the change-over relay RL2 using additional pins where CF4 would have been fitted.

After that, fit and solder the MOSFETS Q1 and Q2 keeping the leads as short as possible and taking sensible anti-static precautions (see below). First pre-form the leads downwards 2mm from the body so that they will pass through the holes on the PCB.

Fit 'tails' of 22SWG Tinned copper wire about 1in (25mm) long to the pins adjacent to the IF socket and 50mm long to the ANT and GND Pins adjacent to RL2 by making one turn tightly around the pin then soldering. Bend these tails up at right angles to the PCB. They will be used to connect to the IF and antenna sockets when the PCB is fitted in its box.

Finally, make a trial assembly of the PCB into its case, using the spacers provided and fit the IF BNC socket. Leave the other BNC socket off at this stage as it will prevent removal of the PCB from the box once fitted. Next, fit R29 in place, bolt it firmly in position and 'tack solder' its leads on the top of the PCB. Remember that it will need to be un-bolted from the box when the PCB is removed. Now remove the PCB from the box and trim and solder R29's leads on the underside of the PCB. It is recommended that initial testing is carried out before final assembly into the box.

To test the transverter you will need a multi-meter with an input impedance of 20kΩ/V or better and either an oscilloscope with an input sensitivity of from 10mV per division (bandwidth immaterial) or a millivolt meter. A frequency counter, a general coverage receiver and a 70MHz signal source are also helpful, but not essential. In addition to the basic equipment mentioned, you will need two very simple and useful pieces of test equipment, an RF 'sniffer' and a diode probe to turn your oscilloscope or millivolt meter into a wideband RF level indicator.

Before final assembly into the box, connect a red 14/0.2mm lead to one of the pins at either end of LK1 and a black lead to any convenient point on the ground plane. Carefully apply 13.8 volts DC from a current limited supply. If possible limit to 100mA to avoid damage if there are any serious errors in the construction. If you do not have a current limit on your supply or if it cannot be set to 1 amp or less, connect the supply through a resistor of 47 to 100 ohms.

With a voltmeter having an input impedance of 20kΩ/V or more, set to a range which extends to at least 15 volts, check the following points:

Local Oscillator supply.	TP1	12V
Local Oscillator Emitter.	TP2	1.8V
Q1 Source	TP3	0.55V
Q2 Source.	TP4	0.8V

With a diode sniffer, tune L5 for maximum deflection, then tune L6. There will be slightly less deflection for L6 but the meter should move significantly. These two circuits interact so repeat the process. Set VC1 so that it is about half engaged and re-tune L5 for the correct frequency. If you have a frequency counter, couple it loosely to L6 using a single or two turn loop and re-tune L5 for the correct crystal frequency. If you do not have a counter don't worry as the oscillator is unlikely to be far out and can be trimmed to frequency using an off air signal later. For fine trimming adjust VC1.

Now increase the current limit on your power supply to 500mA, or reduce the series resistor to about 22 ohms and connect the EXT PTT input (R26) to the +ve supply. You should hear the relays click. Now check the following voltages:

TX Supply	TP5	13V
Q4 Emitter	TP6	1.8V
Q5 Emitter	TP7	0.9V
Q6 Emitter	TP8	0.3V

Satisfactory results to these tests give us some confidence that resistors and semiconductors are correctly placed and there are no disastrous short circuits between tracks!

As taking the PCB out of the box to correct errors is very tedious, it was found it best to run through the tune-up procedure with the PCB out of the box first and carry out a final re-tune later after fitting the fully tested PCB into the box.

First, however, fit the 10mm spacers in the box and screw the PCB in place temporarily, Next, fit R29 in place and bolt it firmly to the box before soldering it in place. Now un-bolt R29 and remove the PCB for initial set-up and test. Fit 'tails' of 22SWG tinned copper wire to the IF and Antenna pins and adjacent ground pins then connect the loose BNC sockets to these tails with the ground 'tails' soldered to the large solder tags supplied with the sockets.

Having already tuned the oscillator, the front end, mixer and IF stages (L1, L2 and L3 and L4) can be tuned by connecting a receiver (or transceiver) to the IF socket, and using a strong local signal.

Before tuning the transmit section, you should adjust the output of the driving transceiver to a level compatible with the attenuator R29, R30 and R31. If you are using a CB transceiver, set it to the low power (0.4W) power setting.

The input level to the transmit mixer is just 0dBm (1mw). If you are using an HF transceiver which might be capable of 100 watts or more, test the power level into a dummy load first. Once you have got the level about right, a simple way to check it is by measuring the voltage at TP9. For 4W input it should be about 15-20V and for 0.4W about 6-8V. Remember however that if you apply much too much power you could damage the transverter and you definitely will produce a poor quality signal.

With enough power applied to operate the relays and switch the transverter to transmit, work through the tuned circuits in the transmit path using the diode probe and an oscilloscope, millivolt meter or a 50μA meter movement. Start at the junction of L7, C22 and C23 and tune L7 for maximum. Next move on to the base of Q4 and tune L8 for Maximum. Repeat the process for the base of Q5, tuning L9 and L10 for maximum.

After this, sufficient power should be present in the self supporting coils to give a reading on the RF sniffer. Hold its loop near the circuit to be tuned and tune for 'maximum smoke'. At this stage, a dummy load should be fitted to the output. If you have a power meter, put this in the output circuit. To tune the driver and PA stages a number of variable capacitors need to be tuned and some will interact with others so you should expect to adjust each several times to achieve maximum output. Next, check that the output is at the correct frequency within the 4m band ie between 70.000 and 70.500MHz. This can be done using a frequency counter and a wavemeter, taking a wide sweep on the later to ensure there are no measurable spurious products. Also, check that the output falls to zero when the drive is removed while the transverter is held in transmit mode using the PTT. If it does not, there is self oscillation. Under no circumstances transmit until this problem is solved!

Assuming all is well, you are ready to go on the air for a test.

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