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VHF/UHF Receivers, Transmitters and Transceivers

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. Chris Lorek, G4HCL, has written a section on choosing a commercial rig or a PMR rig for conversion to the amateur bands. There are references in the bibliography to help you locate equipment and component suppliers.

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), 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.

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. Single-mode 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 on

the VHF/UHF bands 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 and the equipment is designed to run 24 hours a day without a break. 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 bandwidth of the IF filters? Is it 50kHz, 25kHz or 12.5kHz? The latter two are preferable, whereas the 50kHz bandwidth would indicate that the equipment is many years old and 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*, 6th edition, 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

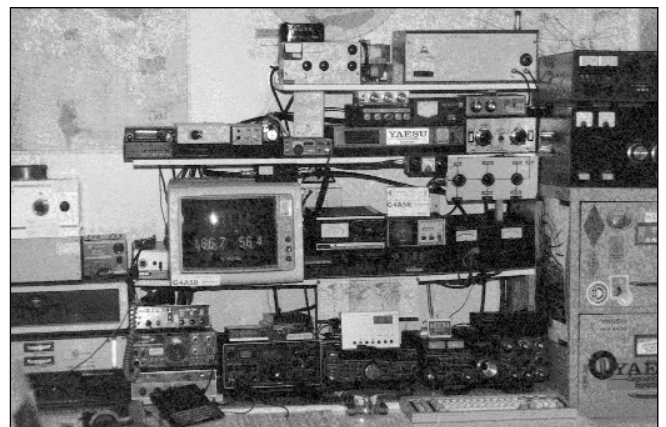


Fig 9.1: The VHF station of David Butler, G4ASR

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 70cm. 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 vertical antenna, although you might consider using a small 4 or 5-element beam 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.

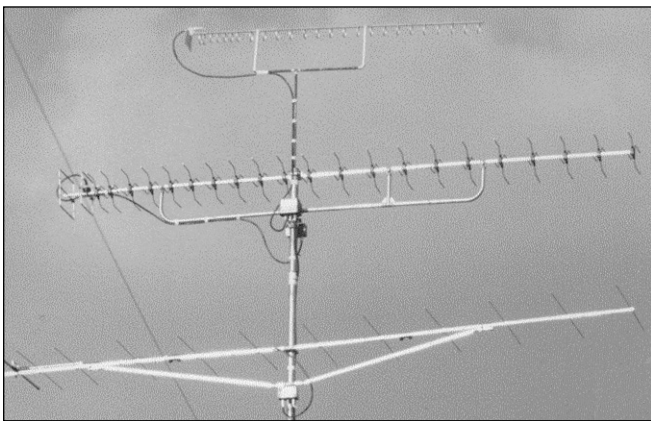


Fig 9.2: VHF/UHF antenna system of Richard Girling, G4FCD (circa 1991)

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. 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 QTH, the best location 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 70cm 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 $0.5\mu\text{V}$ 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 6-15dB 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.

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 ($1\text{nV} = 1\text{V} \times 10^{-9}$). 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

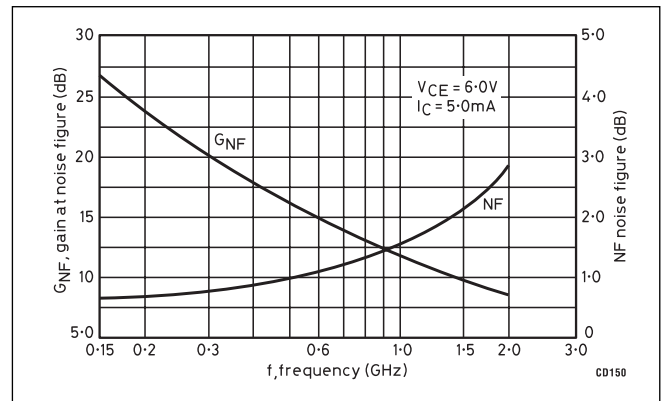


Fig 9.3: Gain at noise figure and noise figure versus frequency

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 (OK). 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 9.3 and 9.4 illustrate this feature. The transistor is capable of operation up to 2.0GHz. As Fig 9.4 shows, however, the gain and noise figure falls but the actual noise figure increases with increasing frequency. This characteristic is also shown in Fig 9.4 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

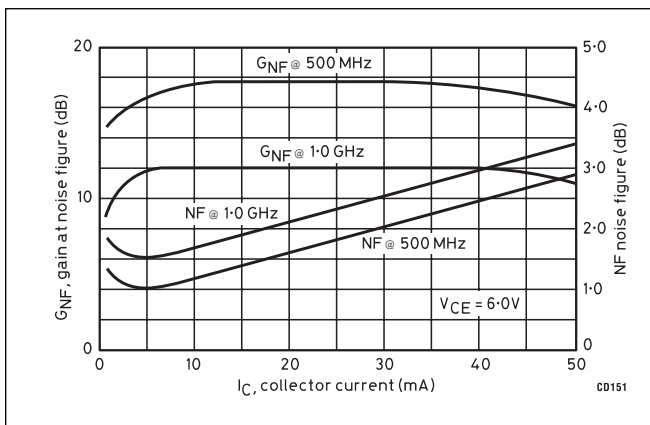


Fig 9.4: Gain at noise figure and noise figure versus collector current

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 9.3 and 9.4 applies only when the device is connected to a 50Ω 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 these 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 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, e.g. 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 drive 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 is allowed to produce 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 a 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 con-

trol 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 $V_1, V_2, V_3, \dots, V_8$ and the stage gains by $G_1, G_2, G_3, \dots, G_8$, the total noise present at the receiver detector will be:

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

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 degenerative 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 $G_4 \dots G_8$ 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.

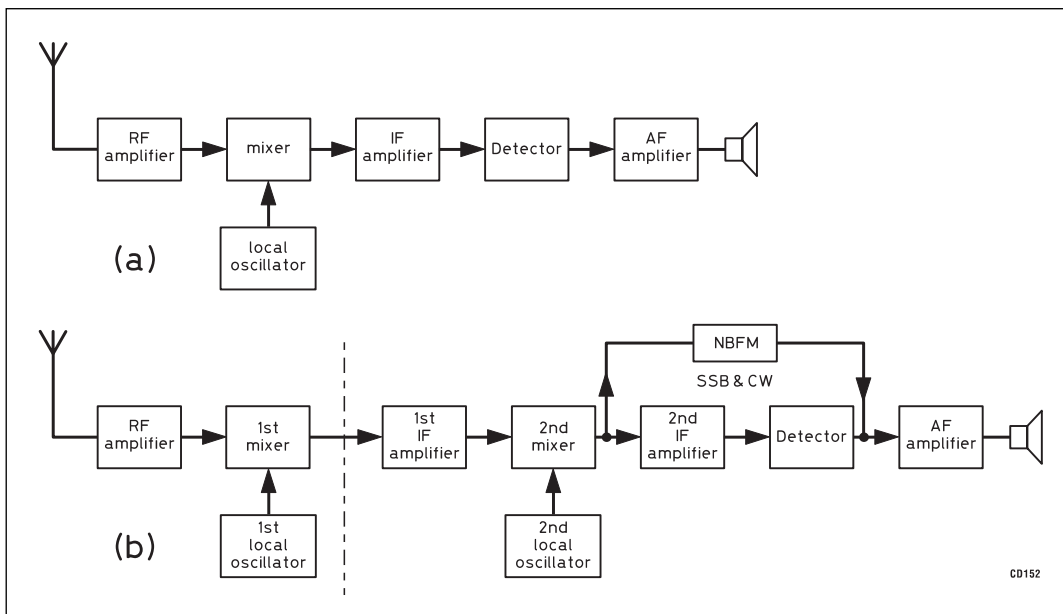
(b) Closely sited input and output circuits.

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 collector/drain) on

Fig 9.5: 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



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 9.5. Fig 9.5(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 9.5(b) shows a double superheterodyne. This can be a purpose-built receiver (or as illustrated in Fig 9.5a) 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 (x10 to x30).

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 (x3 to x4). 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. 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.

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

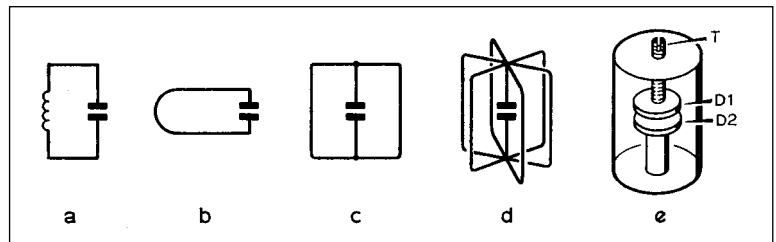


Fig 9.6: Progressive development of tuned circuits from a coil to a cavity as the frequency is increased

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 9.6 illustrates how progressively lower inductances are used to tune a fixed capacitor to higher frequencies. In Fig 9.6(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 9.6(c). This halves the inductance and can be very convenient for filters.

Fig 9.6(d) represents a multiplication of this structure and in Fig 9.6(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 rumbatron, otherwise it is a coaxial cavity.

The simple hairpin, shown at Fig 9.6(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_u of the resonator is dictated by the size of the shield, which can be round or square. Q_u is made higher by silver plat-

Parameter	HRW (231MT-10001A)	HRQ (232MT-1001A)
Centre frequency (MHz)	435	435
Bandwidth at 3dB (MHz)	12 min	11
Attenuation (dB)	20 min at ± 30 MHz	25 min at ± 15 MHz
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

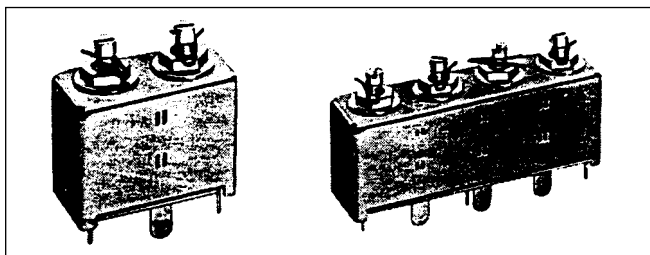


Fig 9.7: (left) Toko resonators. (left) HRW and (right) HRQ

ing 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_u . 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 15MHz; with a ripple factor of 2dB and insertion loss of 4dB (see Table 9.1). Fig 9.7 shows pictures of the helical filters tested. The test circuit is shown in Fig 9.8, and Figs 9.9 and 9.10 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 is 50 Ω 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 pre-

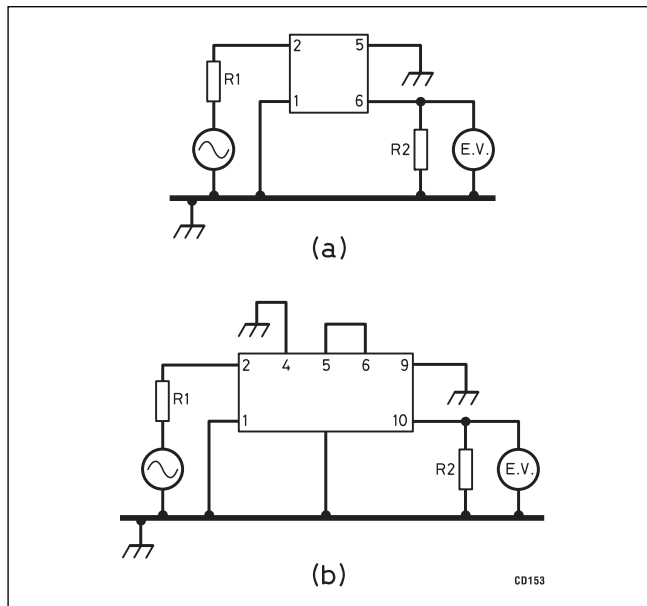


Fig 9.8: Test circuit for (a) Toko HRW and (b) Toko HRQ filters. The case lugs must be grounded. $R_1 = R_2 = 50\Omega$

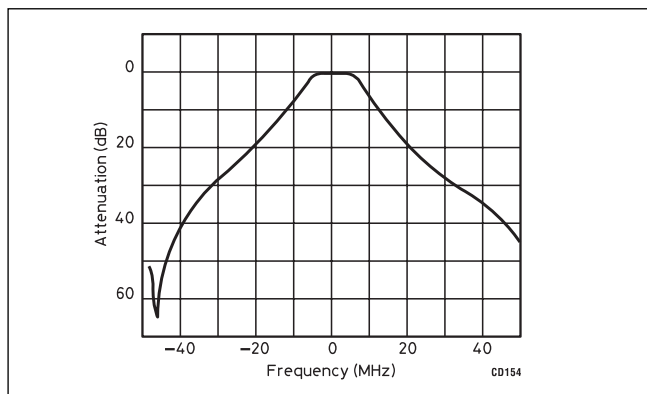


Fig 9.9: The Toko HRW frequency response

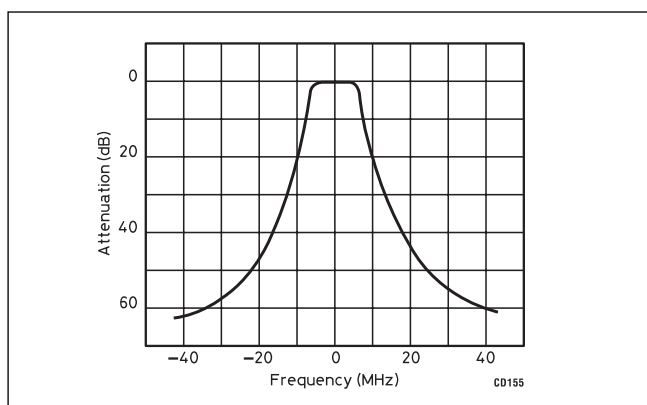


Fig 9.10: The Toko HRQ frequency response

amplifier. Such an amplifier should have the lowest possible noise figure and just sufficient gain to ensure that the overall performance is satisfactory.

Fig 9.11 shows the improvement to be expected from a pre-amplifier 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

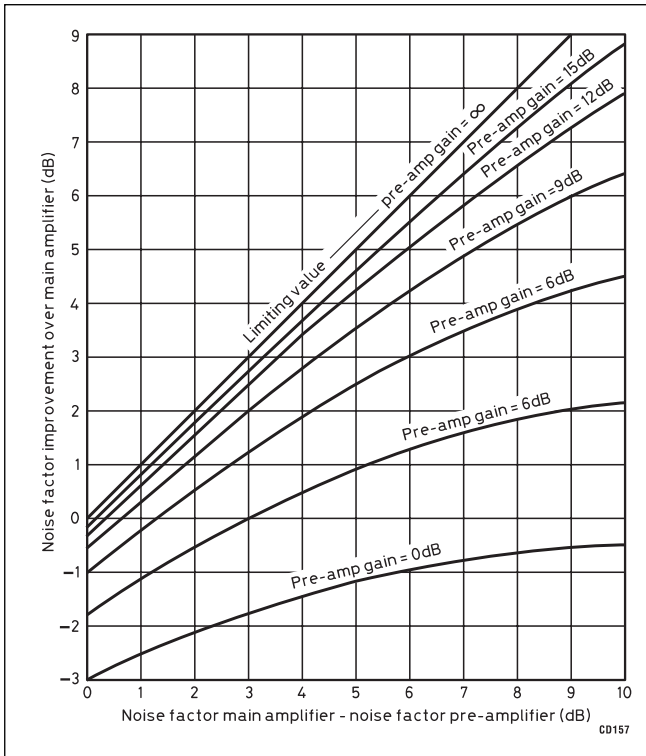


Fig 9.11: Receiver noise figures

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 9.11, if the preamplifier has a gain of 10dB, the overall noise figure will be better than 4.1dB. Increasing the gain of the pre-amplifier 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.

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 9.12 shows how such an RF stage for a 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.

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, ensur-

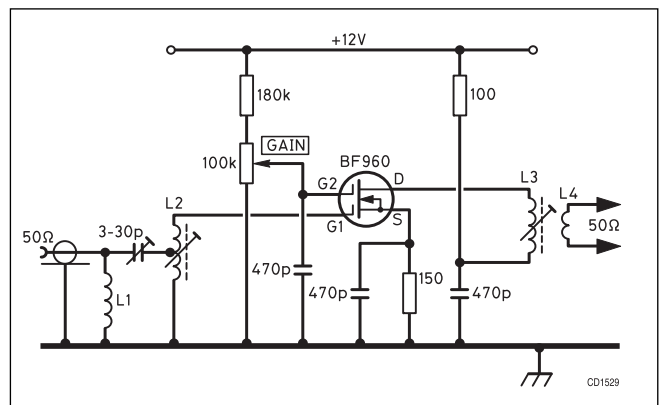


Fig 9.12: Dual-gate MOSFET RF amplifier with gain control. L1: 3 turns 1.0mm enamelled wire, 6.0mm ID, 8.0mm long. L2: 2 turns 1.0mm tinned copper wire on 8.0mm former 10mm long, tapped at 1½ turns and tuned with dust core. L3: 6 turns 1.0mm enamelled wire on 8mm former tuned with dust core and coupled to L4. L4: 2 turns 1.0mm enamelled wire on same former, close spaced to capacitor end of L3

ing, however, that the gates are never positively biased with respect to the channel.

The circuit for a GaAsFET preamplifier [1] that incorporates a self-biasing arrangement is shown in Fig 9.13. 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 enamelled copper wire
432MHz: 5 turns 0.5mm enamelled copper wire centre tapped as a 4:1 transformer on an Amidon T-20-12 toroid Core

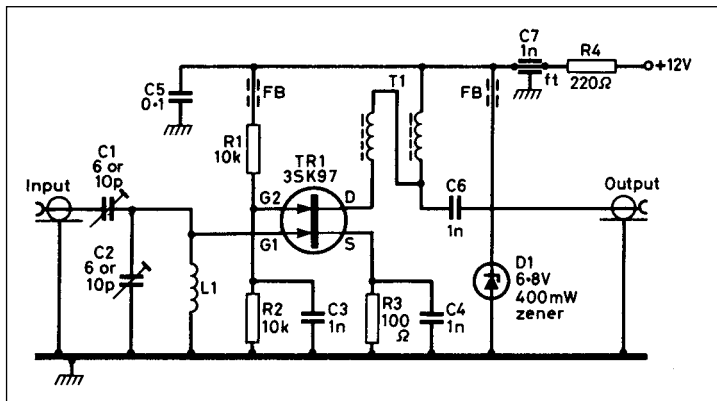


Fig 9.13: GaAsFET preamplifier for 144 or 432MHz

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.

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 9.14(a), or by the switching action between forward and reverse conduction as shown in Fig 9.14(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 9.14(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.

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

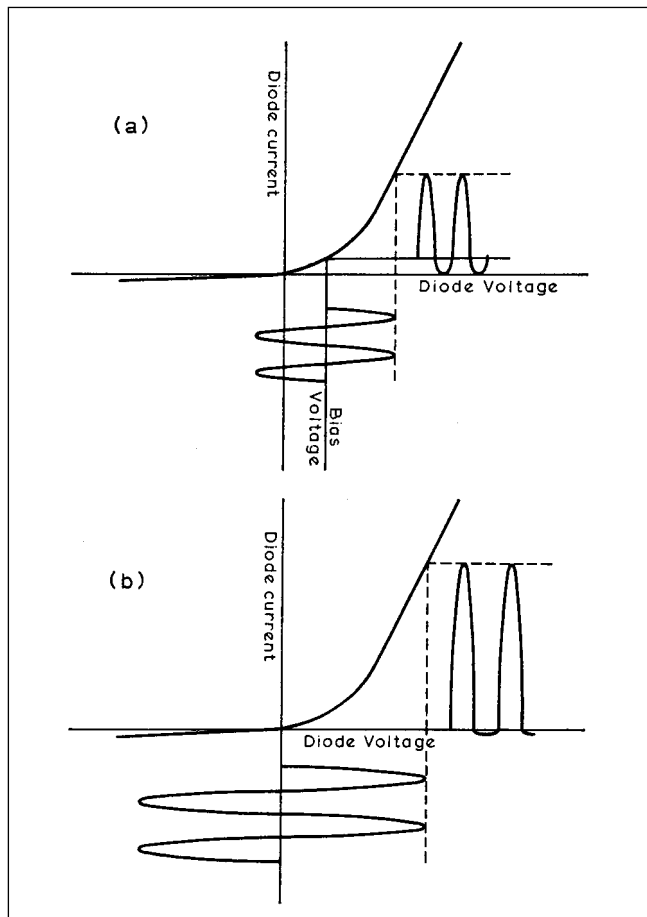


Fig 9.14: Working conditions of diodes mixers. In (a) forward bias is required to provide an optimum point. In (b) the local oscillator is higher and no bias is required

noise performance. It is always necessary to include a low-pass filter between the mixer and the antenna to reduce LO radiation from the antenna.

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 mix-

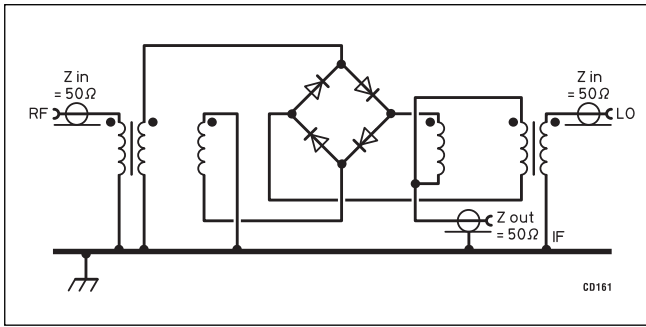


Fig 9.15: Typical diode ring mixer circuit (ARRL Handbook)

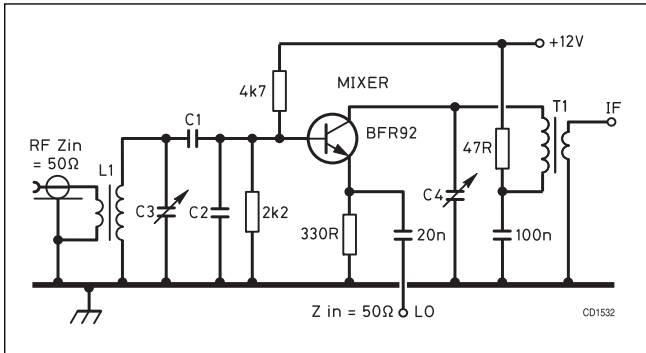


Fig 9.16: Bipolar mixer circuit (ARRL Handbook)

ers operating at these frequencies. The diode ring mixer as shown in Fig 9.15 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 Q 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 9.16 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 C5 has a low reactance at the IF. 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.

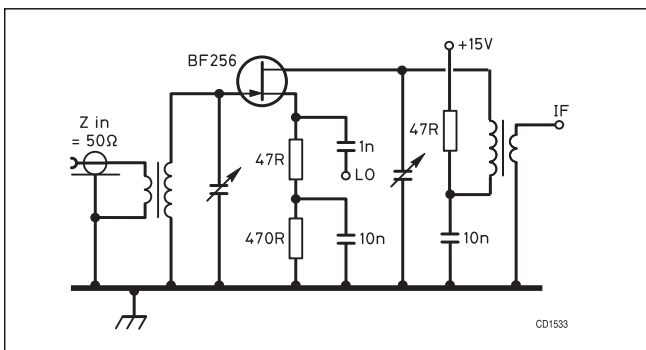


Fig 9.17: A JFET mixer (ARRL Handbook)

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.

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 9.17. 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 9.17. 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 9.18 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.

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

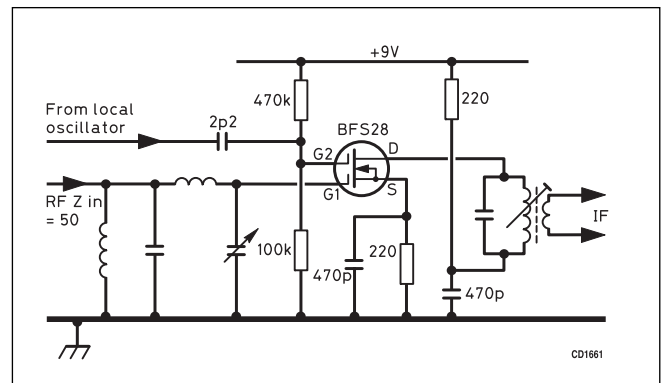


Fig 9.18: A dual-gate MOSFET mixer

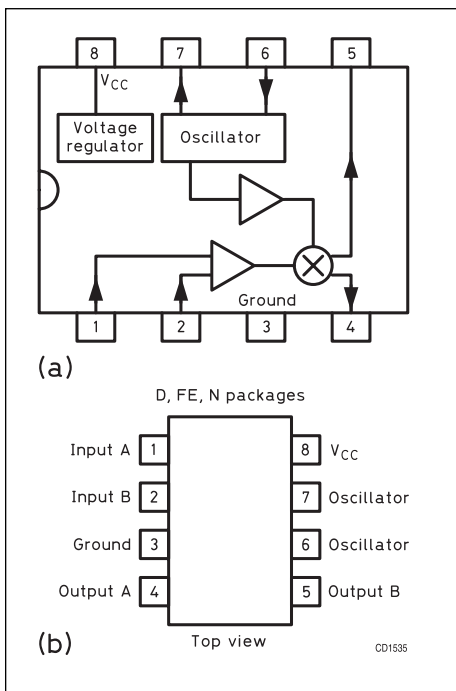


Fig 9.19: NE612 double-balanced mixer and oscillator. (a) Block diagram. (b) Pin configuration (Signetics)

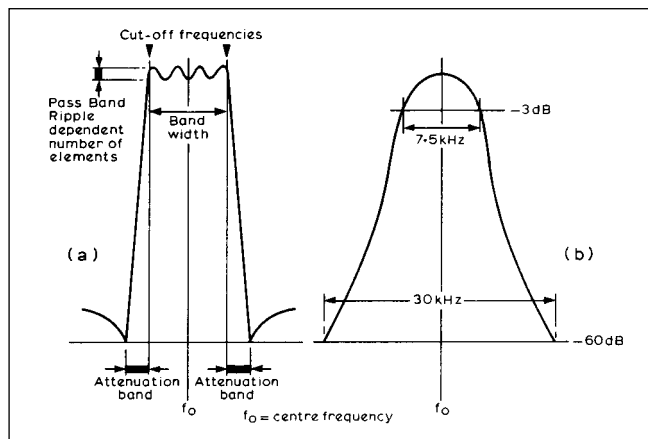


Fig 9.21: Block filter characteristics (a) SSB, (b) NBFM

ferred). The low power consumption, typically 2.4mA at $V_{cc} = 6V$, 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

These filters are available in two, four or six-pole versions; a six-pole filter will give the best attenuation to out-of-pass-band signals. Good selectivity must be achieved before the first IF amplifier and the crystal filter will give superior selectivity compared to an LC filter. Crystal filters have low input and output impedances, the actual impedance being dependent on the number of poles. These impedances vary from 3kΩ/3pF at 10.7MHz to 500Ω/1pF at 75MHz and are given in the maker’s literature.

The preferred matching mechanism to the first mixer is the pi-network as illustrated in the mixer circuits. This network has an impedance inversion property. The input impedance of the crystal filter appears as an open-circuit to out-of-pass-band signals while the load presented to the drain (of a MOSFET mixer) appears as a near-short-circuit. This will enhance the mixer IM product attenuation.

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. The pass-band of the filter characteristic can possess ripple and will not introduce pre-detector audio modulation distortion. 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 modulation distortion. The -60dB bandwidth for a typical

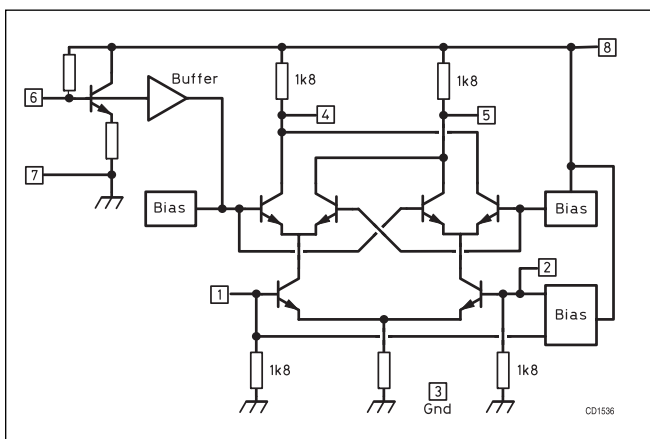


Fig 9.20: Equivalent of NE612 double-balanced mixer and oscillator (Signetics)

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 9.19 and the equivalent circuit in Fig 9.20. 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.

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 pre-

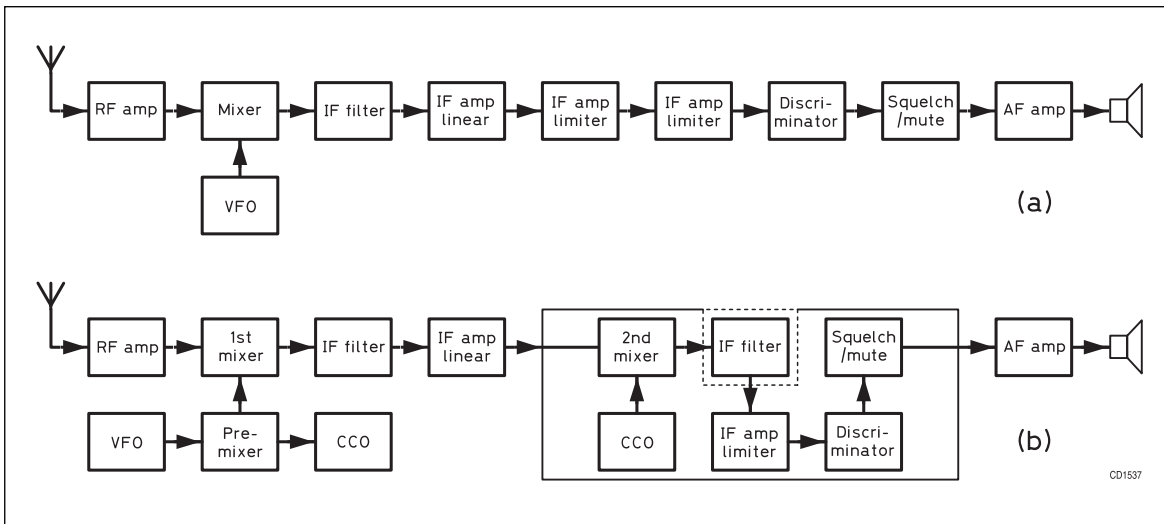


Fig 9.22: Comparison of the essential stages of receivers for (a) NBFM with discrete circuits and (b) NBFM with integrated circuits

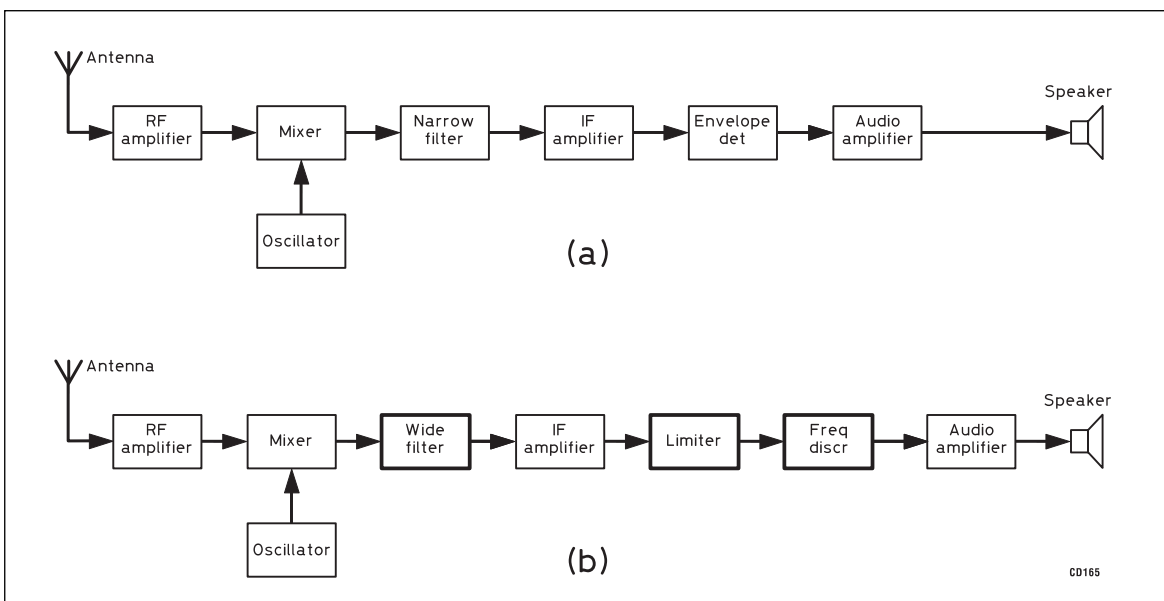


Fig 9.23: Block diagrams of (a) an AM and (b) an FM receiver. Dark borders outline the sections that are different from the AM set (ARRL Handbook)

six-pole NBFM crystal filter can be 30kHz for 12.5kHz channel spacing and 60kHz for 25kHz channel spacing, giving a shape factor of 4:1. 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 9.21** 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.

The FM Receiver

The block diagrams of an FM receiver and AM/SSB receiver are shown in **Figs 9.22 and 9.23**. 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 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.

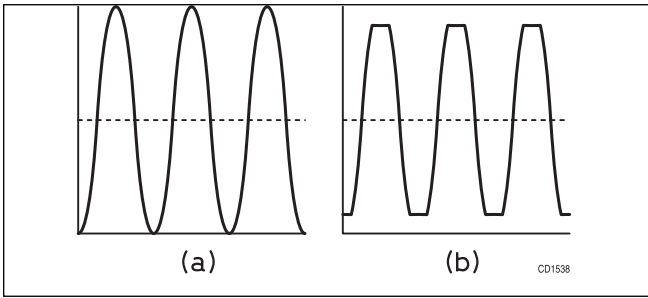


Fig 9.24: (a) Linear amplifier output waveform. (b) Limiting IF amplifier 'clipped' output waveform

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. With modern NBFM transmitters this 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 9.24 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

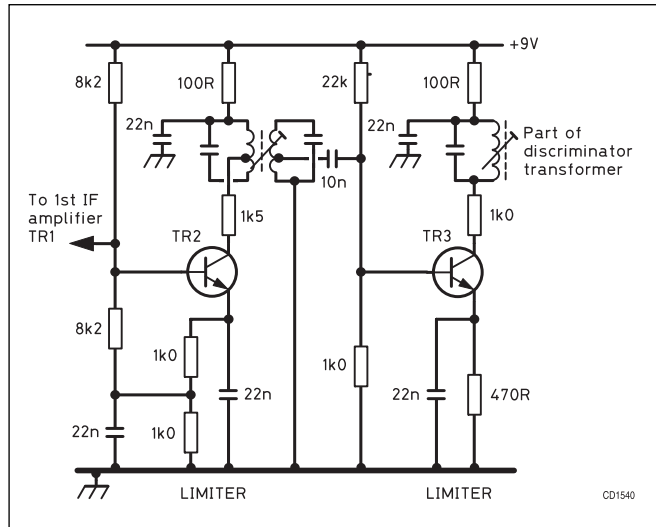


Fig 9.26: A two-stage limiter developed from the circuit of Fig 9.25

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 9.25 and 9.26. Linear amplifiers precede the limiting amplifiers. The base bias on the final IF amplifier in Fig 9.25 is varied to set the required limiting input level. In Fig 9.26 the base bias on the two final amplifiers is varied. This 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 9.26 gives an improved limiting performance compared to the circuit in Fig 9.25.

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.

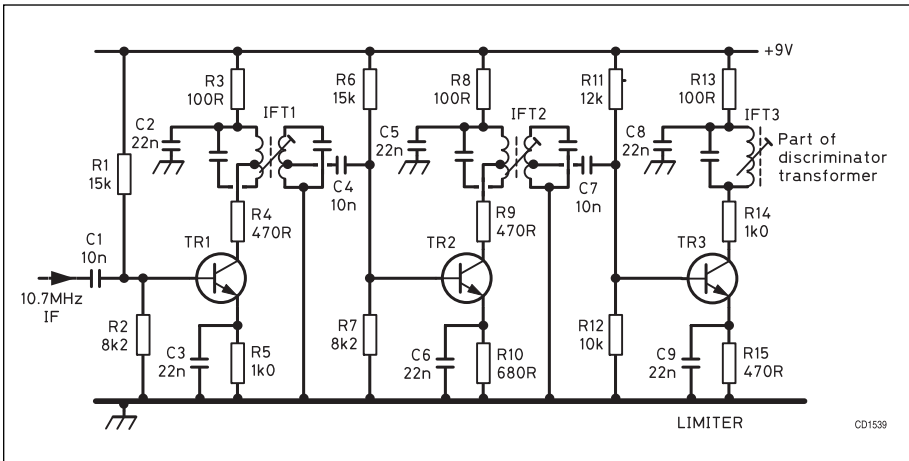


Fig 9.25: A three-stage IF amplifier and limiter. Transistors are BF194 or similar

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

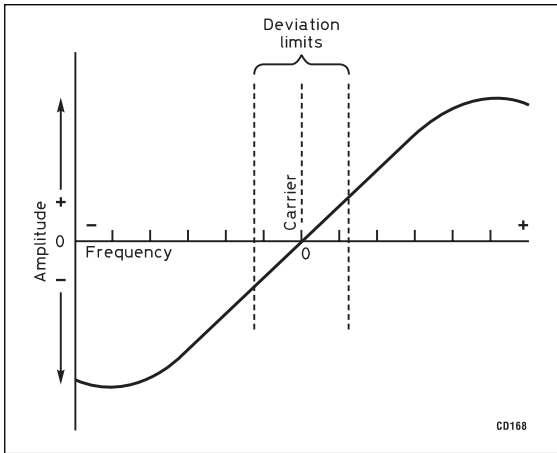


Fig 9.27: The characteristic of an FM discriminator

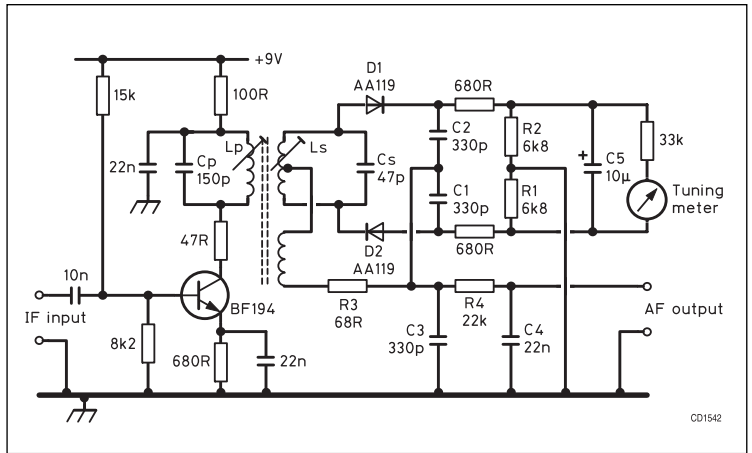


Fig 3.29: A practical ratio detector circuit

deviation) corresponds to the amplitude of the audio signal applied to the transmitter modulator.

For example, if the transmitted carrier is deviated by $\pm 2.5\text{kHz}$ by the modulator, the received carrier will be changing in frequency by $\pm 2.5\text{kHz}$.

This frequency 'swing' is detected by an FM discriminator. Fig 9.27 shows clearly the 'S' curve characteristic of the discriminator. Provided the swing is in the linear (straight) portion of the curve, ie about $\pm 6\text{kHz}$, very little distortion will be present in the recovered audio signal.

Maximum frequency deviation in amateur FM equipment is limited to $\pm 5\text{kHz}$ compared to $\pm 75\text{kHz}$ 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 9.28. 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 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.

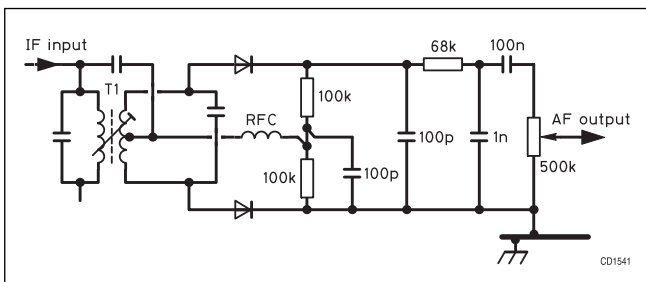


Fig 9.28: The Foster-Seeley discriminator

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 9.29. 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

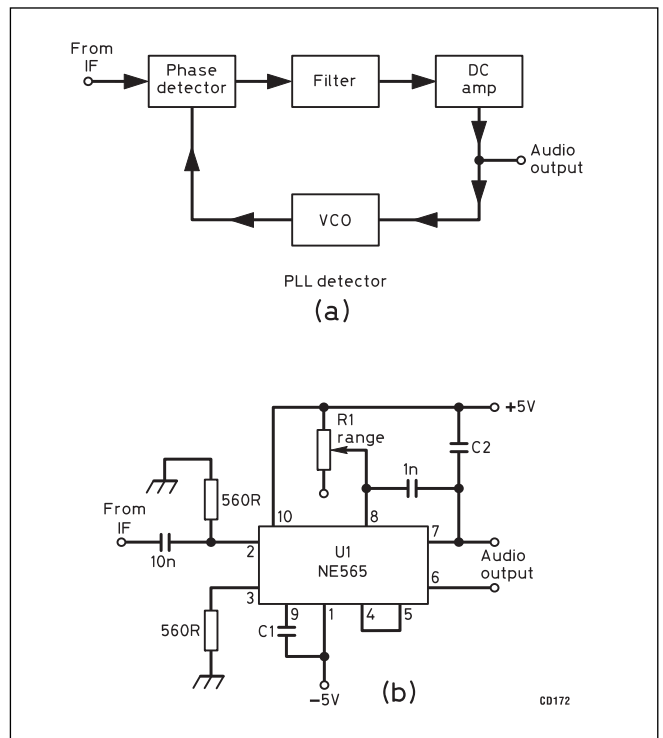


Fig 9.30: (a) Block diagram of a PLL demodulator. (b) Complete PLL circuit

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 9.30. The block diagram is shown in Fig 9.30(a) and the circuit including external components in Fig 9.30(b).

Referring to Fig 9.30(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 9.30(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.

Quadrature detectors

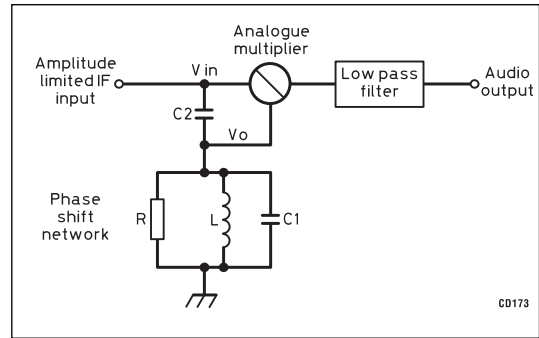
This detector is also known as the quadrature discriminator (sometimes as the coincidence detector). The symbolic circuit is shown in Fig 9.31. It is found in virtually all NBFM subsystem ICs. Alignment is simple, there being only one coil to align for maximum audio output.

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 to recover the 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 9.31: Symbolic circuit of a quadrature detector



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 from 3kHz to 1kHz.

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 9.32 and in a practical circuit in Fig 9.33.

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°

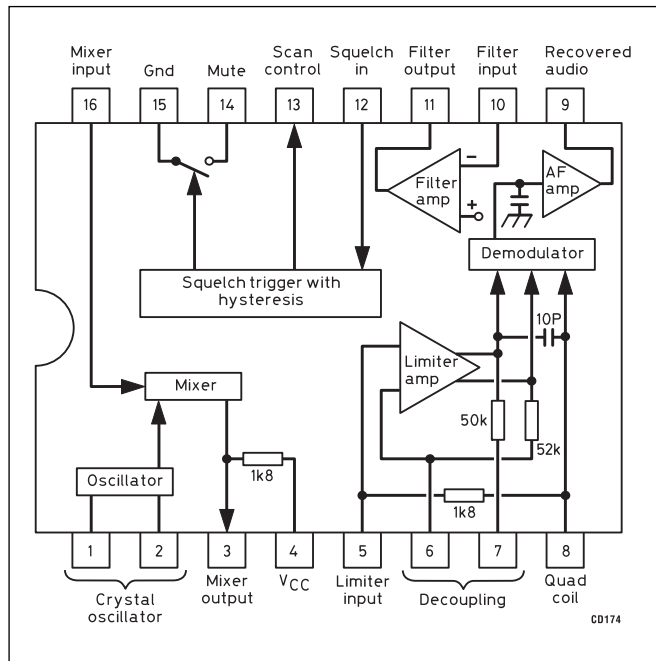


Fig 9.32: Symbolic circuit of a quadrature detector

Fig 9.33: Practical circuit using the MC3361 (Motorola)

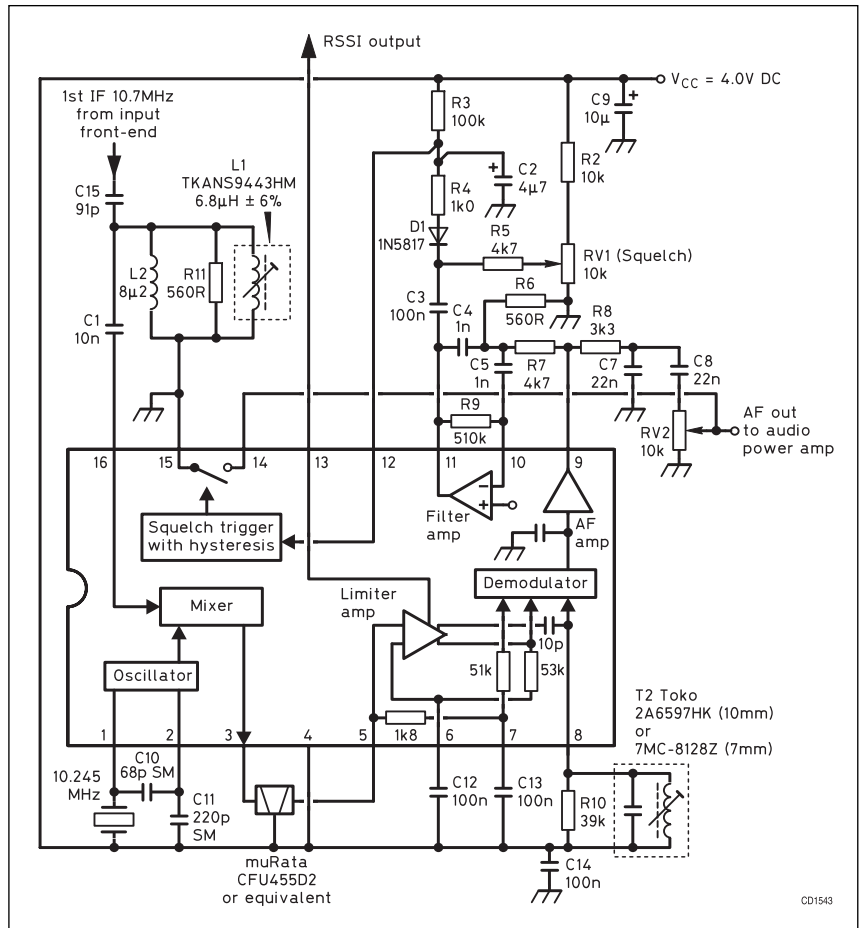
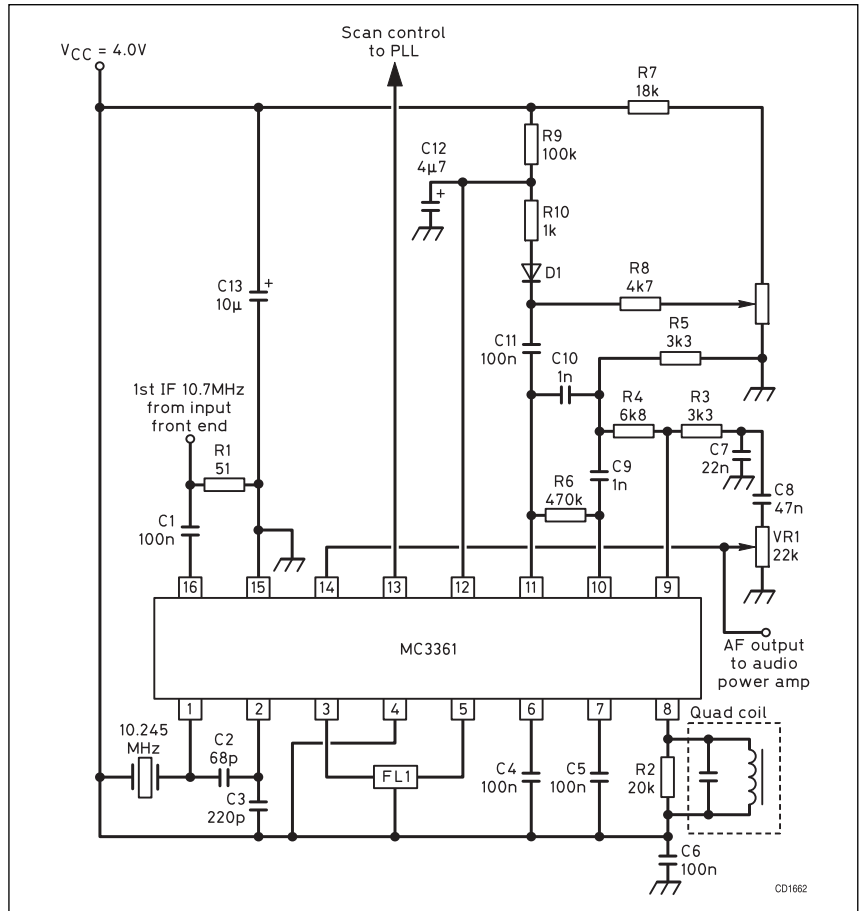
phase shift between one output port of the final limiting amplifier and one input port of the quadrature detector. The quadrature coil is shunted by parallel resistor R2. This controls the linearity of the detector swing from centre frequency, hence the harmonic distortion in the recovered AF output 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 10kHz, ie above the normal audio pass-band of 300Hz to 3kHz.

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 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 RV1. 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 9.33 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.

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

Fig 9.34: Typical application for the MC3371 at 10.7MHz with a parallel LC discriminator (Motorola)



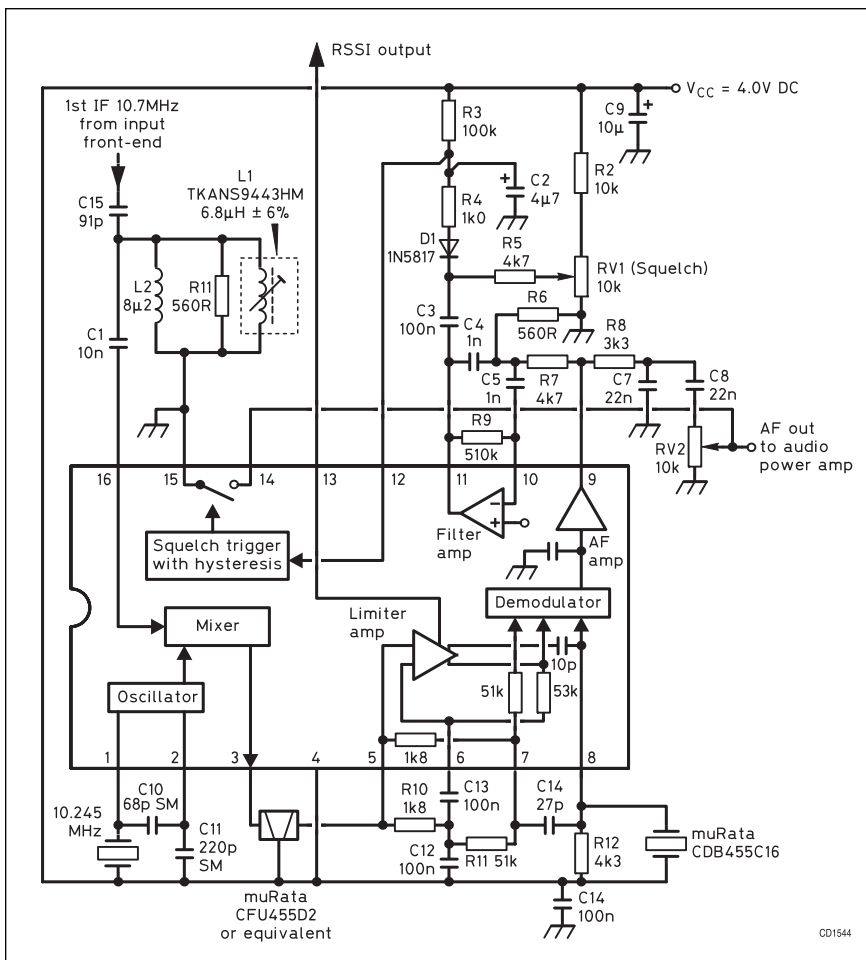


Fig 9.35: Typical application for the MC3372 at 10.7MHz with a ceramic discriminator (Motorola)

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.

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 and 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 9.34 and 9.35**.

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 $V_{cc} - 1.0V/60\mu A$, 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 [2] and show that bipolar transistors rather than FET transistors still have their place. The design for 70cm comes from Ole Nykjaer, OZ2OE [3] uses a modern E-PHEMT transistor.

Preamplifiers for 6m, 4m & 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 [2]; 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 [4 - 7]. 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 inter-modular distortions 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 recognize 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 aerial 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 Z_{in} and S_{11} values are relatively close to 50Ω , 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

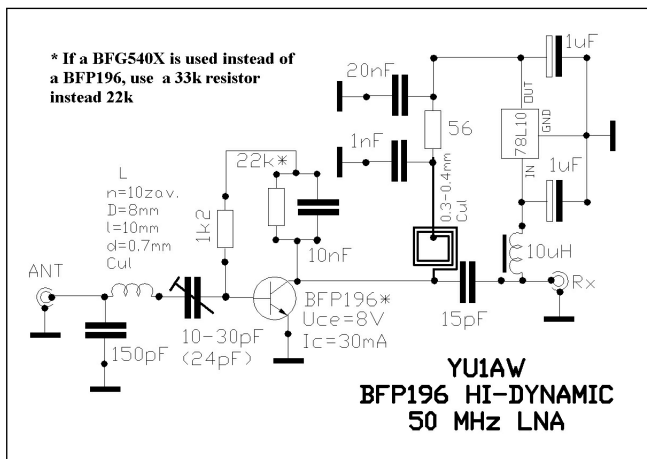


Fig 9.39: Circuit diagram for the 6m low noise amplifier

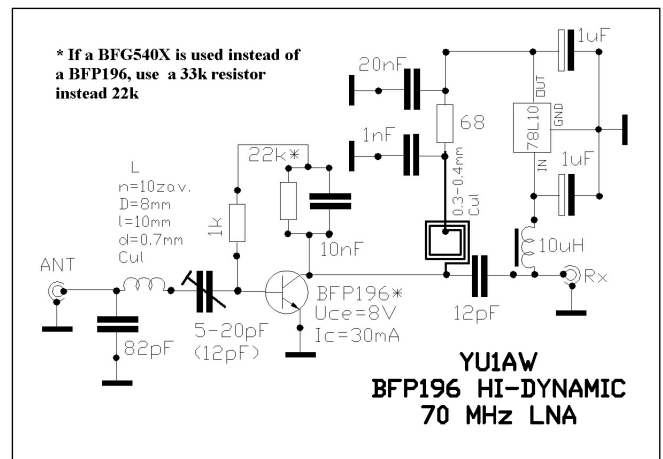


Fig 9.40: Circuit diagram for the 4m low noise preamplifier

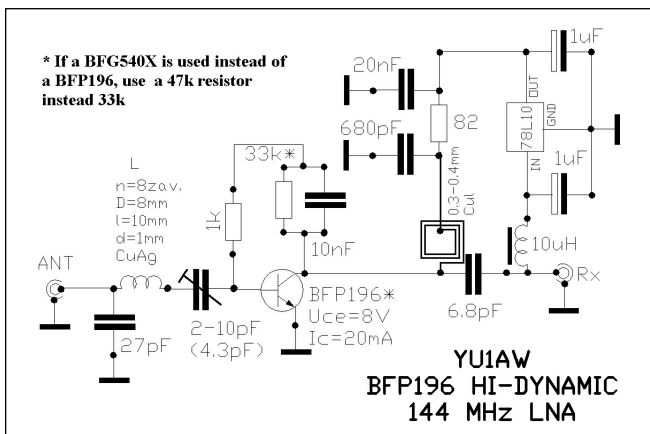


Fig 9.41: Circuit diagram for the 2m low noise amplifier

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 9.36 for 6m, Fig 9.37 for 4m and Fig 9.38 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.

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 9.39 for 6m, Fig 9.40 for 4m and Fig 9.41 for 2m), with silver plated copper wire, thickness 'd' and 'n' turns with a body diam-

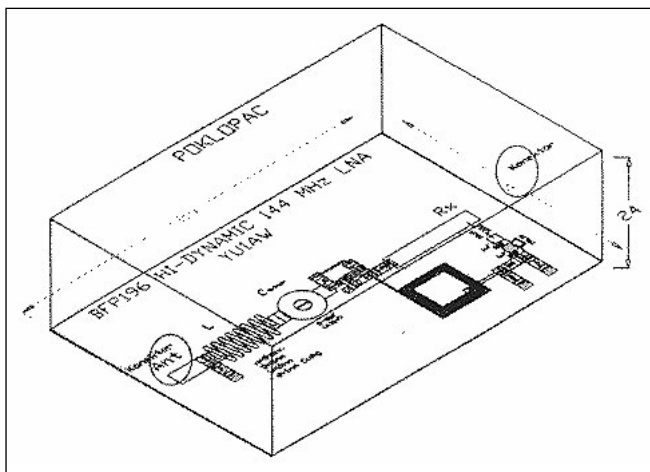


Fig 9.42: Mechanical layout of the low-noise amplifier

eter '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 9.42. 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 aerial to the amplifier input and the receiver to the output and adjust the trimmer capacitor for maximum received signal using a weaker 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 on the aerial pole and connected to the aerial with the shortest possible cable, using coaxial relays to switch the aerial 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 9.43.

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 [4 - 7], 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.

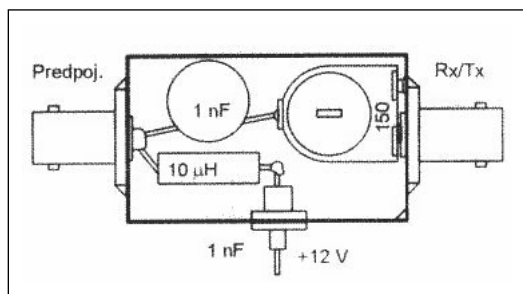


Fig 9.43: Power feed for the pre-amplifiers

As a comparison, **Figs 9.44 and 9.45** 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.

The component layouts for the amplifiers are shown in **Figs 9.46 - 9.48** (in Appendix B) and the predicted performance is shown in **Figs 9.49 - 9.56**. 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.

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.

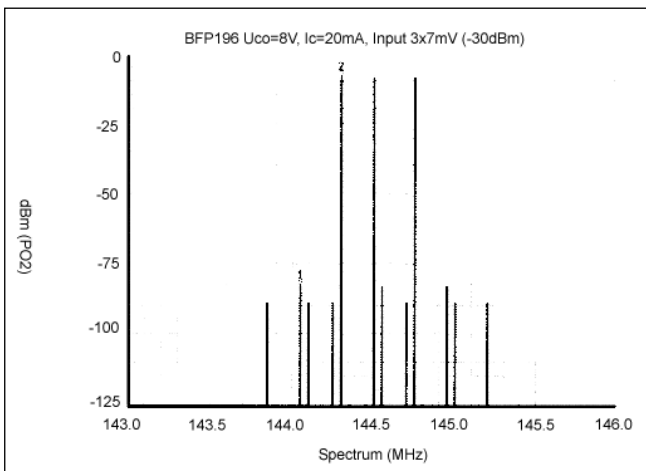


Fig 9.44: IMD products from a BFP196 preamplifier

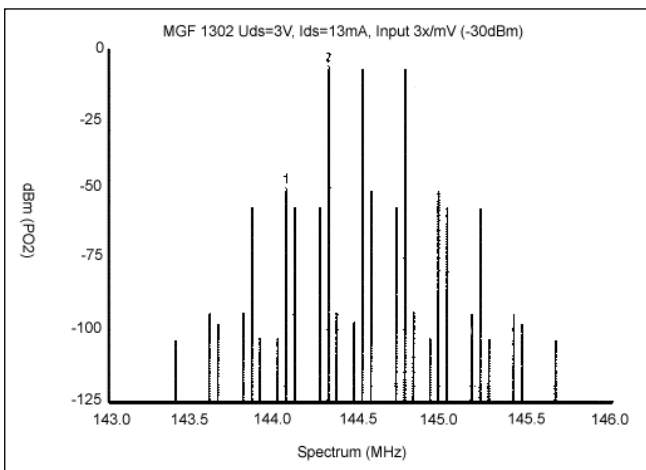


Fig 9.45: IMD products from a MGF1302 preamplifier

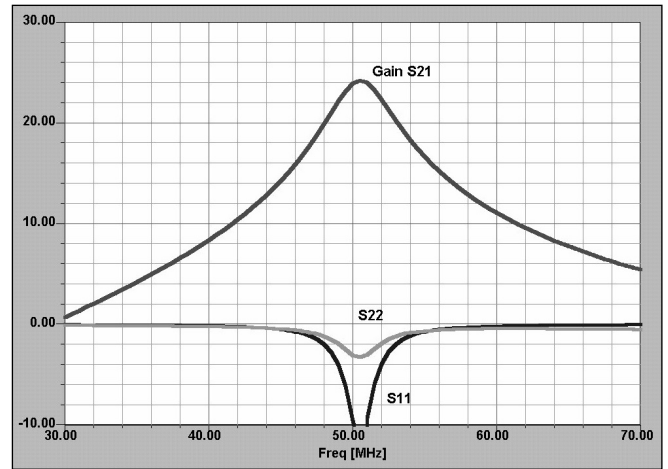


Fig 9.49: Amplification, input and output adjustment for the 6m low noise preamplifier

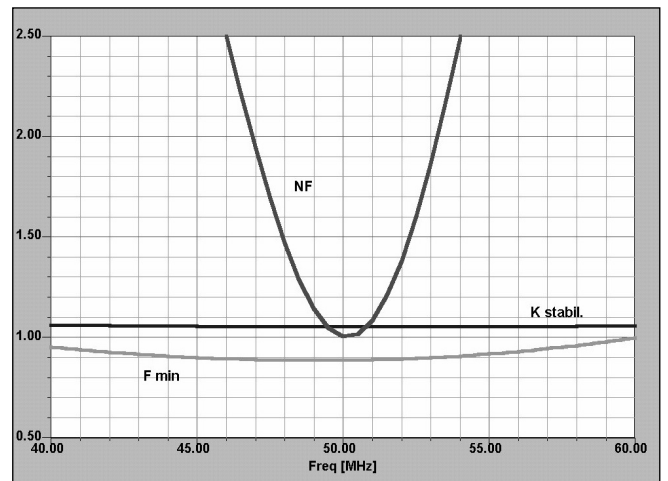


Fig 9.50: Noise figure, minimum noise and stability of the 6m low noise preamplifier

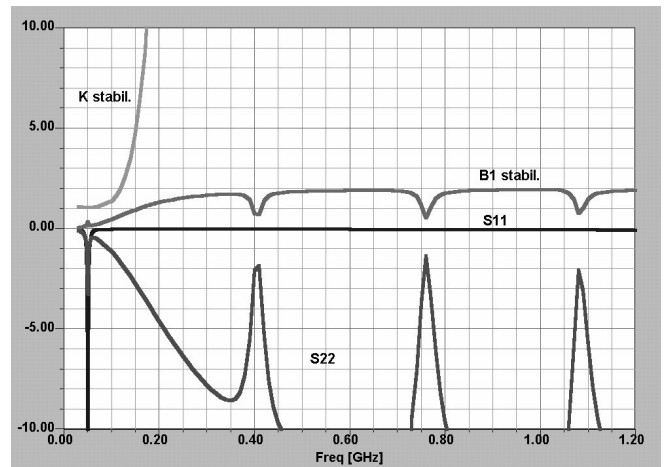


Fig 9.51: Stability factor and adjustment for the 6m low noise preamplifier

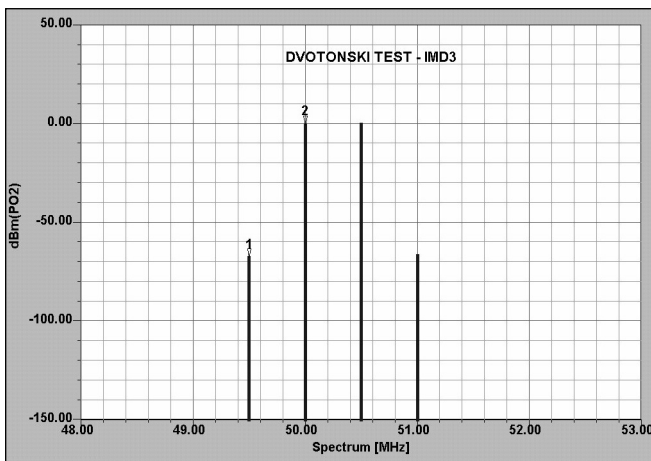


Fig 9.52: Two tone test and IMD products for the 6m low noise preamplifier

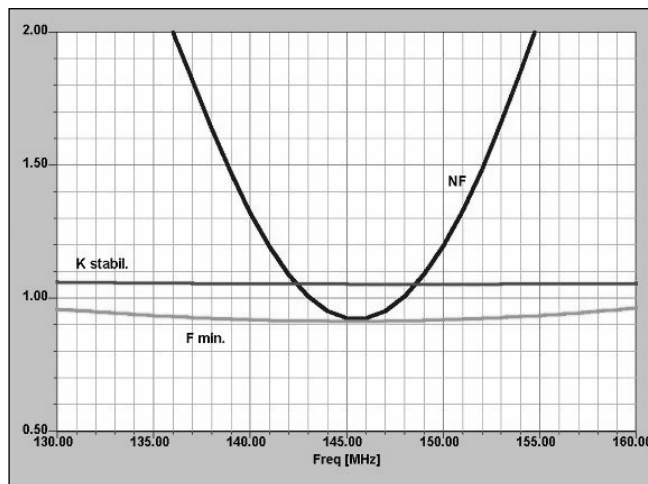


Fig 9.55: Stability factor and adjustment for the 2m low noise preamplifier

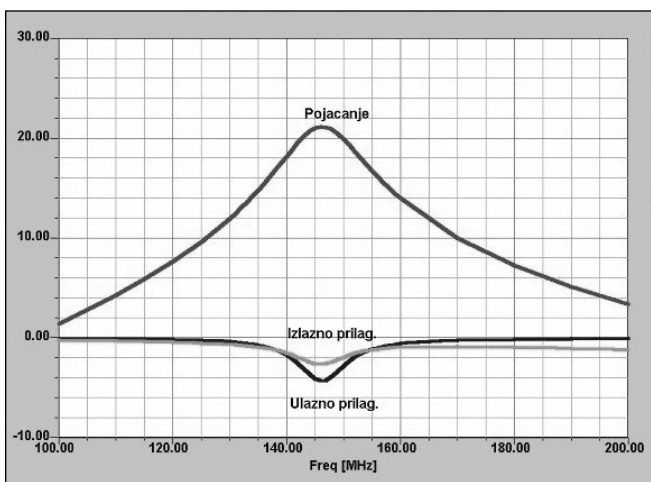


Fig 9.53: Amplification, input and output adjustment for the 2m low noise amplifier

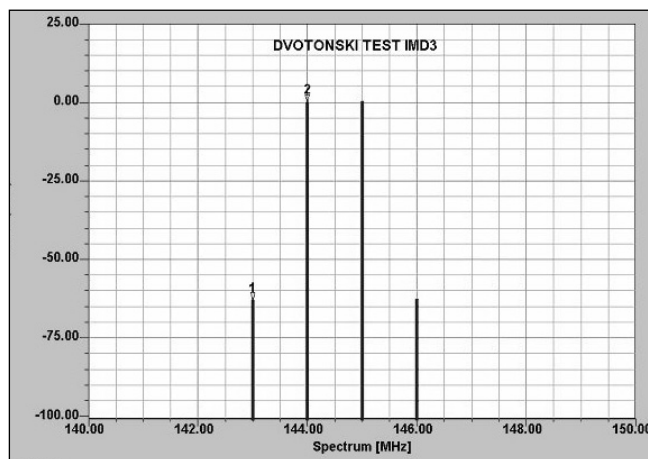


Fig 9.56: Two tone test and IMD products for the 2m low noise preamplifier

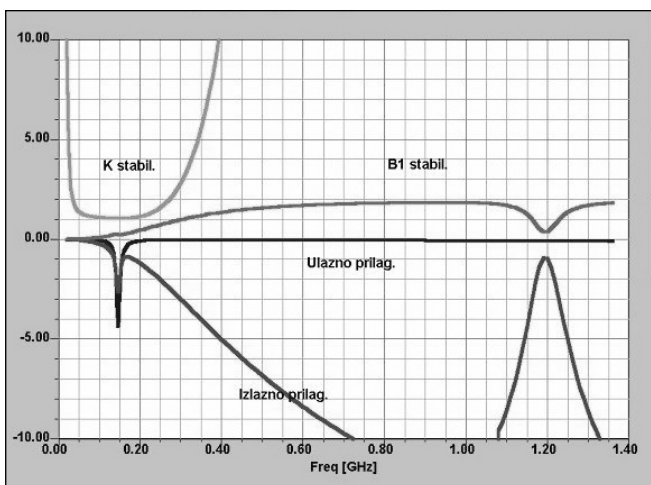


Fig 9.54: Noise figure, minimum noise and stability of the 2m low noise preamplifier

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 atten-

uated 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 100-500Ω trimmer potentiometer connected to the receiver input can be used in place of a variable attenuator. This can be built into the amplifier supply adaptor box as shown on Fig 9.43. 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.

70cm Preamplifier

The ATF-54143 is a new E-PHEMT from Agilent (formerly HP), developed for use in low noise amplifiers from VHF to 6GHz. Typical specifications at 2GHz and 3V, 60mA:

- low noise figure 0.5 dB
- high linearity 3rd order IP 36.2 dBm
- high gain 16.6 dB

Other useful features:

- enhancement mode - no negative bias!
- easy matching for best noise figure
- low cost

Fig 9.57: Pin connections and package markings for an Agilent AFT-54143 E-PHEMT

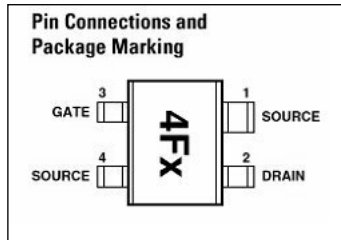
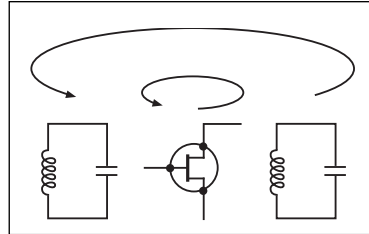


Fig 9.58: Illustration of stability in a transistor amplifier. Feedback can occur inside or outside the device



Not so useful features:

- high gain (>30 dB at VHF), can give instability
- small size 1.2 x 2mm - you need a steady hand

The package and pin connections are shown in **Fig 9.57**. For more information look up the datasheet on Agilent's homepage [8], select Products and RF & Microwave and find ATF-54143.

An amplifier must be stable, not only around the working frequency, but also on frequencies far away; **Fig 9.58** illustrates this point. Stability can be expressed in terms of k-factor. The k-factor is calculated from the transistor S-parameters. With $k > 1$ the circuit is unconditionally stable - so whatever you put on the input and output terminal, it will not oscillate. Using S-parameters for ATF-54143 and with a little help from the 'ARRL Radio Designer' circuit simulator, we get a plot of k-factor from 0.1 to 14GHz (**Fig 9.59**). As you can see, the transistor is potentially unstable below 3.5GHz - this is a wild animal!

To increase stability, some attenuation is put directly on the drain terminal. This can be in the form of a ferrite bead or a resistor. Resistors on the gate terminal will also improve stability, and must be placed after the UHF decoupling to avoid any impact on noise figure. Another stabilising technique is using just a little inductance in the source (1-2mm of source lead).

The designer used a computer program 'ARRL Radio Designer' for designing the amplifier. It ran under Windows and was based on the professional Super-Compact circuit simulator program - at a fraction of the cost. (Unfortunately ARRL has stopped selling this program so get it 'surplus' if you can). The design procedure is this:

- load S-parameters (from internet)
- run stability analysis and then increase stability with resistors and/or source feed-back as required
- next calculate input and output matching for optimum noise (a special feature of 'ARRL Radio Designer')

The program can now calculate gain, noise figure, stability, input VSWR and a lot of other parameters for the finished amplifier. After that, all that is needed is to build the thing!

The input circuit transforms the 50Ω antenna into the optimum generator impedance for the transistor. Looking at a Smith Chart we have to go from the centre of the chart to a point towards the periphery. The more transformation, the more complex (and lossy) the input circuit. Unfortunately FET transistors - at low UHF frequencies - have rather high impedances for optimum noise match, making it difficult to achieve the very low noise figure the device itself is capable of.

A great advantage of ATF-54143 is an optimum noise match close to the centre of the Smith Chart. To give some low fre-

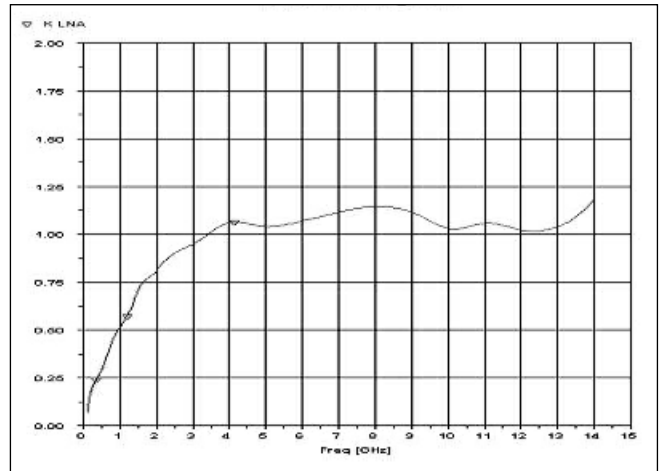


Fig 9.59: Stability factor versus frequency for an Agilent AFT-54143 E-PHEMT

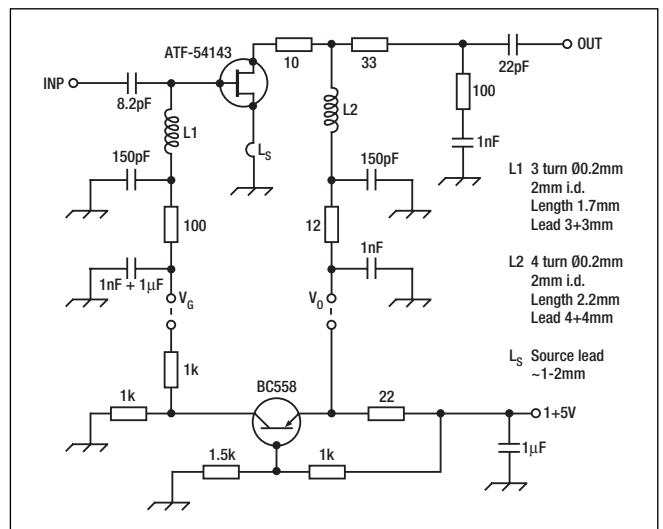


Fig 9.60: Circuit of the 432MHz preamplifier using an Agilent AFT-54143 PHEMT. L1 is 3 turns of 0.2mm wire, 2mm ID, length 1.7mm with a lead length of 3 +4mm. L2 is 4 turns of 0.3mm wire, 2mm ID, length 2.2mm with a lead length of 4 +4mm

quency cut-off the transformation is carried out by a series capacitor and a parallel inductor.

Similar to the input matching, the output match transforms the 50Ω output impedance into the load impedance for maximum gain (or max. output power depending on what you want). Again transformation is done by a parallel inductor and a series capacitor.

The ATF-54143 works with a drain voltage of 2 to 4V. Maximum drain current is 120mA. The noise figure is almost independent of drain current and (as you might expect) large signal handling improves with drain current. The designer selected 3V and 60mA as bias. (but with 10Ω drain resistor I end up with 0.6V less voltage).

Being an enhancement mode device the transistor needs a positive voltage on the gate in order to draw drain current - typically + 0.59V on the gate to give 60mA drain current. If the gate is put to ground potential the current will drop to a few microamperes of leakage current. This is more like biasing a bipolar transistor! Therefore we have the possibility of making a simple resistor bias as well as building an active bias. The circuit is shown in **Fig 9.60**.

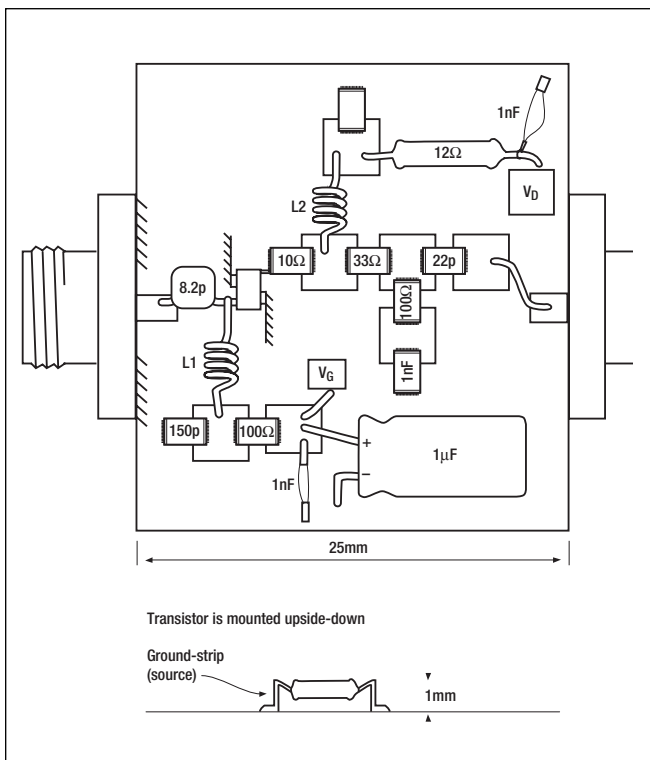


Fig 9.61: Component layout for the 432MHz amplifier using an Agilent AFT-54143 PHEMT

To test the bias circuit before using it, you can connect an ordinary NPN transistor (BC547) instead of the HEMT (V_g on base, V_d on collector). If everything works OK, you should be able to bias the collector to 60mA at 3V with active bias (but only 30mA 4V using passive bias due to different DC characteristic)

Prototypes have been built on breadboard, the layout is shown in Fig 9.61 and a picture of the prototype is shown in Fig 9.62. The material used is piece of 0.8mm epoxy board (FR4). Islands were cut out, about 2 x 2mm, for mounting the components. The SMA connectors are soldered to copper 'ground'. SMD components were used around the UHF signal and decoupling path.

Simulation shows the 432MHz amplifier to be more stable with some source self inductance. This is achieved by mounting the transistor upside down on two small ground strips each 1mm high by 4mm long. These strips add about 0.5nH to the source. The gate (and drain) is in free air, so there is no loss due to the epoxy PCB.

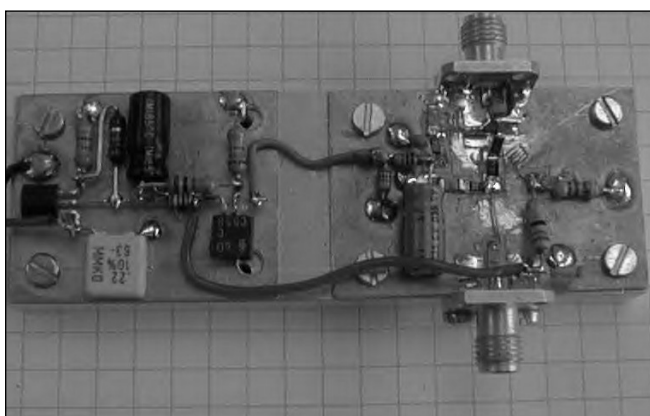
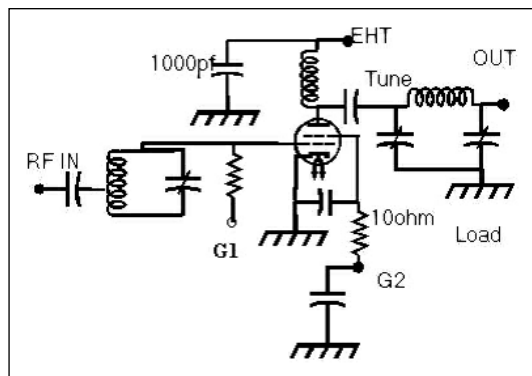


Fig 9.62: The prototype 432MHz preamplifier using an Agilent AFT-54143 PHEMT

Fig 9.63: Circuit diagram of the 6 metre power amplifier



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 6m power amplifier was designed by Geoffrey Brown, G4ICD [9] and the 70cm power amplifier is from *VHF Communications Magazine* 4/1998 [10]. 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 from 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 transistor amplifiers by Dragoslav Dobricic, YU1AW [2], for 6m and 2m are a good example of discrete semiconductors being used, and give details of several suitable devices.

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 9.63 and the parts list is shown in Table 9.2.

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 9.2: Parts list for 6m power amplifier

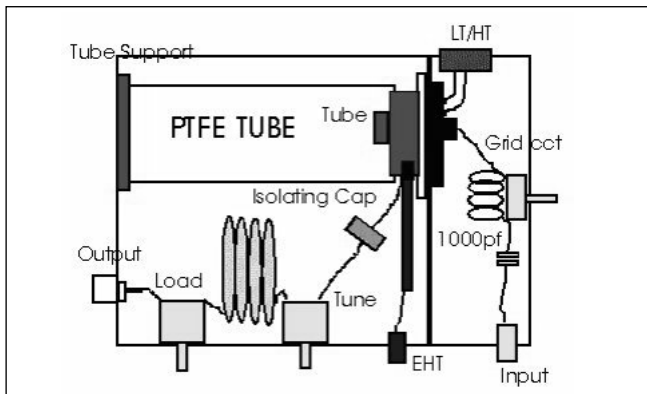


Fig 9.64: Component layout for the 6m power amplifier



Fig 9.65: The 6m power amplifier

You will need a diecast box, fan, SK600 or SK610 surplus socket for the valve, a 4CX250B valve, a couple of tuning capacitors, 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 9.64). 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 chimney, it has a hole cut in it for the exhaust and a brass shim



Fig 9.66: The A200 amplifier. (Picture supplied by the Pye Museum [24])

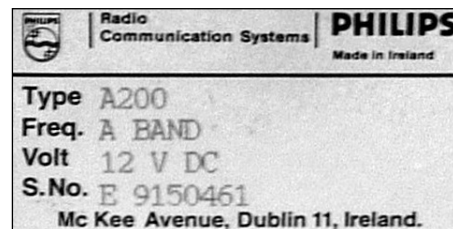


Fig 9.67: The identification plate on the A200 amplifier (Picture supplied by the Pye Museum [24])

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 Figs 9.63 and 9.65. 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 unlikely that you will buy one that does not work.

It is easy to spot the A200 by the chunky black case shown in Fig 9.66. 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 9.67. They are marked "Cat No.

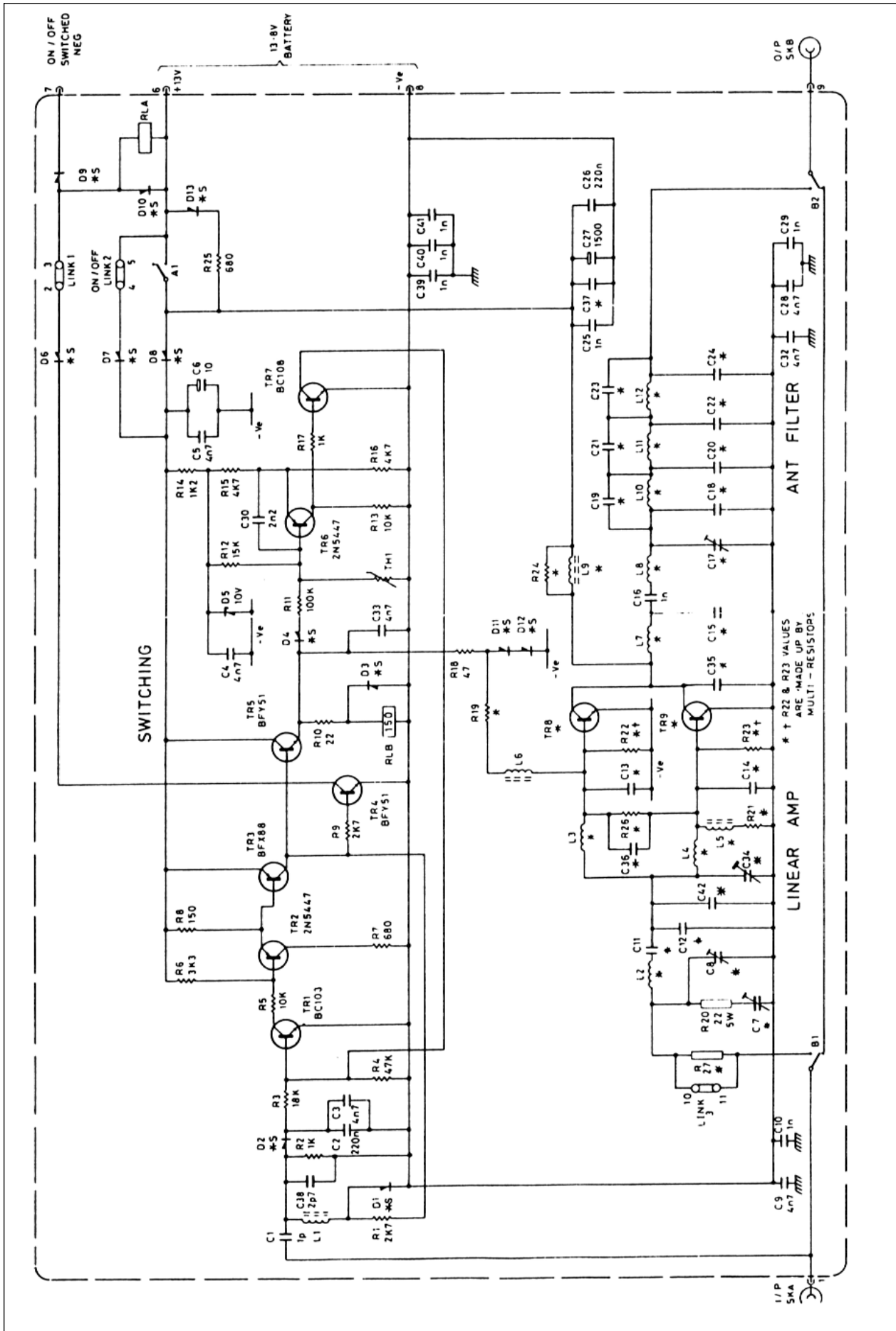


Fig 9.68: Circuit of the A200 amplifier. Reproduced with the permission of Pye Telecommunications

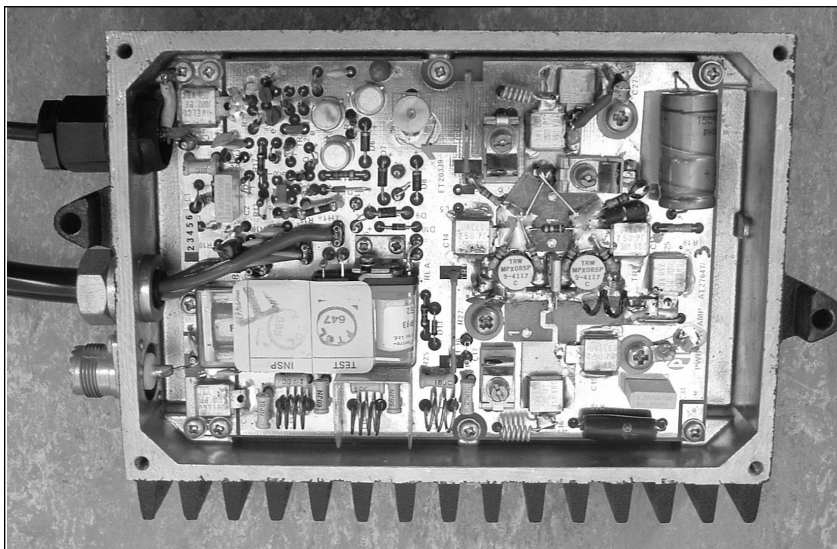


Fig 9.69: Internal view of the A200 amplifier. (Picture supplied by the Pye Museum [24])

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 [11].

Fig 9.68 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 9.69** shows an internal picture of the amplifier and **Fig 9.70** 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 10W drive to produce 60 - 70W output and will 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

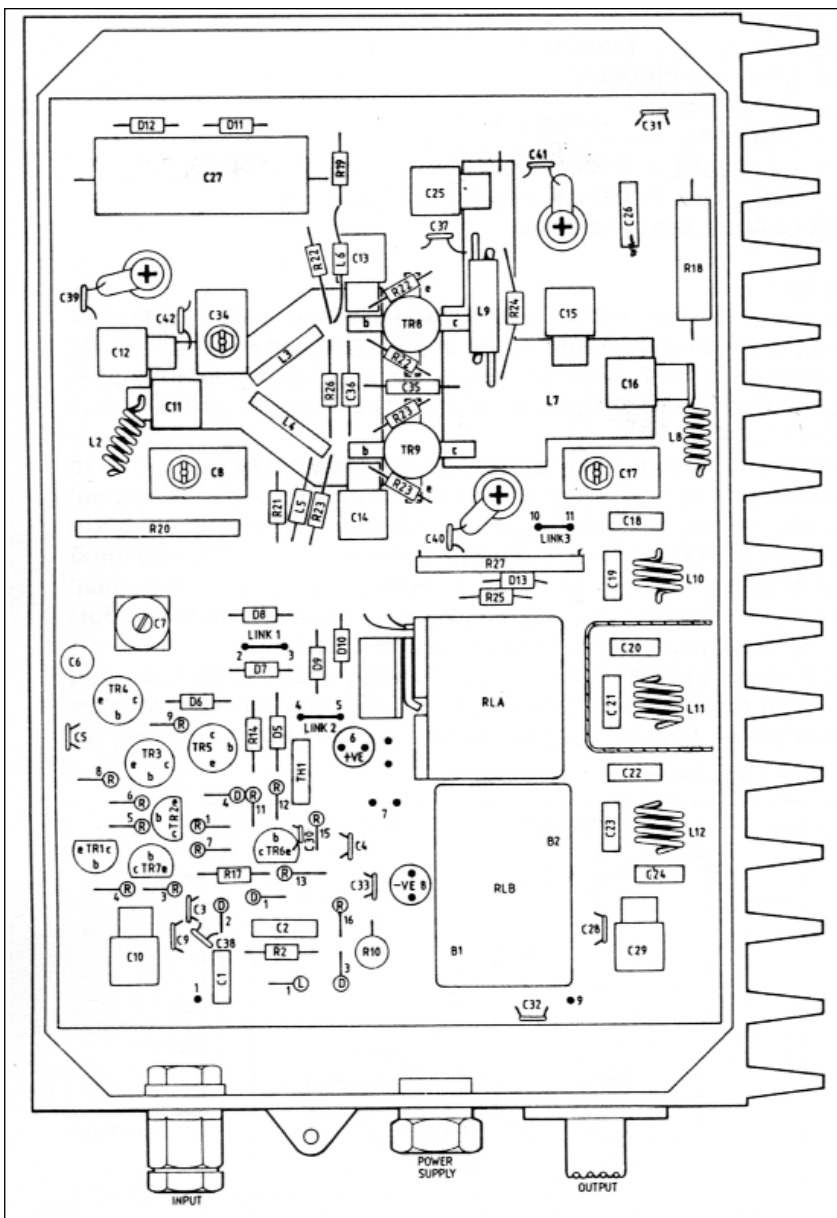


Fig 9.70: Component layout of the A200 amplifier. Reproduced with the permission of Pye Telecomm

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 over-driving the amplifier. This will still mean that you get 45-50W PEP output with the third order IMD products at least 28db down.

6m and 2m Transistor Amplifiers

Several transistorised VHF power amplifiers are described here; examples are given for the 50 and 144MHz bands. Two types of amplifiers will be considered, amplifiers with one or two transistors. Assembly instructions are given as well as explanations of some general issues with regard to the construction of VHF power amplifiers.

The first step is to choose a transistor, this ultimately dictates what will be achieved. The quality of an amplifier is defined by several parameters that define its quality. An amplifier can be deemed to be a high quality amplifier if it has the following characteristics:

- High efficiency level
- High power amplification
- Good suppression of harmonics and low level of IMD products
- Good electric and thermal stability
- Simple design and adaptability

Some of these are contradictory and often have to be solved by various compromises, while others are compatible and by creating the first we automatically create the other. High efficiency is the most important factor, maximising this ensures that the majority of the other parameters approach their optimum. Beside an arbitrarily chosen working point or class in which the transistor will operate, the matching as well as the circuit losses influences efficiency. For highest efficiency it is necessary to:

- Strictly respect AC and DC current and voltage recommendations by the manufacturer for the class of operation
- Match the transistor to the output conditions recommended by the manufacturer for maximum efficiency. Also losses in the matching circuits should be minimal
- Keep the temperature of the transistors under the maximum allowed temperature, this is achieved by adequate cooling

By fulfilling these conditions a high degree of efficiency is obtained, as well as highest output power, relatively low operating temperatures and slight harmonic and IMD distortions.

The second condition, high power amplification, is achieved by matching the input to ensure stable operation. Losses in the matching circuits directly influence the power amplification. Another extremely important condition that has to be fulfilled for high amplification is extremely good grounding of the emitter or source in FET transistors. Leads should be extremely short and therefore of extremely low resistance!

Stability of the amplifier depends on the transistor itself, but also in a large measure on other factors such as: matching, mechanical construction and separation of electric paths in the input and output circuits. This will eliminate unwanted feedback between

input and output. The most important issue for the stability of any amplifier concerns the low frequency performance where the transistors have enormous gain. It is important to make sure that the base and collector 'see' several dozen ohms. If a collector or base 'sees' a short circuit via a large capacitor or open circuit via a high inductance at low frequencies, an amplifier oscillates very easily.

The third condition, good suppression of higher harmonics and low level of generated inter-modular products (IMD) also depends on many factors. We will list just a few:

- choice of the optimal value of efficiency (Q) of output circuit
- linearity of a transistor's chosen operating point
- output matching
- small output resistance of the base bias circuit
- value of excitation power

By correct design, the unwanted secondary products generated by the amplifier could be reduced to a minimum. Output filtering can reduce these even further. Filtering cannot reduce some of the odd order inter-modulation products because they are very close to the operating frequency. They can only be influenced by correct design of the amplifier; making it linear generates a very small level of these unwanted products. Even harmonics and even series of inter-modulation products can be efficiently decreased by push-pull amplifier design. This is an advantage in comparison with amplifiers operating with two parallel transistors.

For amplifiers operating in linear classes (A and AB) the correct choice of the base bias circuit is also very important and it provides good linearity of the input signal. The stability of collector voltage also has an influence, it should be stabilised for amplifiers up to 20V, and for voltages above these it is sufficient if it has good regulation. For the linear operation of an amplifier, the level of excitation power is crucial. It should not be too high, because it 'pushes' the transistor into non-linear operation.

Thermal stability is provided by the correct choice of a heatsink and more importantly, by correctly mounting the transistor. In order to compensate for the thermal movement of the operating point it is necessary to provide thermal feedback that

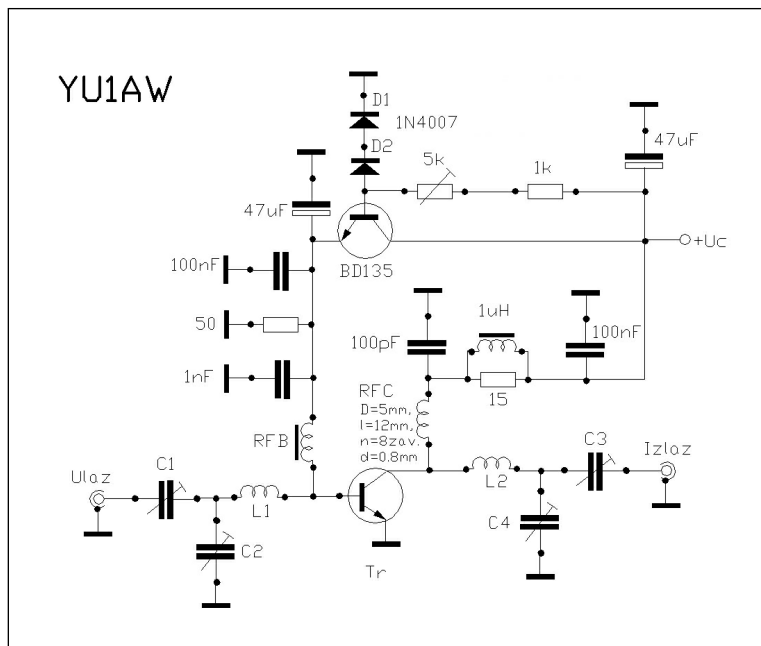


Fig 9.71: Circuit diagram of the one transistor power amplifier

Transistor	Freq MHz	C1 pF	C2 pF	C3 pF	C4 pF	L1					L2				
						Ind nH	D mm	L mm	n turns	d mm	Ind nH	D mm	L mm	n turns	d mm
BLW76	50	36	200	21	46	44	6.5	6	3	1.2	157	12	10	4	1.5
BLX15	50	56	363	20	37	23	6	4	2	1.2	195	15	12	4	1.5
MRF317	50	40	210	24	57	47	10	10	2.6	2	140	15	15	3.7	2
2xBLW84	145	12	33	5	3.3	15	6	3	1.5	1.2	113	10	10	4	1.5
2xMRF245	145	21	50	21	33	7.5	7	3	1	1.5	14	10	4	1	2
2xMRF247	145	21	50	21	33	7.5	7	3	1	1.5	14	10	4	1	2
MRF317	145	6.7	18	13	46	48	10	10	2.7	2	22	15	5	1	2
MRF247	145	18	110	18	80	8	7	3	1	1.5	13	10	4	1	2
MRF239	145	13	67	10	31	14	8	4	1.3	1.5	29	12	5	1.5	2

Table 9.3: Components required for various transistors

Transistor	Frequency MHz	V _C Volts	I _{CO} mA	I _{Cmax} A	Pdrive W	Pout W	Efficiency %
BLX15	50	50	50	6.5	15	150	65
BLW76	50	28	50	8	6	90	60
MRF317	50/145	28	10	6.5	12	100	60
MRF238	145	13.8	20	4	3	30	55
2xMRF245	145	13.5	2x30	2x18	35	200	50
2xBLW84	145	28	2x30	2x2	6	60	55

will, by monitoring the temperature of the tran-

Table 9.4: Transistor parameters

sistor, change the base bias to maintain the DC operating point stability.

All these conditions have to be fulfilled if an amplifier can be described as a high quality amplifier; they are only discussed briefly here. They are elaborated in literature [12] and [2].

One Transistor Power Amplifiers

The circuit of the one transistor power amplifier is shown in Fig 9.71. A PCB layout is shown in Fig 9.72 and the component layout in Fig 9.73 - these are both in Appendix B. The PCB may have to be varied slightly depending in the actual RF transistor chosen. Table 9.3 shows the component values for various transistors and Table 9.4 shows the expected performance. A circuit consisting of C1, C2 and L1 performs input matching. The Q of this circuit is about 10 - 15, which is the optimal value in this case. Base bias is fed via an RF choke, which is a VK 200 ferrite bead with six holes. The output circuit that matches the output of the transistor to 50 ohms is performed by L2, C4 and C3. All capacitors in the matching circuits are ARCO, ceramic-mica trimmers of corresponding capacity. The values of the matching circuit elements are shown in Table 9.3. The number printed on the trimmer gives the range of certain trimmers (Fig 9.74). Types of trimmers and the corresponding range of capacity are given in Table 9.5. The voltage values, currents, power and efficiency for each separate type of transistor are illustrated in Table 9.4. The collector supply is fed via an RF choke which instead of a ferrite is open wound. The diameter of this choke is 5mm, it is 12mm long, and has 8 turns of 0.8mm wire. After the choke, a fairly

small value capacitor decouples VHF frequencies to the ground. The 100nF capacitor decouples low frequencies via a 15Ω resistor to ground. The choke in parallel with the resistor is a VK 200 ferrite or a similar one that can carry the high currents that flow through the transistor.

A BD135 transistor connected as an emitter follower with extremely low output resistance carries out base bias, this ensures a stable operating point. Thermal monitoring and compensation of the operation point is done by diode D1, positioned so that it has physical and thermal contact with the transistor (not with the heatsink!). This ensures stable operation over a large temperature range. The other diode, D2, monitors the ambient temperature. Both diodes are of the 1N4007 type or similar. Transistor BD135 is mounted on the same heatsink as the RF transistor. Adjusting the operating point is performed by 5kΩ trimmer potentiometers. Apart from the electrolytic (or tantalum) capacitors all the others are disc ceramic or of some similar VHF quality.

Type of ARCO trimmer	Range
404	7pF - 60pF
423	7pF - 156pF
426	37pF - 250pF
462	5pF - 80pF

Table 9.5: Types of ARC trimmer

Two Transistor Amplifiers

As can be seen in the circuit diagram shown in Fig 9.75 there are two identical amplifiers connected by wideband baluns, they operate in a symmetrical anti-phase or push-pull connection. A PCB layout is shown in Fig 9.76 and the component layout in Fig 9.77 - both are in Appendix B. The PCB may have to be varied slightly depending in the actual RF transistor chosen. Table 9.3 shows the component values for various transistors and Table 9.4 shows the expected performance. Circuits for supplying the base and collector of the transistors are the same as the one transistor amplifiers. Even the input and output matching circuits are similar. The only novelties are two pieces of 50Ω coaxial cable 95mm long that transform the asymmetric input and output of the amplifier onto the symmetric connection of two



Fig 9.74: An ARCO trimmer

transistors. Teflon cable several millimetres thick, RG-142, type or similar should be used. The coaxial cable can be wound in a coil, or even better on some small ferrite toroid or onto a larger bead. Ferrite with two small holes can be used for this purpose. Using two such ferrites, two windings could be wound. If you do not have these, you could thread two small ferrite beads with one hole each onto the cable so that there is one bead on each end, but it is even better if you thread several beads and place them along the cable length. This will improve the symmetry of the transformer, especially at low frequencies, which can be significant for a stable amplifier.

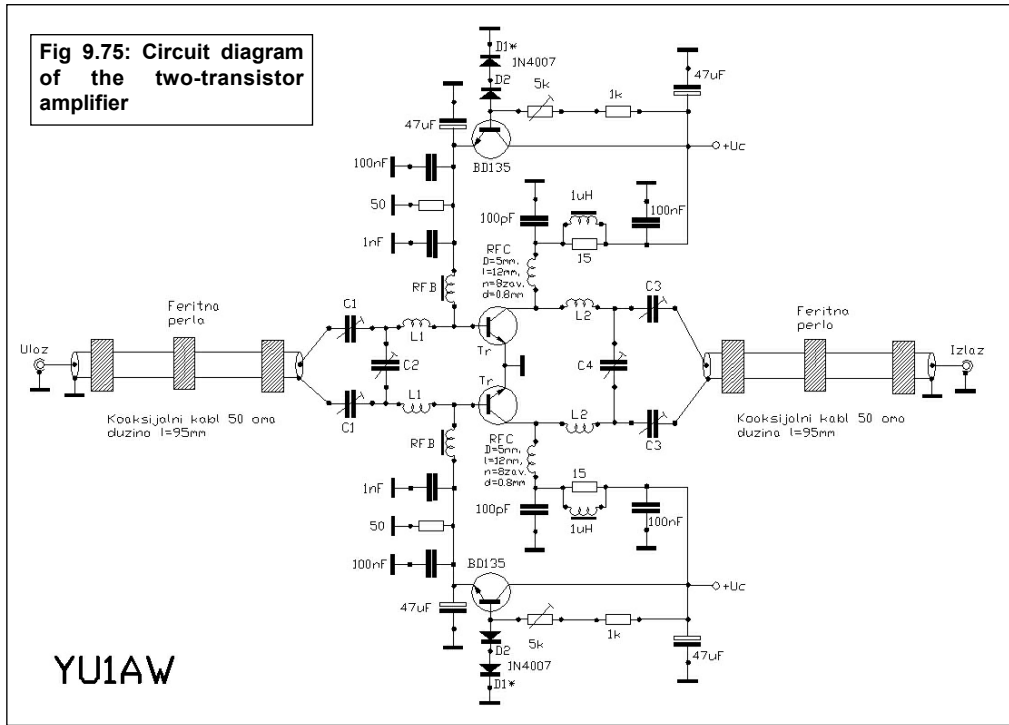
A self-resonant balun has been found to give an improvement over the wideband balun made from 95mm of coax cable. A 144MHz resonant balun can be made from 620mm of 5mm diameter PTFE coax cable, RG142 or similar. Close wind into 6 turns with 30mm inner diameter and about 30mm long with no space between turns. For a 50MHz balun, the coax cable length is 1770mm, coil inner diameter is 50mm, and coil length is around 55mm and 11 turns.

It is very important that the entire amplifier has to be totally symmetrical with regards to the mechanical layout of components and electrical parameters (values of elements, currents and voltages, etc), so that more power, amplification, efficiency and suppression of even harmonics can be achieved. While adjusting, it is very important to maintain the same capacitances of C1 and C3 in both amplifiers.

Mechanical Construction

The whole amplifier has to be built on a relatively small piece of a single layer FR4 printed circuit board (Fig 9.76 in Appendix B). Emitter leads are soldered as short as possible onto the ground of the board. The transistor should be mounted onto a large heatsink using thermal paste as illustrated in Fig 9.78. The surface of the board should be small to enable the fins of the cooler to be as close to the transistor as possible. To provide better cooling it is important to make sure that the hole drilled for mounting the transistor onto the heatsink is big enough to allow a screw to pass through it. The transistors should be mounted with their entire surface on to the heatsink. The surface of the heatsink where the transistors are mounted has to be smooth. Input and output connectors can be fixed onto the board but also onto the heatsink or eventually the box and connected with coaxial cable to the board with its grounds soldered on both ends. The coils, L1 and L2, in the matching circuits should be mounted so that their axes are at a right angle, to decrease interactive coupling.

Once everything is connected, as illustrated in Fig 9.77, check once again to ensure that there are no mistakes and short circuits to the ground, and adjust the potentiometers for maximum resistance. Connect the supply to one transistor and adjust the collector current to the value from Table 9.4. The same proce-



sure should be carried out with the second transistor. Even more important than the exact value of quiescent current is that they are identical in both transistors! Then connect both transistors to the supply and connect a 50Ω dummy load to the output via a wattmeter or SWR meter. Supply minimal excitation and by measuring the output power alternately adjust trimmers until maximum output power is achieved. Repeat the adjustment several times, gradually increasing the excitation power. Finally, with full excitation, which should not exceed the permitted output power, adjust all trimmers to the highest output power. At the same time the transistor's current should be measured so as not to exceed the maximum allowed value. If input trimmers, C1 or C2, need to be at maximum or minimum capacity during adjusting, it is necessary to change the length of the cable between the exciter and the amplifier itself. The optimal length of the cable should be determined experimentally to obtain adjustment with approximately the same values of C1 and C2. This experimentally determined cable should always be used when operating the amplifier. A change of exciter could occasionally require a new length of cable to be determined. In

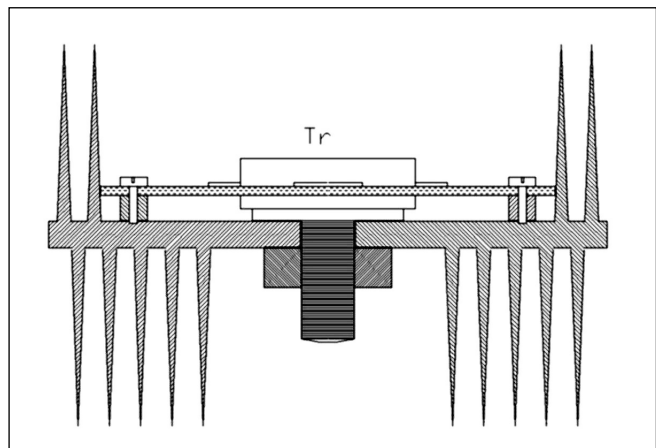


Fig 9.78: Mounting the power transistor on a heatsink for the transistor power amplifiers

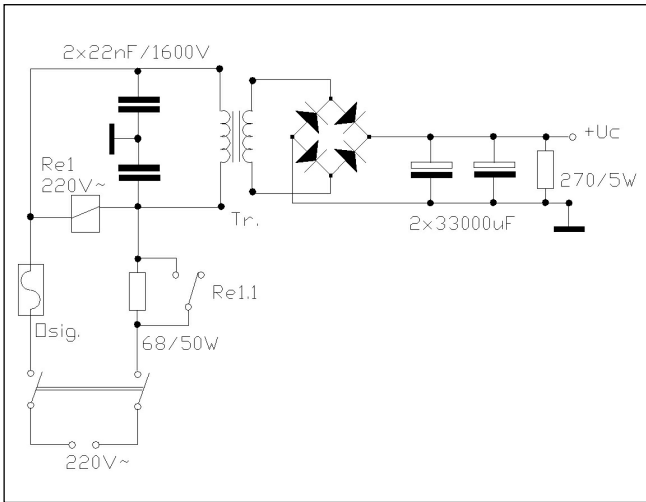


Fig 9.79: Circuit diagram of a power supply for the transistor power amplifiers

push-pull amplifiers it is extremely important to perform adjustment so that the corresponding trimmers on each transistor are adjusted simultaneously to ensure that they have approximately the same capacity during adjustment. By maintaining symmetry during adjustments, extremely dangerous situations are avoided which can cause the amplifier to self-oscillate.

For transistors with a power supply higher than 20V, a non-stabilized supply can be used, but it is constructed so that it has very good voltage regulation as illustrated in Fig 9.79. It should be well specified with good quality electrolytic capacitors. The transformer should be slightly over specified. To avoid blown fuses caused by the charging current of electrolytic capacitors, it is necessary to build in a delayed switching device. This is performed simply by a 220V relay connected as shown in the circuit. At the moment when power is switched on, the transformer is connected to a power supply via a resistor that limits high charging currents. When the capacitors are charged and transients in the transformer settle down, the current through the resistor decreases, the voltage on the primary increases and the relay that bridges the resistor with its contacts is switched on. It is also possible to use relays that switch on via some electronic timer after a couple of dozen seconds. Although this appears to be a more elegant solution, it is a far worse solution, for two reasons: first, in the case of a very short interruption of supply voltage the timer has not been reset, voltage is switched on without delay; and second: in the case of a fault that causes high current consumption, when the relay would not be switched on, the entire system would protect itself, whereas the timer would switch on the relay and subsequently provide full power.

Transistors with 12-18V collector voltage must be supplied with stabilized voltage. In this case it is vital to build in a thyristor device for over voltage protection. This device must connect the voltage via a thyristor to the ground and thereby induce a forced blown fuse if the voltage exceeds some preset value; this

prevents the RF transistor blowing up. Any type of relay protection is not recommended in such cases as it simply cannot react quickly enough and by the time the relay is switched, the transistor is already blown up!

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 9.80, 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.

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 loaded 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 9.81, should have the following calculated values for optimum operation:

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 9.81.

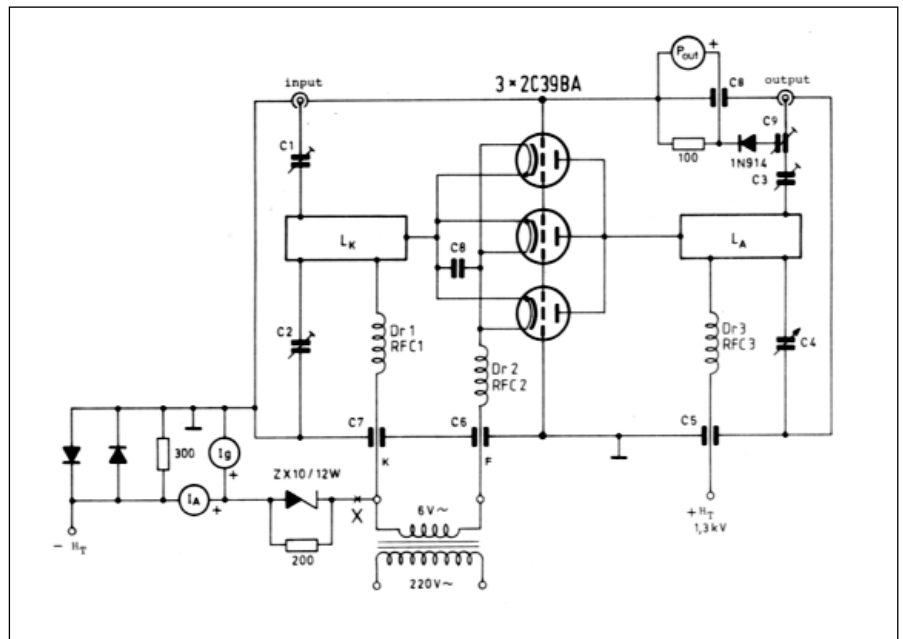


Fig 9.80: Circuit diagram of the 300W 12dB gain, 2C39A power amplifier for the 70cm band

The construction is very simple as can be seen from Figs 9.82 to 9.84. 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 9.82, 9.83 and 9.84. The sides are soldered together.

Tuning capacitors C_2 and C_4 are made from 0.5mm thick brass plate (Fig 9.81) 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 C_4 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 9.82) 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 (C_8) 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 9.80. The coupling (C_9) to the detector is adjusted by varying the 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.

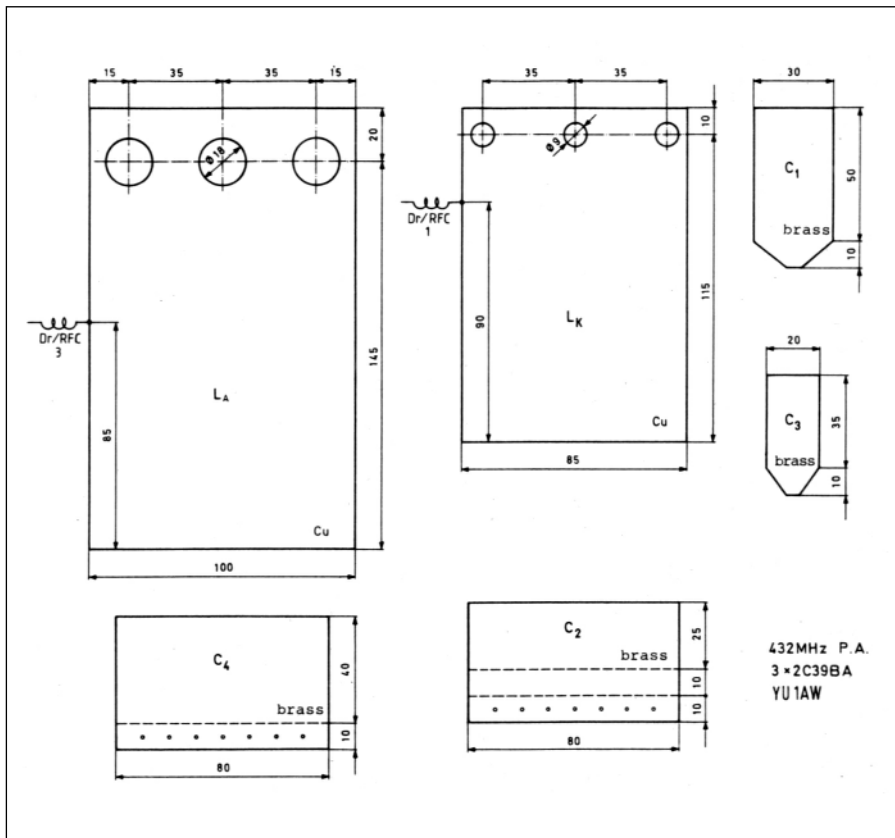


Fig 9.81: Dimensions of the housing parts for the 2C39A power amplifier for the 70cm band

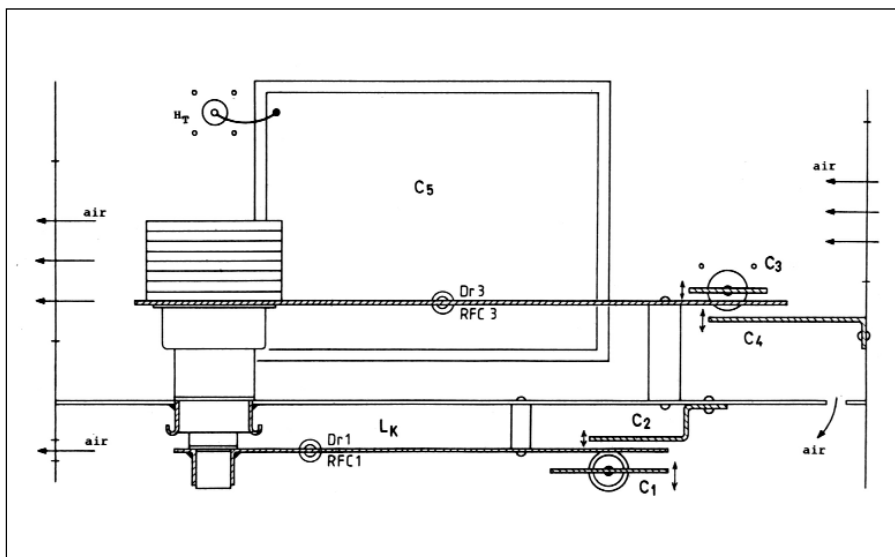


Fig 9.82: Side view of the 2C39A power amplifier for the 70cm band

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.

RECEIVERS

With the large number of commercial receivers available for the VHF and UHF bands, not many amateurs build their own. This design by Andy Talbot, G4JNT, from *RadCom* shows a new type of receiver [13].

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 9.85. In the RF path two MMICs, a MAR-6 and a MAR-3, amplify the RF; there is a two stage bandpass 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 [14]. 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 spuri 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 the microwave bands, and allow for odd LO frequencies. The multiplier output level is +10dBm 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 approach-

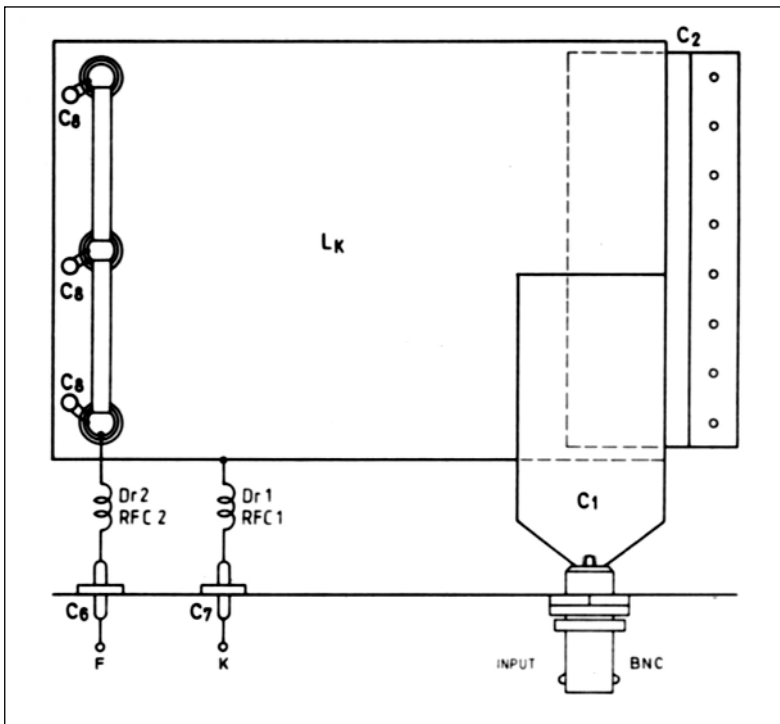


Fig 9.83: Bottom view of the 2C39A amplifier for the 70cm band

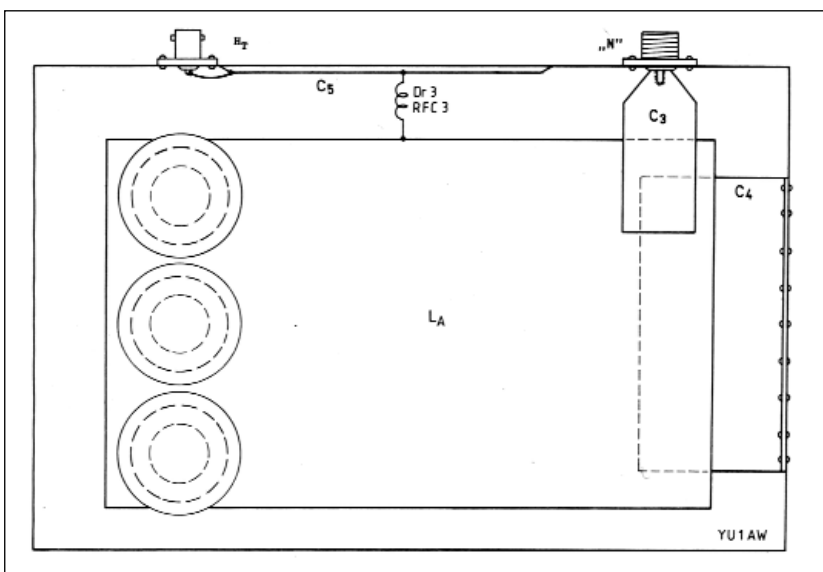


Fig 9.84: View of the anode enclosure of the 2C39A power amplifier for the 70cm band

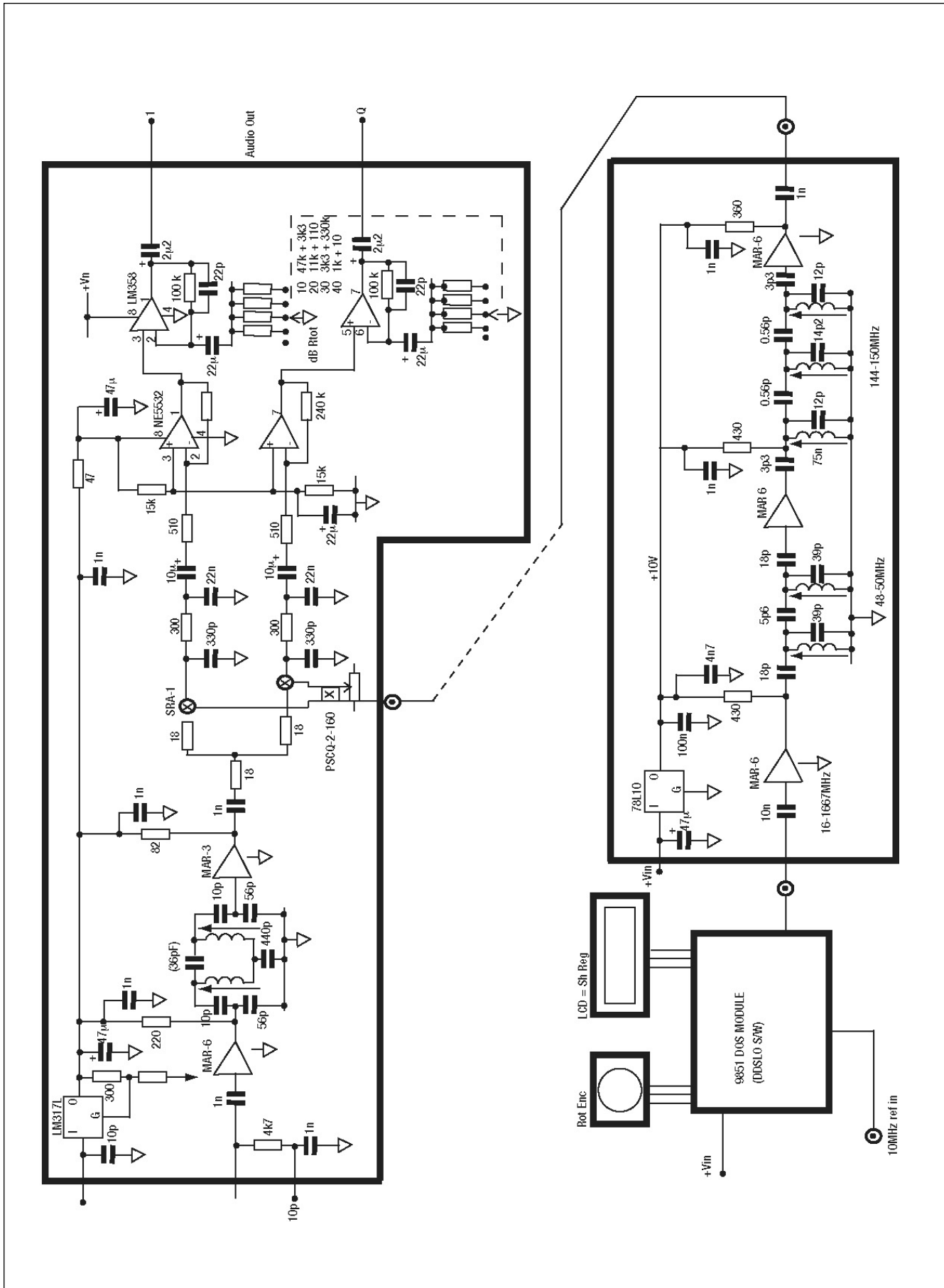


Fig 9.85: Circuit diagram of the 144MHz direct conversion receiver

ing 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 800Ω 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 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), an 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 tinfoil 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 50Ω microstrip line into a double-sided PCB. To make a 50Ω 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 50Ω 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 50Ω 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 50Ω 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 [15] 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 a look at the Software-Defined Radio software [16] from I2PHD, for another solution

The DDS module as described in [14] 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 [17].

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!

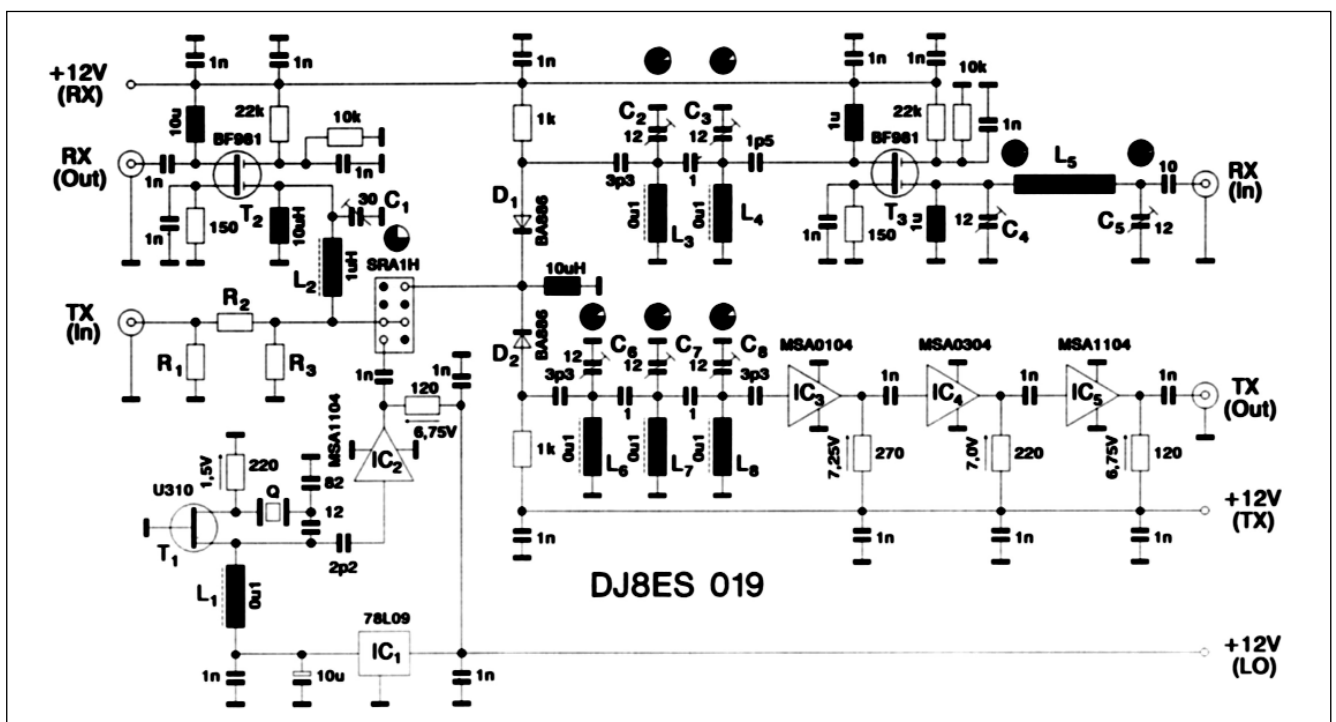


Fig 9.86: Circuit diagram of the 2m transverter

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.

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 and 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 [18].

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 [19] and [20]. The following design is the resulting 28/144MHz transverter [21]. 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.

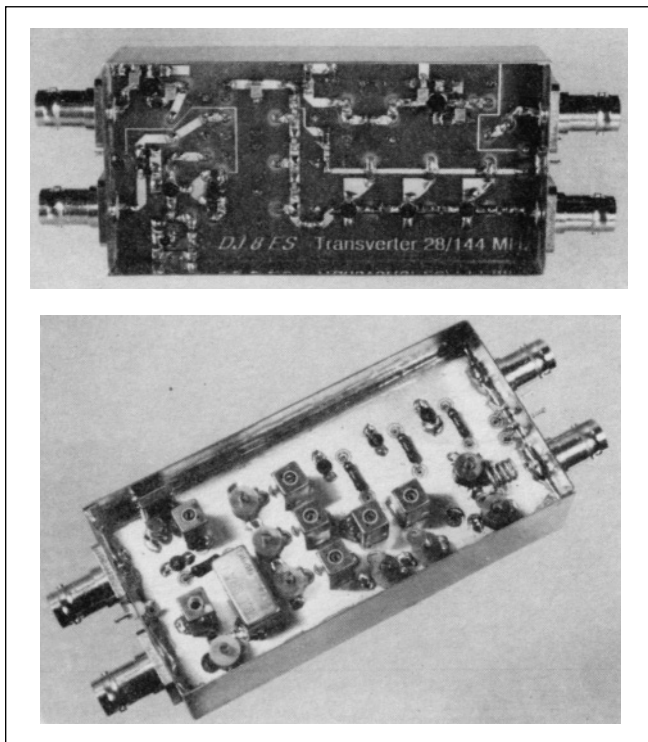


Fig 9.87: The completed 2m transverter

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 9.86 shows the circuit diagram of the 28/144MHz transverter. The local oscillator uses a tried and tested U310 crystal oscillator at 116MHz.

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 BF981 low-noise tran-

IC1	TA78L09F SMD voltage regulator
IC2, IC5	MSA1104 (Avantek)
IC3	MSA0104 (Avantek)
IC4	MSA0304 (Avantek)
T1	U310 (Siliconix)
T2, T3	BF981 (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.)
<i>All other components in SMD format</i>	
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.

sistor to give the required intermediate frequency amplification.

The 2m received signal is fed to the gate of the BF981 RF amplifier through a pi filter from the 50Ω 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 tinfoil housing of 55.5mm x 111mm x 30mm, Fig 9.87 show pictures of the completed transverter. The board 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 9.88 and 9.89 (see 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.

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 9.90 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 9.91. The current consumption of the module is directly proportional to this ratio.

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 9.92 shows what a typical module looks like from the inside. The 50Ω input and output matching circuits are clearly visible.

A low-pass filter (Fig 9.93) on the output provides the harmonic suppression required. Amazing suppression is obtained

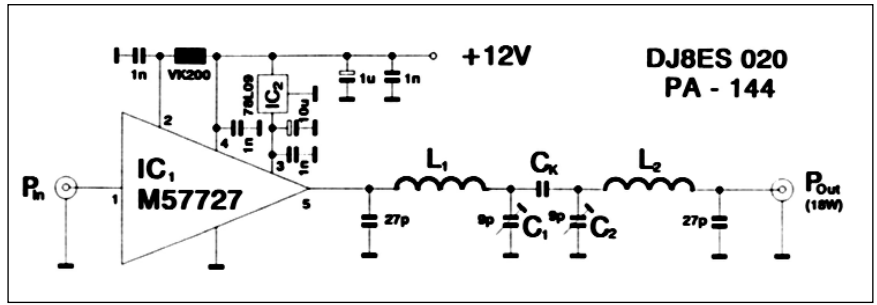


Fig 9.90: Circuit of the 2m power amplifier to be used with the 2m transverter

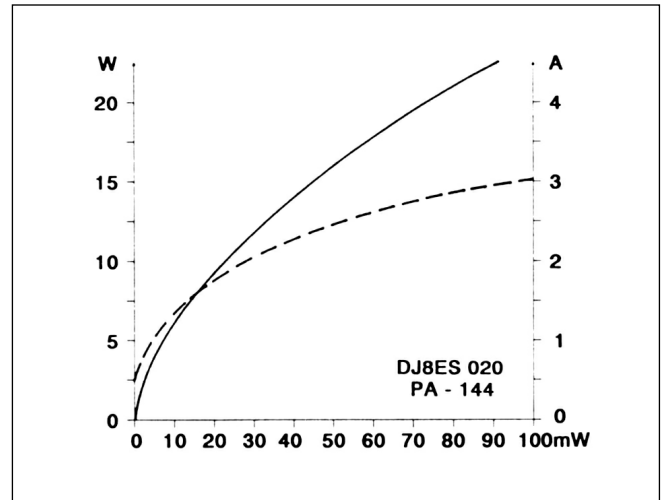


Fig 9.91: Power transfer characteristics of the Mitsubishi M57727 hybrid module

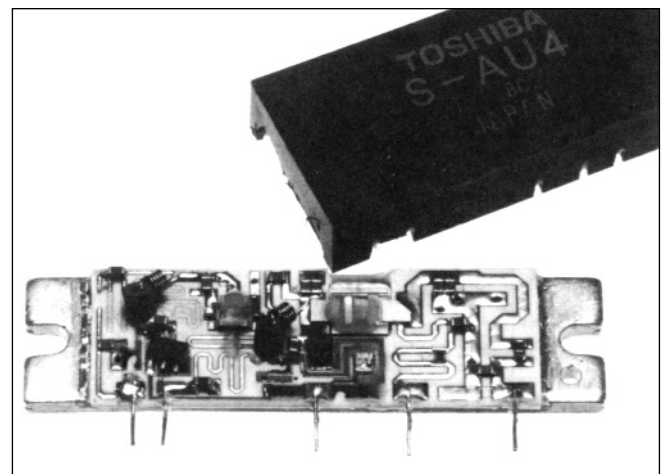


Fig 9.92: An internal view of a hybrid module. This is the Toshiba S-AU4 which is a 70cm amplifier

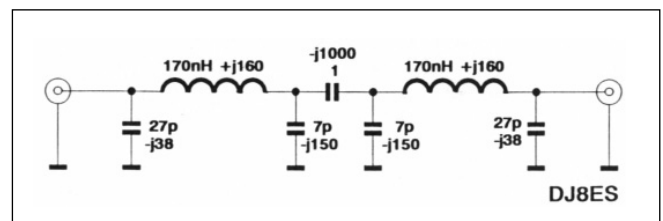
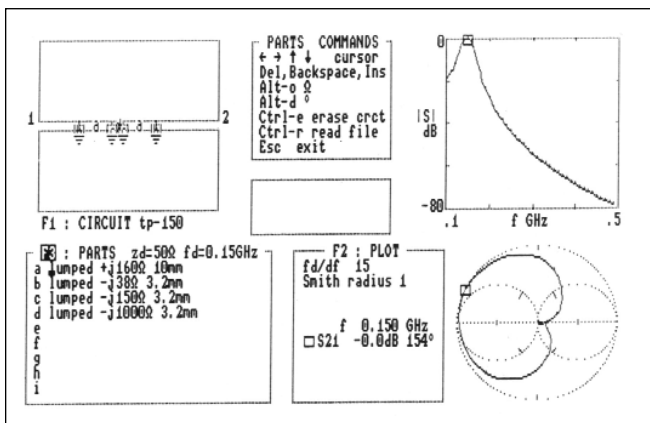


Fig 9.93: Circuit of the low pass filter used in the 2m power amplifier



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
<i>All other components in SMD format:</i>	
	1 x 1µF/20V tantalum
	1 x 10µF/20V tantalum
	2 x 27pF, ATC chip
	3 x 1nF, ceramic capacitor

Fig 9.94: PUFF CAD software output used to design the 2m low pass filter

Table 9.8: Parts list for 2m hybrid amplifier.

using only two pi filters wired together. Fig 9.94 shows the output of the PUFF CAD design software used to design the filter.

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 used here, due to the relatively high power level. A thin copper plate is soldered between the two pi filters for screening, see Fig 9.95 (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.

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 9.95 (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.

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.

The two coils (L1, L2) and the coupling capacitor, C_K, are hand made. The coils are 8.5 turns of silvered copper wire with a

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).

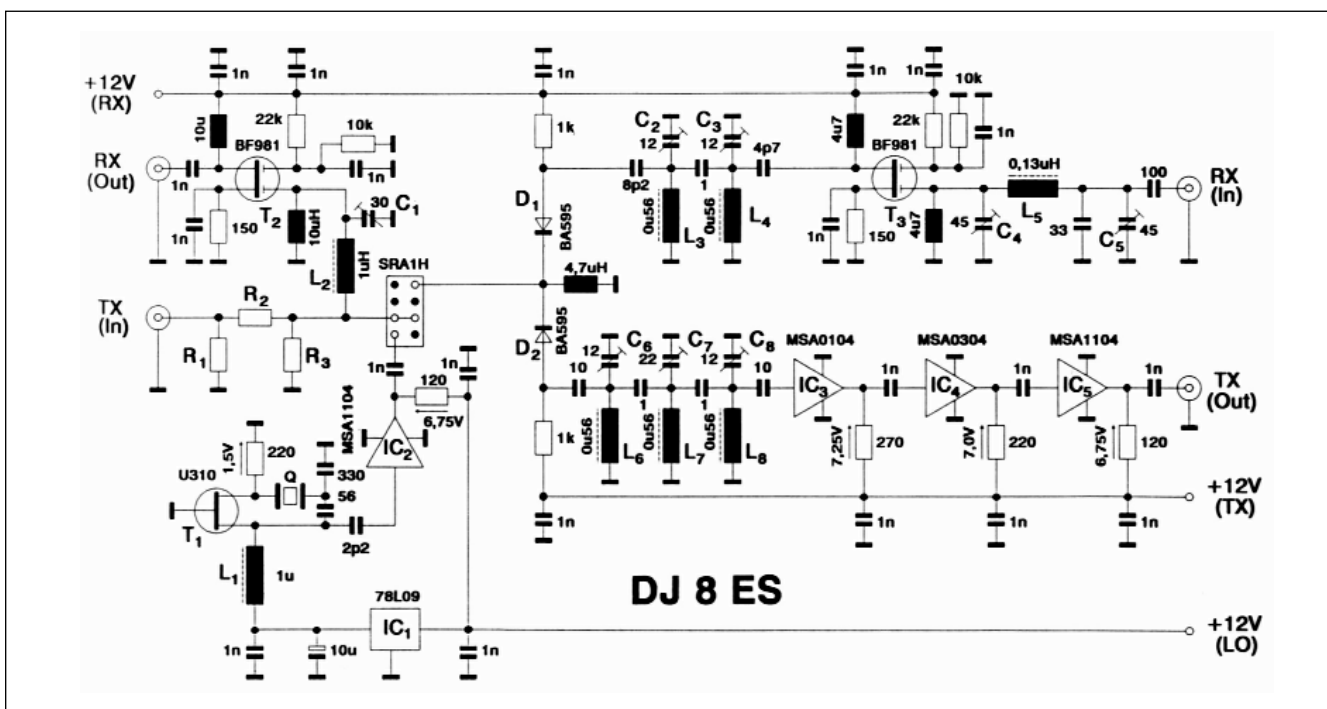


Fig 9.96: Circuit diagram of the 6m transverter

IC1	TA78L diameter 9F voltage regulator (SMD)	2 x 120Ω / 0.5W Carbon film
IC2, IC5	MSA1104 (Avantek)	1 x 220Ω / 0.5W Carbon film
IC3	MSA0104 (Avantek)	1 x 270Ω / 0.5 W Carbon layer
IC4	MSA0304 (Avantek)	4 x BNC flanged connector (UG-290 A/U)
T1	U310 (Siliconix)	3 x Teflon bushing
T2, T3	BF981 (Siemens)	1 x Tinplate housing 55.5 x 111 x 30mm
D1, D2	BA595 PIN diode (SMD)	9 x 1.5mm dia. Copper rivets
L1, L2	BV5048 Neosid coil, 1 μH, yellow/grey	<i>All other components are SMD format:</i>
L3, L4	BV5036 Neosid coil, 0.58μH, orange/blue	Ceramic capacitors
L5	BV5063 Neosid coil, 0.58μH, blue-orange	3 x 1pF
L6, L7, L8	BV5063 Neosid coil, L8 0.58μH, orange/blue	1 x 2.2pF
C1	30pF foil trimmer (red) 7.5 mm grid (Valvo)	1 x 4.7pF
C2, C3	12pF foil trimmer (yellow) 7.5 mm grid (Valvo)	1 x 8.2pF
C4, C5	45pF foil trimmer (violet) 7.5 mm grid (Valvo)	2 x 10nF
C6, C8	12pF foil trimmer (yellow) 7.5 mm grid (Valvo)	1x 33pF
C7	22pF foil trimmer (green) 7.5 mm grid (Valvo)	1 x 56pF
Q	22MHz crystal, HC18U or HC25U	1 x 330pF
1 x SRA1H ring mixer		17 x 1nF
		Resistors etc
		1 x 150Ω
		2 x 220Ω
		2 x 1kΩ
		2 x 10kΩ
		2 x 22kΩ
		1x 10μF / 20VTantalum
		3x Choke, 4.7μH
		2x Choke, 10μH

Table 9.9: Parts list for 6m transverter

6m Transverter

A transverter for the 6m band can be produced based on the 28/144MHz transverter described above [22]. All that is required is modification of the oscillator and the filter.

Fig 9.96 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 9.97 (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 9.98 and Fig 9.99 (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 9.100. 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.

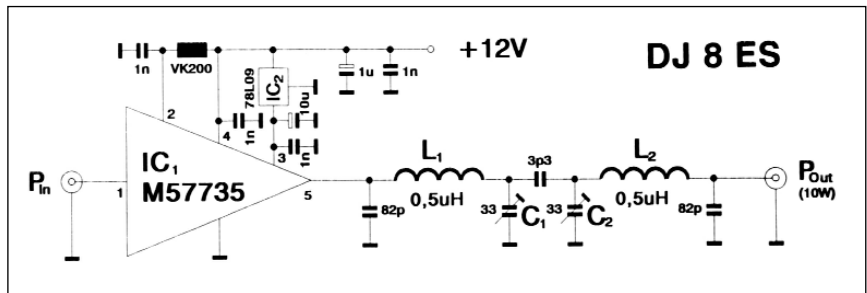


Fig 9.100: Circuit of the 6m power amplifier to be used with the 6m converter

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 [23] is similar to the 28/144MHz transverter described above. It uses two boards; the oscillator and the transverter Fig 9.101 shows a pic-

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

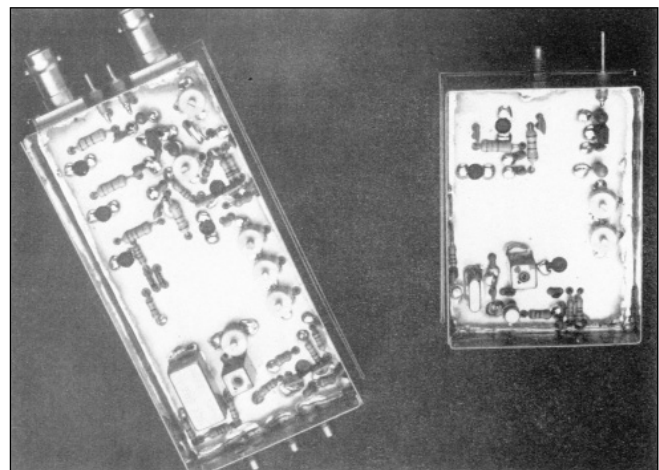


Fig 9.101: The completed local oscillator and transverter units of the 70cm transverter

Fig 9.102: Circuit of the local oscillator used in the 70cm transverter

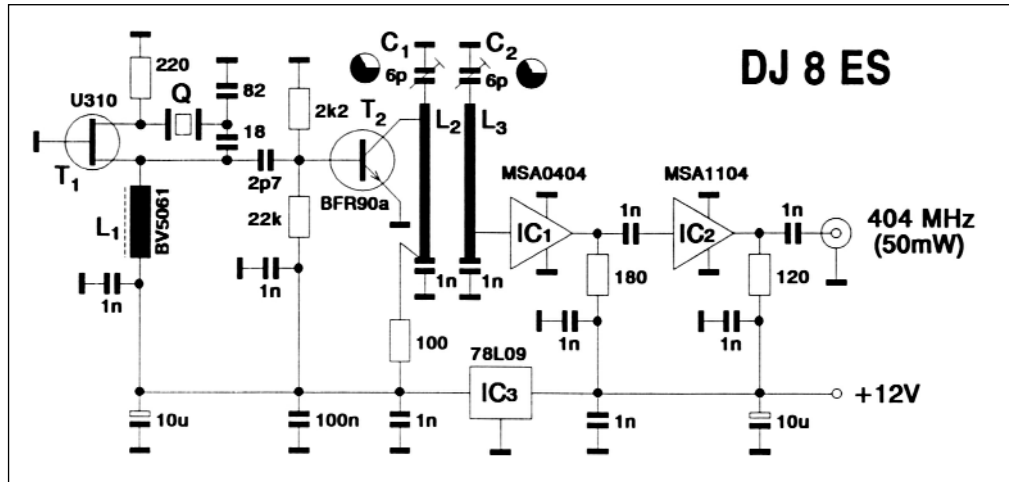
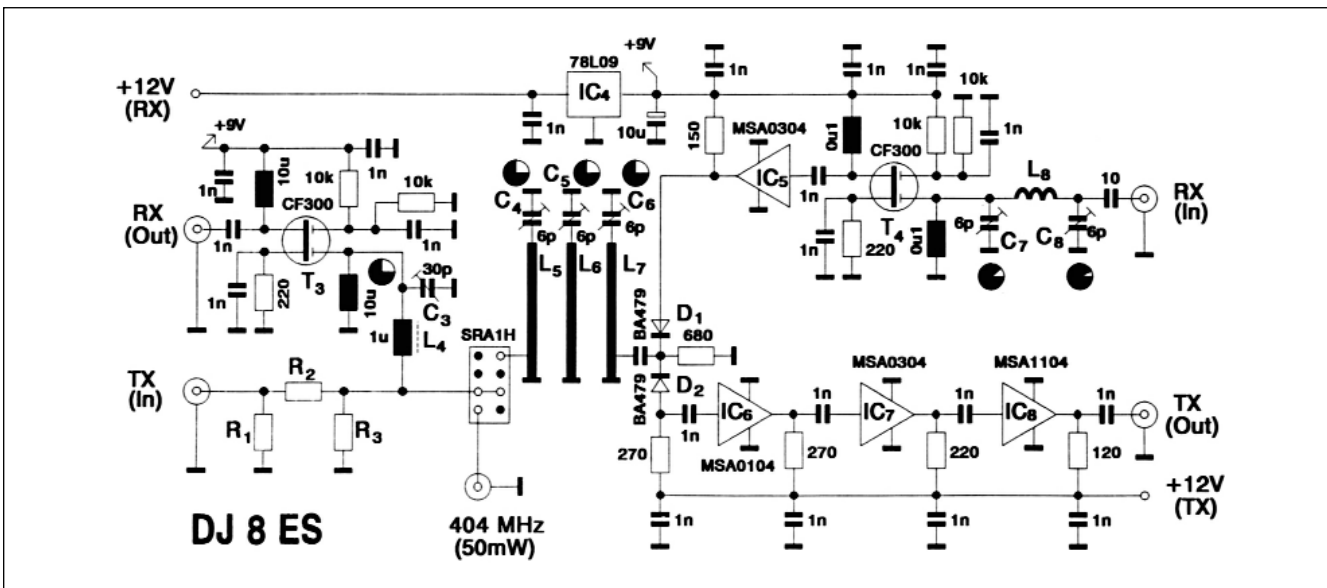


Fig 9.103: Circuit diagram of the 70cm transverter



ture 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 9.102 shows the circuit diagram 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 9.103 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 9.10 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 50Ω 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 CF300 (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.

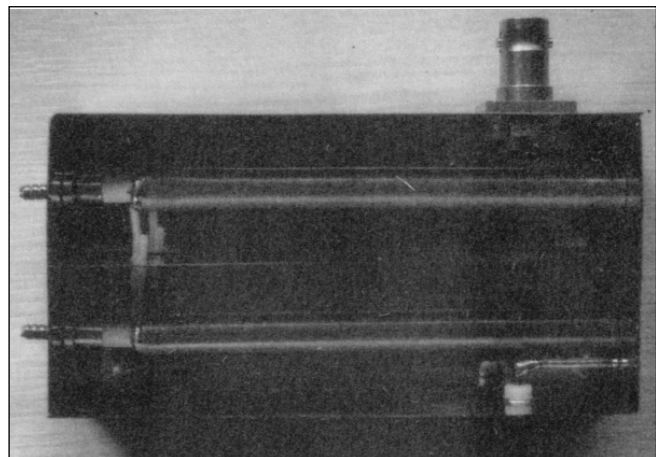


Fig 9.104: 70cm bandpass filter that can be used with the 70cm transverter

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 9.104 shows a possible two-pole bandpass filter. It can be assembled as an air core construction using a standard tinfoil 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 9.105 and 9.106, and for the transverter in Figs 9.107 and 9.108 (all four of these are in Appendix B). The PCB are mounted in standard tinfoil 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 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)

IC1	MSA0404 (Avantek)
IC2	MSA1104 (Avantek)
IC3	78L09 voltage regulator
T1	U310 (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	Tinfoil housing: 55.5mm x 74mm x 30mm
2 x	1nF trapezoid capacitor
2 x	10µF 20V tantalum capacitor
<i>Ceramic Capacitors (2.5mm grid)</i>	
1 x 2.7pF	<i>Resistors (1/8W, 10mm)</i>
1 x 18nF	1 x 100
1 x 82pF	1 x 220
6 x 1nF	1 x 2.2k
1 x 100nF	1 x 22k
<i>SMD Capacitor (model 1206 or 0805)</i>	
2 x 1nF	

Table 9.11: Parts list for 70cm transverter local oscillator

IC4	78L09 voltage regulator
IC6	MSA0104 (Avantek)
IC5, IC7	MSA0304 (Avantek)
IC8	MSA1104 (Avantek)
T3, T4	CF300 (Telefunken)
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
1 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	Tinfoil housing 55.5 x 111 x 30mm
4 x	1nF trapezoid capacitor
2 x	0.1 µH choke, 10mm grid, axial
2 x	10µH choke, 10mm grid, axial
1 x	10µF 20V tantalum
<i>Resistors (1/8W, 10mm)</i>	
2 x 220Ω	<i>Ceramic Capacitors (2.5mm grid)</i>
1 x 270Ω	1 x 10pF
1 x 680Ω	12 x 1nF
4 x 10kΩ	SMD Capacitor (model 1206 or 0805)
	6 x 1n

Table 9.12: Parts list for 70cm transverter

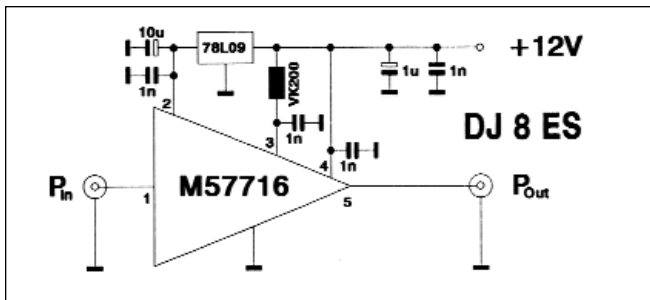


Fig 9.109: Circuit diagram of the 70cm power amplifier to be used with the 70cm transverter

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 factor 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 9.109 shows the circuit for such a module using a Mitsubishi M57716. A 2C39 valve PA can be fully driven using this 10W output.

4m Transverter

This transverter was designed as a club project for the Andover Radio Amateur Club [18]. 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 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 9.110 and 9.111.

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 9.112 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.

These doubly balanced mixers require about +7dbm of local oscillator injection and 0dbm (1mW) of Drive (IF). 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

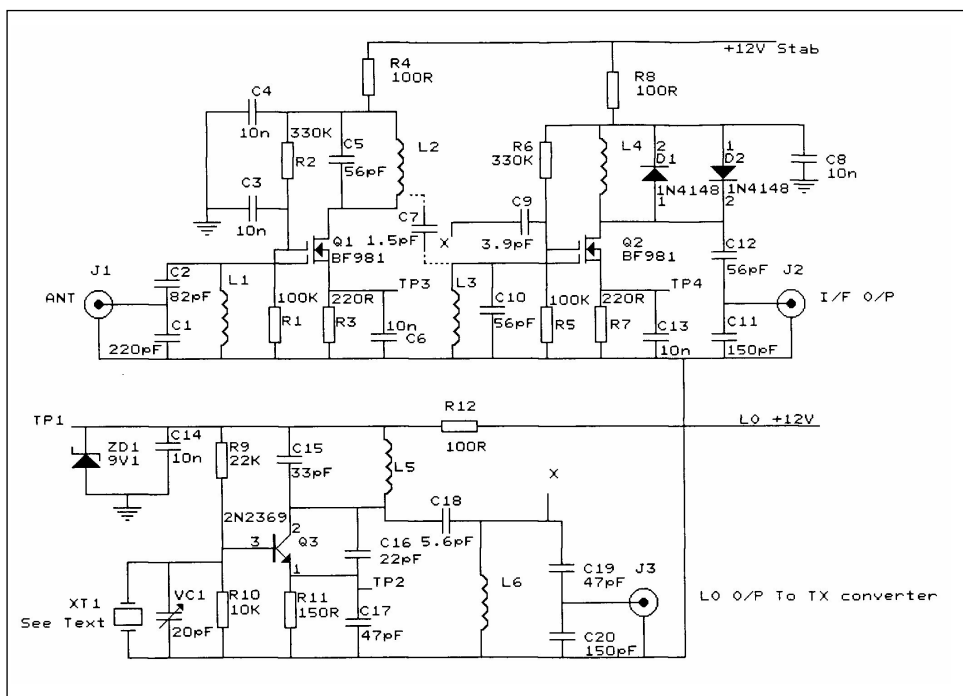


Fig 9.110: Receive converter circuit diagram for the 4m transverter

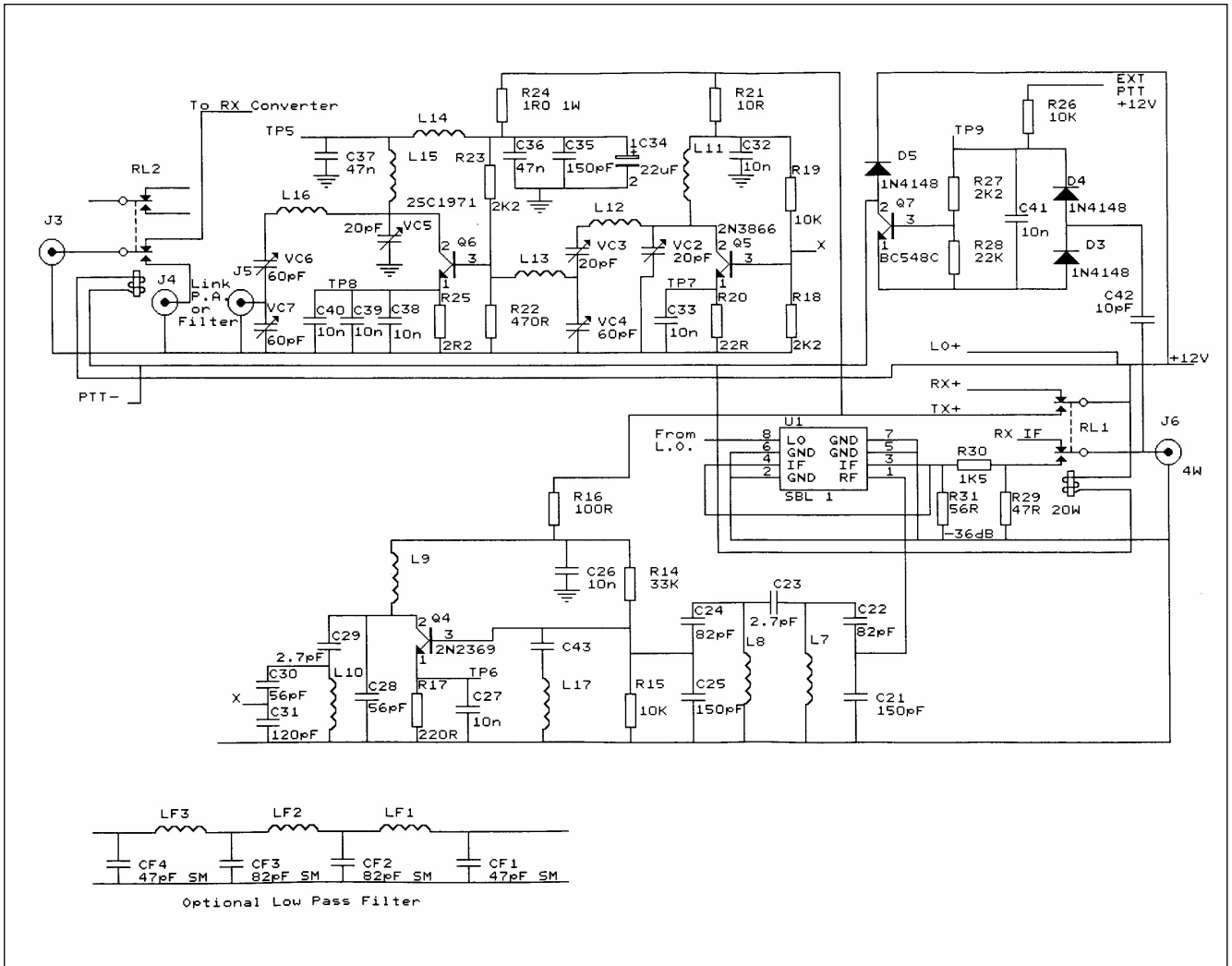


Fig 9.111: Transmitter converter circuit diagram for the 4m transverter

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 volt-

age 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 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

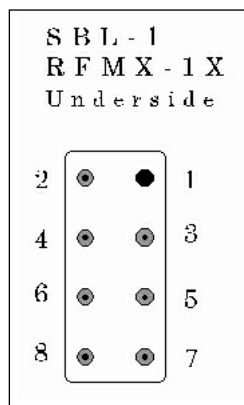


Fig 9.112: Pin connections for the mixer U1 used in the 4m transverter

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

Table 9.14: Parts list for 4m transverter

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.

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.

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 9.113 and 9.114 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. It is helpful to 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. 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 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 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.

Next, 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 20k Ω /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 20k Ω /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 50 μ A 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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Choosing Commercial Equipment for the VHF/UHF Bands

When choosing equipment, everyone will naturally have a budget in mind. But within this budget, it's often totally useless in asking "what's the best transceiver to buy?". This is because each person will have his or her own particular interests and operating restrictions, or indeed lack of them. For those just starting out on the hobby, the best advice would be to go for a general coverage multi-mode receiver initially, possibly a second-hand model so that its resale value, for when eventually 'trading up', isn't compromised too much. Once licensed, or on the verge of getting one, a transceiver would be a better bet, but in this case it is useful first to decide where your interests primarily lie and then your operating possibilities.

If you're very restricted in terms of the size and height of antennas you can erect at home, yet want to work long distances, then aim to get a transceiver that can also be used in portable and mobile situations, such as from a local elevated site using either a mobile antenna or a temporary portable mast-mounted antenna. If however you're only interested in operating through your local 2m or 70cm repeater, a simple single-band or dual-band FM transceiver will suffice. But beware, once the 'bug' bites, you may quickly want to move on to bigger and better things.

One of the best improvements you can make to your station is to get a better antenna system; it's little use having a £5000 transceiver operating into a compact mobile whip mounted in the loft or on a window ledge when you've a garden or rooftop at your disposal. You'd usually be a lot better spending somewhat less on the transceiver and rather more on improving the antenna system, provided of course that you can physically get something more substantial erected without falling foul of planning laws or the neighbours!

Take some time to sit back and think what you're interested in and where and who you'd like to contact. If you enjoy a challenge, then either low-power 'QRP' operation, possibly with a home-built transceiver (but again with the very best antenna system you can manage) is one way to go. Another could be the challenge of gaining operating awards and competing against other amateurs around the world, in which case you'll need the utmost in transceiver performance and transmit power level, again not forgetting a good antenna system. If however you're keen to use the hobby more as a social outlet and technical discussion scenario, typically having 'raghews' with local amateurs and exchanging information, then an appropriate transceiver system that's easy to operate from a comfortable position, without scores of tiny knobs and buttons on the front panel, would be more appropriate.

If you'd like to be able to have a wide choice of operating bands, modes and facilities, there are a growing number of reasonably priced multimode transceivers offering HF, VHF and UHF all combined in a compact portable 'carry around' unit which can also be used at home and in the car. Once you get more experienced and you possibly find your interests lie in a more defined aspect of the hobby, a transceiver and antenna system dedicated to this can be obtained, with your earlier transceiver being used either as a 'trade-in' or as a secondary transceiver for other modes and bands.

Whichever transceiver you go for, bear in mind one final but very important point. Even if you're not considering integrating computer control into your station right now, do make sure that whichever transceiver you eventually choose for your base station or even portable / mobile station has this facility built in. There will eventually come a day when you'll want to at least try this, and there is a plethora of very powerful freeware and shareware programs available for download or on CD/DVD. These offer everything from simple transceiver frequency control right up to complete station management including automatic log keeping (taking the frequency, mode etc automatically from your transceiver), QSL card printing (again with time, date, callsign etc. entered for you), DX Cluster handling and automatic QSY to the frequency of that rare station on your 'wanted' list, automatic antenna rotor control to the identified station, satellite tracking and automatic Doppler shift correction, even down to automatic generation of award sheets for you after a contest.

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