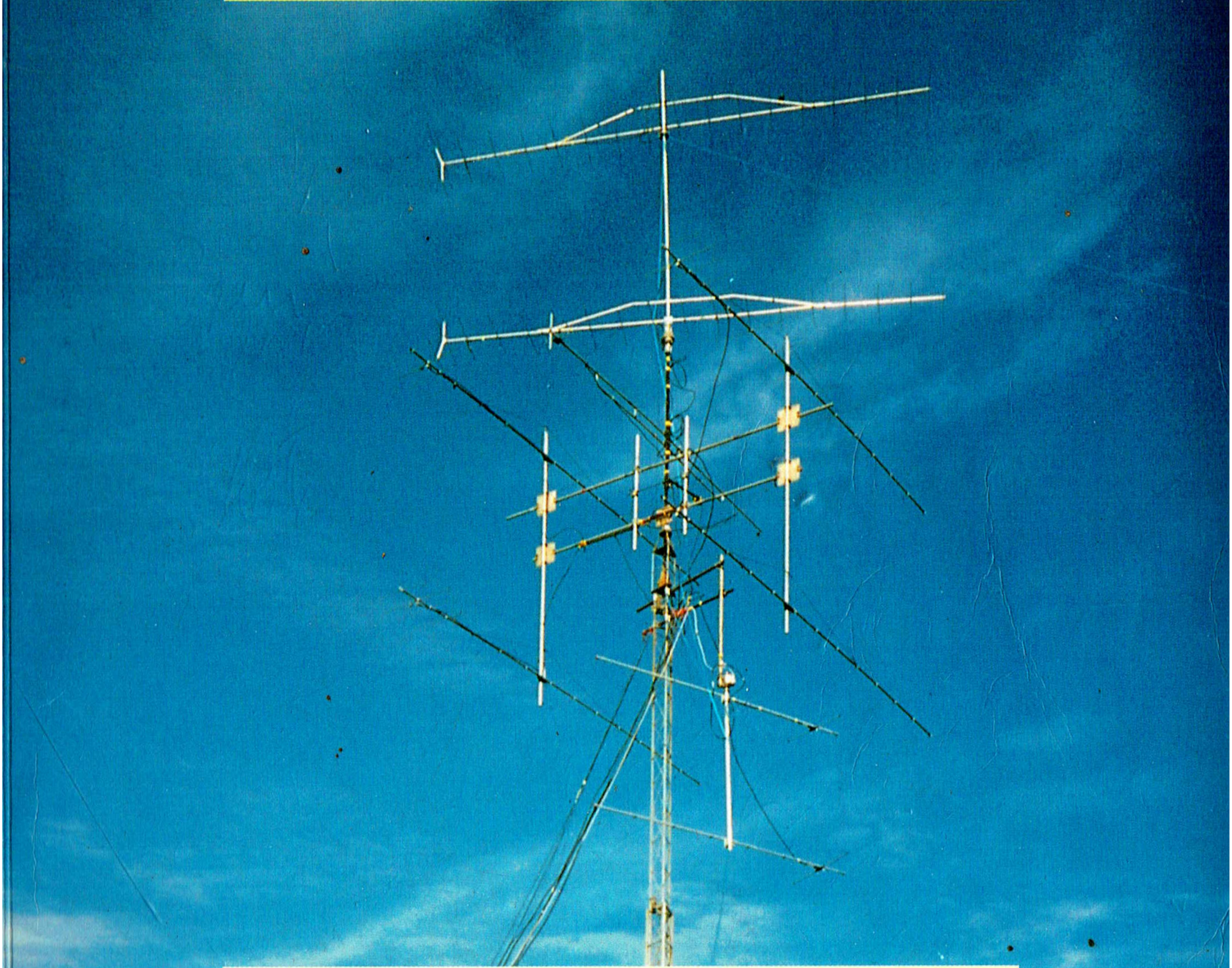
EDITED BY DICK BIDDULPH, G8DPS



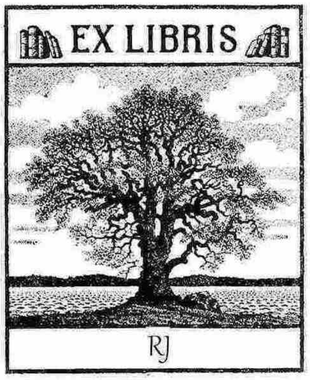
VHF/UHF HANDBOOK

RSGB

VHF/UHF Handbook

Editor: Dick Biddulph, G8DPS

Radio Society of Great Britain



Published by the Radio Society of Great Britain, Cranborne Road, Potters Bar, Herts EN6 3JE.

First published 1997

ISBN 1 872309 42 9

Publisher's note

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Cover design: Jennifer Crocker.

Illustrations: Derek Cole and Bob Ryan (Radio Society of Great Britain), Ray Eckersley, and Jean Faithfull.

Production and typography: Ray Eckersley, Seven Stars Publishing.

Printed in Great Britain by Bell & Bain Ltd, Glasgow.

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Acknowledgement is also made to the authors of articles published in *Radio Communication* and other magazines from which extracts have been made. In this connection, thanks are due to the publishers of *Ham Radio*, *Practical Wireless* and *VHF Communications* for permission to reproduce copyright material. Acknowledgement is also due to Jaybeam Engineering Ltd for permission to reproduce certain charts appearing in Chapter 5. Other acknowledgements are made in the text where appropriate.

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Preface

SINCE the last edition of the *VHF/UHF Manual* was published in 1983 there has been rapid progress, particularly in the hardware associated with VHF and UHF. For example, it is now possible to build a legal-limit linear amplifier using only solid-state devices. In reception, devices with very low noise levels have appeared and, furthermore, integrated circuits containing most of the 'works' of a receiver have also arrived on the scene.

There is one major departure in this book from its predecessor – no microwave chapter. This is because there is now the *Microwave Handbook* published by the RSGB. There is a little overlap as both books contain information on the 23cm (1.3GHz) band. Another area left out, apart from its operation, is mobile work since nearly everyone uses commercial equipment now.

The only disappointing trend since the *VHF/UHF Manual* was last published is the reduction in interest in home construction. One of the aims of this book is to promote that facet of amateur radio by including tried and tested circuits for receivers and transverters as well as for building blocks for receivers, transceivers and transmitters. It is the editor's belief that any amateur ought to be able, and should be encouraged, to build something, even a complete receiver, transmitter or transceiver. These should be at least as good as the commercial articles but may not be as neat or as small. By doing this, he or she can incorporate only the 'bells and whistles' that are needed.

Dick Biddulph, G8DPS

1 Introduction to VHF/UHF

WHAT IS VHF/UHF?

The electromagnetic spectrum cannot be infinite due to limitations imposed by the finite age of our universe and by there only being a finite amount of energy in it. Nevertheless, the range of frequencies at which something, somewhere, is happening is immense. The (relatively) much smaller range of frequencies used for deliberate communication is still too big to contemplate all at once.

VHF (very high frequencies) is the name given to the 30 to 300MHz part of the electromagnetic spectrum. *UHF* (ultra high frequencies) is the name given to the 300 to 3000MHz part.

The whole of the 'radio' part of the electromagnetic spectrum is divided into a series of named ranges. The boundaries between ranges are at 3, 30, 300, 3000MHz (or not quite exactly at 100, 10, 1, 0.1 metres wavelength). Fig 1.1 shows the broad picture of the electromagnetic spectrum. Fig 1.2 focuses on the part of the spectrum used for radio communications.

The boundaries were just chosen arbitrarily at nice, round, numbers. Nothing abrupt happens at them. It is well known that there are differences in the propagation characteristics of radio signals of different frequencies, but these are caused by the environment – our atmosphere and terrain – rather than being direct consequences of the frequency of the signal. As a result of changing weather and upper-atmospheric conditions, the frequencies where changes in propagation effects occur are not fixed. They can move by at least a factor of 10, and the changes may be abrupt or gradual. Without any fixed natural landmarks, any fixed divisions have to be arbitrary. Any attempt to fix boundaries based on the different technologies we use would fare no better, as they too would shift as new technologies evolve.

30 to 3000MHz is still a very large chunk of spectrum, and while someone studying propagation may be interested in the whole of it, most amateur interest naturally focuses on the bands where our licences permit us to transmit. Table 1.1 shows the worldwide VHF and UHF amateur bands.

Table 1.1. VHF/UHF amateur bands in the IARU Regions

Band (MHz)	Region 1	Region 2	Region 3
50–54	UK 50–52MHz*	Yes	Yes
144–146	Yes	Yes	Yes*
146–148	No	Yes	Yes*
220–225	No	Yes*	No
430–440	Yes*	Yes*	Yes*
902–928	No	Yes*	No
1240–1300	Yes*	Yes*	Yes*
2300–2450	Yes*	Yes*	Yes*

*Shared band

THE BORDER WITH MICROWAVES

Although the formal definition of UHF extends up to 3000MHz, traditionally some of the higher-frequency part of the range has been considered to be 'microwaves' and treated separately, with its own technologies, literature and specialists. The word 'microwave' does not feature on the HF, VHF, UHF, SHF, EHF etc scale. It's really an older name from an earlier nomenclature of the spectrum and overlays part of UHF upwards, right up to infra-red light. It doesn't have any definite frequency boundaries, but you will find that contests, specialist groups and specialist publications have drawn their own (sometimes different) arbitrary lines and class everything higher in frequency as 'microwaves'.

Historically, microwaves were treated separately from the rest of radio technology because of the very different techniques that had to be used at such frequencies. Waveguide structures, dishes, horns, klystrons, and cavity magnetrons made up the armoury of the microwave engineer. Active devices were unusual in that the dimensions of their internal components were chosen to be resonant at their intended operating frequency. Conventional radio techniques of the time were simply not capable of approaching these frequencies. Lumped components and pin-based valves worked well into the VHF region, while microwave components had an effective limit to how low a frequency they could be made for – imposed simply by what physical size could be tolerated.

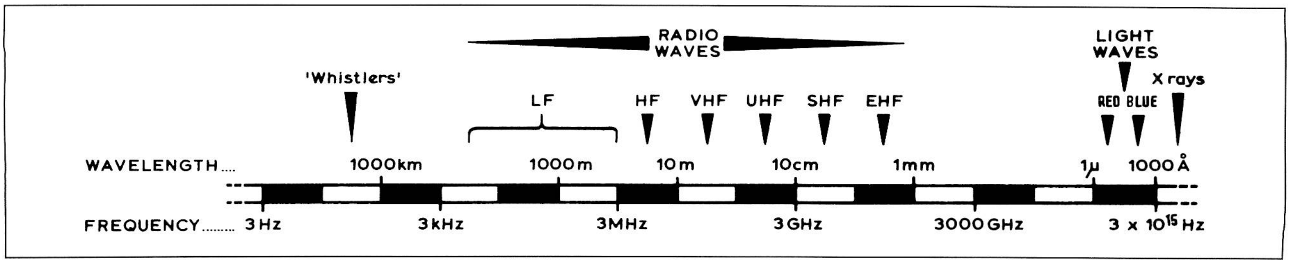


Fig 1.1. The electromagnetic spectrum

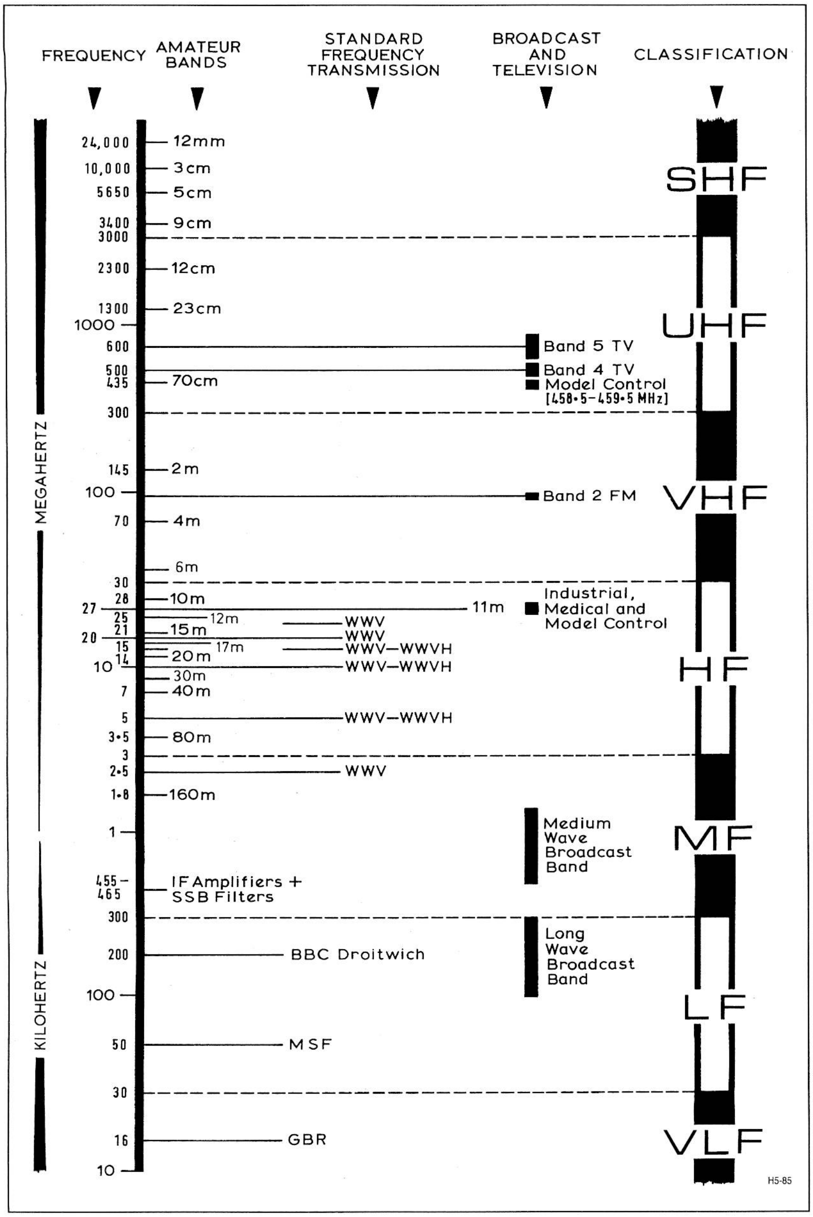


Fig 1.2. Amateur bands in relation to other services

Progress is rarely smooth, and the frantic development of centimetric radar in the 'forties was a sudden jump to much higher frequencies than had ever before been used. It opened up a gap in the VHF/UHF region where available techniques were much less advanced than those used at both lower and higher frequencies. Subsequent advances have filled in the gap, and the whole of the VHF/UHF range is now fought over as prime territory for many different uses. Modern semiconductors, surface-mount components, and stripline techniques allow printed circuit type construction to be used well into microwave territory. The 1296MHz amateur band used to be considered to be a microwave band, but advancing technology has meant that techniques used at 144 and 430MHz can now be used at 1296MHz, so it is now firmly in the UHF fold. The appearance of commercially made equipment for this band has reinforced this. There is a still-higher UHF band at 2310MHz, but it is currently thought of as being a microwave band even though it's definitely in the UHF range. Technology continues to advance and the '13cm' band at 2310MHz may eventually be poached from the microwave people . . .

THE BORDER WITH HF

Radio signals do not halt obediently at national borders, so the governments of the world have long been forced to meet in order to plan the usage of the electromagnetic spectrum. The level of chaos that would result without meticulously detailed planning is a powerful incentive. Even those nations who frequently refuse to sit around the same table send their representatives, who sit down and get on with some good, pragmatic compromising. The ITU is the overall international co-ordinating body for all communications, covering wired and optical fibre systems as well as radio. The CCIR is its main committee on radio matters. The large spectrum management meetings are called the *World Administrative Radio Conferences* (WARCs), and are usually given a year number suffix. WARCs are not held regularly, but are called when there is sufficient need. The amount of information gathering and planning needed to put together proposals for such meetings mean that several years notice is normally given, though it may be less if the amount of spectrum scheduled for review is limited. WARC79 was a major one which reviewed the entire spectrum and gave us three new HF amateur bands. There have since been a number of smaller conferences but WARC97 is the next major one.

The HF-VHF boundary, at 30MHz, is the major breakpoint in this planning process. Signals below 30MHz are treated as if they could propagate over the entire planet, and thus need to be co-ordinated globally. Those above 30MHz are treated as if long-distance propagation is rare, and so need only relatively local co-ordination. This is rather simplistic and fails to take into account the large variations in propagation that occur in this area, but it has been made to work reasonably well to date. Consequently, changes to the plans above 30MHz can be made readily by more frequent and smaller meetings than the global-scale ones that have to be called to make decisions below 30MHz. Above 30MHz, the ITU regulations give a lot more freedom for individual administrations to create their own variations to suit their particular countries. In the UK, these freedoms can be seen in the existence of the 4m and 6m bands as well as the Class B licence.

PREHISTORY

Faraday, Oersted, Gauss and others had discovered that there was some sort of relationship between electric charge, motion and magnetism. They had performed a variety of experiments and evolved a number of theories, but there was something missing. The ideas of forces that acted at right-angles to motion and action at a distance seemed bizarre compared to classical mechanics. James Clark Maxwell set out to tidy up the theories, and produce something simpler that would collect them all together and cover all the known phenomena. Between 1864 and 1873 he published four very small equations that did just that. They were written in partial-differential form, in a dialect of mathematics he had learned at Cambridge. They were quite general and covered all conditions, but when they were used to study what would happen if an electric charge was oscillated, they predicted something new. They predicted that it would cause a pair of moving waves, one of electric field, the other of magnetic field, to radiate away from it. They also predicted that the waves would be in phase with each other, moving at the same speed, the speed of light. They showed that the two fields would be oriented at right-angles to each other, and be transverse to the direction of propagation. The concept of electromagnetic waves had been born.

Theory is fine, as far as it goes, but experimental verification does a lot to help make it believable. The first experiment looked like a disaster. Maxwell had thought of how sound waves are carried in air, and how waves of the nautical sort are carried in water, and proposed that there had to be a medium that carried electromagnetic waves. It was called the *ether*. Michelson and Morley devised a neat experiment in 1881 to prove the existence of the ether. Surprisingly, it proved that the ether did not exist. This destroyed the concept of any sort of carrier medium for electromagnetic waves, and caused a great upset that eventually resulted in the theories of relativity. It did not destroy Maxwell's equations, though. They described moving waves, irrespective of whatever they did or did not move in. Maxwell's equations similarly survived the theories of relativity, and are still believed to be generally valid today.

Heinrich Hertz set out to make some electromagnetic waves. Michelson and Morley had needed waves of extremely short wavelength to make their experiment practicable, and had chosen light (wavelength about 0.0005mm). Hertz wanted to show the relationship with electricity and used a gapped metal ring which he excited with a spark. His receiver was a second gapped ring which would spark across its gap if it was close enough when the transmitter was excited. Using amazingly simple apparatus, he showed that electromagnetic waves did indeed travel at the speed of light. He demonstrated polarisation and the fact that the electric and magnetic field components lay in orthogonal planes, transverse to the direction of motion. With different-sized rings he made signals of different wavelengths, and showed that a similar-sized receiving ring was needed to be able to detect a signal.

Many people believe that radio started on long wave, and only much later expanded into the VHF, then UHF, regions. This is completely wrong! Hertz had chosen to make his apparatus a comfortable size for bench-top experimentation. His early experiments were at UHF, about 800MHz, as a consequence. Just for fun, it could be argued that because Hertz's receiver produced light as its output, then the first-ever radio link was in fact UHF television. Admittedly, it was only of one pixel resolution, but at least its intensity could be varied!

Hertz had no interest in long-distance communication. He had been inspired by the work of Helmholtz and Maxwell, and had produced the necessary experimental support for their theories. 'Action at a distance' was no longer dubious, it was demonstrable. Hertz's work, in its turn, was the inspiration for others. Marconi saw great possibilities in developing Hertz's laboratory toys. After a start at UHF, radio technology entered a period of rapid evolution with antennas getting higher, power levels getting higher and frequencies getting far lower.

HISTORY

The history of the development of long-wave communications and broadcasting, followed by the revolutionary discovery of the capabilities of short waves is widely known and oft repeated. Comparatively unknown, perhaps simply lost in the tumult of new LF/HF developments, people were experimenting at even higher frequencies.

In 1919, Marconi was experimenting at 150MHz, pushing the limits of what valves could then do. He used dipole antennas with the oscillator and detector placed right at their centres, and used parabolic reflectors to form the emission into a narrow beam. Also in 1919, Barkhausen discovered that ordinary, cylindrical anode, triodes could be made to oscillate internally. Frequencies up to 900MHz could be created under very peculiar bias conditions, with the anode slightly negative and the grid at a high positive voltage.

In 1920, Hull applied a strong magnetic field to a simple diode valve. He could make the electrons spiral out from the cathode to the anode because their trajectories were bent by the magnetic field. With a strong enough field, the anode current cut off as the electrons went into circular orbits. With anode current flowing, useful oscillation could be achieved to several hundred megahertz. A later refinement, splitting the cylindrical anode into two pieces with separate connections, made it much easier to extract a signal from the valve, and the magnetron was born. George Jessop reported that one commercial type was capable of 50W output at 144MHz. This was before strong, compact, permanent magnets had been developed, so early magnetrons were rather unwieldy – just recall the loudspeakers of the period with their huge permanent magnets or large field-coils.

The discovery, by amateurs, that the short waves were not after all useless, but instead were dramatically better than long waves for long-distance communication, made people wonder what might be found at still higher frequencies. If a move to higher frequencies produced wonderful results, then maybe going still higher might produce still . . . people had to try it. It was not long before commercial use of the short waves started. This fuelled the development of improved valves for high-frequency use.

One proud tradition of amateur radio – give an amateur a new device and he'll soon extract more power and higher frequencies from it than the designer ever thought possible – goes right back to the beginning. HF ionospheric propagation had been explored with signals coaxed out of valves intended for LF use. Small 'short-wave' triodes with the electrode connections brought straight out of the bulb to avoid the stray capacitance and inductive coupling of conventionally-based types proved useful up to 50–70MHz.

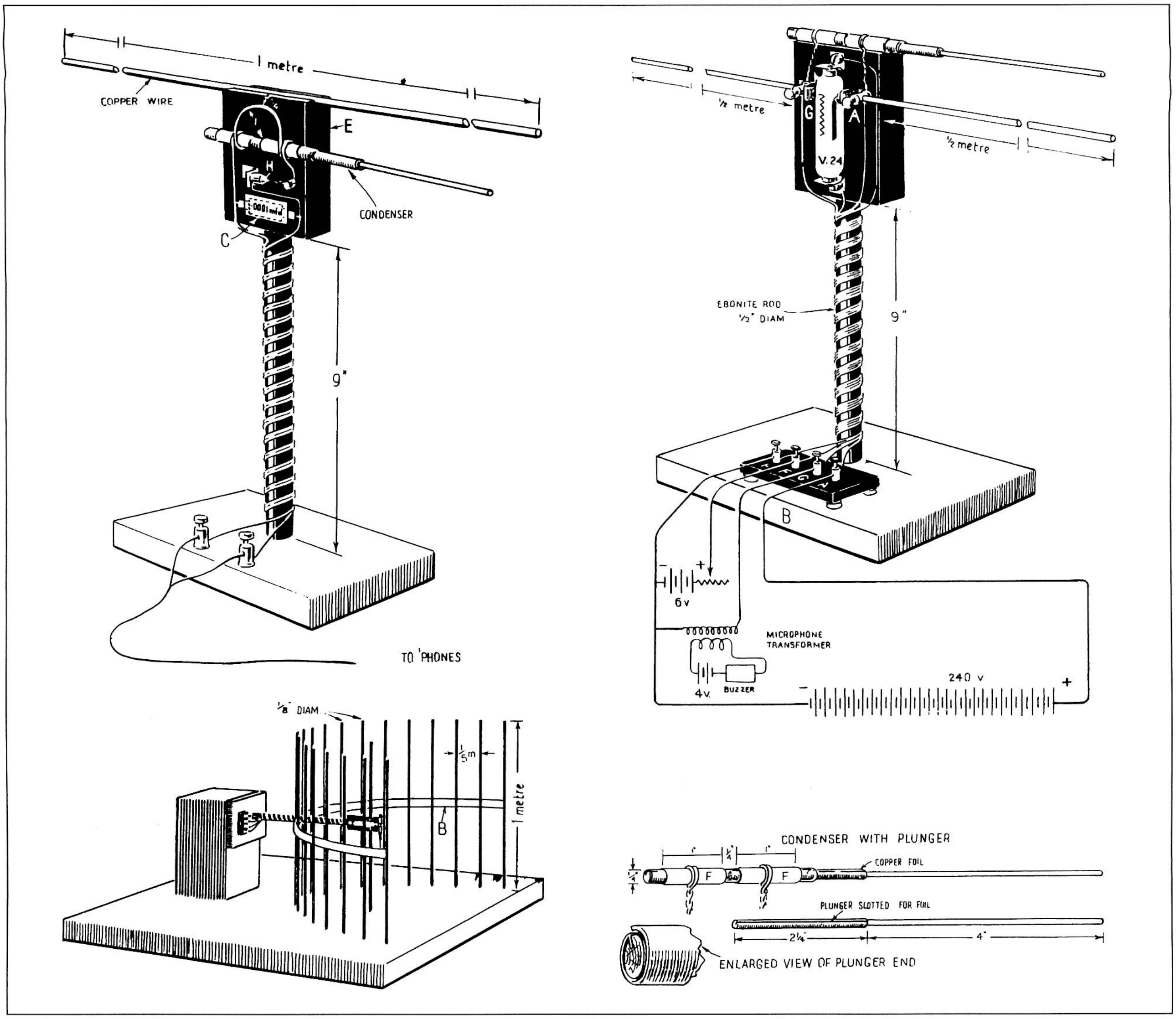


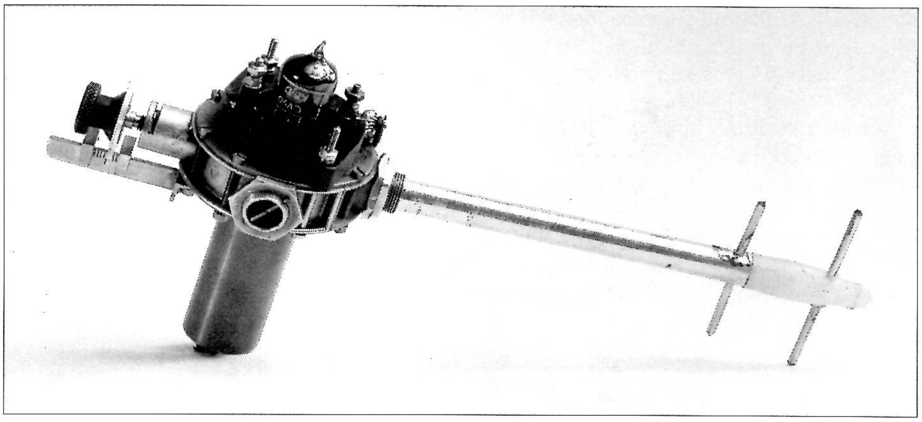
Fig 1.3. Marconi's 2m receiver and transmitter, made in 1919, which demonstrated the production of a beam using a parabolic reflector

All valves were very expensive in this period, so minimising the number needed was a very high priority. A typical transmitter would have been a pair of triodes configured as a push-pull oscillator, while super-regenerative receivers provided the necessary sensitivity with the minimum number of valves. Wire antennas were used, based on those used at lower frequencies.

In Japan, Yagi and Uda were experimenting with arrays of dipoles and had discovered the ability of non-driven dipoles to reflect or direct the signals from a driven one. From this, in the late 'twenties, they developed the *Yagi-Uda array*, much better known as the *Yagi antenna*. The ability of this new antenna to concentrate the radiated power in a narrow beam made up for the difficulty of generating appreciable power at higher frequencies, though only in the direction it was pointed. Marconi experimented with Yagi antennas and special valves at 500MHz and proved that communication beyond the visible horizon was possible. Up to this time, VHF and UHF signals had been assumed to behave like light, and the first sign of a flaw in this belief helped to stimulate increased interest in these frequencies.

Commercial research is done with a view to eventual profit. Academic research is done with a view to the publication of new knowledge. Both have to justify their existence; there has to be some possible outcome that can be explained to the controllers of their sources of funding. We amateurs are free to explore where our interests take us. We may not have industrial-strength budgets to spend, but we do have a unique level of freedom. In the late 'twenties, SSB was invented by a telephone company as a means of multiplexing several telephone channels onto one wire. It was considered to be esoteric and rather fragile in view of the quality of cable it needed to make it work acceptably. No-one thought that a radio connection could be good enough to carry it until amateurs demonstrated reliable intercontinental SSB contacts. Earlier, it was the sheer number of amateur stations, their geographic diversity and their operation at all hours that allowed the totally unexpected propagation effects of HF to be discovered.

The amateurs exploring VHF were developing new techniques and setting new records. Almost everything that could be thought of was being tried. The small size of efficient VHF antennas must have made trials from aircraft seem natural. One VHF DX-pedition of 1935 ran an extensively equipped 5m station at the top of Snowdon and logged a 205-mile contact with Romford, Essex.



An early 2.3GHz klystron oscillator and receiver

Higher-frequency bands were tried and techniques were developed to suit them. Coaxial cavities and troughlines gave the frequency stability needed to get results on 112MHz. 'Acorn' valves like the 955 triode appeared and were pressed into service up to 500MHz. Their extremely direct electrode connections and small size made them ideal.

WAR

The 1939–1945 period saw some of the most intense development of radio and electronics ever. The skills of the radio amateurs were quickly put to work as operators, technicians, instructors and researchers. Because of the amateur's insatiable desire for long-distance contacts, much of the commercially made equipment for the amateur market represented the state of the art. The American Hallicrafters S27 receiver covered from 28 to 142MHz and proved to be the only available instrument that could be used in the hunt for the first VHF radio navigational aids that had been revealed in notes from crashed aircraft. According to an unsubstantiated legend, the entire stock of S27s at Webbs Radio, Soho, was bought up by an RAF signals officer – on credit! One lesson learned from this was that even line-of sight frequencies can be used over large distances when one station is on a high-flying aircraft, without needing any help from unusual propagation. The navigational beams were simple extensions of the Lorenz landing aid that had been developed in the mid-'thirties. The ILS system still in use today is a direct descendant.

The possibility of making directional antennas of convenient size made the VHF part of the spectrum the natural choice for beam-type navigational systems. Throughout the war a cat-and-mouse game of progressively more sophisticated navigational systems and their subsequent countermeasures was played. Perhaps we ought not to get too focused on the interesting technology of it all – we must always remember that these 'games' were literally deadly to people on both sides. As fast as new valves could be developed to make high power at higher frequencies, radar moved up in frequency to take advantage of the increased directivity of the antennas that could be made in the available space. Night fighters were larger than their daytime equivalents and had arrays of Yagi beam antennas pointing forwards from their noses. Imagine a respectable 2m or 70cm moonbounce array bolted to the front of an aircraft! Ground-based radar could detect nocturnal bombers, and could direct fighters to their rough location, but the fighters needed their own radar to close the remaining distance.

In the late 'thirties, a system of electronically-scanned television had been developed at EMI to challenge Baird's mechanically scanned system. This had created a whole new sort of electronics: sawtooth 'timebase' oscillators, cathode-ray tubes, pulse amplifiers, synchronisation and triggering circuits. Previously the only signals that anyone had had much experience with had been sine waves, speech and modulated sine waves. Britain's lead in the strange new circuits of electronic television was a great help to the radar pioneers. Because several megahertz of bandwidth were needed to transmit what was then called 'high-definition' 405-line television, operation in the deserted parts of the VHF band was inevitable. Not only did television create the signal-processing circuits needed by radar, but it also created the high-power VHF transmitter technology. After the war, radar repaid its debts because the new valves and circuits that had been developed for it gave a flying re-start to the infant TV industry.

Ships' masts are the obvious mountings for sea-borne radar antennas, and for an uninterrupted view, the radar antenna has to go on top. This means that the size of the antenna has a large effect on the stability of the ship as well as on its visibility. Rayleigh's criterion from optics works for all wave phenomena, so the Royal Navy was able to estimate what angular resolution it needed, and what size of antenna it could tolerate, in order to decide on the minimum frequency they needed. They commissioned a team at Birmingham University to develop radar at 3000MHz. This was a huge step from the existing systems which had been in the 30, 50 and 200MHz regions. Radar requires very high peak power pulsed transmitters and the increase in frequency required a step far beyond the limits of any known technology. One group at Birmingham based their work on the klystron valve that was known to be able to oscillate at 3GHz, but they could not achieve anything like the power output needed. Instead, the klystron was to prove useful as the receiver local oscillator. In one of those jokes that Nature sometimes plays, John Randall and Harry Boot were working on the Barkhausen-Kurz oscillator (the mis-biased triode described earlier) to develop it as the receiver local oscillator. Eventually, they gave up hope of getting enough power to drive a mixer, or good enough frequency stability, and switched their attentions to the split-anode magnetron. This was at least known to be capable of moderate power, although the frequency stability was known to be poor.

Randall and Boot attacked the frequency stability issue by replacing the simple split anode with a solid block of copper, drilled to make a group of resonant chambers, sized for the wanted frequency. Their prototype had six resonator cavities, each linked to the central cavity by a slot. It was reputed to have been made by using the cylinder of a revolver as a drilling jig. It burned out their dummy load and easily produced several hundred watts, close to the intended frequency. Permanent magnets had been greatly improved by this time, and the quest for a receiver LO had produced a marvellous high-power transmitter. Small, robust, and efficient, the cavity magnetron was simply ideal, and could easily produce peak powers of many kilowatts in pulsed radar duty. Klystrons, which had started out being groomed for the power output job, made great LOs for superhet receivers. Centimetric radar had arrived, right at the upper limit of UHF.

Just after the war, many people were using beam antennas, crystal-controlled transmitters and superheterodyne receivers. The first trans-American contact had been made at 50MHz, showing that VHF was sometimes anything but 'optical'.

WAR SURPLUS

At the end of hostilities, the immense military organisations could not be dismantled instantly, and time was needed for the world's economies to change back to peacetime activities. Military service and rationing continued for several years in some countries. Huge quantities of hardware became surplus to requirements. Cryptographic equipment and other highly secret things were stockpiled or were carefully destroyed to protect their secrets, but tremendous amounts of communications equipment and components were sold off. Disposable incomes were very low before the economies recovered, so technical treasure could only be sold for a tiny fraction of its original cost.

It took until the middle of the 'sixties for the flow of war-surplus goodies to dry up – there was simply so much of it. In the space of a couple of decades we had gone from it taking many weeks worth of the average income to buy one valve, to amazing devices, being sold by the tea-chest-full, at disposal prices. Large numbers of people had been trained as makers, operators and repairers of all manner of communications and radar equipment. They, too, had become surplus to requirements. Suddenly dumping large quantities of people or things onto an open market is usually a precursor to disaster, but in this case the things and people were complementary and were just what the emerging civilian markets needed. Radar experience was very appropriate to television, and the work done in developing CRTs and wide-band valves was not wasted. Many home-brew TV receivers were founded on the VCR97 electrostatic CRT and the EF50 VHF pentode.

Most amateurs were primarily interested in HF activities, so the demand created a scale of prices, with radios like AR88s at the top, then down through the HRO, CR100, BC348, R1155 etc. The 19 set tank radio was one of the bargain-basement jobs, and although it had a crude VHF transceiver built into it, the VHF section was usually unceremoniously ripped out to make space for an audio amplifier. The period of homebrew television construction created a market for some VHF bits and pieces, but it was relatively short lived as commercially made televisions became available. Perhaps black-and-green pictures on 5in screens just didn't fit in with the entertainment needs of other members of households? Specialised VHF or radar equipment and components went for junk prices.

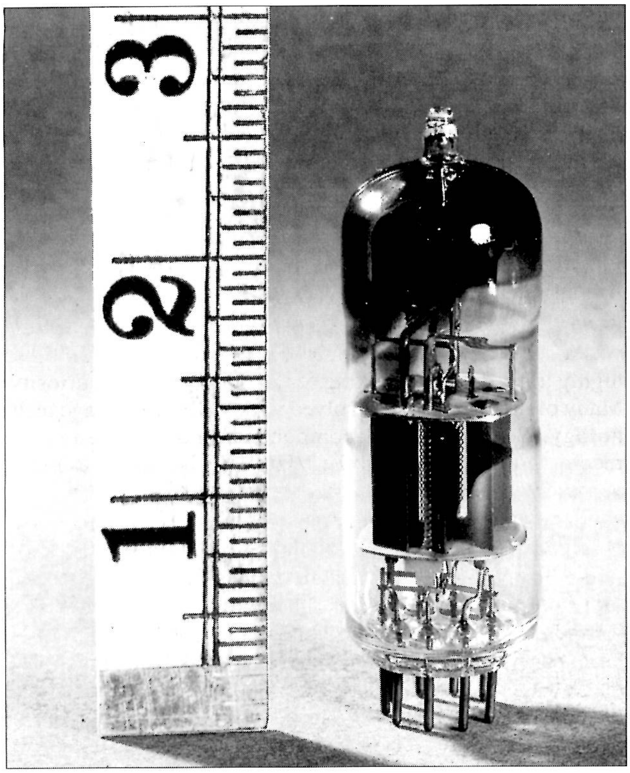
No true amateur can ever turn down something for nothing or even close to nothing, so a lot of strange things wound up in the hands of people who then wondered what they could do with them. In addition to the dispersal of all this bounty, there was the dispersal of the knowledge to go with it. There was a big influx into the hobby caused by some of all those highly trained people looking for an outlet for their curiosity. Many of them had been involved with VHF/microwave technology. With advanced components and knowledge the records previously set on the VHF/UHF bands were highly vulnerable, and new frequencies with no records had become accessible. The 144 and 432MHz bands were allocated in the late 'forties, virgin territory for new explorers.

As had happened earlier on short waves, it was the creation of a large number of stations spread across the world, operating at all hours, looking for long-distance contacts, which led to new discoveries. Two American stations made the first moonbounce contact in 1953. One bunch of radio astronomers-to-be tried to use the surplus mobile radar unit they had obtained to detect their own lunar echoes. They included Bernard Lovell, and they went on to build the Jodrell Bank radio telescope. Terrestrial propagation studies got a boost in 1957, the International Geophysical Year, with the opening of GB3IGY, the first of the VHF beacons. The potential of meteor-scatter and sporadic-E propagation were beginning to be discovered. In 1959, a sporadic-E opening gave contacts between the UK and Italy. In America, 400 miles was achieved on 1.3GHz, then California and Hawaii were linked on 220MHz. Records and 'firsts' were there for the taking.

THE RISE AND RISE OF THE TRANSISTOR

Typical VHF/UHF stations of the time consisted of a crystal oscillator-multiplier-PA chain as the transmitter, with a crystal-controlled converter ahead of a standard HF receiver as the receiver. In the 'sixties, commercial VHF transceivers became commonplace in emergency services vehicles, with utilities, and with taxi firms. There had been experiments with mobile 'two-way' radio much earlier, on HF and VHF, but this was the full-scale deployment of private mobile radio (PMR). Successive generations of PMR mobiles were rapidly developed, as new types of transistors became available, capable of supplanting valves in progressively more awkward applications. Some of these earlier generations of PMR sets found their way into amateur hands when they were prematurely pensioned off to make way for each successive smaller and cooler replacement. It was not difficult to re-crystal some of them for 144MHz, though some people who acquired 'low-band' units had to put in a lot of work rebuilding their RF stages to convert them into 'high-band' units. Up to the appearance of this equipment, anyone who wanted to operate mobile had to build their entire station.

On HF, the arrival of the transistor coincided with the rapid growth of high-power short-wave broadcasting, so the



Manaufacturers introduced smaller and smaller valves for VHF mobile radio, such as this 6W double tetrode (Mullard QQV02-6), but all-transistor technology was already on the way

sudden worsening of the ability of receivers to handle nearby big signals was rather badly timed. It took over two decades to make up for this great step backwards, and to be able to make semiconductor-based HF receivers that were in no way inferior to the best of the valved units. On VHF/UHF, conditions were more favourable, receivers were not being 'tenderised' to the same extent, and semiconductors offered progressively lower noise figures, a great boon where noise limitation was far more likely than overload problems.

In 1961, just four years after the first-ever artificial satellite, Sputnik-1, the first amateur satellite, Oscar 1, was launched. It transmitted telemetry on a low-power beacon in the 144MHz band. Later amateur satellites offered communications facilities for short periods. While these were marvellous advertisements for the hobby, and a great deal was learned from them, satellite operation in the 'sixties was very much a minority interest. Many people listened to them, but only the very dedicated built the multiband stations with azimuth-elevation mounted antenna arrays needed for two-way working through something that might only last weeks, if it made orbit at all. It was Oscar 6, which was launched in 1972 and survived for four years, that broke the pattern and showed that satellite DX was a long-term proposition. Four years was more than long enough for the exploits of the pioneer users, and the details of how to follow their lead, to reach print and become general knowledge.

CLASS B LICENCES

The ITU Regulations call on member states to require proof of proficiency at Morse code from anyone to be awarded an amateur licence giving access to the bands below 30MHz. This was a simple consequence of some MF/HF amateur bands being shared with higher-priority users, who were only equipped for CW transmission and might need to request an amateur station to close down. Without this requirement, we would probably not have those bands at all. However, because signals above 30MHz were not expected to travel far, countries were left free to make their own decisions about Morse proficiency and access to these bands. The UK took advantage of this and created a new type of licence limited to 430MHz and above, and issued it without requiring the usual Morse test pass slip. Later, this 'B' licence was extended to include the 144MHz band, and more recently, to cover all bands above 30MHz. An experiment in the 'eighties to allow Class B licence holders to use Morse code on the air proved very successful, and it is now permanent.

The Class B licence (and similar licences in other countries) has had a very large effect on amateur radio. Once upon a time, the usual mode of entry to the hobby started with a period of short-wave listening with one of those war-surplus HF receivers, while reading and re-reading every book and magazine that could be found on the subject. Local amateurs offered encouragement, help, and the loan of books and magazines from their joist-threatening 'archives'. After enough of this, the beginner would be ready to attempt the Radio Amateurs' Examination, which then had a small number of questions, each requiring a mini-essay as an answer. The candidates were also expected to be able to draw circuit diagrams from memory of typical sections of transmitters and receivers. This was not as fearsome as it might sound today – radios were a lot simpler then, and there were only a few circuits that were ever called for. After plenty of listening to Morse transmissions, especially the low-speed training broadcasts, the beginner would take the dreaded Morse test at a shore station or other Post Office radio establishment. Despite a few stations that were reputed to have had at least one ogre on their staff, many amateurs have described examiners making great effort to be welcoming, encouraging and to help candidates overcome their nerves.

The freshly licensed beginner went on the air after some years of familiarisation with operating practices. The 1.8MHz band was the most popular starting point because the receiver used for the short-wave listening period could be used with a cheap and easily built low-power transmitter. On the 1.8MHz band, everyone was limited to low power, so the beginner was at little disadvantage with a simple transmitter.

The Class B licence now allows someone to go on the air after passing just the RAE, and the only extra limitation on them is the 30MHz barrier. If someone wishes to learn Morse code, either for VHF use or to gain access to the bands below 30MHz, they can now get a Class B licence, and join in one of the Morse training nets on 144–146MHz. This is a lot less lonely than just listening, and a lot more encouraging. Copy can be read back for checking, and the instructor can choose the speed to suit the progress of the group. Due to the relative price and availability of commercially made radios for the different bands available to the Class B licensee, the FM segment of the 144–145MHz band has replaced the 1.8MHz band as the 'nursery slopes' of amateur radio, where the majority of newcomers make their first contacts.

The Class B licence has now been with us for a few decades, and, contrary to the predictions of some of its critics, the world has not ended. It has been the entry point for many good people. It can justly be called a success. The USA has recently introduced its first-ever no-Morse-required VHF-and-above licence and is still suffering the inevitable flare-up of its branch of the great Morse debate.

ENTER THE 'BIG THREE'

In Japan, amateur radio had long been practised by a much larger fraction of their population than had been the case in the West. This proved to be a fertile market for the evolution of a number of fiercely competitive manufacturers, each large enough to support mass-production methods. Once their home market approached saturation, they looked to exports for further expansion. The Western manufacturers never knew what hit them.

The radio amateurs in the West were abruptly confronted with shiny, new, feature-packed radios at much more affordable prices than ever before, and the move from home-built to purchased equipment was fully underway. The first equipment from Japan concentrated on HF SSB operation and rode the change from AM to SSB on those bands. Soon afterwards, the Japanese companies started to export their VHF models. The VHF bands had never before been served by specifically manufactured equipment, and affordable off-the-peg radios were an attractive alternative to scratch-building or modifying surplus PMR equipment. In the UK, the appearance of these radios fuelled the rise in popularity of the Class B licence.

REPEATERS

The amateur world was awash with affordable, commercially made, 2m band FM transceivers. Some were retuned, re-crystalled commercial PMR sets, others were purpose made for the amateur market. The 144–146MHz band had been fairly popular for local natter using AM PMR equipment, but the switch to FM accompanied a big increase in occupancy. The majority of PMR sets had been designed for mobile or portable operation, and the Japanese manufacturers produced portable and mobile models. To enhance the coverage of all these mobiles, an extensive network of repeaters has been built. With the part of the 145–146MHz bandplan assigned to repeaters pretty much full, most new repeater development has shifted to the 430–440MHz band. Not all repeaters are FM, just the majority, so the rarer SSB and TV repeaters are especially interesting.

The repeaters have acted as magnets for a very small number of disaffected people who try to jam them. This is antisocial, illegal, and damages the reputation of the hobby. Their motivation seems to be a mixture of wanting to annoy people so that they can listen to the resulting anger, and of wanting to get attention. If you do encounter a repeater jammer, the best advice is to do nothing that acknowledges their existence in any way. Do not rise to any taunts. Any sign whatsoever that they have been noticed seems to encourage them. Quietly log the times, the signal strength on the input frequency (and direction, if possible) as this may later help. Never discuss such things on the air as this may either encourage them or tip them off. Ultimately, the Radiocommunications Agency has the power to prosecute, leading to confiscation of equipment and fines. Finding theperpetratoris only part of the problem; sufficient evidence has to be collected in a way that is admissible in court. All this work goes on behind the scenes, and all that is ever seen is a small piece in the news columns of *Radio Communication* magazine announcing the result of a court case. The number of jammers is very small indeed (though they try to make as much nuisance as possible), and the number of prosecutions indicates an excellent clean-up rate. Some have tried jamming while mobile, but now their vehicles are at risk of confiscation.

Based on the number of QSOs and the number of users, the repeater network is very successful. Very few of the QSOs may contain matters of direct importance to amateur radio, but they are useful indirect tools that allow the communities around them to organise various activities.

THE 6m BAND

The creation of a 6m band in the UK came as a large surprise, and was solid evidence that the DTI looks favourably on the Amateur Service. This part of the spectrum was freed up when the 405-line TV broadcast system was closed down and, although it is assigned as an amateur band in the Americas, in our region it is not. The first amateur use was by a few stations with special licence variations as an experiment. Part of the experiment was to investigate the effect on French TV which was still active on low VHF. As the French TV service had coped with our old 405-line TV transmitters, each blasting out many kilowatts across the same band, some care over our power output and antenna configurations ensured a satisfactory demonstration of co-existence. The experiment turned into a permanent new band. It is ideally situated, with interesting propagation effects to be explored and a heavily equipped continent just an ocean away to form the other end of long-range contacts.

DATA MODES

The radio-teleprinter techniques used on HF have always been directly applicable on VHF, but the appearance of mass-market home computers in the 'eighties prompted widespread exploration of new possibilities. Just using the computer to simulate the function of a classic electro-mechanical teleprinter may not look like a bold step forward, but the reduction of size, noise and weight in a domestic environment did a lot to increase the numbers of people prepared to try RTTY.

The X.25 protocol of the packet-switched data technology, used in some parts of the public telephone network, was modified for use on the air (AX.25) and grew into the packet radio network. The originators of X.25 never considered it suitable to be deployed over links made out of simple speech-type radios, but amateurs love making the unexpected work. To the basic packet radio network, bulletin board servers have been added, offering news and electronic mail facilities. Specialised bulletin board servers have been linked to make the DX clusters which rapidly distribute news of interesting band openings or the appearance of rare stations. They may be a mixed blessing, for while the news of a band opening can stir up activity on that band, the 'spots' of rare stations do tend to create beautifully synchronised pile-ups – though mostly on HF! The packet format is well suited to VHF/UHF conditions, and does not work anywhere near as well when tried on HF.

Here Belgian and Dutch radio amateurs are exchanging TV pictures via the British TV repeater GB3LO

IMAGE MODES

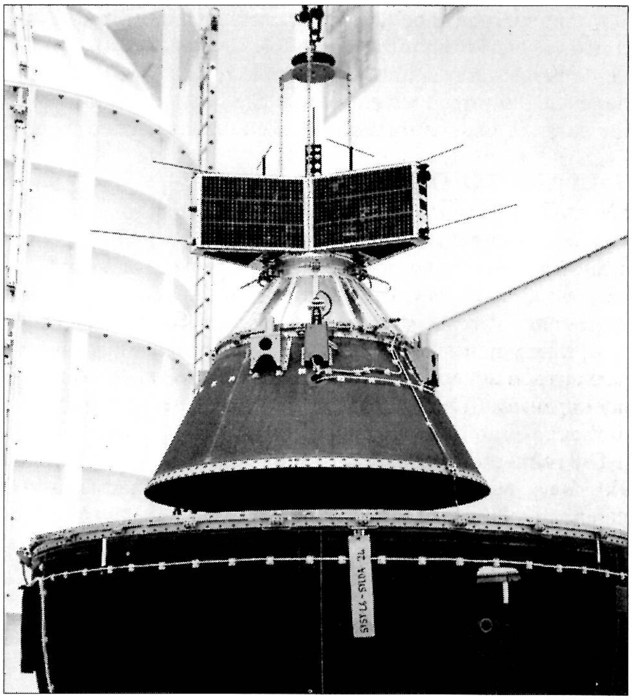
Slow-scan TV works well on VHF/UHF and, compared to HF, the clearer channels give less interference on the pictures. Once again home computers have made things much easier and have replaced the long-persistence CRTs, photographic drum printers and scanners that were once the only way of handling slow-scan and facsimile images. The UHF and microwave bands offer the space needed to carry a conventional TV signal. TV repeaters have also been built. Digitised images can be treated just as any other data files and sent by the normal data modes.

SPACE

There are amateur satellites acting as very-wide-area repeaters for SSB, while some provide images of the Earth and others carry packet mailboxes. Send your long-distance email up when the satellite is over you, and the recipient can read it when the satellite passes over him or her. Astronauts and cosmonauts have taken out amateur licences, and taken up amateur transceivers. There have been many contacts with Mir and the Space Shuttle. One group onboard Mir actually did their exam in space to get their licences before the end of their mission, and had equipment sent up on a routine supply rocket. Amateur radio has become a popular hobby with spacemen and spacewomen everywhere. It is probable that most future missions will have some amateur radio content. As part of the first-ever amateur operation from orbit, Owen Garriott also sent slow-scan TV images of the interior of the Space Shuttle in the 2m band. Average VHF stations were sufficient to receive these. Few of us will get the chance to try radio actually *in* orbit, but we can all be Earth stations if we wish and join in the funding, design and development of future amateur satellites.

THE PRESENT AND THE FUTURE

History is not 'bunk', it is simply inaccessible, and we can neither visit it nor change it. Living in the present, we can see the past clearly, yet our influence is over a future that we cannot see. This seems like a recipe for disaster. Given the unfair advantage of the certainty of the past over the uncertainty of the future, it is easy to believe that most of the wonderful things have already been done, and that there is little left worth doing. This is very probably not so – such things



The Phase 3C amateur radio satellite being packed for transit to the launch site

have been said for centuries by people who ought to have known better.

Amateur radio is now quite a mature interest and, with the advances in communications in general, it must now look not at all magical to the layman. One of the greatest challenges of the future must be the attraction of new people.

The future holds great threats to the survival of the hobby, especially on VHF/UHF. In the 'eighties, the liberalisation and privatisation of telephone companies in various countries has led to a boom in radio-based communications services that is still continuing. The analogue cellphone networks may be close to their peak, but the digital networks will replace them and grow much larger still. Public appreciation of the relative security of their conversations on different types of telephones can only increase. Some companies are trying to sell cellphones not just as mobile devices, but also as alternatives to wired phones in homes. The network of wires leading from the telephone exchanges to their subscribers is something that cannot be duplicated to allow free-market choice without astronomical cost, so there are plans for competitive phone networks using fixed radio links to homes and businesses. Add in the growth of cordless telephones, cordless burglar alarms, remote control devices, and car key-fob transmitters, and it can be seen that the demand for space in the RF spectrum is accelerating dramatically. Proposals for global spectrum allocations for use by low-orbit satellite-to-person communications have included requests for the reassignment of all of our prime VHF and UHF bands. The idea of a wristwatch phone that will work anywhere on Earth (or in nearby space) with everyone having a single personal telephone number, without geographic codes, is now very close to being achieved. Such progress cannot and should not be stopped, but nor should it be allowed to simply destroy other services.

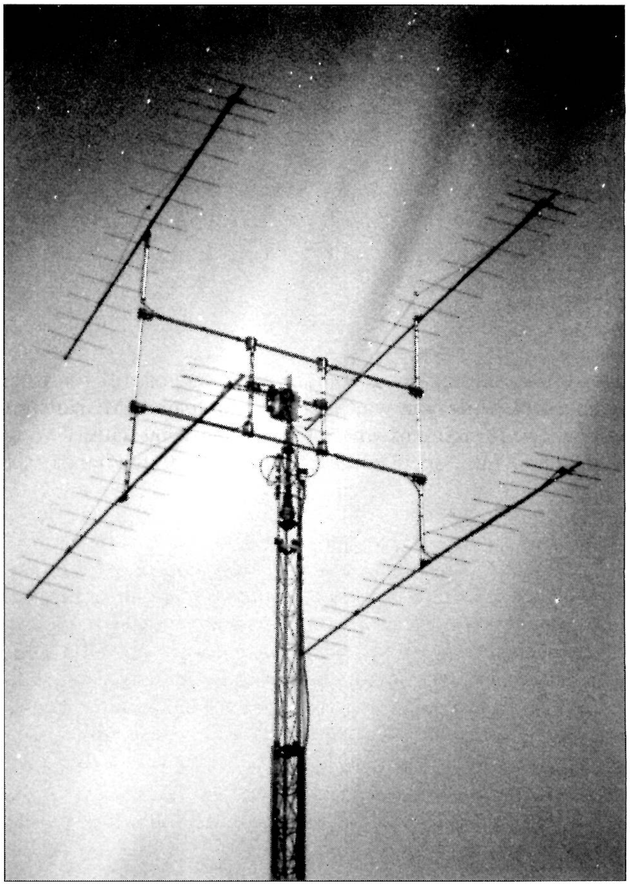
Amateur radio above 30MHz is facing the most severe threats ever. To fight these pressures we need to be able to demonstrate large numbers of people regularly using all the bands, and we need to be able to make good presentations of the value of the continued existence of these amateur bands.

THINGS TO DO

There are plenty of things waiting to be done that will advance the individual and/or the hobby.

On 2m or 70cm FM, there are new people to be welcomed and encouraged. This could range from informal help when requested, to organised on-air RAE tuition for novices in your area. The voice repeater network is still expanding but it is fairly mature now. The receiver sections of many repeaters are suffering from overload from adjacent paging transmitters, so a high-dynamic-range receiver design would solve many problems. Some groups are experimenting with solar or wind powered repeaters to take advantage of low-cost isolated sites, and also with linking groups of repeaters. A few TV repeaters already exist, but more of them would stimulate TV evening nets.

DXing can be done via sporadic-E openings, meteor-scatter or moonbounce. The moonbounce station at GM4JJJ is a typical medium-sized installation for the 144MHz band, with four long-boom Yagis driven by a pair of 4CX250B valves at about 900W, allowed by a special experimental licence. A small moonbounce station might have a single long-boom Yagi fed from a 100W solid-state amplifier. There are big-league stations with monster antenna arrays that can make CW contact with small stations, or SSB contacts with medium stations. Occasionally some amateurs get the temporary use of a really big dish and moonbounce contacts become possible for quite modest stations.



The moonbounce antenna at GM4JJJ

The packet network is there to be explored, but it is now bound by two limitations. Our licence conditions covering message content and purpose prevent it becoming open to full Internet traffic, because of the wide-open nature of that network. More open to being fixed is the speed limitation forced by channel bandwidths originally planned for narrow-band FM speech. The progressive replacement of the network with high data-rate microwave links could yield a great improvement in network capacity and response speed.

There is very little activity on the higher and microwave bands other than during contests. On the use it or lose it principle, we need to devise attractive new uses for these bands. TV and fast data links are obviously suited to the great bandwidth available, but what other possibilities are there?

In a market flooded with commercially made equipment, there is still a place for home construction. Much of the commercial gear tries to do everything at the cost of doing nothing particularly well. If you want a competition-grade system, think in terms of masthead pre-amps and home-made transverters into HF transceivers, if not entirely home-built transceivers. Much commercial equipment has been compromised – the current fashion for wide receiver coverage has made the receiver sections of many transceivers resemble scanners, with a subsequently increased likelihood of intermodulation and overload problems from out-of-band signals. Hand-held transceivers have responded to market demands for smaller size, longer battery operation, as well as wide range scanning. The tiny 'rubber duck' broad-band antennas are very inefficient, but the receivers have been compromised by their wide range and low power consumption design goals, such that the use of any better antenna risks overload problems. Much better performance can be got from a radio specifically optimised for amateur-band-only use. One approach would be to convert one of the cheap ex-PMR handhelds that appear at all rallies. It would be unfashionably big, but it would also be tough enough to knock nails in with and would also maintain a QSO under conditions that would overload something more fashionable. Kitchen-table construction can still produce things that can outdo the big firms if the constructor targets his specific needs and tolerates a less 'consumer' appearance.

WELCOME

Well, that's a brief introduction to the VHF/UHF bands, a place of threat and opportunity. It offers wide open spaces in which to try out things that no-one has yet thought of, and it offers many known activities that may still be adventurous for the individual. The Brendan Trophy still awaits the first transatlantic 2m QSO.

Anyone fancy a QSO with the first manned mission to Mars?

2 Getting started

OK, you've done it! The pass certificate for the Radio Amateurs' Examination is in your hand and perhaps the Morse test on the way – you are ready to post off for your licence but while you wait for that anticipated 'M' callsign you will be thinking of where to start and what with.

You may have spent some time listening to the bands as a short-wave listener and have a pretty good idea of where your interests lie but there are those who do not have this preconceived idea of their immediate future.

The best place for your first stop is your local club – if you are not sure where your nearest club is located then a phone call to the RSGB will give the required details of dates, times and contacts. Most clubs give newcomers a hearty welcome and this will be your chance to talk to other amateurs about what they prefer. Many clubs have a club station where you can get 'hands on' experience of some of the available transceivers and get some answers to those difficult questions. Many of the local amateur radio retail outlets will also offer advice but remember that they may be tempted to push the higher-profit radios your way.

TRANSCEIVERS

The first real decision is which transceiver to obtain. There are many on the market but which one is best for you? Should you go for a hand-portable one ('hand-held'), or a more expensive base unit to sit on the table at home? Would a mobile unit be preferable and should it be just for FM or for SSB and CW use too? Which bands do you want?

The questions are easy to ask but much more difficult to answer, and of course the answers will be constrained by the money you have available for the hobby or the units you are willing and able to build. Let's take a look at each in turn – it should be noted that all comments made here refer to the 2m band but are also applicable to all VHF and UHF bands too, from 6m upwards to the microwave bands.

A great percentage of radio amateurs own a 'hand-held' transceiver – most are on the 2m (144MHz band), some are only for the 70cm (432MHz) band and some are dual-banders for 2m and 70cm together. They are typically FM only and will only give an output power of about 1W when run from the supplied battery. (Some may give higher power when run from 12V, though.)

These hand-helds can be used while walking the dog, while you are strolling around the local rally or even while mobile or sitting in the armchair at home. When in use out walking, the 'rubber duck' antenna supplied with the transceiver will be used – these are helically wound in most cases and will radiate a signal (of sorts). For base use or in an mobile environment some form of outside antenna will be beneficial. It must be remembered that the 1W output of these rigs will be heavily attenuated if poor-quality coaxial cable is used. Even in short lengths a good proportion of the signal can be wasted. Having said that, a good antenna can make all the difference to any radiated signal, but more of that later.

So, we have our hand-held radio clipped to our belts or in a pocket. It will prove very useful, but not so good for longer-distance contacts except perhaps through a repeater or when standing on the top of a hill. Many of these repeaters also require slightly more power than the typical hand-held can deliver.

There are many, many hand-helds available on the market, both new and used. It all depends on what you require of your rig, how much you can afford or more importantly how much you are willing to spend. If this is your main interest than a call into the local shop or a browse through the adverts in a few magazines will point you in the right direction.

The older type of thumbwheel-tuned type such as the IC2E has been used by many amateurs over the years. There are even a few modern look-a-likes available too. They are simple, easy to use and above all cheap. The more modern type of hand-held with its built-in processor may at times make you feel that you need a degree in programming to understand the way they work.

If your interest lies more further afield you may wish to consider something a little different. While FM is a good chat mode for stations fairly close to each other it is not as efficient as single sideband working for longer-distance contacts. In this mode the carrier is removed and thus more of the available power is used to concentrate the voice into a narrower bandwidth than FM.

Single sideband is a much more efficient way of using the available spectrum when conducting a contact – the FM signal is typically 12kHz wide, while the SSB signal is typically only 2.5kHz. Thus the output power of the transmitter is more efficiently used. It is because of this that an SSB signal will be heard over a greater distance and stations that would be impossible to work when using FM can be contacted with ease when using SSB.

So, which transceivers are available for SSB and FM working? Without doubt the most famous of all the simple, low-power transceivers ever built were the FT290R and the FT790R from Yaesu. This is not meant as an adverse comment about all the other manufacturers, but these rigs stand out. Throughout the late 'seventies, the 'eighties and the early 'nineties almost every amateur owned one of these at one time or another. They were used mobile, portable, at home, and on planes and ships – truly versatile rigs. It is true that they only gave 2.5W out on 2m or 1W on 70cm but they worked and worked very well.

These radios gave the newly licensed amateurs their first taste of longer-distance working (DX) on VHF/UHF and the addition of a small linear amplifier made the chance of those 200km-or-more contacts a distinct possibility. Like all good things the early FT290R was superseded by a Mk 2 version which in many people's opinion never came up to the standard of its elder brother.

So, in the case of the hand-held transceiver, the advantages are: portability, low weight, ease of use (mainly), and low power consumption. The disadvantages are: low power output, the need to recharge the batteries, and the ease of losing the set if you are not careful.

The trusty hand-held with its 1W of power may seem very basic, but under certain conditions can work very long distances – the chapter on propagation should be studied in detail to give more information on this. Suffice it to say that I have worked from my house in Folkestone, Kent to Paris, France on FM with 'flea power' – just 2W into a vertical antenna. On VHF you should be able to work all over Europe with a couple of watts of CW.

Let's look a little more closely at the larger transceivers available for base use. These typically run off a mains supply but after the introduction of the EMC regulations in 1996 most of the big manufacturers changed over to transceivers that required an external 12V supply.

During the late 'seventies and the early 'eighties the champion VHF rig was the FT221R. This transceiver was used by most of the big contest groups and when fitted with the famous replacement muTek front-end became the 'be-all and end-all' of contest rigs. These transceivers changed hands at huge amounts of money. Because of trends they have become less popular. These rigs didn't have too many bells and whistles but did work extremely well. They had analogue tuning too – no digital frequency readout here. The advantage for us now is that because of these supposed disadvantages the transceiver is considered 'old hat'. The value has dropped considerably but they still work very well when compared with modern equipment. The later contest radios didn't become classics like the FT221R did, but some became firm favourites. The Icom IC251E and the later IC271E were also very good and the later IC275E was used by many contest groups.

The moral of this story is that even some of the older rigs have good receivers and that is what is important. Remember that 10W of RF from an cheap old transceiver will sound much the same as that from its multi-thousand-pound rival. It's the receiver that counts!

For those who wish to have only a simple, inexpensive FM rig for VHF/UHF the ex-PMR (private mobile radio) types are an invaluable source of cheap rigs. Most will require some work to get them onto the amateur frequencies but many of these modifications have been published in the amateur press. Chris Lorek, G4HCL, has published a profusion of them. His *PMR Conversion Handbook* (and its predecessor, *Surplus 2-Way Conversion Handbook*) gives many. These transceivers may be found at silly prices at your local rally. There are always a selection available if you look around.

These days, there is a profusion of rigs that can be used by the newly licensed, the only proviso being whether your interests lie in only FM local contacts and/or SSB with longer-distance contacts, packet use with the radio connected to your computer or even the receiving and transmitting of ATV (amateur television). There is of course a world of difference in the way the various modes are operated. There is usually a huge difference in cost, too. The multitude of transceivers available will baffle most, but eventually one will stand out above the rest. Listen to what the locals say about it and try to listen to other owners of the same model. See what they think about it. Are they happy with their purchase?

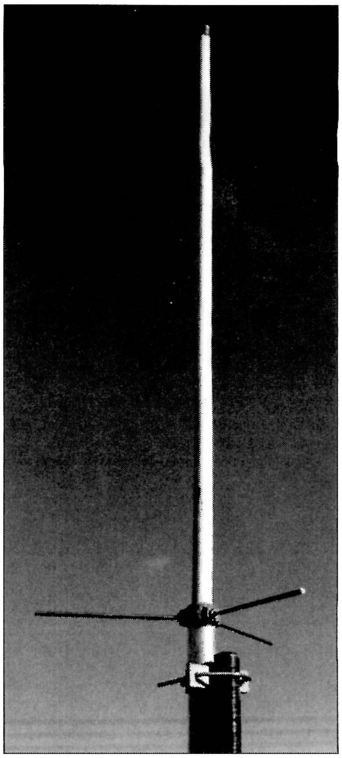


Fig 2.1. Typical collinear antenna

Reviews of current and older transceivers are always available. Finding which magazine did the review will not be difficult as they generally all do a review of each one. Read the reviews and see what they say. Be careful and read between the lines – sometimes a lot can be said by what is *not* written about a radio.

So, we have discussed the basic radios that you may consider. There is much more than this of course. The old adage of "if you can't hear them, you can't work them" makes sense.

If the long-distance station (DX) is inaudible at your station then you will never work him. What can be done to remedy this situation?

CHOOSING AN ANTENNA

No matter how much you spend on your shiny new or dusty second-hand transceiver it will always be improved by a better or more suitable antenna.

By now the newcomer should have a vague idea of where his or her main interest lies – if it is for local chatting on the FM simplex channels on VHF/UHF then even a modest outside antenna will give the transmitted (and received signal) quite a boost.

Many amateurs keep their antennas in the loft of the house, while others prefer to have them on the roof. Whichever you choose, the higher the better.

The standard quarter-wave antenna will radiate an omnidirectional signal, that is a signal of the same strength in all horizontal directions. This is what we need for our hand-held. Convention has it that when using FM the antennas will be vertically polarised. This means that the radiating element will be in a up/down configuration. Don't assume that the antenna must always have the feeder at the bottom. There are circumstances, such as when the antenna is fitted under the eaves of the house, that it can be fitted with the feed point at the higher point and the antenna is then 'hanging upside down'. This will not make for any difficulties in transmissions – the antenna will still work very well. Fig 2.1 shows a typical collinear antenna.

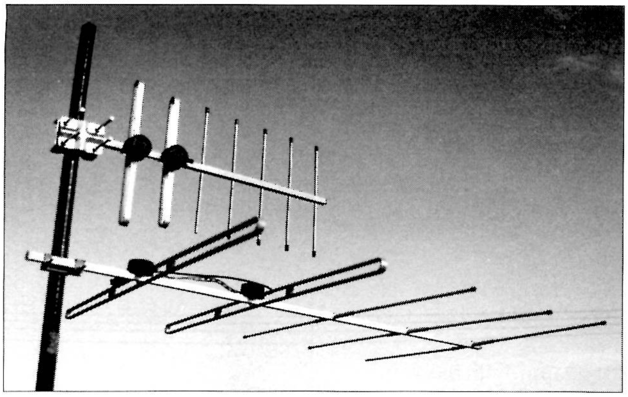


Fig 2.2. Vertically and horizontally polarised antennas on the same mast. The upper one is a vertically polarised 70cm antenna, the lower one is a horizontally polarised 2m antenna

By looking at the roofs of houses around us we will see several antennas used for the reception of TV. Most of these types are called *Yagi beams*. These have directional properties in that they 'fire' the signal in one particular direction to the detriment of other directions. So, if that repeater you wish to access is too far away for the omni-directional antenna, put up a beam. Remember, for FM it should be mounted so that the elements are vertical and for SSB work it should be mounted horizontally. Do not do as one operator I know who mounted the Yagi pointing to the sky and wondered why his signals were not so strong! Fig 2.2 shows both vertically and horizontally polarised antennas.

You don't have to have two antennas if you swap between these two modes. You can buy antennas specially made for the job, where a single boom has two sets of elements running along it, one for vertical and the other for horizontal polarisation. You will need either two lengths of feeder to the shack and a switch box to select the polarisation required or you may decide to have a relay at the antenna and just a single line of feeder to the radio.

Of course the gain of any beam antenna works on receive as well as transmit. While we mention 'gain' in an antenna it must be remembered that if we supply an RF power level to an antenna of, say, 10W, there is no way that this antenna can increase this 10W level to a higher power level. What it can, and does, do is 'steal' a little of this power from one (unwanted) direction and push it in an other direction. We still have 10W being radiated from the antenna but not in every direction. Look at Fig 2.3. It can be seen that the loss of apparent power in one direction can give an apparent increase in another. In this case the 'loss' is to the rear of the beam and the 'apparent' gain to the front.

For those who prefer to sit and chat on the simplex FM

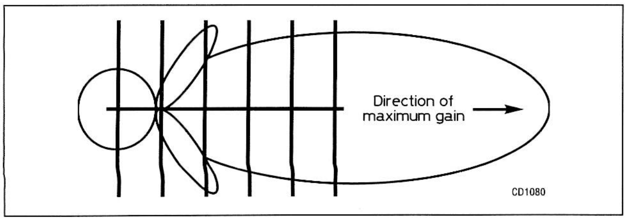


Fig 2.3. Yagi beam antenna showing direction of maximum gain

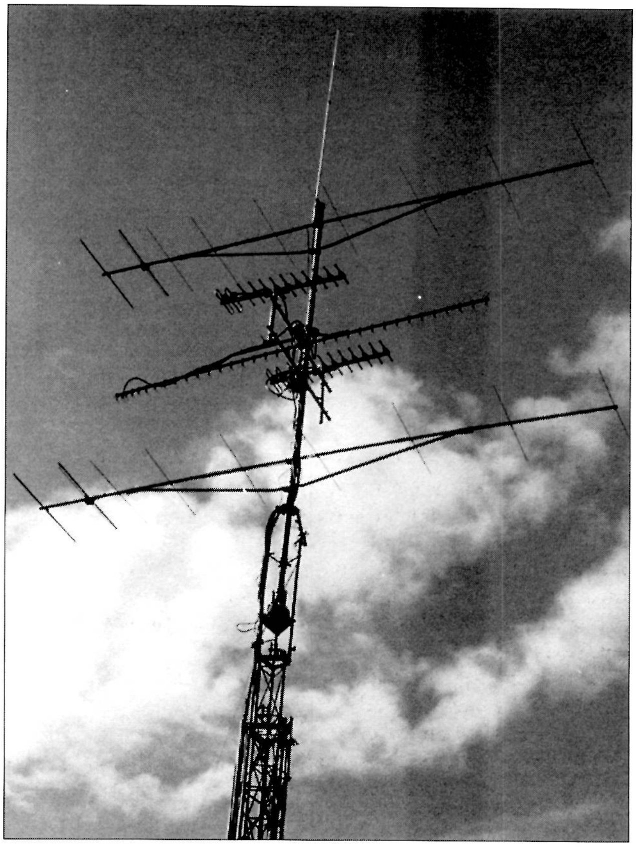


Fig 2.4. Typical VHF/UHF antenna system featuring two 2m, two 70cm and one 23cm beam

frequencies then some form of vertically polarised antenna will be required. This may be simply a vertically mounted dipole, a simple ground plane or a collinear array. For those who need to get a bit further in one particular direction then a beam may be required. The photograph (Fig 2.4) shows the antenna systems in use at my house.

The biggest and best antenna that can be put up is the one to go for. A careful look at antenna design will give guidance on which one will suit your particular location. But initially look for one that has a good forward gain and a good front-to-back ratio. This means that it radiates most of the signal towards the front and little from the rear. This front-to-back ratio is important as sometimes you may have a strong station behind you and by turning the beam a little you may be able to null them out. (Again, check the antenna chapter for more information.)

On VHF and UHF, although most will start out with just one beam, the wish to add another may come later, especially on UHF where many stations use two or more. A pair will increase the forward gain from the antennas by as much as 2.5dB. It has other advantages too. The receive capability will increase too as the gain of the antennas works both ways. If they are *stacked*, that is one mounted above the other, the radiation pattern will be narrowed in the vertical plane. This may help stop any potential EMC problems.

Having bought the antenna is not the end of it all – some means of turning it must be found. The local TV shop may well have a cheap rotator for sale in the window but these are usually intended for small TV beams only. They may last a year or more with a small VHF beam but it would be much better to get a proper rotator capable of handling the loading. The instructions that come with the antenna should tell you its wind loading. The rotator used should be capable of handling this.

How is it all going to be held up in the air, though? The dream of most amateurs is the 60 or 80ft tower with a cradle of antennas bunched at the top. Of course, some of us can get these up, but often neighbours and the local planning authorities don't see towers as much of a thing of beauty as we amateurs do. Many will use a chimney lashing kit to bolt a short stub mast to the chimney. The small rotator carrying the beam will be safe with this. But just for safety, how about using a double lashing kit and making sure, especially if a pair of antennas are intended?

If you do intend to erect a pair of beams, it is no good siting them a foot apart – they must be set the correct distance apart to get this 2.5dB gain. Check out Chapter 5 – 'Antennas and transmission lines' for more on this subject.

Antennas don't have to be out in the open of course. Many estates have restrictions on mounting antennas on houses above the eaves or the ridge line. All is not lost. Many operators use a small rotatable beam in the loft. If a rotatable one cannot be fitted, consider a pair of delta loops at 90° so you can select one or the other. The radiation pattern will be off the side of each. Not perfect but again a signal will be radiated.

Many amateurs use scaffold poles to mount the antennas. One easy way to do this is to beg, borrow, or buy two poles, one of about 2m and one of 7m in length. The shorter one is concreted into the ground and used as the support for the longer one. A hinged scaffold clamp is used and thus the whole can be lowered for work to be carried out on the antennas. If guy ropes are used even two poles maybe used. See Fig 2.5. Even unwanted windsurfer masts can be pressed into service but these should be used with care when brackets are bolted on.

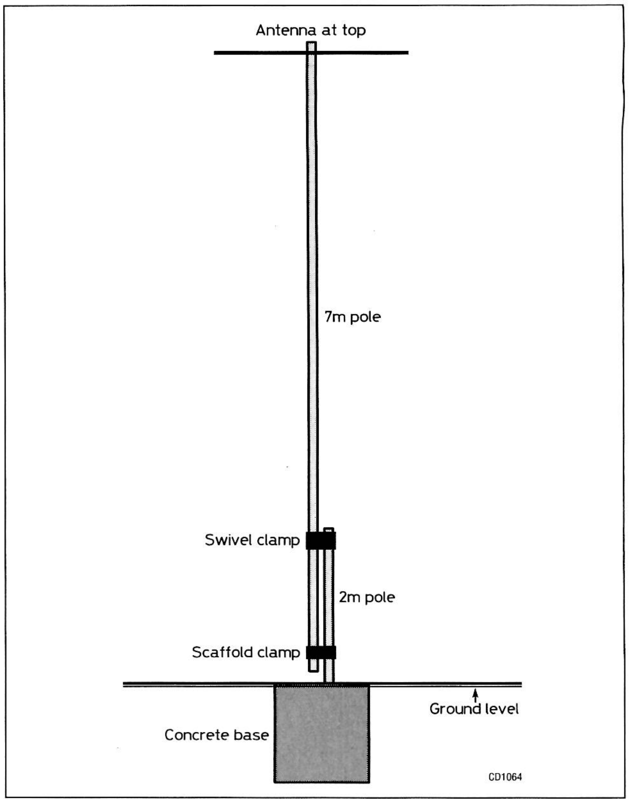


Fig 2.5. Using scaffold poles to mount antennas

FEEDER

There is no point in putting a good-quality vertical antenna up in the air and using a coaxial feeder such as RG58U. This feeder has a loss of 4.65dB per 100ft at 100MHz, so if your transmitter gave out 1W, and assuming no other losses, only about 200mW would be radiated from the antenna. By changing to a better feeder these losses can be reduced by a huge amount. The 'common or garden' RG213 has a loss of 1.9dB at 100MHz per 100ft, about a third of that of the RG58U. See Chapters 5 and 12 for more data on coaxial cables.

The benefit of choosing good-quality feeder is obvious. There are many types available on the market today, ranging from the very expensive Heliax type to the cheaper, but still very good, Pope H100 and Westflex. Remember that, although the transceiver gives you the results of the signal and the beam radiates it, it is the feeder that gets it to the rig and the connectors that join them together.

The station must be thought of as a whole, not as separate parts. If one part fails, the whole will fail.

OTHER EQUIPMENT

Several other items of other equipment will be found in most amateur shacks. Some of it will be essential, some may be just useful. Some may be borrowed from friends if they are only required from time to time.

One essential item when putting the antennas up is either a directional power meter or an VSWR (voltage standing wave ratio) bridge/meter. The intention of these is to measure the effectiveness of your antenna(s). Our requirement is that the minimum amount of power is returned to the transmitter. The directional power meter can be used to measure this actual amount of returned power. The VSWR (usually abbreviated to just 'SWR') bridge/meter will enable the user to measure the ratio of power levels in the system. Usually we try to tune for an SWR of unity or 1:1. This is often unobtainable, so anything under 2:1 will be acceptable.

The power supply to drive all this amount of equipment will be an essential item. The small 5A CB power supplies found cheaply are often not regulated to the extent we require. I have seen many trip into 'over-voltage supply', causing problems to the transceiver.

It may seem an extravagance but I would recommend buying the biggest power supply you can afford. It should be capable of delivering a very stable 13.8V under load. It is no point having a supply that 'sags' to 11V on load. A good one that will deliver 20A will suffice until a bigger one comes along.

Another essential is an accurate clock or watch. The type that relies on Rugby MSF time clock signals for accuracy is an excellent one. Some operators even have a pair in the shack, one with CET (Central European Time) and the other with local time.

The standard microphone supplied with the transmitter will almost certainly provide good audio on the transmitted signal but a base (free-standing) microphone may prove beneficial and permit some hands-free use. The log book and notes can then be completed during the contact. Later on a complete headset (earphones and incorporated microphone) may be used with a foot switch to give totally hands-free operation.

Initially the only Morse key will be the hand key required for the Morse test. These can be used to provide good-quality Morse code for long stretches but this takes practice. A good electronic keyer with a twin paddle can provide easy sending without fatigue over long periods. The only thing that moves with the keyer is your two fingers. With the hand key the whole arm from elbow to fingers moves. Remember, when wiring your paddle and learning to send, that convention has it that the thumb sends the dots and the finger the dashes!

The transceivers we have discussed so far have been run 'barefoot', ie without the addition of any form of amplifier to boost the transmitted or received signal. There are three types of amateur when it comes to amplifiers:

1. Those with a 'big mouth and no ears' (big linear transmit amplifier and poor reception facilities)
2. Those with 'big ears and no mouth' (no big transmit amplifier but perhaps a masthead preamplifier). This is a much better configuration.
3. Finally, there are those who have both a 'big mouth and big ears'. These are the full contest-style stations who have possibly full legal output power available with multiple antennas and a masthead receive preamp.

Many commercial linear power amplifiers incorporate a preamplifier. While these will help a little with low signal levels they also tend to amplify the noise on the feeder too. The best place to fit the preamp is at the masthead where it amplifies the signal at the antenna and passes it to the receiver.

Most commercial linear power amplifiers are excellent but several will have been 'got at' by previous owners. Always check the signal quality of your new amplifier before going on the air and *never try to overdrive it.*

WHERE TO PUT IT ALL

So, the antenna(s) are up in the air and we have a pile of other bits in their boxes – where in the house is the best place to operate from? This will very much depend on the space you have available and the amount of equipment you have accumulated.

For the VHF and the UHF station feed length may be a little critical and the attic a good place to consider. Remember though that if it is in the attic sleeping children may hear your signals and during the summer it may get very hot.

Many operators use a spare room for the shack; others prefer a garden shed. Wherever you put your equipment make sure it is safe from burglars and away from small hands when not in use.

Remember, if you plan to operate for any length of time, that the operating position should be comfortable. There is no point in entering a contest when you have to sit awkwardly. Set up the desk so that each item is within reach without stretching. The most-used items such as the tuning control should be closest to hand.

OPERATING TECHNIQUES

For ease of use the whole of each amateur band is split into segments, and a list of these can be found in the *RSGB Yearbook* and other publications. On VHF and UHF in the UK we keep the bottom section for moonbounce (or *EME* as it is known from 'Earth-Moon-Earth') where amateurs try to bounce signals off the Moon back to Earth. The next section is for dedicated Morse code users. There are separate sections for those interested in meteor scatter (MS) operation using Morse code and another for those using MS SSB. The next section is for those using SSB in normal terrestrial modes. A small all-mode section is followed by the packet section, the beacon subsection and the FM simplex channels and finally the satellite section. There are also sections for amateur TV (ATV), QRP activity, and many more. Let's look a little more closely at a couple of these. (Check the current *Yearbook* for more details of the band plans.)

In the UK the main 'calling frequency' for FM work on 2m at present is known as 'S20' (simplex channel number 20). It is on 145.500MHz (on 70cm it is 433.500MHz and is currently known as 'SU20'). Operators should use the established calling frequencies to only make initial contact and agree another frequency to move to (QSY). Don't stay and chat on the calling frequency even if the band appears dead. Keep all conversation on calling frequencies to an absolute minimum.

The band is at the moment split into 25kHz sections so that 144.475MHz is known as 'S19', 145.525 is 'S21' and so on. However, a change to 12.5kHz spacing is currently taking place and the names for channels are being revised. For 145MHz the letter 'V' is used, and the number is the channel, counting in 12.5kHz steps, above 145.000MHz. So 145.500MHz ('S20') is now being called 'V40'. For 435MHz the letter 'U' is used, and the number is the channel, again counting in 12.5kHz steps, above 430MHz. Thus 433.500MHz (formerly 'SU20') is 'U280' in the new system.

In some areas such as London it may be difficult to find a clear channel but in the outer regions it will be easy.

On SSB, the calling frequency is 144.300MHz and this has often been the cause of contention. In the old days, it was the norm to call CQ and state that you were 'tuning high (or low) for a contact'. Modern rigs have very stable VFOs and this practice has now vanished.

Most UK and European mainland amateurs listen on 144.300MHz (50.150MHz, 432.200MHz) for an opening or the band to liven up. Again, if this frequency is used to establish a contact then a move of frequency to complete the contact should be made immediately away from 144.300/432.200.

We have covered part of the band plans and learnt to move away from the calling frequency, but what actually is a valid contact? If I hear a station calling CQ such as:

"CQ CQ CQ, HB9FAP, over"

and I answer with:

"HB9FAP, you're 59, roger? "

He then replies with:

"Roger roger, you're also 59 here, name Fabio, QSL?"

Is this a valid contact? No, of course not – let's look at this once again, only this time done properly . . .

"CQ CQ CQ, HB9FAP, over"

"HB9FAP, you're 59 from G0BPS, over"

" G0BPS from HB9FAP. Many thanks, you're also 59. My name is Fabio. G0BPS from HB9FAP, over"

In this case both callsigns have been given by both parties so there can be no confusion. In the first instance he may have been answering another station, not you. In the latter there is no mistake at all who he is talking to. A valid contact is where both callsigns have been exchanged and also signal reports in the usual RS(T) configuration. In some cases, especially on VHF and UHF, the designated worldwide locator should be given too. This comprises a set of two letters, two numbers and two letters. For example, my locator is JO01oc.

Contests are a great way to make some longer-distance contacts on all bands, but before jumping in listen for a while and find out just what information is being exchanged. On some contests the exchange of region may be required, in others your county or even your age. In most, though, it will be a signal report and a serial number. Your first contact in the contest will be 001 (zero zero one) and increase one by one from there. In most VHF/UHF contests the worldwide locator will also be required. You may well hear something like:

"DJ5VE from G0BPS, you're 59085 in JO01OC. QSL?"

and followed by "73 and good luck".

Above all, listen before jumping in – you'll be VERY unpopular if you ask the other station what information he requires if he has a pile-up going.

Never, never . . .

There may also be the occasion where you may hear a station calling CQ but in a specific direction. On the VHF/UHF bands you may hear a Dutch station perhaps (or even me) calling "CQ Oscar Kilo or Sugar Papa, CQ Oscar Kilo or Sugar Papa" (calling Czech Republic or Poland). It means that the station wants to speak to Czech and Polish stations *only*.

I have often heard newly licensed stations hearing this type of call and then calling the Dutch station. They do *not* want to hear from you – if they did they would be calling CQ G! The station calling a specific CQ as above *only* wants a call from that specific area. If he wanted just anyone he would call CQ (a general call for a contact), or CQ DX for further afield. On the VHF and UHF bands DX is relative – for those who have never worked further afield than the next county, the next country will be DX. However, for those who can regularly work 400–500km on a flat band, only a 500km+ contact will be considered by them to be DX.

The basic rule is *never* to call a station that is calling a specific CQ *unless you are in that country or area he is calling*. If you do answer them from a closer area, expect the proverbial 'flea in the ear'!

Often the station calling CQ OK or SP etc will be 'big', ie running full legal power to several rotatable antennas, masthead preamp, quality feeder etc. Your 10W to a five-element or so antenna does not make you DX, but if he slips up and just calls CQ you can justifiably call him.

Remember a band devoid of signals does not mean that the band is flat. It just means that there are no operators – often there are people just listening for someone like you to call CQ. So, a short CQ call may well bring forth a station that may surprise you.

One of the most-misunderstood controls on many rigs is the 'clarifier'. This control is often referred to on HF as the 'RIT' (receiver incremental tuning). Its purpose is to enable the operator to move the receive frequency away from the transmit frequency.

Often we hear two newly licensed stations meeting and agreeing to move to, say, 144.330MHz. Both change frequency so that the dial on their rig shows this frequency and call each other. It is most unlikely that the true frequency of each rig exactly matches the other so the two stations are not *netted* (on the exact same frequency). In with the clarifier . . . and they can hear each other now. We now have in effect a duplex contact with both receiving on the other's transmit frequency which differs to their own.

The simple answer is to agree who will call who when you arrive at the agreed frequency. The other person should listen and use their ears to get to the frequency that makes the other station sound right. Ignore the tuning dial. It may well show '144.332' but if both stations avoid using the clarifier they will both be on the *same* frequency and be operating in a simplex mode or, as we say, 'netted' to each other. Best of all, ignore the clarifier except for very exceptional occasions.

Talking English

Having established a contact you will wish to exchange some information. Remember that if you bumped into a stranger at a party you would introduce yourself perhaps with: "Hello, my name is Dick and I live in Folkestone". How many times do we hear on the air "Personal here is Harry, QTH is Lyminge" or similar? You wouldn't talk at the party like that so why on the air? Use plain English and avoid the royal 'we', eg "We are running 100 watts . . ." when the operator is alone in the shack. The use of Morse abbreviations on the air should also be avoided – they were designed to speed up CW communications and should remain there. The exception to this rule, perhaps, is the use of 'QSL' when referring to the contact-confirmed card!

When passing the contact to the other operator do so clearly – listeners may want to know who is there, especially if you have hooked a rare one. "HB0QS from G0BPS back to you" or similar will suffice. If you are both a clear, strong signal with each other, ie 57 or so, then you may wish to avoid phonetics, but if conditions are difficult it may be beneficial to both to give full callsigns in phonetics. Stick to the proper ones too. The oddball ones may help in rare occasions but most operators' ears are tuned to the correct ones and can pick them out of the noise.

The pile-up

If you happen to hear a huge amount of noise on one frequency it will be almost certain that a rare station has appeared on the bands. This can provide you with two options. Sit on the frequency and shout your call in amongst all the others and hope, or . . . cheat!

Tune around the edges of the pile-up listening carefully as you go. Very often the loudest rare station will have another from the same area who is not so strong but is also audible. Those in the pile-up won't hear them, but those tuning the edges will snap up the rare one and grin as they pass the noise again.

If you do decide to go for the pile-up the first thing is to listen, listen and then listen. Unless you are running a 'big' station the only option is to get all the information you need such as callsign, locator and possibly their name first. When you have this information you can try for the contact.

Many times we hear an operator in the pile-up calling the DX station, getting answered and then asking for his callsign first. Oh dear . . . We say listen and listen again – is the DX station working simplex or on *split frequencies*? (That means calling CQ on one frequency and listening a little higher, or lower, in frequency.) If they are working 'split' *never, never* call him on his transmitting frequency – the wrath of all and sundry will fall upon you if you do. A transceiver with twin VFOs is essential if the other operator is working 'split'. Never try to work them by tuning between the two frequencies – it just won't work.

Having found his listening frequency, which may even be a small section of the band. you may then call him. If you are competing with some high-power stations then skill, not power, will be required. Listen for the style of his operating. Yes, I know this requires a lot of listening, but I get more hits than misses with my calls by listening carefully first. Having established his rhythm give your callsign quickly and just once without giving his. It will be assumed that all are calling the DX station. Try and find the gap between all the others. I know this will be difficult but that is the reason for listening so much first.

If the full call doesn't work try just giving the last two letters of your callsign – use phonetics and call twice in a gap. "Papa Sierra, Papa Sierra" is my way. Often the DX operator will call back "Papa Sierra again". You can then give his callsign and yours, his report and your locator and name. Do not give your station details etc unless they are asked for. You will get his information and then maybe just "73 . . . QRZ?" without waiting to see if you got the information. A good operator will do a check though and often give his callsign, locator and QSL manager where appropriate.

QSL cards are often exchanged to confirm contacts. It is not essential that you send a card for every contact made. Often you may see written that "a QSL is the final courtesy of a QSO". If you want a QSL card to confirm the contact, say so, asking politely, and you will probably get one. However, if you have worked that square many times you may not want it again, so say so. It is not essential to exchange cards for every contact – they should be treated for special use only, not as a matter of course for each contact.

Many stations that hold or use rare callsigns may prefer to use a QSL manager – this is usually a friend who is willing and able to handle the large amount of cards arriving for the DX operator and enable them to spend their time on the air rather than filling out cards. If a manager is given it is usual to use that route.

If a card is required by the direct route then it is usual to enclose one or two IRCs (International Reply Coupons) which are available from the Post Office. You may instead prefer to use a 'green stamp' – the name used by all amateurs throughout the world for a one US dollar bill. Needless to say, from one UK station to another a stamped, self-addressed envelope (SSAE) will usually suffice.

G0BPS

G0ROO & M1ABJ

Is pleased to confirm a two way QSO with

Located in JO01OC, WAB TR23 51°05'50"N 01°10'50"E
Seaview, Crete Road East, Folkestone. CT18 7EG. UK.

Dick Pascoe G0BPS

Date (dd/mm/yy)

Time UTC

Mode: CW/SSB/

Band MHz

Ur RST

Pse / Tnx QSL
Via Buro

To:

TRx IC275 or

Ant 2x / 4x 9 Ele Vargarda

Power : 10/25/100/400 W
170m ASL, 15m AGL

73 de Dick

Fig 2.6. Front and rear of typical double-sided QSL card

QSL cards can be a pleasure – I have received many delightful ones in full colour but most are on thin card and quite plain. This is fine because I only need to have confirmation of the new square – it doesn't have to be a work of art. Talk to the members at your local club and ask them what they have on theirs.

Check out the adverts and send for samples. These will give you a fair idea of what you may wish to see on yours. As a minimum it should show your callsign, name, your locator, your power used, QSL PSE/TNX. You may also wish to add your address, your WAB information and your latitude/longitude. Simply add any information that you think the other operator may find interesting. See the copy of mine shown in Fig 2.6.

Single-sided cards are fine – double sided ones will cost much more. Check out your local printer too. They will often prove to be quite competitive when compared to those advertising in the magazines.

Finally, the best advice I ever heard for the newcomer to the bands is: "Listen, listen and then listen some more".

REPEATERS

A large number of repeaters are scattered across the UK and these are usually on the 2m and 70cm bands. Repeaters are just a way of getting your signal a little further. They are typically set on a high site with a range dependent on the others in the area. Each repeater is on a different frequency to those others nearby. They will require either a short 1750Hz tone burst of usually about 400 milliseconds. Most also have a CTCSS tone access. This is a sub-audible tone that 'opens' the nearby repeater for you to make a call. Each one will have

AMATEUR RADIO STATION LOG

DATE	TIME (UTC) start	TIME (UTC) finish	FREQUENCY (MHz)	MODE	POWER (dBW)	STATION called/worked	REPORT sent	REPORT received	QSL sent	QSL rcvd	REMARKS
2 Nov '88	0800	0810	3	J3E	20	GMSABC	59^{+10}	59^{+5}			Bert
"	0811	0820	145	F3E	16	G7XYZ	57	56			Terry first G7
"	0825	0830	14	J3E	20	CQ					No reply
"	1725	1735	145	F2D	16	GB7XYZ					Local packet mailbox
"	1740		Station closed down								
4 Nov '88	1030	/P	from 73 Antenna Lane, Squelch-on-Sea								
"	1031	1036	50	J3E	10	G1ØXYZ	55	56			Jim, Bridgetown
"	1036	1045	50	J3E	10	G7XYL	58	58			Anne, Nr Squelch-on-Sea. QRM
"	1205	1215	433	F3E	13	G2XYZ	46	47			
"	1220	Station closed down									
5 Nov '88	0945	/P	from 73 Antenna Lane, Squelch-on-Sea								
"	0950	1005	144	J3E	16	GB2GUY	56	56	✓		Catherine Fawkesville
"	1010	1015	144	A1A	16	GD5ZZZ	542	541	✓		QSB! QSL via WF9XYZ
"	1526	1530	144	A1A	16	G7CW	579	589			Good keying!
"	1535	Station closed down and dismantled									
7 Nov '88	1810	1902	435	C3F	10	G7ZZZ	P3	P3			Ted, first ATV contact!
"	1930	1945	21	J2B	16	VK2ABC	559	569	✓		RTTY, Sid at Bandedge
"	1946	2005	21	J2B	16	ZL3ZZZ	569	559	✓		1st ZL on RTTY
"	2010	Test for TVI/Harmonic Radiation - Nothing noted									
"	2020	Station closed down									
8 Nov '88	1735	1737	7	J3E	20	CQ					
	1738	1805	7	J3E	20	GIØZZZ	58	58			Nobby - chatted about G5RV ant
	1930	1945	51	F3E	10	GØSIX	55	55			Allen, wanted WAB ref.
	1950	Station closed down.									
NOTES											

Fig 2.7. Typical hand-written log page

a different CTCSS tone to stop operators opening those repeaters not required.

European repeaters in the 2m band use a 600kHz offset between the transmit and receive frequencies, and in the UK the user transmit frequency is the lower of the two. For example, the repeater outside Folkestone in Kent is GB3KS (Kent South) with an input frequency of 145.025MHz and a repeater transmit frequency of 145.625MHz. On 70cm the offset is 1.6MHz and on 23cm it is 6MHz except for ATV where it is variable (see the current *Yearbook* for details).

Repeaters were first set up to enable mobile stations to have contacts but of course base stations can use them, too. However, it is courteous to let any mobile station use take preference over base station use.

Often there will be gatherings of users in the mornings and evenings as amateurs travel to and from work.

There will be times during enhanced conditions when it may well be possible to chat through other repeaters. During one memorable opening a UK station was chatting to a Danish station through a UK repeater. Often stations in the Midlands will be able to access Southern ones. The use of CTCSS will help to stop this, though.

So far we have discussed terrestrial repeaters, but the amateur also has access to satellites that are solely for amateur use. These sometimes require two radios. Check the chapter on satellites for more information on these.

Finally, you may also hear the request for a QSL card for a repeater contact. You may of course exchange cards but they will not be accepted for any awards or contests. Why? Because you have been in contact with a repeater, not the other station direct.

MOBILE OPERATING

For those who do not have the facilities for an operating position in the house a parking spot with a good take-off can provide hours of fun. Just because you are operating from a vehicle doesn't mean that you are restricted to FM only. Lots of fun can be had with the mobile whip on the gutter when using SSB. Yes, we know that the preferred polarisation for SSB is horizontal, but the difference between vertical and horizontal is only about 3dB.

What about your mobile whip? Will it lift out of the base and hang sideways, horizontally? Many do, and by rotating the whip around the base you can have a slightly directional antenna. Yes, it is cheating but it does work – try it!

Before I had an antenna on the roof of the house I would often take to the hills with a small beam antenna on a short pole attached to the side of the car and the roof rack. I would point the beam and call CQ either on FM or SSB from my old FT290R. Only 2.5W out but I could have hours of fun. I well remember taking to the hills at one of the better VHF/UHF spots in a country lane for one of the QRP contests, never dreaming that I would some years later buy a house only a few hundred yards away on that same lane.

LOGGING

Before the advent of computers, and their acceptance by most, all logging was done on paper. Many would keep a rough scrap of paper beside the operating position and make notes as time went past. They would then write up the log in a nice neat hand after the operating session was over.

A paper log must not be loose leaf, must be indelible, and must contain a certain amount of information. The minimum requirement is that you record: date, time, callsign of station worked or called, CQ calls made, frequency and the mode of transmission used.

Many, if not all, operators prefer to add to this essential list by including the reports exchanged, the power level used, the name of the other operator, and whether a QSL card has been sent and/or received. Fig 2.7 shows a typical written log page.

Since the advent of computers the rules have been relaxed and the log book may be kept on the computer. I use the hard disc drive for my main log and also every time I shut down the log program I save a copy to floppy disc. I even use two discs and alternate them. Fig 2.8 shows a computer printout log.

Apart from the essential information required by our licences we can add any information we want to our logs. Where possible, I always keep a note of names – it is so nice to hear a station, check the log for their name and call them with a "F6PBZ from G0BPS. Hello Mike, how are you?"

Amateur radio station GOBPS   Page 1

QSO	DATE	UTC	MHz	MODE	CALLSIGN	RSTs	RSTr	LOCATOR	REMARK	QSL
1091	25.07.96	2103	144	SSB	LC3LAT	59	59	JO59HJ	BJORN	S-
1092	25.07.96	2105	144	SSB	LA7TJA	57	57	JO59FG	JAN	S-
1093	25.07.96	2121	144	SSB	SM6VQW	59	59	JO67FA	OLA	--
1094	25.07.96	2126	144	SSB	SM6USS	59	59	JO67AT	DENNIS	--
1095	25.07.96	2200	144	SSB	SK6HD	59	59	JO68SD	FRED	S-
1096	25.07.96	2203	144	SSB	OZ1ABA/P	59	59	JO55LR	BOB	S-
1097	25.07.96	2208	144	SSB	DG2LBF	55	59	JO54BH	BERND	--
1098	30.07.96	2037	144	SSB	EA2AGZ	59	59	IN91	------------	S-
1099	30.07.96	2044	144	SSB	EA1DDO	57	57	IN53UI	MAX	S-
1100	30.07.96	2047	144	SSB	EA1OS	55	53	IN53TI	AL	--
1101	30.07.96	2117	144	SSB	EA1DKV	59	59	IN53TJ	JOSE	S-
1102	30.07.96	2121	432	SSB	EA1DKV	55	55	IN53TJ	JOSE	S-
1103	30.07.96	2205	144	SSB	EA1BCB	57	58	IN63ID	SENEN	S-
1104	30.07.96	2254	144	SSB	EA1DDU	55	55	IN73FM	MICK	--
1105	30.07.96	2255	144	SSB	EB1EWE	00	00	------	LOST	--
1106	30.07.96	2300	144	SSB	EB1HAL	53	59	IN63UN	------------	S-
1107	30.07.96	2313	144	SSB	PAOGHB	59	59	------	GERARD	--
1108	30.07.96	2333	144	SSB	PAOGHB	59	59	------	GERARD	--
1109	30.07.96	2321	144	SSB	EA1OS	59	59	------	TRYING 70CMS	--

FASTLOG 3.2   Signature ____________________

Fig 2.8. Typical printout from a computer log

The particular opening shown was a great tropo opening covering most of Europe. In my listing will be seen the number of the contact, date, time in UTC, frequency, mode, callsign, report sent, report received, the worldwide locator of the other station, his or her name and the QSL status.

This program that I use is widely available for the PC on a share-and-enjoy basis from I0XGR and can be found from many software sources such as Venus Electronics. There are many other sources of logging software and each will be slightly different. There are also many programs specifically for contest logging and these may also be found at the various amateur software sources.

3 Propagation

INTRODUCTION

In the early 'twenties communication engineers decided that wavelengths of less than 200m were useless for serious message handling – so they generously gave them all to amateur operators. This was an action they were soon to regret, for by the end of 1923 two amateurs using a wavelength of about 100m had carried out two-way exchanges with another in France. The following year, transatlantic working was becoming quite commonplace down at 100m, using considerably less power than the 'big boys' were needing on their medium-wave circuits.

The authorities acted swiftly and predictably. They called a conference. At the end of it the amateurs found that they had lost all of their gift apart from a sequence of harmonically related bands – 80, 40, 20, 10 and 5m.

Amateurs began to progress their way through the list of assigned bands, tackling each in turn when it became practicable by the state of the art. Their professional counterparts were surprised to find that each time they seemed to work longer ranges with lower powers, all the way to the 10m band. Beyond that were 'ultra short waves', and there their luck appeared to run out. The 5m band really did seem to be a desert, with little prospect of any DX working at all.

Then came the war. Amateur activity ceased but many future amateurs were fortunate enough to acquire a solid grounding in radio communications while serving in the armed forces, where they had been able to use up-to-date (and expensive) commercial equipment rather than the customary pre-war home-brew.

A lot of that sort of equipment came on the market after the war at knock-down prices that people could afford. Many post-war amateurs began their activities using modified war-surplus gear and they were eager to find out just what it was capable of doing, given a free rein. At that time everyone started off with an interest in propagation, whether they admitted to it or not. A good proportion of those took it very seriously indeed.

In the UK, television swallowed up the old 5m (56MHz) band. So, in a way, everyone venturing on to VHF and UHF after the war was starting with a clean slate. In the years that followed they discovered that many of the features of propagation that they were taking for granted were things that rocked the foundations of previously held textbook theories. They were working far greater distances than their limited power should have allowed; they worked paths crossing mountain ranges that, at VHF and UHF, should have been blocked completely by the terrain; they were finding that the ionosphere could, on occasion, reflect signals of up to at least 200MHz – about four times higher in frequency than had been expected – and amateur experiments had revealed an unexpected mode of propagation, trans-equatorial, that had stimulated a wide range of interest, particularly among broadcasters. They had projects covering a variety of other modes, including auroral, meteor-trail, moonbounce and propagation via field-aligned irregularities, all active on a truly international scale and all producing results which would have been unobtainable by other means.

As a result the Amateur Service enjoyed a rewarding relationship with professionals in the field of propagation research. But, although the two sides were working towards the same end, their reasons for doing so were poles apart. To an amateur a 'tropo opening' is like a heaven-sent reward for eating up all his spinach, whereas to a professional it signals a period of frustration, when co-channel interference creates interruptions in his data flow, patterning or breaking-up of his television signals and dents in his reputation for reliability. Despite that, both groups were interested in knowing such things as: "When?", "For how long?", "Where?", "To what extent?" and those were just the sort of questions that amateurs found themselves in an ideal position to answer.

Sadly, those golden periods of close co-operation are becoming less frequent nowadays. There are basically three reasons for that. One is that the professional research interest has moved beyond the VHF and UHF parts of the spectrum to much higher frequencies, well outside the province of this handbook, where point-to-point working involves extremely narrow pencil beams. Our large and dense networks of potential observers can contribute little to that sort of situation. Secondly, many of the former problems have been avoided by the routing of commercial traffic through satellite transponders. Thirdly, professional and commercial organisations are cutting back on expenditure to such an extent that research requirements are currently being shelved unless they can show an immediate financial return against the investment supporting them.

That last consideration affects our activities also. To analyse our own original signal records in terms of cause and effect we need an ongoing supply of solar and geophysical data from official sources. In the past we have been very fortunate in being able to get what we need either on an exchange basis or through various 'old boy' networks. But now, in many cases, such information has to be paid for and some of the rates are extremely high. As a result, amateurs are gradually being priced out of individual research. In that we are not alone; amateur meteorologists face a similar situation and frequently voice their frustration in that respect.

The bottom line of this is that when it comes to citing current sources of information this propagation chapter has had to be much less precise than we would have liked. It may well be that, during the time that this book is kept on sale, the situation will improve and new sources will become available. At the time of writing certain rumours abound, but nothing is certain. But do not let that stop you from undertaking research of your own. When the time comes to do something with it you can always contact the RSGB for some up-to-date advice on where to get the information you need.

But that, of course, is jumping rather far ahead. At this stage you will be wanting to see what VHF/UHF propagation is all about . . .

OUR FIELD OF INTEREST

The term 'radio wave propagation' really covers two objectives. The first is to determine the nature of the mechanism involved in getting signals from point A to point B, the second to explain the route in terms of physical quantities. To do this may require some knowledge of meteorology, the Earth's magnetic field and even of events taking place on the Sun. There will be one or two close encounters with mathematics, but surely nothing more complicated than is needed in order to gain a pass in the Radio Amateurs' Examination.

In the context of this handbook we are supposed to be concerned only with those amateur bands that fall between the lower limit of VHF (30MHz) and the upper limit of UHF (3GHz). But the various modes of propagation do not sit easily within those confines and, at some time, you should take a look at the wider picture that includes frequencies below (HF) and above (SHF etc) where similar principles apply but the end results may be very different. For example, there are several references to the ionosphere in this chapter but we shall not be concerned with the part it plays in everyday worldwide communications.

Long, long ago, the Ancients believed that the world had been formed on the top side of a disc, carried, it was said, on the back of a giant turtle. From the propagation point of view that would have simplified things considerably. Many of our problems stem from the fact that we live on the surface of a sphere. Electromagnetic radiation (which includes, among other things, visible light and radio waves) travels in straight lines so, if their rays are to be persuaded to follow the curvature of the Earth, they must get themselves bent by some means in order to achieve any significant distance and remain near to the ground. Fortunately for everyone concerned with radio communications, Nature has thoughtfully provided four alternative ways of getting beyond the horizon: reflection, refraction, diffraction and scatter. We shall see how they operate later on.

As you turn the pages you may be surprised to find references still to the 'old' form of QTH locator. That is not something left over from the last edition of the *VHF/UHF Manual* that ought to have been edited out. The so-called 'squares', 1° of latitude by 2° of longitude, which happen to be common to both the old and the current systems, represent a network of areas which are ideally sized for the needs of propagation research. Although the IARU Maidenhead locator (which is worldwide in coverage) has replaced the QTH locator in most logs by now, many research workers continue to use the two-letter designators when it comes to report storage and analysis. There are two good reasons for that, one being that only four characters serve to define a path instead of eight, the other that it is easier to familiarise oneself with a single grid of 26 × 26 lettered squares than it is to deal with

Table 3.1. Conversion between the old locator system and the IARU locator for main squares

Longitude (first QTH locator letter)

1st IARU letter	1st IARU figure 0	1	2	3	4	5	6	7	8	9
I	—	—	—	—	U	V	W	X	Y	Z
J	A	B	C	D	E	F	G	H	I	J
K	K	L	M	N	O	P	Q	R	S	T

Latitude (second QTH locator letter)

2nd IARU letter	Second IARU figure 0	1	2	3	4	5	6	7	8	9	
M	Q	R	S	T	U	V	W	X	Y	Z	South
N	A	B	C	D	E	F	G	H	I	J	
O	K	L	M	N	O	P	Q	R	S	T	
P	U	V	W	X	Y	Z					
P							A	B	C	D	North

Examples:

1. To find QTH locator equivalent to IARU locator JN18. Enter longitude table with first letter (J) and first figure (1) to find first letter (B). Enter latitude table with the second letter (N) and the second figure (8) to find the second letter (I). Required locator is BI.
2. To find IARU locator equivalent to QTH locator GP. Record indicated IARU first letter and the first figure (J-6-). Find second letter (P) within the boxed section of the latitude table. Record second letter and second figure (-O-5). Combine. The required IARU locator is JO65. Refer to the text for the use of the letters outside the boxed section.

six or seven grids of 10 × 10, even though the areas covered turn out to be much the same.

The propagation chapter in the RSGB *Radio Communication Handbook* contains tables showing how to convert between either the old locator or the new and latitude and longitude (or vice versa, of course), and nowadays there are computer programs which perform the same task. Here, what is most often required is a simple conversion between the two locator systems at the basic square level and that is provided by Table 3.1.

A word of caution, however. For locations at latitudes below 40°N (roughly the heel of Italy) and above 66°N (the north end of the Gulf of Bothnia) there is an ambiguity because the lettered squares repeat. But that is easily resolved by reference to the callsign of the station concerned. No trouble if you are doing it manually, but a point to watch if you entrust the job to a machine.

RECOGNISING VHF/UHF MODES OF PROPAGATION

At frequencies above 30MHz (following the definition of the terms 'VHF' and 'UHF') propagation by the regular layers of the ionosphere takes place but rarely and then generally only around times of maximum sunspot activity.

The usual mechanism governing the day-to-day performance between two Earth-based stations has its origin in the lower part of the atmosphere, at rarely more than 4–5km above the ground. *Tropospheric propagation* is descriptive of this mode and the fundamental properties of the air which have the most influence are the vertical distributions of temperature and water vapour, both of which tend to decrease with height and, in so doing, cause elevated radio rays, such as might otherwise escape into space, to bend back down towards the ground, and to reach it beyond the normal visible horizon. At times, when dry warm air overlays cool moist air, usually in the presence of an anticyclone, ranges extend dramatically and signals from distances up to about 2000km may be expected. At the same time the strength of nearer signals may be enhanced, effects which extend throughout the VHF and UHF parts of the radio spectrum. During a *tropo opening*, as it is often called, signals generally rise slowly, accompanied by a progressively slower rate of fading. At peak strength, fading may be absent altogether. A long period of enhancement generally ends when a cold front reaches one end of the transmission path.

Tropospheric scatter depends on the presence of small-scale refractive index irregularities and dust or cloud particles in a relatively small volume of the atmosphere towards which both the transmitting and the receiving antennas are directed. High power is required at the transmitter and good signal-to-noise performance at the receiver. Scattered signals are weak, spread in frequency by up to 1kHz either side of an unmodulated carrier, due to the differing motions of the scattering particles, and several rates of fading may co-exist, often giving the impression of a rough modulation. The rate at which intelligence may be sent is limited by *blurring*, introduced by the range of signal path transit times possible within the upper and lower limits of the scattering volume.

At the top end of the UHF band *atmospheric absorption* effects become noticeable, for beyond 3000MHz, in the SHF part of the radio spectrum but outside the scope of this book, *attenuations* due to oxygen, water vapour and precipitation (rain, snow etc) become increasingly important. These affect not only transmission paths that are wholly within the troposphere, but paths originating within and terminating without – ground to satellite, EME etc – although there the effects tend to diminish with increase of beam elevation as the length of that part of the path which contains the absorbers and attenuators decreases.

Although many textbooks still imply that the ionosphere has little effect at VHF and above, a number of very important events have their origin there. Nearly all of them are associated in some way with the level around 100km above the ground, which is generally occupied during the day by the regular E layer.

Of these the most important is *sporadic-E,* which radio amateurs have studied particularly at 144MHz for many years, despite the fact that its presence there, according to our professional colleagues, ought to be impossible. In 1980 the Amateur Service was invited to contribute to a symposium on sporadic-E held at the Appleton Laboratory, and it was clear that at that time the amateur activities concerning this mode of propagation came as a surprise to many of the distinguished authorities present. It is now acknowledged by them that such a mode does exist at frequencies that may exceed 200MHz for short intervals of time, although the feeling is that it may not be sporadic-E at all but an entirely different mechanism as yet unidentified. In this chapter, it will still be referred to as sporadic-E (or E_s) until such time as its true identity is discovered. VHF sporadic-E signals generally begin suddenly and unpredictably (hence their name), bring in stations from distances of 1000–2000km at excellent strength and clarity for periods of up to several hours, and then, with a rapid decline, they cease. The duration of an opening decreases with increasing radio frequency, the higher frequencies starting later and finishing sooner than the lower ones. During the event the locations heard gradually progress from one area to another. Sporadic-E events at VHF within Europe are generally confined to the months of May to August.

Operators living in southern Europe make use of another VHF mode, which depends on the presence of *field-aligned irregularities* (FAI) in the distribution of free electrons in the ionosphere at E layer heights (around 110km). It has a similar seasonal variation to VHF sporadic-E but differs from it in that signals do not follow the direct path between stations but appear to originate from a scattering volume which is often situated near to the Swiss Alps or close to other mountainous areas at about the same latitude.

Another ionospheric mode is associated with the appearance in the northern sky of the aurora borealis (or 'Northern Lights'), which is caused by the interaction of streams of charged particles from the Sun with the Earth's magnetic field. Signals reflected from the very mobile *auroral-E* curtains, which usually accompany visual displays often seen in the Northern Isles and the north of Scotland but less frequently further south, are readily recognisable with their characteristic tone, variously described as "rasping", "ringing" or "watery", and the fact that beam headings for optimum signal strength are commonly well to the north of the great circle path joining the two stations in contact.

Short-lived trails of ionisation due to the entry into the Earth's atmosphere of small particles of solid matter (seen at night as shooting stars) can be responsible for *meteor scatter,* where two stations, usually widely spaced, can establish contact in intermittent bursts ranging in duration from several seconds down to periods which afford little more than occasional 'pings' of signal. Meteor-scatter signals should be looked for at times of meteor showers, which are listed later in this chapter. Duration of meteor reflections and their frequency of occurrence decline with increasing frequency. Meteor-scatter propagation has been used professionally at operating frequencies of between 30 to 40MHz for communication purposes.

Trans-equatorial propagation is usually confined to paths in which transmitter and receiver are situated approximately equal distances either side of the magnetic equator (eg the Mediterranean area and Zimbabwe). 144MHz openings seem to require high solar flux and low geomagnetic index; frequency spreading is apparent at 144 and 432MHz, with flutter fading, often giving the signals a quality similar to that of signals reflected from the aurora. On the Zimbabwe-Cyprus path openings were centred on 2000 local time at Cyprus. It is believed that extensions to TEP via E_s or tropo may be possible.

For many years any involvement of the F layers of the ionosphere with TEP was disputed but it now seems likely that some, if not all, of the extreme ranges that have been recorded may have come about as a result of double reflection at those heights, without intermediate contact with the ground.

Without question there is an involvement of the *regular F2 layer* at 50MHz around the peak period of the solar cycle. There should be no difficulty in recognising such signals because they ought to bear all the characteristics of normal DX working at HF. At the appropriate time the likelihood of

Table 3.2. Working frequency bands of various VHF and UHF propagation modes (megahertz, unless shown otherwise)

Aurora	50, 70, 144, 432
F2 layer	50
FAI	144
Meteor scatter	50, 70, 144, 432
Moonbounce	50, 144, 432, 1.3GHz, 2.3GHz
Sporadic-E	50, 70, 144
TEP	50, 144
Tropospheric	50, 70, 144, 432, 1.3GHz, 2.3GHz
Troposcatter	70, 144, 432, 1.3GHz, 2.3GHz

FAI = field-aligned irregularities in the E layer
TEP = trans-equatorial propagation

regular layer propagation at 50MHz and the paths concerned should be signalled in monthly ionospheric prediction tables.

Diffraction is a mechanism that is associated with signal paths that cross sharp mountain ridges (it is sometimes referred to as *knife-edge diffraction*). At the ridge a small degree of bending occurs, acting in the direction towards the ground. It is the likely reason for the ability of near-mountain stations to work out over seemingly impossible paths. However, it may be difficult to rule out assistance by tropospheric refraction, particularly for places where the mountain crest supports a blanket of snow which may be undergoing sublimation, that is, going directly from the solid state to vapour. As a rule, signals may be considered to have been diffracted when they have travelled along a path which has crossed a mountain ridge and similar contacts have been possible between the same two stations on a fairly regular basis.

Table 3.2 shows a summary of the bands in which the various VHF and UHF modes play a part in propagation.

TROPOSPHERIC PROPAGATION

The propagation of light

It may be found helpful to begin this study of tropospheric propagation by considering first some comparable aspects of the propagation of light. In most cases the analogy is a close one because radio and light are both forms of electromagnetic radiation, differing only in wavelength (or its inverse, frequency). However, light has the advantage of being readily detectable by its direct action on one of our senses and most of us have had many years of experience working with it. We do not usually think of a torch bulb as being a transmitter, nor our eyes as being receivers but they are nevertheless, and all the perturbing effects to which a radio wave is subjected within the troposphere have their visual counterparts with which we are very familiar already.

A beam of light normally travels in a straight line unless something is done to deflect it. This can be brought about by *reflection,* as in a mirror or from the surface of a still pond, *refraction,* when light passes from one medium to another causing a straight rod in water to appear to be bent, or by *scattering* as from the dust in a shaft of sunlight. Certain frequencies can be made to suffer *attenuation* by inserting one or more filters in the path of the beam, and a very important filter which occurs naturally is provided by a layer of ozone in the upper atmosphere which prevents harmful amounts of ultra-violet light from destroying life on Earth. Mist and fog are visible counterparts to attenuation and scatter.

It will be seen later that most tropospheric radio events of any importance are manifestations of refraction. In terms of light it is refraction which provides the lens with its well-known properties, whereby light leaving one medium, such as air, and entering another, such as glass, suffers a deflection. A Dutch scientist named Willebrord Snell discovered in 1621 that the sine of the angle made by the incident ray with respect to the normal, divided by the sine of the angle made by the refracted ray with respect to the normal, was a constant for a given pair of media. The property possessed by each of the materials involved is known as the *refractive index,* and Snell's constant (sin *i*/sin *r*) *is* equal to the inverse ratio of the refractive indices of the two media.

Changes in refractive index also occur in the atmosphere, due to variations in density, usually as a result of the juxtaposition of two unmixed layers differing greatly in temperature, or due to the presence of a steep gradient of temperature within a single layer. This is the origin of the optical *mirage.* When air near the ground is heated, as over hot sand in the desert or sometimes along a straight road, a line of sight directed downwards is refracted upwards, giving an unexpected (and usually unsuspected) view of the sky which appears as a shimmering pool some distance ahead. Conversely, where cool air underlies warm air a line of sight directed slightly upwards is bent down, so that objects which are in reality well beyond the normal horizon appear to be on it, or even above. There was a famous occasion in 1798 when the whole of the French coast from Calais to Dieppe became visible one afternoon from the cliffs near Hastings.

Effects such as these are even more pronounced at radio frequencies because the radio refractive index contains a term which is dependent upon the amount of water vapour present, and this is a parameter which is subject to considerable change in the lower atmosphere in both space and time.

The radio refractive index of air

There are two basic methods used to determine the refractive index of air; one is to measure it more or less directly using a device called a *refractometer,* the other is to derive it from other, more readily accessible, measurements of atmospheric functions.

Refractometers are beyond the scope of the radio amateur. They are usually airborne or tethered balloon-borne devices constructed and operated by large research organisations. They depend on the fact that the resonant frequency of an open microwave cavity is a function of the dielectric constant of the air within it, and that this is also a function of refractive index.

The more common method is to use upper-air soundings of pressure, temperature and humidity provided by meteorological services all over the world, generally on a twice-daily basis, at midnight and midday GMT. This information is obtained from cheap and simple balloon-borne telemetry devices called *radiosondes,* which have been in regular use since shortly before the Second World War.

The radio refractive index of the air, symbol *n,* is a quantity which is only very slightly higher than unity, but the difference between, say, 1.000345 and 1.000300 is all-important in propagation studies and may have a profound effect on the path of a radio wave. To bring out this importance, and to simplify subsequent calculations, it is usual to subtract 1 from

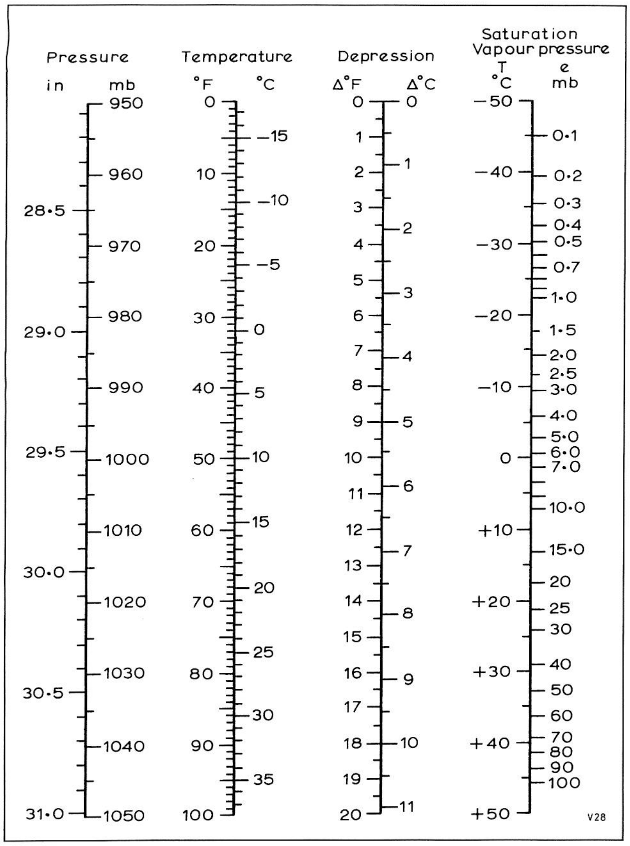


Fig 3.1. Conversion scales based on the following expressions: (a) Atmospheric pressure: inches and millibars. 29.53 inches Hg = 1000mb. (b) Temperature degrees Fahrenheit and Celsius. (°F) = 9/5 (°C) + 32. (c) Depression of wet bulb or depression of dew point. (Δ°F) = 9/5 (Δ°C). (d) Saturation vapour pressure, *e*, in millibars, given the ambient temperature in °C. See p3.6 for an expression for *e*

the refractive index value and then multiply the remainder by one million. This quantity is given the symbol N; in mathematical terms $N = 10^6(n - 1)$.

Before demonstrating how N values may be calculated from meteorological data it will be advisable to define the units involved, and, in some cases, to show how they can be obtained from measurements made at home.

METEOROLOGICAL UNITS

Pressure

According to international agreement the current unit of pressure is the hectopascal (hPa), which is equivalent to a force of 100 newtons per square metre. However, the United Kingdom Meteorological Office (who provide most of the weather observing and forecasting services in this country) have shown themselves to be strangely reluctant to adopt the name 'hectopascal' in their dealings with the public, preferring to stick with the millibar, which has exactly the same value. To avoid confusion that practice will be followed here but at some time in the future you may have to start getting used to the new name. Your home barometer probably still has a scale that is calibrated in inches, which is a relic of much earlier days when air pressure was measured by balancing against it a column of mercury and reporting its height. The units are related such that 29.53 inches of mercury are equivalent to a pressure of 1000 millibars or 1000 hectopascals. For ground level values Fig 3.1(a) provides a rough conversion. Whole millibars are sufficiently precise for most propagation purposes.

Pressure decreases with height in an approximately logarithmic manner. Near the ground the rate of change is about 1mb in 10m, but this should not be presumed to extend over too great an interval because the relationship is actually a function of temperature also.

In meteorological studies it is customary to use pressure rather than height as a measure of vertical displacement and it will be found very convenient to carry over this practice into propagation work, because the physical processes of the atmosphere are a function of pressure, not of height, and any attempt to make them otherwise will complicate normally convenient relationships beyond belief. It requires some adjustment of ideas, not the least being that height is traditionally measured upwards from the ground, whereas pressure is measured from the top of the atmosphere downwards. But the radio wave, once launched on its way from the transmitting antenna, encounters nothing that can be identified directly with height. It 'sees' changes in air density and refractive index, which are themselves functions of pressure, temperature and water vapour content. Height, as such, is not one of the natural properties of the atmosphere, and that is why aircraft altimeters, which appear to measure it, have to be set to read zero at sea level before the pilot attempts to land, for they are really barometers carrying an approximate scale of feet or metres instead of an accurate one in millibars.

Very roughly indeed a pressure level of 900mb may be considered as being equivalent to a height of 1km and the 700mb level as being approximately 3km. Exact equivalents in respect of a given place and time form part of the basic meteorological data used in analysis work.

Temperature

In scientific work temperatures are generally expressed in degrees Celsius (°C, formerly known as Centigrade) or in kelvin (K). Strictly, kelvin are degrees Celsius plus 273.15 but for our purposes the constant may be rounded off to 273 in order to keep the working figures as whole numbers.

Relative humidity

This is a measure of the amount of moisture present in a sample of air, expressed as a percentage of the total amount which could be contained at the given temperature. It can be obtained from the readings of two identical thermometers, one of which has its bulb surrounded by a muslin wick moistened with distilled water.

They should be well-sited in the shade, and preferably enclosed in a properly ventilated screen. The difference between the two readings is the *depression of the wet bulb,* and the percentage relative humidity can be found from Fig 3.2. If the thermometers are calibrated in degrees Fahrenheit it is better to convert their difference using the scale of Fig 3.1(c) than to find the difference of two converted figures.

Dew point

If a sample of air containing a given amount of moisture is allowed to cool it will be found that the wet-bulb depression

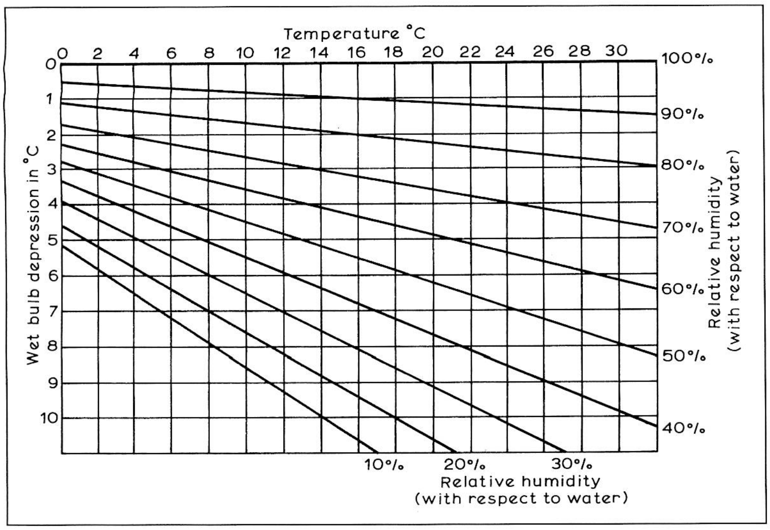


Fig 3.2. Percentage relative humidity as a function of temperature and wet-bulb depression

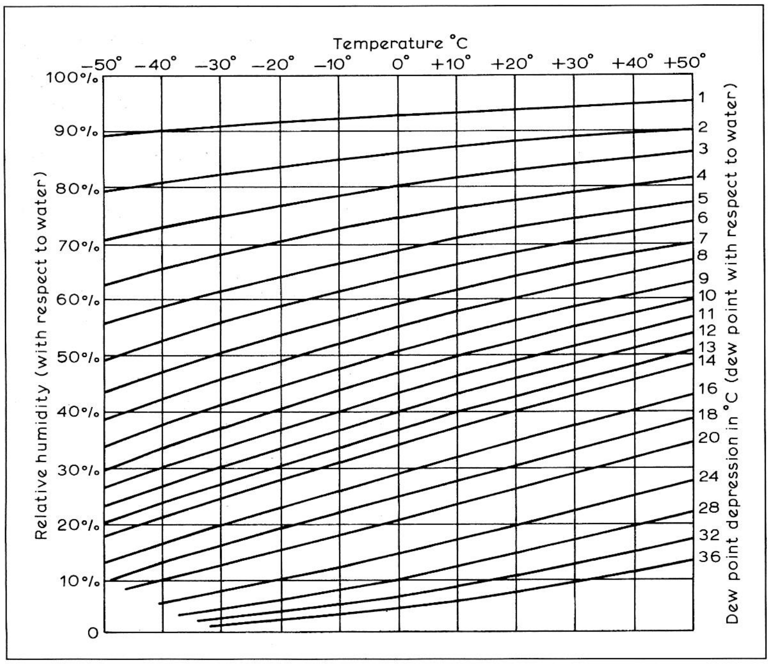


Fig 3.3. Percentage relative humidity as a function of temperature and dew-point depression (may be used to convert radiosonde data published in the 'wrong units')

decreases until eventually both wet and dry bulb thermometers read the same. The relative humidity will have become 100% and the air is said to be *saturated.* The temperature at which this occurs, the *dew point, is* therefore another way of expressing the amount of water vapour contained in a sample of air. Most upper-air reports nowadays show this as *dew-point depression,* the difference between the dry-bulb temperature and the temperature to which the air would have to be cooled in order to reach saturation, but some still refer to percentage relative humidity. The chart, Fig 3.3, can be used to make a conversion either way.

Vapour pressure

The water vapour present in a sample of air exerts a contribution of its own to the total atmospheric pressure. The scales of Fig 3.1(d) show saturation vapour pressures corresponding to a wide range of temperatures, but it has to be admitted that the scale is a difficult one to interpolate. The relationship between temperature and saturation vapour pressure is a complex one and in the past most calculations have involved the use of tables. Recently, however, a number of organisations have tried to find an acceptable approximation, making use of an expression which is within the capabilities of a 'scientific' pocket calculator. The following, which is due to Parish and Purtnam of NASA, has been used elsewhere in this chapter for the machine calculation of refractive index:

$$e_s = T^{-4.9283} \times 10^{(23.5518 - (2937.4/T))}$$

where e_s *is* saturated vapour pressure in millibars and T is the air temperature in kelvin.

When the air is not saturated the appropriate value of vapour pressure, e, can be found from the relationship:

$$e = e_s \times u$$

where u is the relative humidity expressed as a decimal (eg 72% = 0.72), or, more usually nowadays, from the dew point T_d or the dew-point depression, D (where $T_d = T - D$), using the expression:

$$e = T_d^{-4.9283} \times 10^{(23.5518 - (2937.4/T_d))}$$

where T_d is in kelvin (°C + 273) which will be found to be the most practical form for use with a calculator or computer. Of course, the same trick of using the dew point to find the actual vapour pressure as if it was a saturated vapour pressure may be performed on the scale of Fig 3.1(d) to provide the required value of e.

THE CALCULATION OF *N*

The basic equation is:

$$N = \frac{77.6}{T}\left(p + \frac{4810e}{T}\right)$$

where p is the atmospheric pressure in millibars, e *is* the water vapour pressure in millibars and T is the air temperature in kelvin. It is often more convenient to expand this into:

$$N = \frac{77.6p}{T} + \frac{3.733 \times 10^5 \times e}{T^2}$$

because it then separates conveniently into a 'dry' term, corresponding approximately to the optical value of refractive index, and a 'wet' term which contains all of the contribution due to the presence of water vapour. The values which result from these expressions are known as *refractivities,* but they are often referred to simply as *N-units.*

The degree of ray bending which results from refractive index changes can be assessed by calculating the decrease

Table 3.3. Minimum duct thickness for the VHF and UHF amateur bands

Band (MHz)	λ_c (m)	Minimum thickness (m)	Approximate millibar equivalent
50	6.00	317	31
70	4.29	263	26
144	2.08	176	17
432	0.69	96	9
1296	0.23	52	5
2300	0.13	38	4

over unit height change. The normal gradient from the ground may be regarded as being approximately $-40N$-units/km. Should it become $-157N$/km the curvature of the ray becomes the same as that of the Earth, while gradients greater (ie more negative) than $-157N$/km result in *ducting,* where the waves travel for great distances, confined within a relatively shallow range of heights, suffering alternate refractions at the steep-lapse layer and reflections from the ground.

Provided that well-marked contrasts in refractivity exist above and below a ducting layer it is not essential for the ground to be involved at all. Once signals have been trapped in an *elevated duct* they may travel considerable distances without being receivable by stations on the ground below the transmission path. The waves eventually leak out of the duct at some point where the necessary conditions are no longer being fulfilled.

For efficient duct propagation the wavelength concerned must be less than a critical value λ_c, such that:

$$\lambda_c = 1.9 \times 10^{-4} \times D^{1.8}$$

where D is the duct thickness in metres.

Table 3.3 shows the minimum duct thickness and the approximate equivalent pressure difference in millibars, centred on 850mb, the pressure at a typical ducting height. The figures cover all the VHF and UHF amateur bands.

The following example of a calculation directly from basic meteorological data may be found useful:

p = 900mb, $T = -3$°C (= 270K), dew point depression = 8°C.

From this the dew point must be -11°C (= 262K) and the corresponding vapour pressure from Fig 3.1(d) is 2.6mb. Hence:

$$N = \frac{77.6 \times 900}{270} + \frac{3.733 \times 10^5 \times 2.6}{270 \times 270}$$
$$= 259 + 13 = 272$$

Table 3.4 outlines a skeleton program to calculate radio refractive index N from basic meteorological data entered sequentially for each of the available levels. It should be readily adaptable for any type of scientific programmable calculator or a computer. Whole-number answers are adequate for propagation studies and there is nothing to be gained by trying to make the results seem more precise than the data can support. Before undertaking extensive calculations make sure that the test figures yield the result shown.

Table 3.4. Computer or calculator program to obtain radio refractive index, N, from basic meteorological data

S = Store
R = Recall from store

Load stores with constants:
S1 = 273, S2 = 2937.4, S3 = −4.9283, S4 = 23.5518, S5 = 77.6, S6 = 4810

Input data for each level to be computed, pass to the indicated stores:

Enter p = pressure in millibars (or hectopascals). To S7 (p)

Enter t = temperature °C. Add R1 to convert to kelvin. To S8 (T)

Enter D = dew point depression °C. Subtract from R8. To S9 (T_d)

Program

1. Evaluate vapour pressure, e

   R9^{R3} * 10$^{(R4 - (R2/R9))}$ To S10 (e)

2. Evaluate radio refractive index, N

   R5 (R7 = R6 * R10 / R8) / R8

Round off the result to the nearest whole number.

Test

When p = 900, t = −3.0 and D = 8, then N = 272

Note: A step-by-step calculator program (based on the TI58/59 calculator but easily adapted to suit any similar programmable scientific calculator) may be found in the fourth edition of the *VHF/UHF Manual*, p 2.5. It works on the TI66 which may still be available.

CAUSES OF TROPO DX

Having established a method of obtaining refractive index values from standard meteorological upper-air observations it is a natural progression to apply that knowledge to a study of the atmosphere during a well-marked tropospheric 'opening' – probably the main reason why radio amateurs take an active interest in this mode of radio propagation. For that purpose, consider the situation late in the evening of 20 January 1974, when continental Europe was 'wide-open' to the UK. This is a good example of a notable winter event and, as will be seen, it has been used to illustrate various aspects of a single occasion, as is shown in Figs 3.4, 3.5 and 3.6.

Fig 3.4 shows a cross-section of the atmosphere up to 700mb (about 3km in terms of height), from Camborne in SW England to Berlin. The *isopleths* join levels having equal values of refractivity, scaled in N-units. There is no mistaking the concentration formed in the lower part of the diagram. This indicates a steep fall of refractive index with height and is in the correct sense to cause the return to earth of rays which would otherwise have been lost in space above the horizon. Superrefraction of this sort produces bending towards the earth in the case of both ascending and descending rays. Because there is a normal tendency for refractive index to decrease with height, this effect is nearly always present in some degree and this accounts for the fact that radio communication at VHF and UHF is usually possible beyond the visible horizon. The presence in the lower atmosphere of a layer in which refractivity decreases very rapidly with height, as in the case being considered, is always accompanied by enhancement of signal strengths and an increase in working range. However, in the case where very-narrow-beamwidth antennas are used at both ends of the path, received signal strengths may fall, due to energy being deflected away from a path which has been optimised under conditions of normal refractivity.

From a cross-section, such as Fig 3.4, it would be quite possible to calculate the probable paths of rays leaving a transmitting antenna at various angles of take-off, using Snell's Law, as with optical ray-tracing, but this is an exercise which

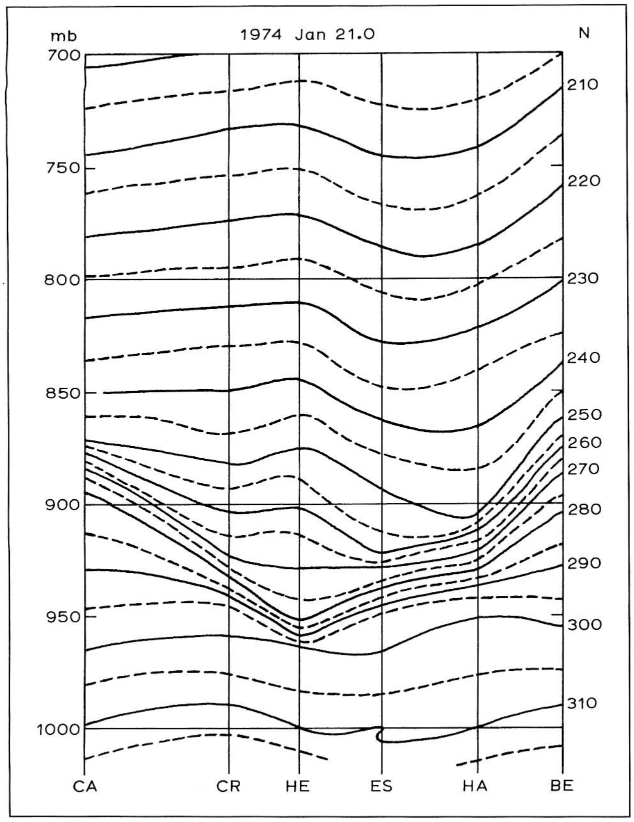


Fig 3.4. Cross-section from SW England to Central Europe at midnight (0000GMT) 21 January 1974, drawn in terms of conventional refractive index *N*. The vertical scale is in terms of pressure. 700mb = 3km approx. CA = Camborne, CR = Crawley, HE = Hemsby, ES = Essen, HA = Hannover, BE = Berlin

is probably outside the needs of most amateurs. It should be noted that the values of refractivity at ground level reveal little of the situation above. For that reason the only really effective study of tropospheric propagation phenomena involves the acquisition of upper-air meteorological data.

THE ATMOSPHERE IN MOTION

It does not require a great deal of experience on the VHF and UHF bands to realise that all the big 'openings' to the Continent occur during periods of high atmospheric pressure. Indeed, some amateurs look upon an aneroid barometer in the home as being their guide to the state of the bands. However, whereas good conditions are accompanied by high-pressure readings, high pressure is not always accompanied by good conditions. Why should that be? The answer lies in an appreciation of the role played by vertical motions in the atmosphere.

In general, rising air becomes cooler and moister, while descending air warms and becomes more dry. Air is sometimes forced into vertical motion by the topography in its path; it rises when it flows over hills and it descends into valleys. However, the present context mainly concerns vertical motion associated with the two main types of pressure system.

Consider first a low-pressure system or 'depression'. Air circulates round it in an anticlockwise direction (in the northern hemisphere), with a slight inclination towards the centre, creating an inward spiral which leaves progressively less room for the volume of air in motion. There is only one escape route available, and that is upwards. So low-pressure systems are associated with rising air.

On the other hand, anticyclones (high-pressure systems) are characterised by light winds blowing clockwise round the centre but with a slight deflection outwards. As the air spirals outwards fresh quantities must be available to maintain the supply and the only source is from aloft, resulting this time in a downward flow. So high-pressure systems have descending air associated with them.

Adiabatic changes

Air in vertical motion changes in both volume and pressure (they are directly related) and in temperature also, although there need be no gain or loss of heat. This may appear at first to be a contradiction in terms, for heat and temperature might be thought to be alternative names for the same thing. In fact, heat is a quantity which can be distributed either over a small volume to provide a large increase in temperature, or spread over a large volume to appear as a small increase in temperature. Thus 1kg of air descending from a height of 3km may begin with a pressure of 700mb and a temperature of –5°C, to arrive at 1.5km with a pressure of 850mb and a temperature of 10°C with no change of heat being involved. Such a process is *adiabatic,* and it is an important principle in meteorology.

A homely demonstration of it at work may be found in the case of the bicycle pump, the barrel of which gets hot in use due to the air inside having been compressed.

When the air is anything other than dry another apparent paradox links the amount of water vapour and the corresponding humidity during the adiabatic process. Going back to the example, at 700mb 3.78g of water vapour would have been sufficient to produce saturation (100% relative humidity) in the 1kg sample of air, whereas at 850mb the same amount would give only 41.5% relative humidity because air at 10°C could hold 9.1g of water vapour. So, air descending adiabatically gets warmer and drier, although the actual amounts of heat and water vapour remain unchanged.

The action is reversible but only up to a point. Ascending air is accompanied by increasing relative humidity, which at some stage will reach 100%. Any further lifting will result in the appearance of liquid water, which will appear either as cloud or larger droplets, which are likely to fall out of suspension as rain. When condensation occurs, the rate of cooling is altered by the appearance of latent heat, and the precipitation will alter the amount of moisture in the sample of air.

No such considerations affect descending air once its relative humidity has fallen below 100%, although there will have been alterations to the rate of change of temperature if liquid droplets of water have been evaporating, again on account of latent heat.

If the sample of air is taken adiabatically to a standard pressure of 1000mb the temperature it assumes is known as the *potential temperature* of the sample. It follows from this that potential temperature is a quantity which remains constant during any adiabatic change: conversely, a change is an adiabatic one if it is associated with constant potential temperature.

Potential refractive index

Referring back to Fig 3.4 it will be seen that, quite apart from the region of interest referred to earlier, there is a general background of fairly regularly spaced isopleths which represent the normal fall-off of refractive index with height. A number of modifications to the standard procedure for calculating refractive index have been proposed from time to time, all with the intention of minimising this effect, leaving emphasis on the features that are of most interest to the propagation engineer.

Opinions have varied on the best way to do this. Most methods proposed have involved some form of model atmosphere and the calculation of departures from it, resulting in complex exercises for which a computer is advisable. Another disadvantage has been the difficulty of recovering the original values of refractive index from the final data (should they be required elsewhere, or at a later date). The method to be described was first proposed in 1959 by Dr K H Jehn of the University of Texas, who does not seem to have taken advantage of the full potential of his suggestion. Curiously, little has been done outside amateur circles to exploit its usefulness; it involves a unit known as *potential refractive index* (K).

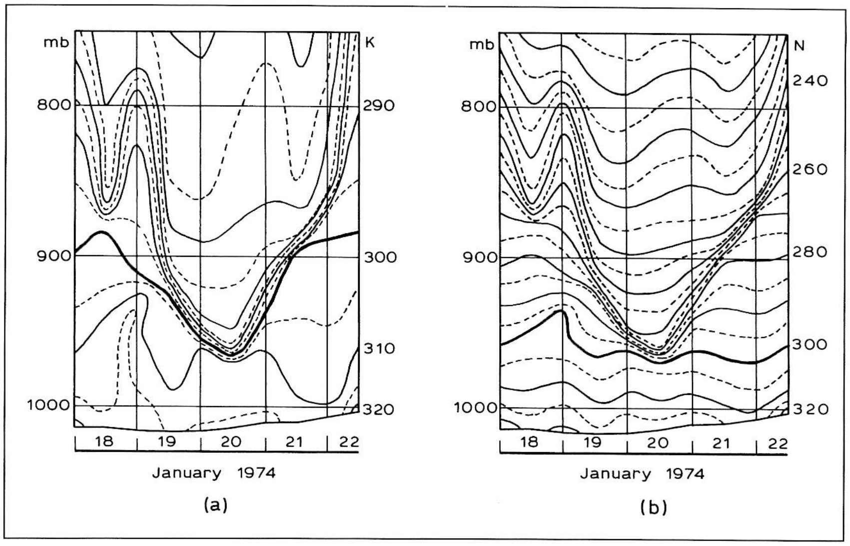


Fig 3.5. (a) Time-section showing isopleths of potential refractive index, *K*. Crawley, 18–22 January 1974. (b) Time-section showing isopleths of radio refractive index, *N*. Crawley, 18–22 January 1974

It may be obtained from upper-air meteorological sounding data in just the same way as has been described for N-units, the only difference being that each sample of air, whatever its true level may be, is presumed to have been transported adiabatically to a pressure of 1000mb before the calculations are made.

The advantages of this form of normalisation are considerable. By adopting a procedure which imitates the natural process of the atmosphere, applying, for example, to the large mass of air which subsides from aloft within an anticyclone, each level of air is effectively labelled with a value of potential refractive index which remains with it during any adiabatic change.

The effect may be seen particularly well in time-sections, such as that of Fig 3.5(a), which shows how the potential refractive index pattern varied from day to day at a single station, Crawley, over a period which included that eventful evening of 20 January 1974.

There is no mistaking the extensive tongue of warm, dry, subsiding air associated with an anticyclone and the steep-lapse refractive index layer built up where it meets the opposing cool, moist air underneath.

Towards the right and left edges of the diagram may be seen evidence of rising air which is associated with two depressions, which preceded and followed the period of high pressure. These potential refractive index isopleths are very sensitive indicators of vertical motion in the atmosphere, and the patterns on cross-sections and time-sections take on an interesting three-dimensional aspect when viewed in conjunction with surface weather charts.

It is interesting to compare the potential refractive index time-section of Fig 3.5(a) with the corresponding section drawn in terms of conventional radio refractive index, N, Fig 3.5(b). Note first that there are fewer lines on the potential refractive index diagram, indicating that the normal fall-off of refractive index with height has been considerably offset. At the steep-lapse layer, the concentration of isopleths has been greatly emphasised in Fig 3.5(a) but it is important to notice that this has not been at the expense of accuracy in indicating either the height at which the effect occurred or its vertical extent.

Because air undergoing adiabatic changes has been shown to carry its value of potential refractive index along with it, no matter what its level, it should not be surprising that the boundary layer across the whole of Fig 3.5(a) is formed of basically the same set of K-values irrespective of changes in pressure (or height). Fig 3.5(b) shows that the same is not true for conventional refractive index. This is not to suggest that the N-values are wrong, but rather to point out that they do not share this very useful attribute of coherence independent of height which appears in diagrams like these. That the same is true of cross-sections may be seen by comparing Fig 3.6 with Fig 3.4.

If values of atmospheric pressure are known (as they always are when radiosonde data have been used) a simple relationship exists between potential refractive index and N. This leads to the conversion chart shown in Fig 3.7, which may also be used as a plotting chart, having the property that an ascent plotted in terms of one of the units may be read off in terms of the other by using the appropriate axes. In this way the potential refractive index values may be converted to N-units for ray-tracing purposes, or compared with N-unit profiles produced elsewhere.

Alternatively, use may be made of the following expressions:

$$N = 0.00731 \times p^{0.712} \times K$$

and

$$K = 136.8 \times p^{-0.712} \times N$$

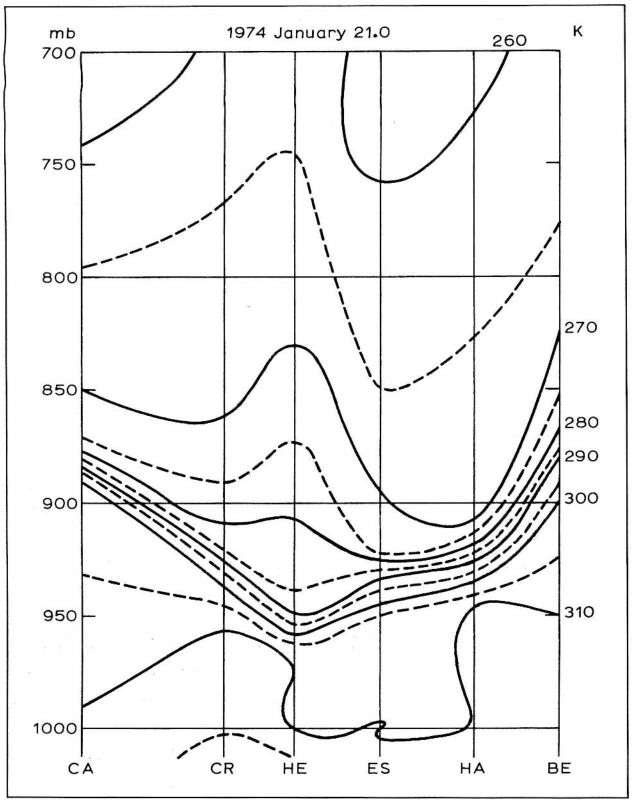


Fig 3.6. Cross-section from SW England to Central Europe, at midnight (0000GMT) 21 January 1974, drawn in terms of potential refractive index, *K*. Compare with Fig 3.4 and note here how the steep-lapse layer contains the same values along the length of the path

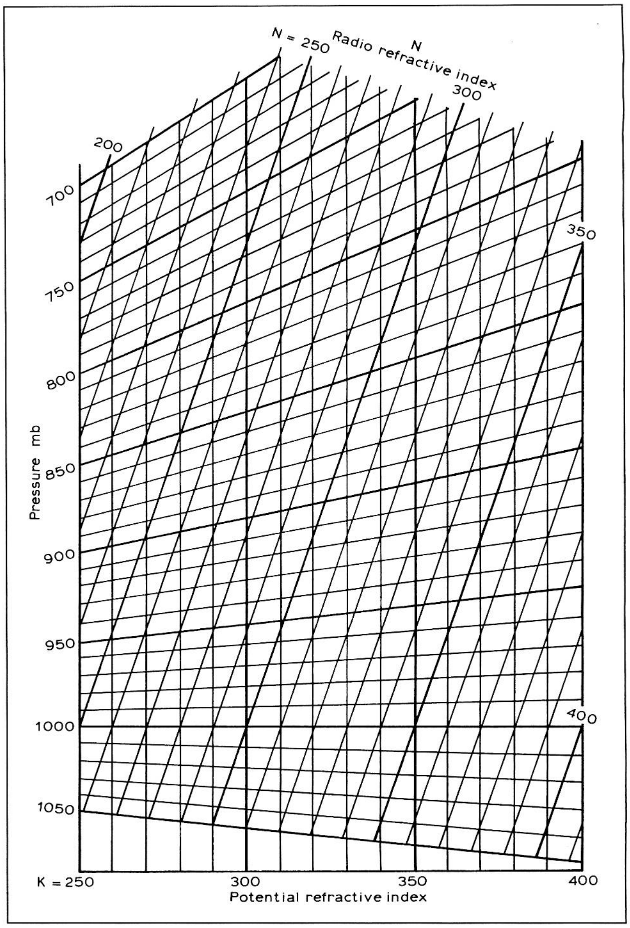


Fig 3.7. Refractive index plotting/conversion chart. Ordinate: pressure (mb). Abscissae: (vertical) potential refractive index, *K*;(slant) radio refractive index, *N*

which may be performed without difficulty on a scientific pocket calculator.

Although this conversion provides a way of obtaining potential refractive index it is usually better to calculate values directly from the radiosonde data. A method of doing that using a calculator or computer will be found later in the chapter, at Table 3.4.

ACQUIRING CURRENT UPPER-AIR METEOROLOGICAL DATA

A much-quoted Victorian lady, Felicia Hemans, began her poem 'Casabianca' with the words "The boy stood on the burning deck / Whence all but he had fled". Those two lines rather neatly sum up the situation as regards the present availability of upper-air meteorological data. Gone are the Morse, RTTY and facsimile broadcasts which used to keep us supplied with current information within an hour or so of the measurements being taken. Gone, too, are the printed daily records that appeared a few days later giving copies of the coded messages. And gone, never to return, are the microfiche summaries that would have served us well had they continued. All that remains for us on easy public access is a daily page of machine-plotted graphs on the back page of the *European Meteorological Bulletin*, published by Deutscher Wetterdienst who, as the name suggests, run the national weather service in Germany. It is to be hoped that the burning deck will continue to hold for some time to come, yet. It is not that the basic messages to which we need access no longer exist.

The trouble is that the whole field of meteorological communications has been depersonalised. Nowadays a machine puts the message together and passes it on to a computer, which launches it on to the global network, by which means it reaches other computers located all over the world. Then, instead of appearing in print for human beings to decode and plot, each message serves its purpose as a set of instructions fed to an XY plotter which not only turns it into a set of graphs but draws the graph paper around them. The message itself never sees the light of day. From the meteorologist's point of view that is progress. For anyone wanting access to the basic information it is a disaster.

The sort of data we are considering used to be freely available as a public service. Not any more. Weather is big business nowadays and the organisations operating in that field jealously guard their products, demanding a high price to retrieve information from the memory banks of their computer. In a letter published in the October 1995 issue of the Royal Meteorological Society's journal *Weather* (page 359) a research worker from the University of Bristol revealed that he had been quoted £35 as the price of supplying just a single copy of a radiosonde ascent.

Fortunately for us, however, copies of the *European Meteorological Bulletin*, mentioned earlier, may still be consulted free of charge at the National Meteorological Library, London Road, Bracknell. Spare copies are also held for loan purposes. If you are unable to attend in person it might be worth the cost of a telephone call (01344 420242, ask for the library loans desk) to see if they would be prepared to meet your requirements by post.

The Bulletin is available daily on subscription from the publisher, Deutscher Wetterdienst, Frankfurterstrasse 135, 63067 Offenbach, Germany. In summer 1997 the price was DM456, plus postage, for a year's supply – rather high a price if you only need the back page. But you would also get seven other pages containing weather maps in colour, some for the whole of the Northern Hemisphere, others for just the Atlantic and Europe, for sea level and several upper levels. Each page is A3 size.

The aerological diagrams are displayed in eight panels, each panel containing two diagrams side by side, making 16 diagrams in all. Each diagram shows machine-plotted graphs of temperature and dew-point depression plotted against pressure for up to three stations, identified by the use of differing symbols. The size of each diagram is such that temperature runs from –40°C to +30°C in 61mm, dew-point depressions from 0°C to 30°C in 26mm and pressure from 225mb to 1000mb in 76mm.

You may find this hard to believe but, with care, it is possible to estimate the plotted values to a sufficient degree of accuracy to be able to provide meaningful cross-sections and time-sections for propagation studies.

Frequently you will find that temperature and dew-point depression turning points do not occur at the same pressure level. Remember that you need all three values to calculate refractive index so you will have to use the plotted curves to interpolate where necessary.

The published selection of up to 48 stations (occasionally one or two are missed) provides excellent coverage over Europe, as may be judged from Fig 3.8, which needs to be studied in conjunction with Table 3.5. All the diagrams show soundings made at 12UT.

Fig 3.8 may be used to select the upper-air stations most appropriate to a given signal path, which could be overlaid using the latitude and longitude scales. It will be found an advantage to extend your cross-section beyond the strict limits of the path, if possible, because two additional soundings help in drawing in the refractive index lines.

The *European Meteorological Bulletin* appears to be all that we have left for upper-air data at the present time, unless, that is, you happen to be, or decide to become, a practising professional meteorologist.

Perhaps there is a way to get current sounding data off the Internet or the World Wide Web. If there is, and you are fortunate enough to be able to tap into it, you will find details of how to use the information later on in the chapter.

Why not use historical upper-air data?

Every researcher feels instinctively that he or she has to work with the latest information available. That is fine if what you need for analysis is easily come by. But when it is in short supply or priced as if it were gold dust it makes sense to consider a more practical alternative.

The monthly VHF and UHF report columns of *Radio Communication* and (especially) the quarterly European report sections in the German magazine *DUBUS* regularly contain details of unusual or exceptional signal events, and have done for many years. Very few of those events have been properly analysed. So, why not look back 20 years or so to a time when the radiosonde network was much more extensive than it is today and the information was readily available? Twice a day, 00UT and 12UT, or, if you go back far enough, four times a day: 00UT, 06UT, 12UT and 18UT.

Why not consider going back to two periods of intense scientific interest the world over – the International Geophysical Year, 1957 and the International Quiet Sun Years of 1963 and 1964? Extremely well documented records covering a wide range of disciplines still exist in scientific libraries. The likes of the efforts that were put into those two periods will never be seen again. Why not put them to good use?

By changing your objectives Mother Hubbard's cupboard could be replaced by Aladdin's cave. To parody a notice which used to be common in general stores before the war – if you don't see what you want in the window, come inside and ask for something else.

Discontinued sources of meteorological data

Until 31 December 1980, the source of data from nine British and Irish upper-air stations was the *Daily Aerological Record*, published by the Meteorological Office.

There was a companion series, dealing with surface observations, the *Daily Weather Report*, which gave six-hourly observations for each of about 50 places located in all parts of the British Isles.

Those two publications had a very wide circulation in their time and copies may still be available in some specialised libraries in various parts of the country. They, and similar publications from other parts of Europe, are certainly available at either the National Meteorological Library in Bracknell or in the Meteorological Office Archives, about half a mile distant. The documents may be consulted without charge and you will find the staff in both places very helpful. Most of the information you want will be on open shelves but you will probably have to ask for some of the continental upper air reports because there is no longer space to keep them in the room open to the public.

EXTRACTING THE DATA

If you are fortunate enough to be able to access the messages circulating on the global network you will need to know that they are headed by an alphanumeric indicator which reveals the type of information concerned and the country or area concerned. Upper-air messages are split into two or more parts, not all of which are of interest in the present context. Printed copies may be similarly split, but the identifiers may have been edited out by then.

The first message, headed with a prefix beginning 'US' (eg USUK for British stations, USFR for French stations etc) and/or by the group 'TTAA', relates to observations at specific levels of pressure. The station number is generally the second of the numerical groups.

Next, look for a group beginning 99. This and the one following are in the form:

99ppp TTTDD

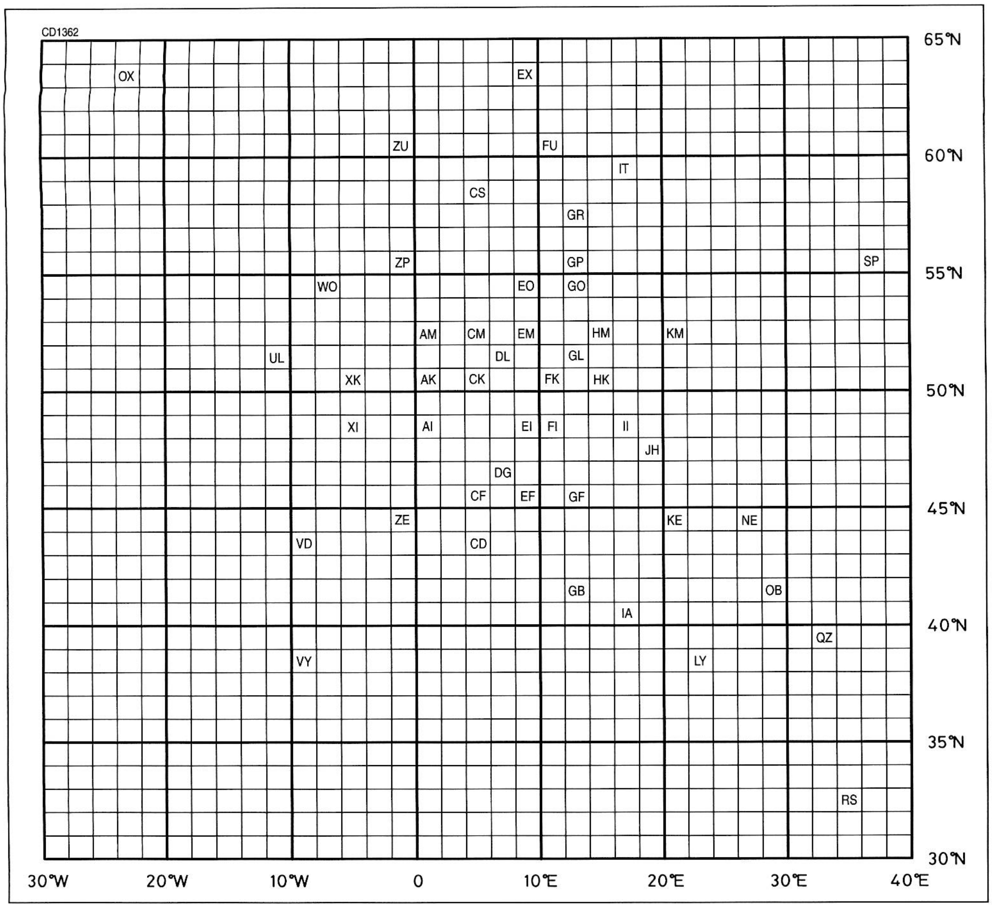


Fig 3.8. Chart showing the positions of meteorological upper-air stations relative to QTH locator squares. The stations may be identified by reference to Table 3.5

where '99' indicates that ground level data follows, 'ppp' is the pressure in whole millibars, with the initial 1 omitted for 1000mb and over.

'TTT' is the air temperature in degrees Celsius. If the tenths figure, the third one, is odd, the whole number is negative (ie 046 = 4.6°C, but 045 = –4.5°C).

'DD' is the dew-point depression in tenths of a degree up to 5°, then in whole degrees with 50 added (eg 46 = dew-point 4.6° below air temperature; 66 = 16° below).

Codes 51–55 are not used.

At regularly spaced intervals there will be further groups beginning 00, 85 and 70, indicators showing that the data which follows are for 1000, 850 and 700mb respectively. These groups and the ones which follow immediately have the form:

00hhh TTTDD . . . 85hhh TTTDD . . . 70hhh TTTDD

Table 3.5. Upper-air sounding stations currently (1997) featured in the European Meteorological Bulletin

QTH locator	IARU locator	Station number	Station name	Country	Height ASL (m)	QTH locator	IARU locator	Station number	Station name	Country	Height ASL (m)
OX	HP83	04018	Keflavik	Iceland	54	EX	JP43	01241	Ørland	Norway	7
UL	IO41	03953	Valentia	Ireland	14	FI	JN5B	10868	München	Germany	489
VY	IM58	08579	Lisboa	Portugal	105	FK	JO50	10548	Meiningen	Germany	453
VD	IN53	08001	La Coruna	Spain	67	FU	JP50	01384	Oslo	Norway	204
WO	IO64	03920	Long Kesh	N Ireland	38	GB	JN61	16245	Roma	Italy	21
XI	IN78	07710	Brest	France	103	GF	JN65	16044	Udine	Italy	53
XK	IO70	03808	Camborne	England	88	GL	JO61	10486	Dresden	Germany	232
ZE	IN94	07510	Bordeaux	France	61	GO	JO64	10184	Greifswald	Germany	6
ZP	IO95	03240	Boulmer	England		GP	JO65	06181	København	Denmark	42
ZU	IP90	03005	Lerwick	Shetland	84	GR	JO67	02527	Goteborg	Sweden	155
AI	JN08	07145	Trappes	France	161	HK	JO70	11520	Praha	Czech Rep	304
AK	JO00	03882	Herstmonceaux	England	52	HM	JO72	10393	Lindenberg	Germany	104
AM	JO02	03496	Hemsby	England	14	IA	JN80	16320	Brindisi	Italy	10
CD	JN23	07645	Nimes	France	62	IT	JN88	11035	Wien	Austria	209
CF	JN25	07481	Lyon	France	240	IT	JO89	02465	Stockholm	Sweden	14
CK	JO20	06447	Uccle	Belgium	104	JH	JN97	12843	Budapest	Hungary	139
CM	JO22	06260	de Bilt	Netherlands	15	KE	KN04	13275	Beograd	Serbia	203
CS	JO28	01415	Stavanger	Norway	9	KM	KO02	12374	Legionowo	Poland	96
DG	JN36	06610	Payerne	Switzerland	491	LY	KM18	16716	Athens	Greece	15
DL	JO31	10410	Essen	Germany	161	NE	KN34	15420	Bukarest	Romania	
EF	JN45	16080	Milano	Italy	103	OB	KN41	17062	Istanbul	Turkey	33
EI	JN48	10739	Stuttgart	Germany	311	QZ	KM69	17130	Ankara	Turkey	891
EM	JO42	10238	Bergen	Germany		RS	KM73	40179	Bet Dagan	Israel	35
EO	JO44	10035	Schleswig	Gernany	48	SP	KO85	27612	Moskava	Russia	156

Note: In the QTH locator column an underlined letter signifies that care is needed to avoid ambiguity.

In more recent years 925mb has been used as an additional standard level.

'hhh' is the height above sea level of the pressure level in metres, omitting the thousands figure. For 1000mb this becomes a negative number when the pressure at sea level is below that value, and this is indicated by adding 500 to the code figure (ie 675 = −175m). The missing first figure is 1 for 850mb and either 2 or 3 for 700mb, whichever puts the value closer to 3000m. 'TTTDD' has the same significance as before.

The second message is headed with an indicator beginning 'UK' (eg UKUK, UKFR etc) and/or the group 'TTBB', signifying that it relates to turning points in the temperature and dew-point profiles. It is the more useful of the two because it contains everything necessary for propagation studies, apart from the relationship between pressure and height for the particular ascent. As before, the station number is generally second of the five-figure groups in the message. To decode the remainder, point off succeeding groups in pairs that begin with the figures 00, 11, 22, 33 etc. The pairs have the form:

NNppp TTTDD

where 'NN' enumerates the data points. '00' always signifies local ground-level. 'ppp' is the pressure, in millibars, at the level of the observation, with the initial 1 omitted if the value exceeds 1000. 'TTTDD' contain the temperature in degrees and tenths and the dew-point depression, coded as before.

For most tropospheric propagation studies there is little point in going beyond the level at which the pressure has fallen to 700mb, unless it is to interpolate a refractive index value for 700mb in order to provide a uniform 'top' to a cross-section.

In a radiometeorological study it is quite likely that all the work will be carried out in terms of pressure rather than height, not only for convenience because that is the form adopted in the radiosonde messages, but because the radio wave, once launched, does not 'see' changes in height but rather changes in air density, a quantity closely related to pressure. In the atmosphere, height, which seems so easy to understand on the ground, becomes a complex function of the integrated effects of temperature and humidity, and of the value of pressure at station height.

There are two ways of finding the heights corresponding to the various pressure levels reported in the Part 2 message. The more accurate, though time-consuming, way is to plot the ascent data on a standard tephigram (obtainable from HMSO, where it is known at Metform 2810B) and then to follow the instructions given on the form. Alternatively, and this may well be accurate enough for the present purpose, refer to Fig 3.9, which assumes an average vertical distribution of temperature and dew point, leaving the height a function only of surface pressure. The diagram is used as follows:

(a) Find the station height from Table 3.3 and draw a vertical line at the corresponding value on the horizontal scale. (Crawley, at 144m ASL, which appears in many of the examples used in this chapter, including this one, is no longer operational. Its place for observations over south-east England has been taken by Herstmonceaux.)

(b) Find the point where that vertical line intersects a horizontal line appropriate to the reported value of ground level pressure.

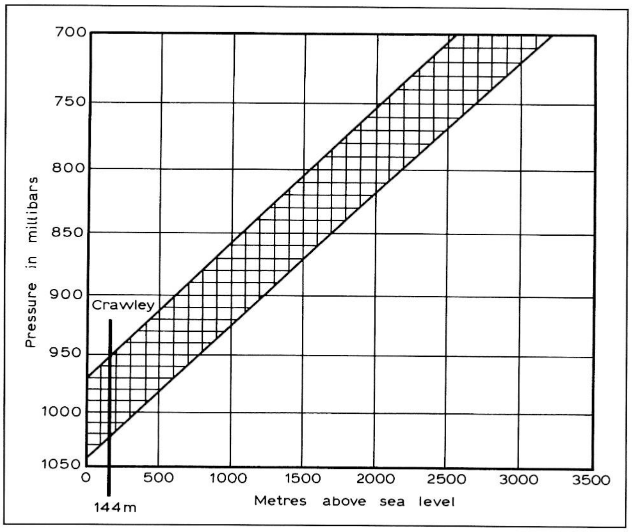


Fig 3.9. Relationship between pressure and height between the surface and 700mb, assuming an average contribution from temperature and humidity

(c) Through that point lay off a line which maintains a constant proportion of the space between the two sloping lines. (An overlay of tracing paper is useful here.)

(d) Approximate heights corresponding to given pressures may now be read from the horizontal scale.

This diagram may be used also to interpolate between the height values reported in the standard-level message.

Full details of all the codes used in meteorological broadcasts will be found in *Met O 920b: Handbook of Weather Messages, Part II, Codes and Specifications,* published by HMSO, London and also in the *Manual on Codes*, WMO No 306, in the section dealing with code FM35.

THE TEPHIGRAM

Meteorologists usually plot radiosonde ascent data on a rather complex thermodynamic chart known as a *tephigram,* (which may be used as a means of calculating potential refractive index) and a knowledge of its properties will help to achieve an understanding of the processes involved in the atmospheric movements we have been considering. Fig 3.10 shows an outline diagram, including a set of *K*-lines which will be explained in the next section. Reference should be made to the small inset diagram which identifies the various axes as they appear at the 1000mb, 0°C intersection:

P–P are *isobars*, or lines of constant pressure.

T–T are *isotherms*, or lines of constant temperature.

D–D are lines of constant moisture content, which are followed by the dew point as the pressure alters during adiabatic changes.

A–A are lines of constant potential temperature, followed by the air temperature during an adiabatic change.

W–W is a saturated adiabatic, which marks the temperature changes followed by ascending saturated air (only one is shown here in order to simplify the diagram as much as possible).

Both temperature and dew point are plotted with reference to the T–T lines.

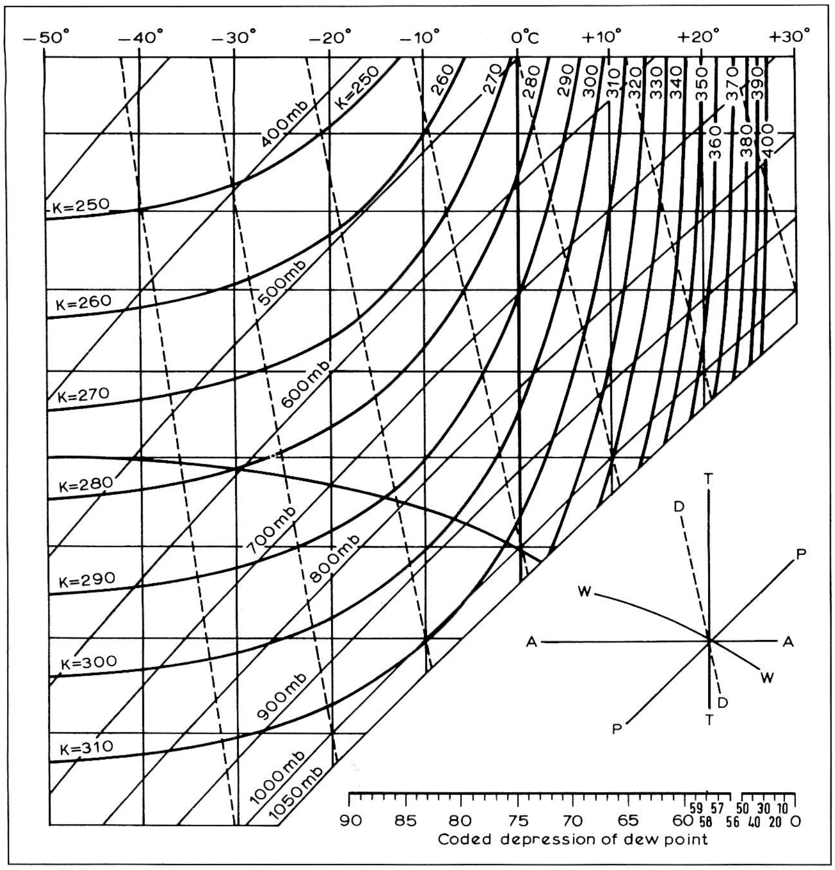


Fig 3.10. Skeleton tephigram (a meteorological temperature entropy diagram) showing the positions of additional curves, labelled K = 250 to 450, used for direct graphical calculation of potential refractive index values from published radiosonde measurements. For practical use it is recommended that the curves should be transferred to a standard full-sized tephigram, available from HMSO as Metform 2810B

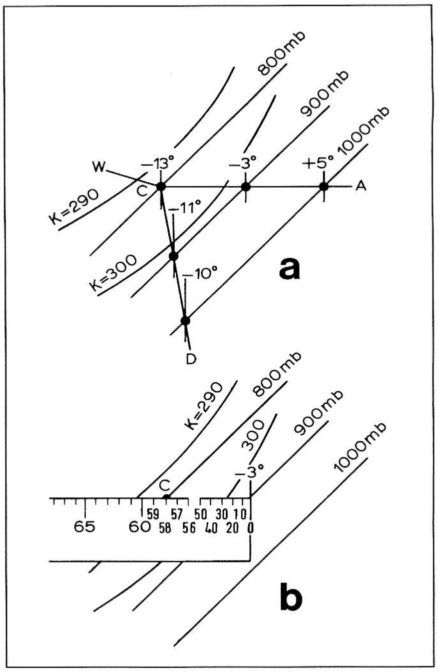


Fig 3.11. Alternative methods of determining values of potential refractive index from reported measurements of pressure, temperature and dew-point depression, using the tephigram modified as in Fig 3.10. (a) Intersection method. (b) Using scale of coded dew-point depression values

An example of the use of the tephigram will help to emphasise the points which have been made earlier in the text. Consider Fig 3.11(a), which shows two points on the 900mb line, representing a temperature of −3°C and a dew point of −11°C. If that sample of air is taken to a pressure of 1000mb adiabatically, the temperature will follow the horizontal line CA, and the dew point will follow CD. At 1000mb the temperature becomes +5°C (by definition the potential temperature), and the dew point becomes −10°C. Lifting would cause the temperature and the dew point to come closer together, and they would become coincident at the point C, which is known as the *condensation level,* where condensed droplets of water begin to appear as cloud. Further lifting will cause the temperature to follow one of the saturated adiabatics such as CW, instead of an extension of AC, due to the liberation of latent heat.

OBTAINING POTENTIAL REFRACTIVE INDEX VALUES

Potential refractive index values may be obtained in one of four ways, the method to be used depending on the resources available.

1. From the expression:

$$K = \frac{77.6}{\theta}\left(1000 + \frac{4810000e}{p\theta}\right)$$

where p is the pressure in millibars at the level of observation and the potential temperature in kelvin:

$$\theta = (T_{^\circ C} + 273) \times \left(\frac{1000}{p}\right)^{0.288}$$

e *is* the saturation vapour pressure at the dew-point temperature.

Example: p = 900mb, T = −3.0°C, dew-point depression = code 58 = 8°C below −3.0°C = −11.0°C, then:

$$\theta = 270 \times \left(\frac{1000}{900}\right)^{0.288} = 278.3$$

and from Fig 3.1(d):

$$e = 2.6\text{mb (at } -11^\circ\text{C)}$$

whence K = 292.8.

2. Using a programmable calculator or computer. Table 3.6 outlines a program which provides K values directly from pressure, temperature and either dew-point depression, dew point or percentage relative humidity.
3. Using a full-sized tephigram based on Fig 3.10 and a two-line construction. The curved potential refractive index lines labelled K = 290, K = 300 etc are so placed that the

Table 3.6. Computer or calculator program to obtain potential refractive index, *K*, from basic meteorological data

S = Store
R = Recall from store

Load stores with constants:

S1 = 273, S2 = 2937.4, S3 = -4.9283, S4 = 23.5518, S5 = 77.6, S6 = 4810, S7 = 0.288, S8 = 1000

Input data for each level to be oomputed, pass to the indicated stores

Enter p = pressure in millibars (or hectopascals). To S9 (p)

Enter t = temperature °C. Add R1 to convert to kelvin. To S10 (T)

Enter either:

D = dew point depression °C. Subtract from R10. To S11 (T_d)

or t_d = dew point temperature, °C. Add R1 to convert to Kelvin. To S11 (T_d)

or U = relative humidity, expressed as a decimal. To S12 (R_H)

Program

1. Evaluate vapour pressure, e
   Either (a) If dew point or dew point depression has been entered:
   $R11^{R3} * 10^{(R4-(R2/R11))}$ To S13 (e)
   or (b) If percentage relative humidity has been entered:
   $R10^{R3} * 10^{R4-(R2/R10))} * R12$ To S13 (e)
2. Evaluate potential temperature
   $R10 * (R8/R9)^{R7}$ To S14 (θ)
3. Evaluate potential refractive index, K
   R5((R6 * R8 * R13)/(R9 * R14)) + R8/R14
   Round off the result to the nearest whole number.

Test

When p = 900, t = −3.0, either D = 8 or t_d = −11.0, or U = 53/100 = 0.53 then K = 293.

Note: A step-by-step calculator program (based on the TI58/59 calculator but easily adapted to suit any similar programmable scientific calculator) may be found in the fourth edition of the *VHF/UHF Manual*, p2.14.

required value of K may be read at the intersection of the dry adiabat through the temperature point and the moisture content line through the dew point, plotted on the appropriate isobar. These points are shown on Fig 3.11(a) at −3°C and −11°C (900mb), as in the previous example. The lines drawn as indicated intersect at the point C which, when referred to the K lines, gives the answer directly: K = 293.

4. Alternatively, the scale labelled 'coded depression of the dew point' may be transferred to the edge of a card and used horizontally on the diagram as shown in Fig 3.11(b), with the right-hand index against the point on the diagram defined by the temperature and pressure. The required K value is read against the coded dew-point depression (58, signifying 8° depression). The scale is the projection on to a horizontal of depressions along an isobar, using the slope of the moisture content lines. This method is strictly correct only on the righthand side of the diagram and the method would suffer a progressive loss of accuracy towards the left (because the moisture content lines are not parallel), were it not for the fact that the dependence of K on dew point diminishes at low temperatures.

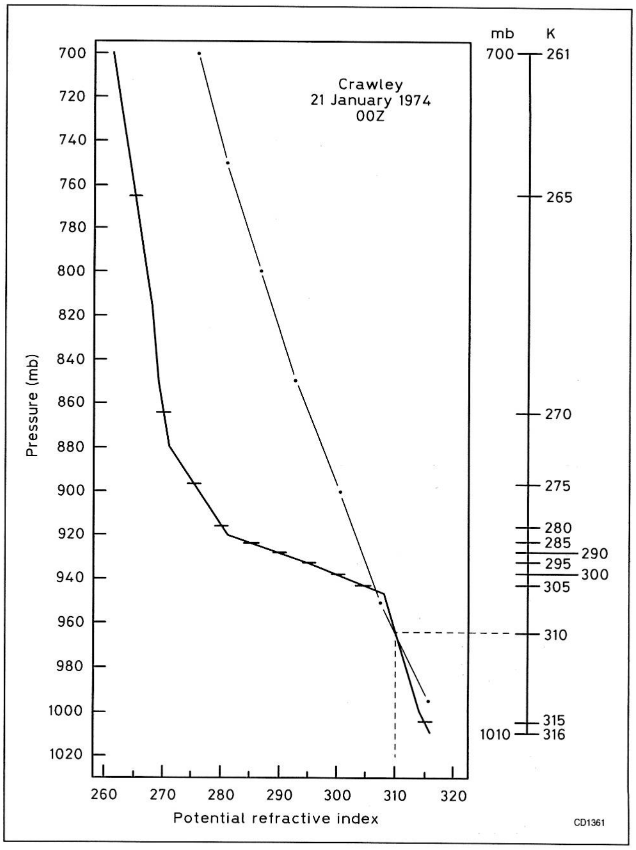


Fig 3.12. Potential refractive index profile from Crawley, 21 January 1974, at 0000GMT. The spaced values along the vertical line on the right have been used in the construction of both Fig 3.5 and Fig 3.6. The thin sloping line shows 'normal' *K* values for the second half of January (see Table 3.8, line B).

FROM PROFILE TO SECTION

Once the calculations have been made all the information necessary to draw a profile will be to hand.

A profile, such as the one shown in Fig 3.12, reveals immediately the presence of warm, dry, low-refractivity air overlying a ground-based layer of air which is cool, moist, and of high refractive index. The more abrupt the boundary between them, the more nearly horizontal will the transition appear on the diagram, and the more pronounced the bending experienced by the radio wave. Some of the occasions when conditions have been most favourable for DX have occurred during periods when there has been anticyclonic subsidence aloft with a contrasting depth of wet fog at the ground.

In some cases the refractive index profile is all that is required. It is much more rewarding, however, to combine it with others in order to make a section.

The first step is to project the pressures at which regular values of K occur across to a vertical line, as shown in the diagram. The spacings which result can then be transferred to form part of a time-section for the station in question (Fig 3.5), or a cross-section for a given path (Fig 3.6) – the one profile forms part of both diagrams.

Table 3.7. Seasonal variation of radio refractive index *N* at Crawley. Six years combined data, 1972–7

Half-month period		Sfc *N*	Sfc *P*	950 *N*	900 *N*	850 *N*	850 *H*	800 *N*	750 *N*	700 *N*	700 *H*
JA	A	318	1001	297	278	260	1465	244	227	213	3001
	B	314	995	296	278	261	1407	244	228	213	2929
FE	C	313	991	296	278	261	1381	245	228	214	2909
	D	315	1002	294	276	259	1468	244	228	213	2995
MR	E	313	999	294	276	259	1444	243	227	213	2966
	F	313	1000	294	276	259	1455	243	228	213	2984
AP	G	312	997	294	277	262	1434	245	229	214	2958
	H	313	1002	294	278	261	1480	244	227	212	3014
MY	J	316	997	297	280	263	1443	246	229	214	2979
	K	319	1000	298	281	262	1483	244	228	213	3035
JE	L	321	1001	300	282	263	1497	245	228	213	3059
	M	326	1001	304	285	265	1510	247	230	213	3091
JL	N	330	1001	306	287	267	1513	247	230	213	3098
	P	331	1000	308	288	269	1503	249	231	214	3096
AU	Q	331	1001	307	288	268	1519	248	229	213	3102
	R	330	1001	306	287	266	1518	246	228	212	3098
SE	S	327	1000	305	285	265	1495	246	229	213	3067
	T	327	999	304	284	265	1479	246	228	213	3041
OC	U	323	997	302	282	262	1455	244	227	213	3014
	V	323	1000	300	280	260	1475	244	227	212	3030
NV	W	320	998	299	281	262	1448	245	228	213	2990
	X	315	998	295	277	260	1439	243	227	213	2968
DE	Y	316	998	296	278	261	1441	244	227	213	2973
	Z	318	1002	296	278	261	1469	243	227	212	3006
Overall six-year mean		320	999	299	281	263	1468	245	228	213	3015

The additional work involved in this type of exercise is amply justified by the sense of continuity which results. Thus the time-section shown in Fig 3.5 reveals in a single glance far more about the formation and eventual dissipation of a subsidence boundary layer than could be gained by a prolonged study of the 10 separate profiles which were combined in its construction.

'NORMAL' VALUES OF *N* AND *K*

It is in the nature of things that refractive index studies, requiring as they do a considerable amount of calculation and detailed graphical work, are carried out on an infrequent basis, as and when periods of interest come to light.

It means, inevitably, that a sense of continuity is lacking whenever an in-depth study is undertaken. It means also that researchers usually see only abnormal conditions and few of them can be bothered to do similar exercises when nothing out of the ordinary has been occurring.

The situation is further complicated by the fact that there is a marked seasonal variation in values of refractive index near to the ground, making it difficult to compare, say, a February event with one that occurred in July, unless one has access to information showing the sort of values that might be considered 'normal' in each of the two cases.

It might seem that monthly mean refractive index values could be obtained by putting mean monthly temperature and mean monthly dew-point data into the computer program in place of individual ascent figures. An inspection of the skeleton tephigram, Fig 3.10, will explain why that will not work. The temperature (T–T) and dew-point (D–D) scales are straight-line functions whereas the refractive index lines have a pronounced curvature which changes character across the chart. Mean figures of refractive index have to be obtained from daily refractive index values, not from means of the basic meteorological data.

That explains the origin of Tables 3.7 and 3.8, which were obtained from 4380 consecutive radiosonde ascents reduced to refractive index and potential refractive index spot values. They represent six years of real-time data gathering, 1972–1977 and are, so far as is known, the only statistics available for either index in terms of atmospheric pressure instead of height. They appeared first in *IERE Conference Proceedings* No 40 (July 1978), to which reference should be made for further details and for other results of the study.

In this chapter refractive index and its distribution in the atmosphere are looked at from a meteorologist's viewpoint, having regard to the fact that the radio wave, once launched, is acted upon by atmospheric pressure, atmospheric temperature and the varying amount of moisture in the air through which it passes. Height, as such, does not come into the basic equations at all and only appears in expressions for refractive index gradients because radio engineers feel that they must have a fixed scale firmly anchored to the ground. Statistics based on specific pressure levels eliminate one of the variables

Table 3.8. Seasonal variation of potential refractive index *K* at Crawley. Six years combined data, 1972–7

Half-month period		Sfc *K*	Sfc *P*	950 *K*	900 *K*	850 *K*	850 *H*	800 *K*	750 *K*	700 *K*	700 *H*
JA	A	318	1001	308	300	292	1465	286	279	275	3001
	B	315	995	307	300	293	1407	286	280	275	2929
FE	C	315	991	307	300	293	1381	287	280	276	2909
	D	315	1002	305	298	291	1468	286	280	275	2995
MF;	E	313	999	305	298	291	1444	285	279	275	2966
	F	313	1000	305	298	291	1455	285	280	275	2984
AP	G	313	997	305	299	294	1434	287	281	276	2958
	H	313	1002	305	300	293	1480	286	279	273	3014
MY	J	317	997	308	302	295	1443	288	281	276	2979
	K	319	1000	309	303	294	1483	286	280	274	3035
JE	L	321	1001	311	304	295	1497	287	280	274	3059
	M	326	1001	315	307	298	1510	290	282	275	3091
JL	N	330	1001	317	309	300	1513	290	282	275	3098
	P	331	1000	320	311	302	1503	292	283	276	3096
AU	Q	331	1001	318	310	301	1519	291	281	274	3102
	R	330	1001	317	309	299	1518	289	280	273	3098
SE	S	327	1000	316	307	298	1495	289	281	274	3067
	T	327	999	315	306	298	1479	288	280	274	3041
OC	U	324	997	313	304	294	1455	286	279	274	3014
	V	323	1000	311	302	292	1475	286	279	273	3030
NV	W	321	998	310	303	294	1448	287	280	274	2990
	X	316	998	306	299	292	1439	285	279	275	2968
DE	Y	317	998	307	300	293	1441	286	279	274	2973
	Z	318	1002	307	300	293	1469	285	279	273	3006
Overall six-year mean		321	999	310	303	295	1468	287	280	275	3017

and provide results that are in the right form for comparison with day-to-day radiosonde measurements.

For an example of the use of the annual statistics consider Fig 3.12, in which the thicker line shows potential refractive index values plotted against pressure for one of the profiles used in the examples of cross-sections and time-sections. The fine line shows the mean values for the time of year and it should be clear that the low values of potential refractive index encountered above 940mb have been the result of air subsiding from above. During an adiabatic change potential refractive index remains unchanged so a parcel of subsiding air brings its original value down with it but, remember, the same is not true of conventional refractive index, *N*.

Strictly, these statistics are valid only for south-east England, but they provide a useful indicator for adjacent areas, for which there are no comparable figures.

SIGNAL STRENGTHS AND RANGES ATTAINABLE

The effect of the boundary layer on signal strength

The time-section is an ideal way of comparing a series of upper-air soundings in terms of refractive index because, once you know the signs, you can see at a glance periods of rising or subsiding air and the formation and dissipation of steep-lapse layers of the sort that lead to periods of anomalous propagation, or 'openings' as amateur operators prefer to call them.

Fig 3.13 illustrates very clearly the way that VHF signal strengths rise during a week of anticyclonic subsidence. When this diagram was originally prepared there was a television station transmitting on 174MHz, located at Lille, in northern France. G3BGL, located at a site just to the west of Reading, used to monitor the strength of the sound carrier using a pen recorder, with the object of compiling statistics for a 300km VHF path (this was before the time that we had beacon transmissions on the amateur bands). The television station operated every afternoon and evening.

The potential refractive index pattern is typical of an anticyclonic opening. The bottom boundary of the isopleths shows that ground level atmospheric pressure reached a peak during 24 and 25 September. The 'inverted pudding basin' centred on the 22nd heralded the commencement of subsiding air, a feature which was to make its presence felt over the next six days. Note the way that the signal level rose as the subsidence boundary layer formed, then lowered and intensified. The highest signals appeared when the gradient was steepest, on the 25th. After that the gradient slackened and the received signal strength fell. On the evening of the 28th a cold front arrived and destroyed, by mixing, all traces of the boundary which had given the good conditions. A low-pressure system moving in is responsible for the rising isopleths of the 29th. The radiosonde station at Crawley was very close to the mid-point of the transmission path.

It is a point worth emphasising that this one time-section

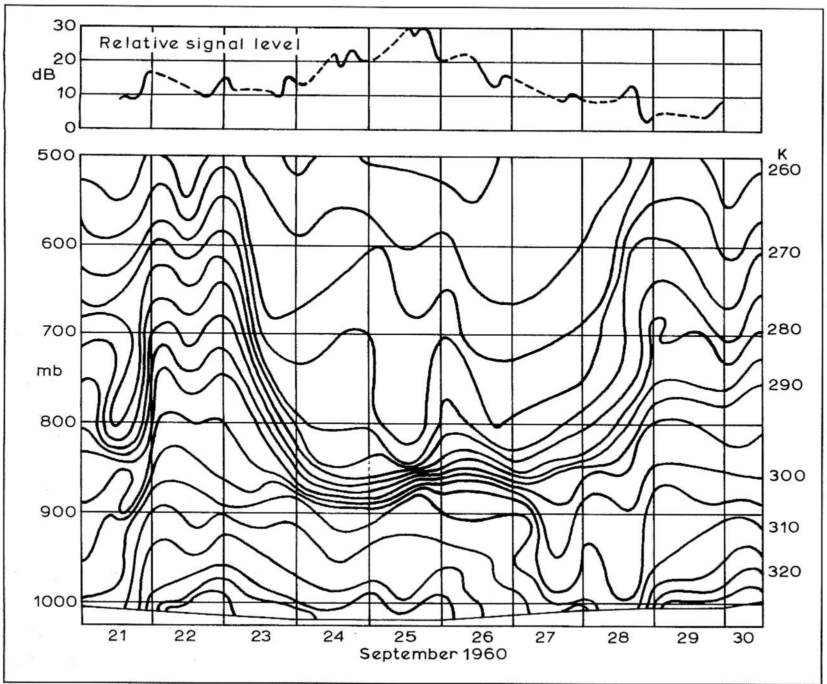


Fig 3.13. Potential refractive index time-section, Crawley, 21–30 September 1960, together with a record of signal strengths, obtained on a chart recorder located near Reading, from VHF TV transmissions originating at Lille in France. Note that the highest signals occur when the steep-lapse layer is most pronounced, and a sudden drop occurs as the anticyclone is replaced by a low-pressure system. With acknowledgements to *J Atmos Terr Phys,* Pergamon Press

shows, at a glance, all the significant information contained in 20 consecutive refractive index profiles.

The subsidence boundary layer may be thought of as being a battlefront. It comes about as a result of descending warm dry air coming up against cool moist air stirred up from the ground. If the two opposing forces are nicely balanced the layer stays where it is and intensifies. If the subsidence increases or the turbulence decreases the boundary falls and may even reach the ground. If that happens the good conditions immediately drop out.

If the low-level turbulence increases the boundary will rise and will eventually be lost through mixing. When conditions are at their best it is the contrast between the two opposing air components which is all-important. Given warm dry subsiding air from aloft in an anticyclone you cannot do better than find yourself surrounded by wet dripping fog.

At their best, tropospheric 'openings' are productive of ranges considerably in excess of conventional textbook expectations. Using an analysis of extreme-range signal reports it has been shown that if 1000km can be exceeded at 144MHz there is a strong probability that 1500km will be reached, for there is a pronounced peak in the distribution at that range. Similarly, if 750km can be exceeded at 432MHz, there is a good probability that 1200km will be reached. Histograms showing these findings are included in a paper in *IEE Conference Publication 195,* Part 2, pp163–167 ('The use of a dense network of amateur radio stations to determine the limits of long-range tropospheric propagation within an anticyclone', by R G Flavell) to which reference should be made for further details.

Table 3.9 gives details of maximum confirmed tropospheric ranges for the six VHF and UHF amateur bands, as at the beginning of 1997.

Table 3.9. European VHF and UHF records (as at the beginning of 1997) – tropospheric

50MHz	GJ4ICD (IN89WF) and OZ5W/P (JO64GX)	1 June 1996	1188km
70MHz	GM3WOJ (IO77WO) and G4KFR (IO90AS)	18 September 1988	774km
144MHz	GM0KAE (IO86CD) and EA8BML (IL27GX)	9 September 1988	3264km
432MHz	EA8XS (IL28GA) and GW8VHI (IO81CM)	5 July 1984	2786km
1.3GHz	EA8XS (IL28GA) and G6LEU (IO70ME)	29 June 1985	2617km
2.3 GHz	EA7BVD/P (IM78JD) and EA8XS/P (IL27GW)	8 July 1984	1481km

Source: John Morris, GM4ANB, for Region 1, IARU

EME (MOONBOUNCE)

Earth-Moon-Earth contacts are, as the name implies, contacts made using the Moon as a reflector. It goes without saying that high power and accurate aiming of a narrow beam are essential requirements.

The paths are line of sight to and from the Moon but tropospheric refraction effects could affect the performance at low elevation angles. Distance (around the Earth, transmitter to receiver) records are made and broken but they are a function of geometry rather than geophysical influences. All VHF and UHF bands apart from 70MHz have been used. Europe to New Zealand has been achieved on 144, 432MHz and 1.3GHz.

FREE-SPACE ATTENUATION

The concept of free-space attenuation between isotropic antennas, or basic transmission loss, L_b, provides a useful yardstick against which other modes of propagation may be compared. It is a function of frequency and distance, such that

$$L_b = 32.45 + 20 \log f + 20 \log d$$

where f and d are expressed in megahertz and kilometres respectively.

Table 3.10 provides a representative range of values against distance for various VHF/UHF amateur bands.

TROPOSPHERIC SCATTER PROPAGATION

Tropospheric scatter propagation depends for its effectiveness on the presence of dust particles, cloud droplets and small-scale irregularities in radio refractive index within a volume of the atmosphere which is common to both the transmitting and receiving antenna beam cones. The height of the bottom of this common volume is a function of distance between the stations concerned owing to the effect of the curvature of the earth. Typical heights are 600m for a 100km path, 9000m for a 500km path. Path losses increase by about 10dB for every degree of horizon angle at each station so that a site with an unobstructed take-off is an important consideration.

Only a very small proportion of the signal energy passing through the common volume will be scattered, and only a small proportion of that will be directed towards the receiving station. Therefore the loss in the scattering process is extremely large and the angle through which the signal ray has to be deflected is an important characteristic of a troposcatter path; for best results it should be no more than a few degrees.

J N Gannaway, G3YGF, has made a critical study of the losses in tropospheric scatter propagation, which appeared in *Radio Communication* August 1981, pp710–714, 717. It should be consulted for a fuller discussion of the mode than can be given here.

Table 3.11 shows the path losses between two stations on a smooth earth, expressed as decibels below the free-space values, for VHF and UHF amateur bands. (Free-space losses have been given earlier in Table 3.10.) To these values must be added losses depending on characteristics of the sites – height, distance to the first obstruction, antenna coupling losses etc – and variables depending on seasonal and weather factors.

Table 3.12, which is taken directly from the work cited, shows the theoretical range that could be expected under flat conditions and from good sites, with the equipments shown against each of the amateur bands. The ranges are for a 0dB signal-to-noise ratio in a bandwidth of 100Hz, representing a weak CW signal; for SSB these ranges should be reduced by 130km on each band.

For distances approaching 1000km the equipment requirements are comparable to those needed for propagation by moonbounce.

Because signal-path transit times vary with height of scatter within the common volume a 'blurring' occurs which limits the maximum speed of transmission of intelligence; narrow beams have the faster capabilities.

Table 3.10. Free-space attenuation

Frequency (MHz)	Distance (km) 50	100	150	200	300	400	500	750	1000
50	100	106	110	112	116	118	120	124	126
70	103	109	113	115	119	121	123	127	129
144	110	116	119	122	125	128	130	133	136
432	119	125	129	131	135	137	139	143	145
1296	129	135	138	141	144	147	149	152	155
2300	134	140	143	146	149	152	154	157	160

Table 3.11. Troposcatter path losses on a smooth earth

Band (MHz)	Distance (km) 50	100	150	200	300	400	500	750	1000
70	47	49	53	55	61	69	75	91	109
144	50	52	55	58	64	72	78	94	112
432	55	56	60	62	69	76	82	99	116
1296	59	61	65	67	74	81	87	104	121
2300	60	63	67	70	76	83	90	106	

IONOSPHERIC PROPAGATION AT VHF AND UHF

Regular layers

The regular layers of the ionosphere play only a small part in the properties of the VHF bands and, according to current thought at least, none at all at UHF and above. Around the time of sunspot maximum and for perhaps a year or two after, there are occasions when maximum usable frequencies exceed 50MHz and cross-band working with North American amateurs becomes possible. The most favourable times for transatlantic contacts at 50MHz occur when the solar flux is high and the magnetic index is low, but the required conditions do not persist for long. On 8 February 1979 G3COJ and WB2RLK/VE made the first 28/50MHz transatlantic contact since 1958. On the other hand, when conditions are good, they are often very good. EI2W, the only 50MHz licensed amateur in northern Europe at the time, succeeded in working 40 states of the USA on 50MHz during 1979–80. Such contacts are made via the F2 layer. It is unlikely that transmissions above 30MHz would ever be propagated by the regular E layer. In the tropics some occasional periods of activity around noon in maximum sunspot years may be possible, using the F1 layer.

The current (1997) European record distance for 50MHz regular F2 layer propagation is 16,076km, set by OZ1LO (JO55VC) and UK3AMK (QF21NT) on 18 October 1991.

Any propagation at VHF which may take place via the regular layers of the ionosphere will have a very strong dependence on the solar cycle.

Non-regular ionisation

Contacts at VHF and UHF are occasionally possible via ionisation which may take the form of sheets, clouds, mobile curtains or long narrow cylinders. Most of these forms are active around E layer height, but they are not directly associated with the regular layers.

Some effects, those involving the equatorial ionosphere, for example, may take place at F layer levels. The varieties

Table 3.12. Theoretical performance between good sites under flat conditions

Frequency (MHz)	Path loss (dB)	Range (km)	Transmitter power (W)	Noise figure (dB)	Antenna	Antenna gain (dBi)
144	240	870	100	3	2 × 16-el Yagi	18
432	247	790	100	3	2 × 25-el loop Yagi	22
1296	258	760	100	3	2 × 25-el loop Yagi	24
2304	262	720	50	3	6ft dish	31

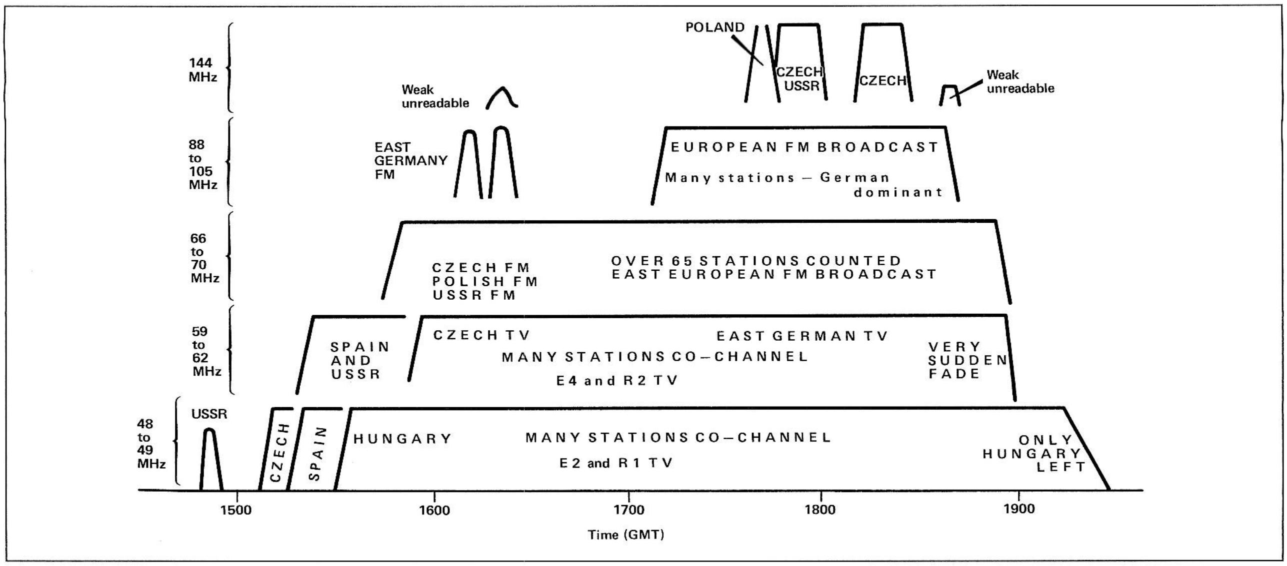


Fig 3.14. Frequency versus time for a major sporadic-E event recorded on 10 June 1980 at GM4IHJ, Saline, Fife

which will be dealt with here are sporadic-E, auroral-E, trans-equatorial propagation and meteor scatter.

VHF SPORADIC-E

As has been mentioned before, there is some question still regarding the nature of VHF sporadic-E. The reason is that sporadic-E at HF has been known about and recognised since the earliest days of ionospheric sounding in the 'thirties but what radio amateurs refer to as sporadic-E – typically extending to the 144MHz band and beyond – just does not show up on ionosondes. Even those times when HF sporadic-E does appear on the record it does not display characteristics which would be sufficient to support reflections well into the VHF band (over 200MHz has been reported).

For many years VHF sporadic-E was regarded with scepticism by professionals but its reality has now been accepted, largely due to the weight of evidence provided by amateur observers in a truly international project co-ordinated by the International Amateur Radio Union. For many successive years the co-ordinator, the late Serge Canivenc, F8SH, collected and collated reports from all over Europe to provide day-by-day summaries, which were submitted to CCIR, the International Radio Consultative Committee (now a section of the International Telecommunications Union).

The project is ongoing and the collection of reports showing which paths are open (and when) remains a task ideally suited to amateur observers.

In its conventional form sporadic-E consists of horizontal sheets about 1km thick and some 100km across, usually at a height of 100–130km. Clouds form in an apparently random manner, although there is an obvious preference for certain times and seasons. They do not behave consistently for, whereas some sheets may travel across continents for several hundreds of kilometres, others remain almost stationary. It has been claimed that there is a general tendency for them to drift towards the equator at about 80m/s. Both scattering and reflection modes are possible in the sporadic-E layers.

Above 30MHz, paths via E_s ionisation are rarely less than 500km. The maximum single-hop range is limited by the geometry of the system to about 2000km and double-hop from a single sheet is relatively rare because it would have to exceed 500km across in order to be able to accommodate the two points of reflection. Two-hop E_s propagation is more likely from two separate sheets, separated by less than 2000km, when the possible maximum range is extended to 4000km.

Sporadic-E at VHF is seasonal, nearly all of it (in Europe) occurring between May and August, although events outside that period are not unknown. The times of maximum activity are generally within the periods of 0700–1300GMT and 1500–2200GMT. The duration of events is an inverse function of frequency: that is, for a particular occasion the event will begin later and finish earlier at the higher of two given frequencies.

John Branegan, GM4IHJ, gave a very interesting analysis of the VHF sporadic-E event of 10 June 1980, as observed from Saline, Fife (Fig 3.14). This demonstrated very clearly how the longest opening – over 4½ hours – appeared on a 48–49MHz monitor, with progressively shorter periods on each of the other frequency bands checked. At 144MHz the event was confined, for the most part, to half an hour either side of 1800GMT. GM4IHJ had also produced a map showing the location of stations which have been positively identified during E_s openings (Fig 3.15). The symbols 'FM' and 'fm' indicate stations in the 70MHz and 100MHz FM broadcast bands respectively, while black dots are Band I TV stations around 50MHz. It may be seen that all the stations received were between 1000 and 2000km distant.

That being so – and, assuming a reflection height of 100–130km, the geometry of the path certainly lends support – how to account for the claimed record distances worked during E_s events (Table 3.13)? Clearly, in the case of 70MHz and 144MHz double hop is feasible, though probably not from a single cloud of ionisation. But, in the case of 50MHz, over 8000km in four hops seems to be stretching credibility a mite too far. Yet this report does not stand in isolation. In the

Table 3.13. European VHF and UHF records (as at the beginning of 1997) – sporadic-E

50MHz	G0DJA (IO93FP) and 7Q7RU (KH74MF)	24 August 1993	8446km
70MHz	GW4ASR/P (IQ82JG) and 5B4AZ (KM64MR)	7 June 1981	3465km
144MHz	OE1SBB (JN88FF) and RI8TA (MM37TE)	21 July 1989	4281km

Source: John Morris, GM4ANB, for Region I, IARU

DUBUS magazine for the first quarter of 1997, the 50MHz E_s Top List showed claimed distances in excess of 4000km (theoretical double-hop) in 36 entries out of a total of 81. If the reason for that is known it does not seem to have received very wide publicity.

The maximum frequency at which sporadic-E has been observed in the European area was found to have been 203MHz, recorded by F8SH on 9 July 1974. It is not known if anyone else has continued his work in routinely seeking a maximum frequency, but it seems certain that 200MHz is reached but rarely and for periods of very short duration.

Several sporadic-E warning nets – some radio, some making use of telephone 'chains' – are in operation in various parts of Europe, including the UK. With their help a random network of several hundred amateur stations may be got on the air in a very short time, and a careful computer analysis of their collected reports, which need consist of no more than time, band, callsigns and QTH locators, is sufficient to provide details of the size, shape and movement of the areas of ionisation responsible.

AMATEUR AURORAL STUDIES

The radio aurora at VHF probably represents the field in which radio amateurs can do most to contribute to present knowledge of radio propagation and the behaviour of the high atmosphere under the influence of solar emissions. A co-ordinated network of stations extending over a continent, each station equipped with nothing more complicated than a well-maintained receiver and operated by a person able to read and log callsigns, QTH locators and accurate times, can establish the existence and movement of areas of auroral ionisation on a scale that is impossible to achieve by any other means. The addition of steerable antennas and two-way communications increase the value of the observations still further, for these enable the location of the auroral reflection point to be established.

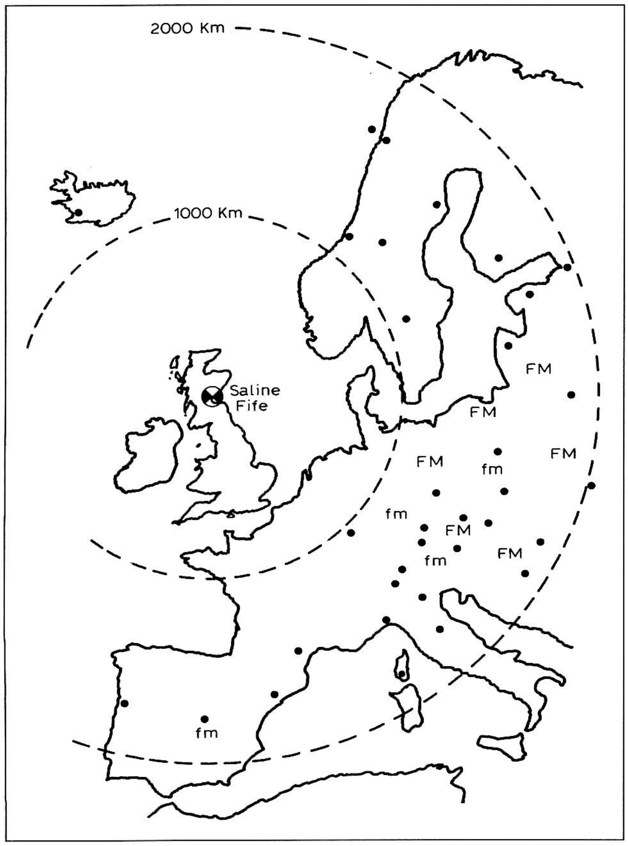


Fig 3.15. Stations received via sporadic-E at GM4IHJ

The geometry of the path is such that no two pairs of stations will reflect off exactly the same point on the radio auroral curtain so a number of near-simultaneous observations from a random network of stations can yield detail of an *area* of ionisation – its position relative to the Earth's surface and the direction of its main axis – and, if the process is continued throughout an auroral event, analysis of successive periods will reveal the motion of the ionisation in both space and time.

Unfortunately the aurora does not present itself as a perfect reflector placed perpendicularly to the surface of the Earth. Its vertical alignment tends to follow the curvature of the geomagnetic field so that a reflection from a relatively low altitude will appear to be above a point on the ground farther north than a reflection at a greater height. The reflection height is a function of the position of the two stations relative to the surface of ionisation, but the vertical and horizontal beamwidths of most VHF antennas are such that a large number of alternative paths are possible without change of beam heading, although not necessarily at maximum strength. There is much to be learned from a study of accurate times and bearings of maximum signal taken as near simultaneously as possible from the two ends of a transmission path. Those stations equipped with two-axis rotators, such as are used for satellite working, can contribute further by rotating in both azimuth and elevation for maximum signal.

From the foregoing it will be clear that, in general, the beam headings in the horizontal plane depart considerably from the great-circle directions between the stations. When amateur auroral studies began in earnest in the 'fifties it was commonly supposed that all stations had to beam their signals towards the north in order to make auroral contacts. During the International Quiet Sun Years (1963–4), when the GB3LER experimental beacon station was first set up beside the magnetic observatory at Lerwick, the beam direction for auroral studies was set at first towards 10° west of true north and nearly all reports of reception via the aurora came from Scottish stations.

A change was made to 25° east of true north, and this brought in reports from many parts of the Continent. That is not to say that 25° east of true north is an optimum direction, even for Lerwick. It is now known that beam headings can vary considerably during an aurora, and from one aurora to

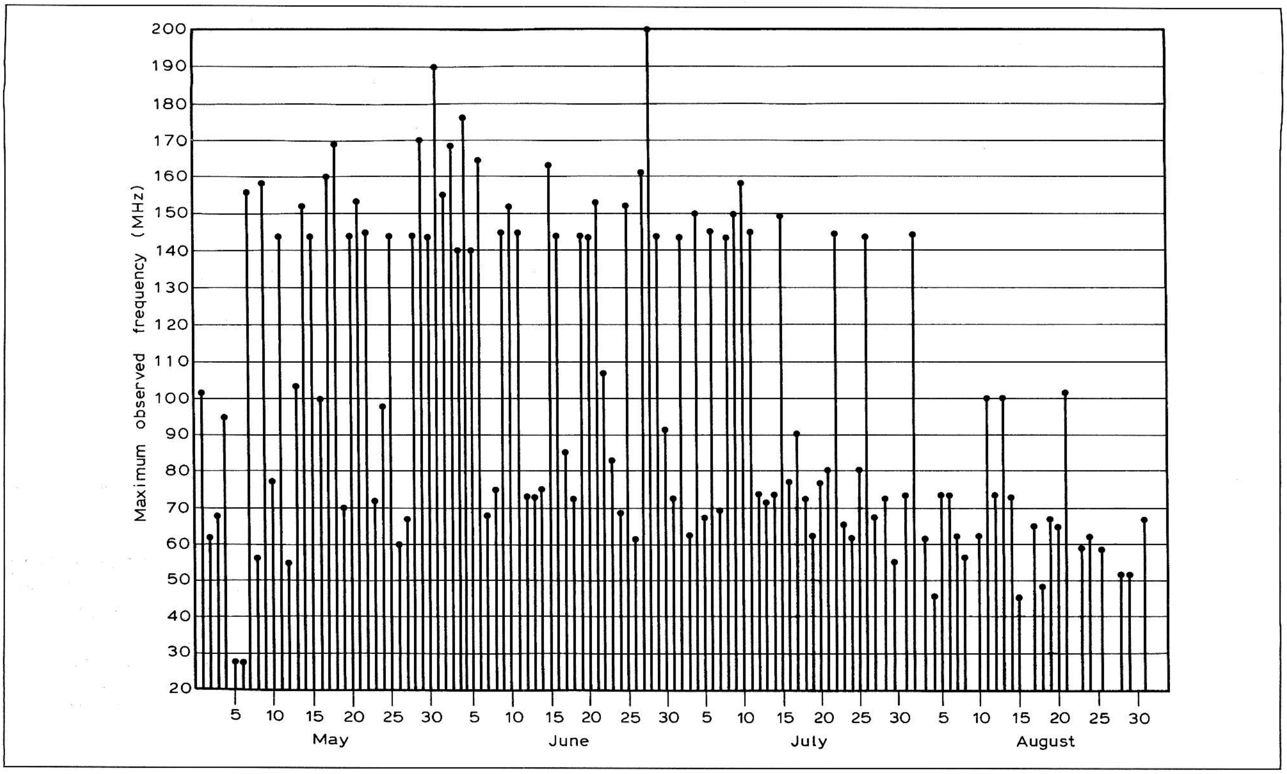


Fig 3.16 VHF sporadic-E activity in Europe during summer 1979

the next. No hard and fast advice can be given on this point, other than to suggest that an occasional complete 360° beam swing during an auroral event may produce results from an unexpected quarter, even when much of the activity appears to be concentrated in one fairly constant direction (Fig 3.17).

Charlie Newton, G2FKZ, who was for many years the IARU co-ordinator for amateur radio auroral studies, has established that, for any given station, there is a well-defined area within which auroral contacts are possible, and he has shown that that area is a function of the magnetic field surrounding the Earth. He has called the perimeter of this area the *boundary fence* and has demonstrated that its shape and extent varies as different locations are considered as origin. As an example, Fig 3.18 shows the boundary fence calculated for SM4IVE (at HT68d); it is approximately elliptical, 2000km from east to west, 1000km from north to south. The large dots on the map indicate the centres of QTH locator lettered squares containing stations heard or worked by SM4IVE via the aurora – the lines are not signal paths; they serve only to indicate the line-of-sight directions and distances to some of the more-distant stations. At any particular time during an event only a small part of the area shown will be accessible to SM4IVE, but the audible 'patch' will move during the progress of an aurora and may differ considerably from one aurora to the next, although all the stations worked will lie within the boundary fence. For stations farther east the area of accessibility is larger; for stations to the west and south it is smaller. Stations in Great Britain suffer from the disadvantage that there are no stations within the western half of the boundary fence.

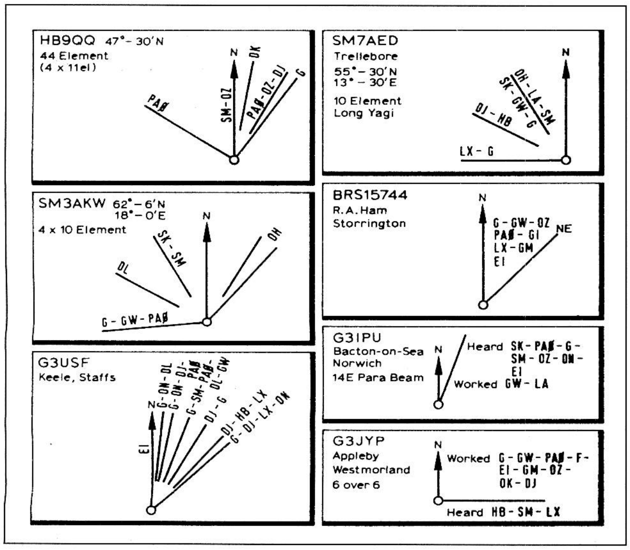


Fig 3.17. Beam headings recorded by certain operators during the aurora of 8 March 1970

There is a fairly close correlation between the occurrence of radio aurora and the three-hourly indices of geomagnetic activity; the greater the magnetic field is disturbed, the further south the event extends. During the International Quiet Sun Years attempts were made to relate motions of the visual aurora at Lerwick to aurorally reflected signals from GB3LER as received on the Scottish mainland, but the results suggested that, on a short time-scale, the two phenomena behave almost independently, although they must stem from a common cause. At times the visual aurora appeared to the south of Lerwick and forward-scatter off the back of the beam was suspected on more than one occasion, although the point was never proved by turning the antenna because it was not accessible enough to be moved at short notice.

A study of pen recordings of signals from GB3LER, via the aurora, to Thurso on the Scottish mainland suggested that it was not the peaks of a geomagnetic disturbance that gave the strongest reflections, but rather the fastest rate of change in the components describing the instantaneous field, as recorded by the observatory magnetometers.

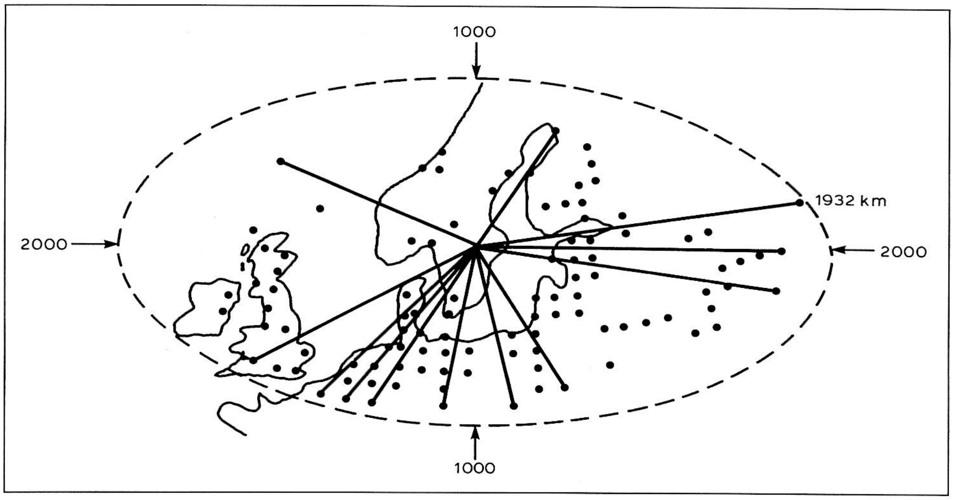


Fig 3.18. Boundary fence calculated for SM4IVE

Table 3.14. European VHF and UHF records (as at the beginning of 1997) – aurora

50MHz	ES1CW (KO29HK) and G3NVO (IO91IK)	6 February 1994	1850km
70MHz	G3SHK (IO90DX) and GM3WOJ (IO89KB)	11 August 1982	904km
144MHz	GM4BYF (IO85JV) and RB5CCO (KN59XG)	1 December 1989	2465km
432MHz	PA0FRE (JO21FW) and RA3LE (KO64AR)	13 March 1989	1851km

Source: John Morris, GM4ANB, for Region I, IARU

Every radio auroral event seems to be unique in some respect but there are characteristic patterns that regularly recur. The weaker or diffuse events, which are often only detected by northern stations, move little and slowly. They are often found to relate to minor irregularities on the magnetometer trace, known as *bays,* when the geomagnetic field deflects for a short while and then gradually resumes its normal diurnal pattern.

An intense auroral event typically opens suddenly with the appearance of signals having a characteristic 'flutter' tone from stations situated to the north or north-east. This often occurs in the early afternoon and contacts from European stations 1000km or more distant are likely. After perhaps two to three hours of activity it ceases and many operators unused to the mode may conclude that the event is over. The more knowledgeable stay on watch, and frequently their patience is rewarded by the appearance of a second phase, usually more rewarding than the first. The motion of the active region often follows the same general movement as the first phase, but reaching several hundred kilometres further south. Finally, when the event seems to have reached a peak, perhaps by late evening, all the activity suddenly ceases as though somebody, somewhere, has 'pulled the big switch' and gone off to bed (Fig 3.19).

Table 3.14 shows the maximum distances which have been worked by auroral reflection on three bands at VHF and one at UHF. It must be pointed out that these distances represent the great circle path between stations, not the path followed by the signals, which is considerably longer.

There is a tendency for a major radio auroral event to recur after an interval of slightly more than 27 days. That is because

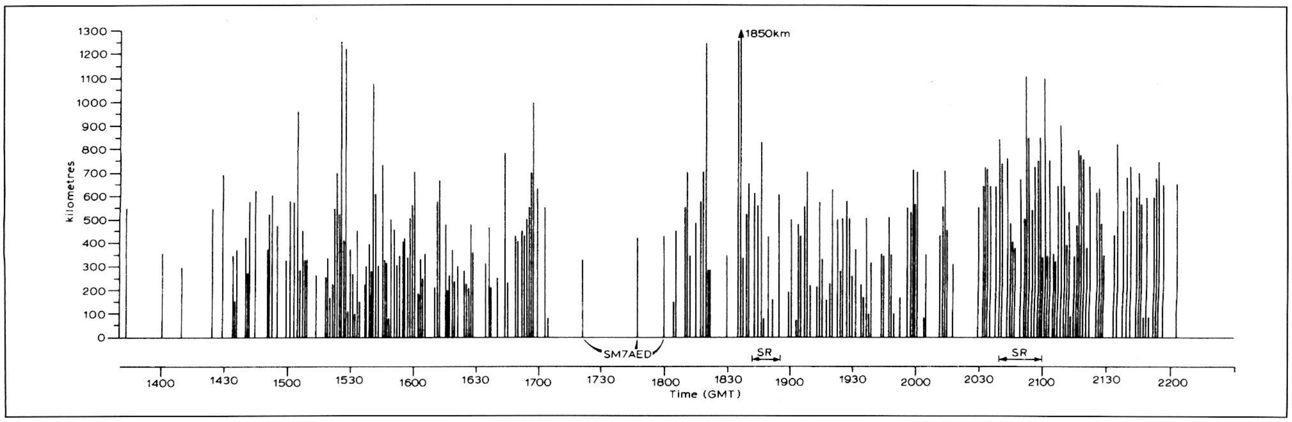


Fig 3.19. Times and distances for all stations heard or worked, based on logs submitted for the 8 March 1970 study. Note the pauses and the bunching of the longer-range contacts. The periods marked 'SR' indicate when a radar at Sheffield University recorded radio aurora to the north west

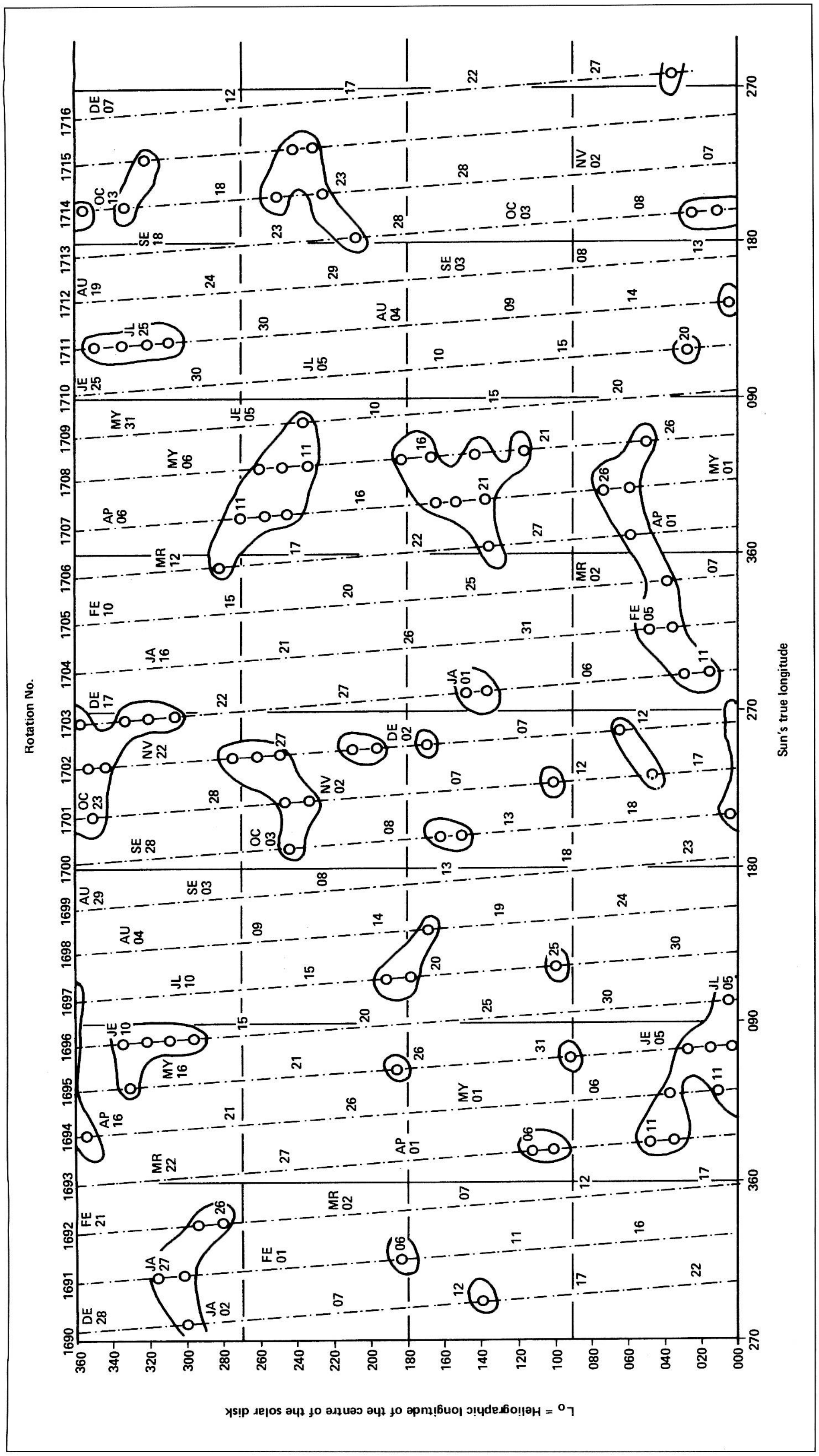


Fig 3.20. Days with reported radio aurora during 1981–2

the event has come about as a result of a disturbance on the Sun, and in that interval an active region has made a complete circuit of the Sun as seen from the Earth, and is in more-or-less the same position again. That position, at the time of the event, is generally about one day past central meridian passage (CMP), or, say, 13° west of the centre of the visible disc.

It is convenient to be able to record the dates of auroral propagation events in some way that links them to the rotations of the Sun. The chart shown in Fig 3.20 does just that. The vertical axis shows longitude on the Sun, L_o, following a sequence begun at Greenwich Observatory nearly 150 years ago. The horizontal axis is, in effect, a measure of where the Earth is around its orbit, passing through 000° at the time of the vernal equinox, 20 or 21 June. The sloping broken lines serve to connect a series of dots, each representing the date on which the indicated longitude, L_o, passed the centre of the disc of the Sun, as seen from the Earth, at 12UT. The sloping lines form a raster such that, if the diagram were to be rolled so as to make the top edge (L_o = 360°) and the bottom edge (L_o = 000°) coincident, they would form one continuous helix.

The chart may be used to record any event suspected of having a connection with events on the Sun. A horizontal trend corresponds to the rotation period of the Sun relative to the Earth; a trend of 45° (in the sense bottom left to top right) corresponds to the rotation period of the Sun relative to the stars. Known dates of radio auroral events are currently included in the propagation news section of the

GB2RS news bulletin service of the RSGB, and the patterns which emerge provide a useful guide to probable active and quiet periods up to about a month ahead. Fig 3.20, which shows the days when radio aurora was reported by UK stations during 1981–2 should be compared with Fig 3.21, which shows the days when the magnetometers at Lerwick recorded a disturbance of 5 or more on the conventional scale of geomagnetic *K*-units. It will be seen that there is a close relationship between the two patterns and this confirms the usefulness of the Lerwick data when dealing with the analysis of radio auroral events taking place in the region of north-west Europe.

Monitoring auroral propagation

In a contribution to the March 1977 issue of *Radio Communication,* Peter Blair, G3LTF, gave practical advice on monitoring distant VHF transmissions, which can be used as a guide to the onset of auroral propagation events on the amateur bands. Many of his remarks were directed towards observers living in the south of England, but his methods were applicable to other locations, provided that suitably placed transmitters can be found. What is required is a signal from a northerly direction in the low VHF region, for this will go auroral before the effects reach 144MHz. These requirements used to be met by one of the Band I BBC-TV sound transmitters. There are now 50MHz and 70MHz amateur beacon transmitters which may well serve the purpose. Two which operate at 100W and are

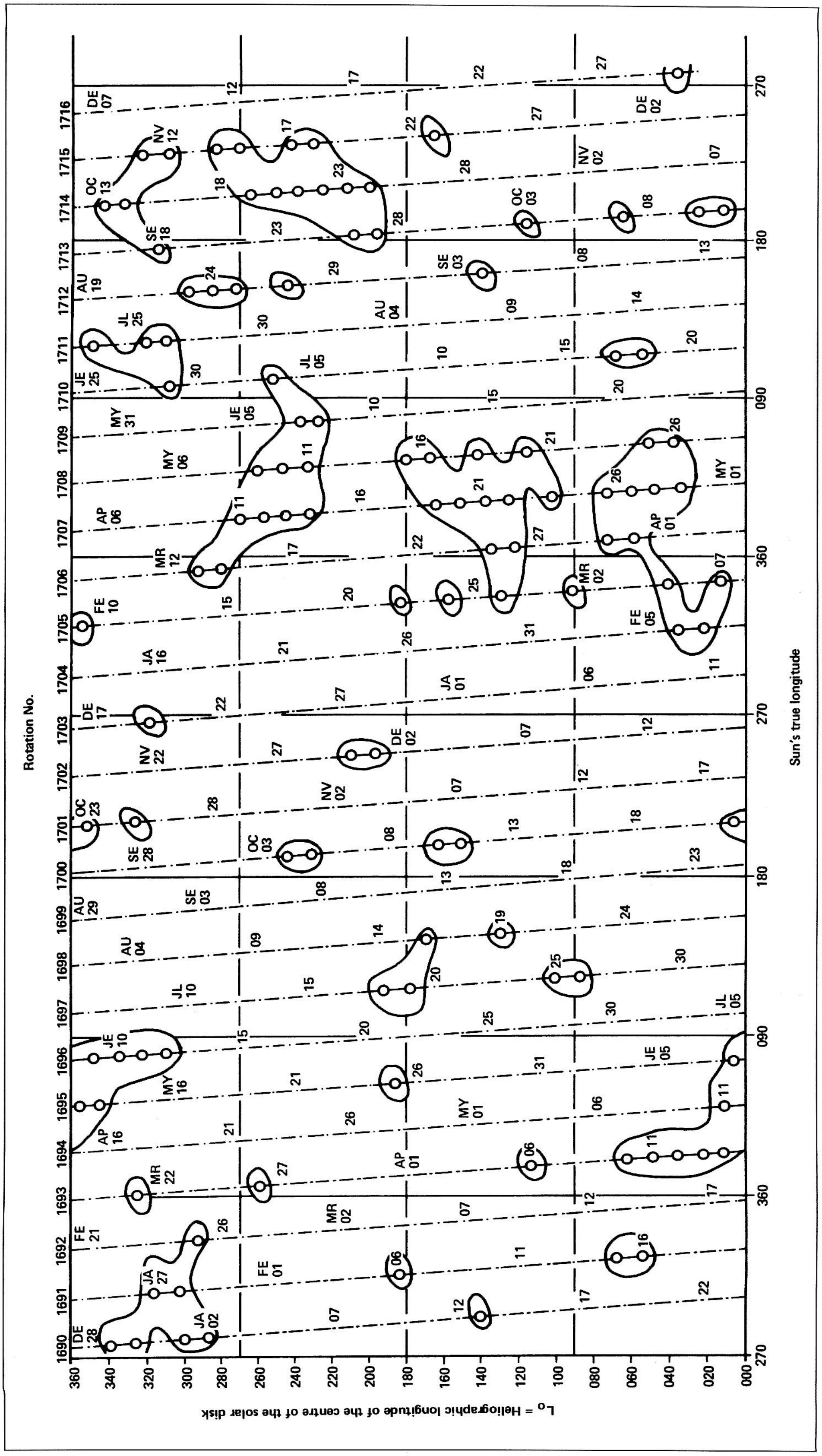
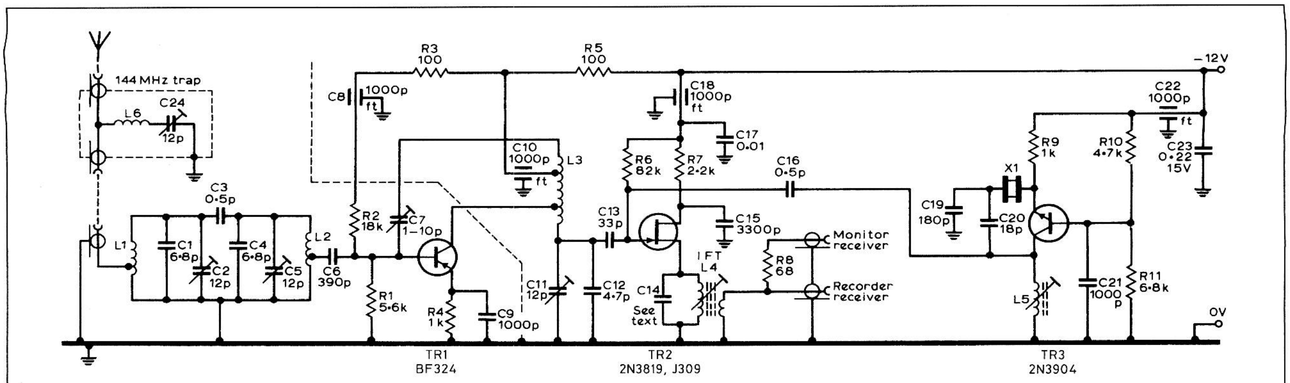


Fig 3.21. Days when Lerwick K-figure was five or greater during 1981–2



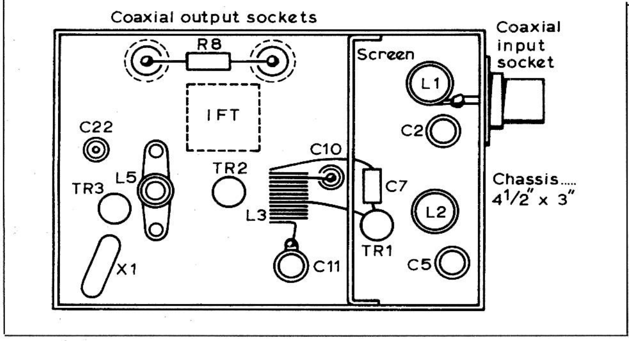


Fig 3.22. Circuit diagram of converter and layout of principal components

suitably placed are GB3LER (IP90JD) on 50.064MHz and GB3ANG (IO86MN) on 70.020MHz.

The suggested equipment for the aurora monitor is relatively simple. The antenna consists of two elements: radiator 95in by 3/8in, reflector 105in by 3/8in, spacing 30in, mounted 18 to 20ft off the ground, pointing north. This feeds a crystal-controlled converter, such as the one shown in Fig 3.22 and Table 3.15, which uses any convenient crystal which will produce an IF of around 7MHz. After alignment in the usual way the circuits should be peaked on the desired signal when the opportunity arises. The two outputs are at around 7MHz; one is intended for a simple fixed-frequency receiver, the AGC voltage of which drives a recording meter through a suitable DC amplifier. When correctly tuned, the 144MHz trap prevents the converter from being blocked in the presence of a local transmission.

Fig 3.23 shows the appearance of some typical auroral signal recordings. According to G3LTF a sudden (say within two minutes) onset of the auroral enhancement usually indicates that the effects will reach 144MHz within 10–15 minutes. A more gradual onset might herald a delay of up to about 30 minutes. It is rare for the effect not to reach 144MHz at all once it has been detected on the monitor. The equipment may also be used as an indicator for 70MHz but the respective time delays are considerably reduced.

The monitor should be left running during the day from about midday onwards. If the operator is unable to attend during the afternoon he or she may come home to find evidence of earlier activity. This should alert him to expect a second phase, and perhaps a third, later on. Where a more-or-less continuous watch is impracticable, the most fruitful times for checking the monitor are mid-afternoon, early evening and around 2200 local time. It should not be overlooked that a broadcast transmitter being monitored may close down each day at the end of the day's programmes, but this is no real disadvantage of the method as it is unusual for an aurora to make its first appearance after midnight.

Auroral warning networks

In the past, very effective auroral warning networks have been organised, both in this country and on the Continent, using the public telephone systems. The warnings are generally initiated by a northern monitoring station who either has firm evidence that an auroral opening has commenced, or who has good reason to believe that one is imminent. Each recipient of the warning passes it on to at least two other stations, working down from north to south, with built-in checks to hold the procedure if the event is a small one or fails to materialise.

At the time of writing it is not known if such a network still exists. It needs to be emphasised, however, that the

Table 3.15. Components list for converter

C1, C4	6p8 ceramic	R1	5k6
C2, C5, C11, C24	12p tubular trimmers	R2	18k
C3, C16	0p5 ceramic	R3, R5	100R
C6	390p ceramic	R4, R9	1k
C7	1–10p tubular trimmer	R6	82k
C8, C10, C18, C22	1000p feedthrough capacitors	R7	2k2
C9, C21	1000p ceramic	R8	68R
C12	4p7 ceramic	R10	4k7
C13	33p ceramic	R11	6k8
C14	22p (to suit IFT L4)	All ¼ or 1/10W carbon	
C15	3300p disc ceramic		
C17	0µ01 15V	TR1	BF324
C19	180p ceramic	TR2	2N3819, J309
C20	18p ceramic	TR3	2N3904
C23	0µ22 15V		
L1	10t 3/8in ID 18 SWG enam close-wound tapped 2t up		
L2	As L1 but tapped 4t up		
L3	11t 3/8in ID 18 SWG enam close-wound tapped at 1t and 5t from cold end		
L4	IF transformer appropriate to crystal chosen		
L5	6t 20 SWG on ¼in slug tuned form (for 65MHz crystal)		
L6	9t ¼in ID spaced wire diameter		

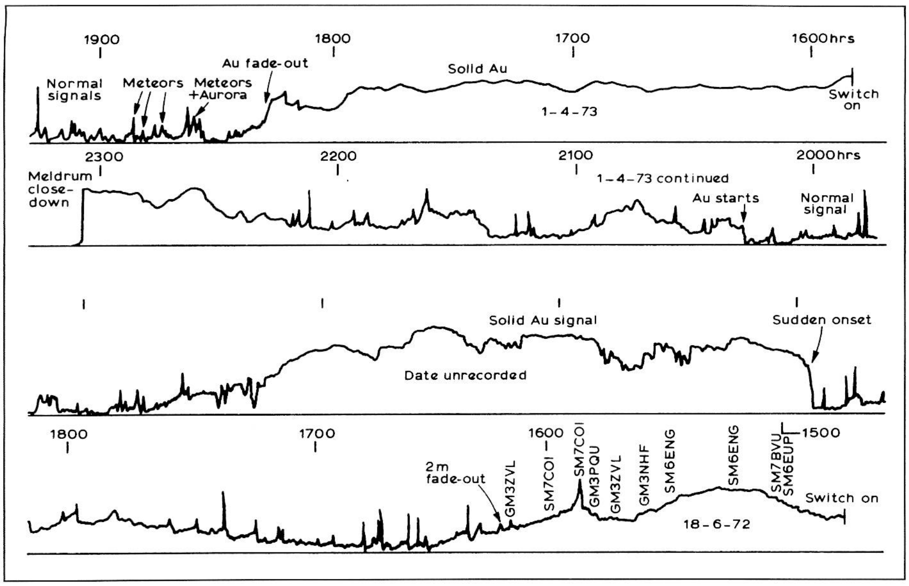


Fig 3.23. Some typical auroral signal recordings

capabilitiesof the system are such that a large-scale aurora can stimulate a level of activity amounting to several hundred transmitters being brought on the air during the early stages of the event. This cannot be matched anywhere outside the Amateur Service and offers a unique opportunity to put the service's talents to good use by doing no more than reporting its successes as soon as possible after the event.

TRANS-EQUATORIAL PROPAGATION

An aspect of research in which radio amateurs can justifiably claim to have played a major part is in the field of trans-equatorial propagation, TE or TEP for short. From its discovery just after the Second World War (between stations in Mexico and Argentina, reported in *QST* for October 1947) to the present day, amateurs have provided almost all the raw material for subsequent study. Some dedicated operators have spent 20 years or more setting up series of carefully controlled experiments designed to test theories proposed or to provide fresh material for consideration. For a particularly useful survey of progress to date of circuits between Europe and southern Africa, reference should be made to an article by ZE2JV and 5B4WR entitled 'Twenty-one years of TE', which appeared in *Radio Communication* June/July, August 1980, pp626–634 and 785–788. Some of the areas of the world which have contributed to the present knowledge of the mode are shown in Fig 3.24, where it will be seen that a prime requirement appears to be that the transmission path shall have the magnetic zero-dip equator (a) approximately at its mid-point and (b) nearly normal to it. The placing of the KP4/ZD8 path with respect to the change of direction of the zero-dip line over the Atlantic Ocean area is particularly interesting evidence in support of requirement (b).

The mode was first observed on 50MHz. Subsequent work has used 28, 50, 144 and 432MHz. During years of high sunspot activity the reliability of a TE path is considerable. On 50MHz the peak time during an opening was found to be 1845–1900. At 144MHz, using 100W RF into 16-element long Yagis, openings between Europe and southern Africa have lasted for up to two hours, centred on 2000 local time in Cyprus; high solar flux and low geomagnetic activity seem essential. Detrimental effects of geomagnetic storms are less evident at 50 and 28MHz than at 144MHz. Fig 3.25 shows the days on which TE signals were observed at 144MHz, plotted on a solar rotation base map, which reveals some tendency towards 27-day recurrences, particularly around the time of the equinoxes. On the American paths, peak occurrences at the equinoxes were noted.

On the Zimbabwe/Cyprus path there was a decline at that time, thought by ZE2JV and 5B4WR to be a peculiarity of the path connected in some way with the southern Africa magnetic anomaly which gives rise to high dip angles at the southern end.

In February 1979, ZS6DN, Pretoria, and SV1AB, Athens, held the world record for a 144MHz contact by the ionosphere, but at time of writing the current holders are ZS3B and I4EAT, who, on 31 March 1979, established transmission and reception in both directions over a distance just under 8000km.

Fading and chopping occur on the signals at rates which increase with transmission frequency. Slow chopping on 28 and 50MHz sometimes makes it almost impossible to read Morse code. At 144MHz the chopping rate is much faster, making the signal sound rough with an apparent raw AC note. Frequency spreading has been observed to 2kHz or more. The character of the signals may change considerably from day to day and from hour to hour in a random manner. Under the

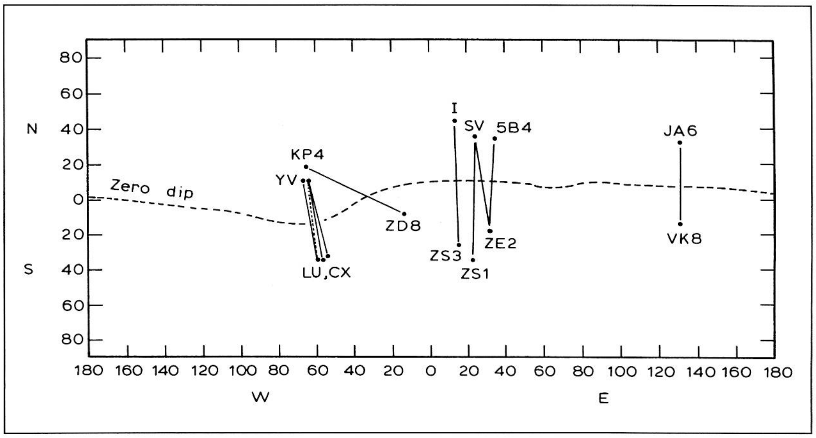


Fig 3.24. Areas of the world where trans-equatorial propagation has been observed

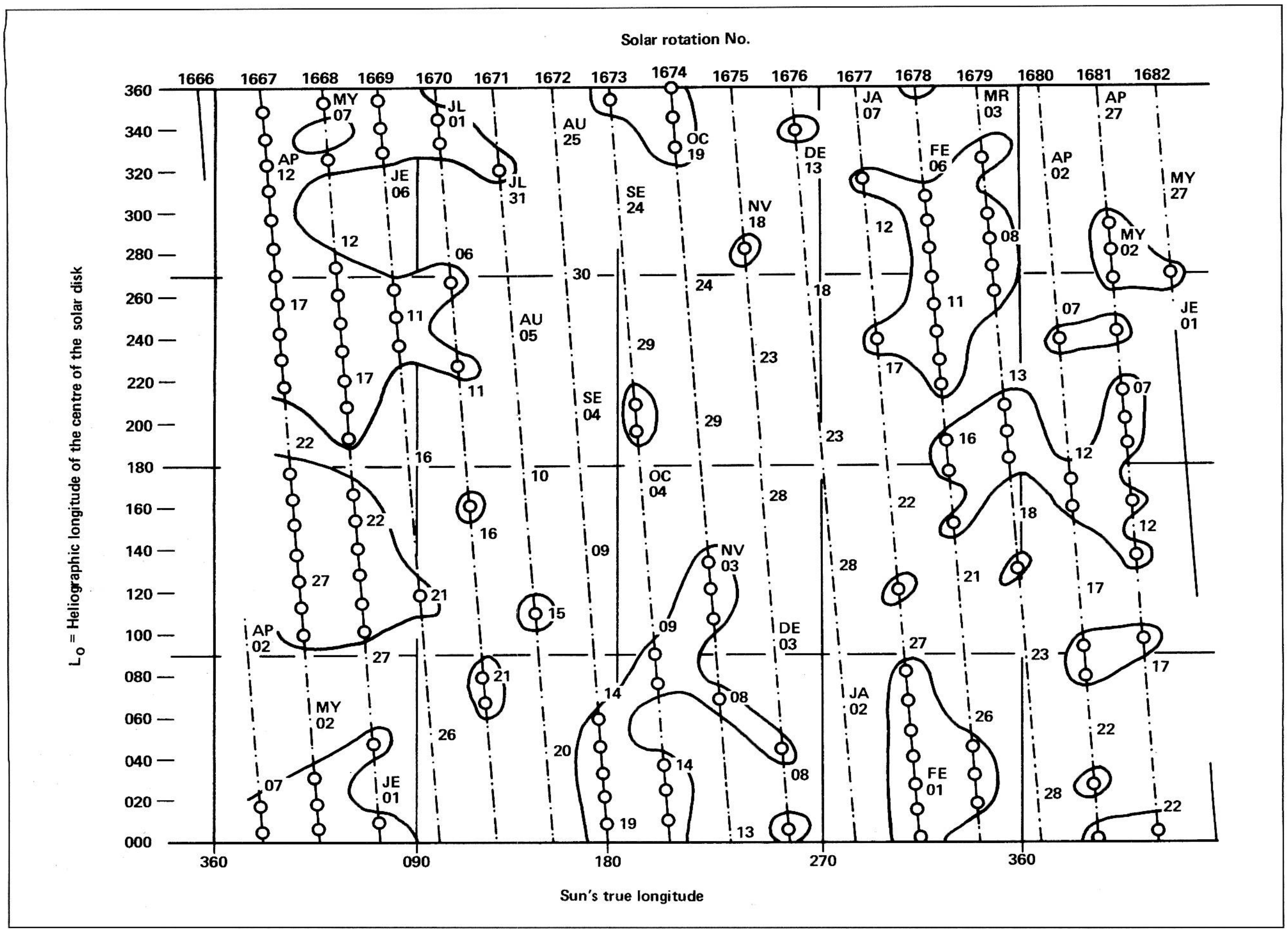


Fig 3.25. Trans-equatorial propagation. Reception of ZE2JV by 5B4WR (144.166MHz), 1630–1900GMT between 1 April 1978 and 31 May 1979

best of conditions 144MHz SSB is just intelligible. At other times the spread is so wide and the flutter is so rapid that no beat note can be obtained with the received signal, which then appears merely as a change in the background noise. At 432MHz the Zimbabwe beacon operated by ZE2JV was heard in Athens by SV1DH and SV1AB between 1816 and 1830GMT on 20 March 1979, and on 13 May 1979 by 5B4WR. Their comments were that the signals were rougher than on 144MHz and spreading more in frequency.

Time-delay measurements made in both directions along the Zimbabwe/Cyprus circuit showed afternoon intervals which at times corresponded with two-hop F2-layer propagation, but evening delays took about 10% longer, which may have been due to an extra ray-path distance of some 600km each way, or was in some way a function of the propagation mechanism. There appears to be no difference in delay time between 28MHz and 144MHz TE; although the character of the received signals differ on account of differences in fading rate.

Table 3.16 shows the European distance records for trans-equatorial propagation as they stood at the beginning of 1997. The mode does not normally extend into UHF.

Table 3.16. European VHF and UHF records (as at the beginning of 1997) – TEP

50MHz	G4IGO (IO80NW) and CE8BHI (FD46)	2 November 1991	13,203km
144MHz	I4EAT (JN54VG) and ZS3B (JG73)	30 March 1979	7843km

Source: John Morris, GM4ANB, for Region I, IARU.

METEOR TRAIL PROPAGATION

Propagation is also possible on an intermittent basis by means of scatter from short-lived trails of ionisation which appear as a result of small particles of solid matter entering the Earth's atmosphere and becoming heated to incandescence by friction. They are usually accompanied by streaks of light, popularly known as 'shooting stars'. These *meteors* (strictly the term applies only to the visible streak, although most writers use it as though it refers to the object itself) fall into two general classes, *shower meteors,* which follow definite and predictable orbits, and *sporadic meteors,* which follow individual paths and are present at all times.

Both the ionisation and the visual display occur simultaneously at heights of around 85 to 120km. Most of the objects responsible are no bigger than a grain of sand and they burn up completely in the upper atmosphere. Occasionally larger ones survive the descent and examples of some which have reached the ground are to be seen in museums, where they are referred to as *meteorites.*

Numbers vary during the year from a maximum in July to a minimum in February, with a ratio of about 4:1. There is a marked diurnal variation, due to the combined motions of the Earth's rotation and its movement around the Sun, leading to a maximum at 0600 local time and a minimum at 1800.

Table 3.17. European VHF and UHF records (as at the beginning of 1997) – meteor scatter

50MHz	G4IGO (IO80NW) and SV1OE (KM17VX)	12 August 1990	2542km
70MHz	GJ3YHU (LN89XI) and GM3WOJ/P (IO89KB)	12 August 1982	1083km
144MHz	GW4CQT (IO81LP) and UW6NA (KN97VE)	12 August 1977	3101km
432MHz	SM2CEW (KP15CR) and PAODZL (JO21HM)	12 August 1989	1869km

Source: John Morris, GM4ANB, for Region I, IARU.

The initial trail of ionisation is in the form of a long, thin, pencil-like cylinder, perhaps 15 to 20km in length. As soon as it is formed it begins to expand radially and to move with the various motions of the air through which it passes. The length of time when the trail is capable of supporting communication is generally very short, often less than a second, although longer persistences of a minute or more occur from time to time. The durations (and the frequency of occurrence) decrease with increase in signal frequency.

Considerations of phase coherence lead to an aspect sensitivity which favours radiation meeting the trail axis at right-angles. Because of this only a small part of any trail acts as a reflector and the orientation of the trail relative to the antennas is of considerable importance because it determines the height and position of the main reflection point. Meeting the right-angle requirement from both ends of a transmission path demands that the trail must lie in such a way that it is tangential to an ellipsoid of revolution having transmitter and receiver antennas at the focal points, and if ionisation is to result this condition has to be met at a level which is within 80 and 120km above the ground. It follows that large numbers of meteors enter the Earth's atmosphere along paths which can never satisfy the tangent condition within the prescribed limits of height and in consequence do not contribute to propagation along a given path.

These requirements suggest that it is unwise to direct very narrow-beam transmitting and receiving antennas (to be used for meteor-scatter work) along the strict line-of-sight between the stations. The only trails which can be tangential to the ellipsoid of revolution in that direction are those lying parallel to the ground, and this is an unlikely attitude to be taken up by a solid body entering the Earth's atmosphere from interplanetary space. The most likely beam directions lie a few degrees to one side or the other of the direct transmission path (both antennas must be deflected towards the same side, of course), and the optimum headings may have to be determined by careful experiment. Where the antennas are less directional the great-circle path between stations may be found to give best results, however, because, although little is likely to be received along the direct path heading, the acceptance angles of the antennas may be wide enough to include the longer but more likely paths on both sides.

The short bursts of signal which result from MS (meteor scatter) can best be observed on stations situated 1000 to 2000km away. In southern England several hundred examples per hour used to be heard carrying signals from the 40kW FM broadcast transmitter at Gdansk, Poland on 70.31MHz.

Table 3.17 shows the European VHF and UHF distance records by meteor scatter, as of the beginning of 1997. It is interesting to note that the record has been set on the same date (but in different years) for each of the four bands. August 12 is the date on which the Perseids shower reaches a peak.

A very detailed treatment of the subject was published in the February 1975 issue of *Radio Communication,* under the title 'VHF meteor scatter propagation'. In it J D V Ludlow, GW3ZTH, has given details of suitable equipments for receiving and recording signals propagated by this mode, and has provided a practical method of calculating beam headings and optimum times in respect of particular showers. For those wishing to follow up the relevant theory there is a very useful list of 25 references. Details of the current IARU Region 1 QSO procedure are given in the RSGB *Amateur Radio Operating Manual.*

Table 3.18 gives a summary of the main meteor showers likely to be of use to stations in the northern hemisphere. Full details, including times of transit and the directions from which the trails appear to radiate (an effect of perspective) appear each year in the current *Handbook of The British Astronomical Association* (see the Bibliography at the end of this chapter).

Commercial use is made of meteor-scatter propagation, particularly at high latitudes where it provides a hedge against the effects of polar cap absorption. The best-known system is the Janet Project of the Canadian Defence Research Board, which is described by G W L Davis and others in *Proceedings of the IRE* December 1957. Another, and later, application of the technique was seen in the 'Snonet' meteorological meteor burst system for collecting observations from remote sensors. Information recorded on magnetic tape at normal speed was played back (and transmitted) in the form of high-speed bursts when a monitor on a slightly different frequency showed that a path was open.

In more recent times the availability of satellite channels has weakened the attraction of meteor scatter for commercial data communications. On the amateur bands, however, interest continues to grow and very detailed logs from stations all over Europe appear in every quarterly issue of the German *DUBUS* magazine. For example, issue 3/96 contained 15 closely packed pages of reports, followed by two pages of operating procdures for meteor scatter QSOs, as adopted by IARU, Region 1. It is a very useful source of data for research purposes.

SATELLITE PROPAGATION EXPERIMENTS

A number of amateurs have suggested propagation experiments using satellites in ways that were not envisaged as part of the basic projects.

Pat Gowen, G3IOR, made use of Oscar 7 and Oscar 8 in its Mode A transponder configuration (145MHz up, 29MHz down) as a guide to conditions on 144MHz. Good tropospheric conditions were indicated by severe attenuation of one's own

Table 3.18. Calendar of the main meteor showers

Start	Dates of Maximum	End	Name	Comparative rate*	Transit Time	Transit Elev
Jan 01	Jan 03	Jan 06	Quadrantids	6	09	90
Apr 19	Apr 21	Apr 24	April Lyrids	3	04	70
May 01	May 05	May 08	Eta Aquarids	3	08	40
Jun 10	Jun 16	Jun 21	June Lyrids	2	01	70
Jun 17	Jun 20	Jun 26	Ophiuchids	2	23	20
Jul 10	Jul 26	Aug 15	Capricornids	2	01	20
Jul 15	Jul 27	Aug 15	Delta Aquarids	4	02	30
Jul 15	Jul 31	Aug 20	Pisces Australids	2	02	10
Jul 15	Jul 30	Aug 25	Alpha Capricornids	2	00	30
Jul 15	Aug 06	Aug 25	Iota Aquarids	2	01	30
Jul 25	Aug 12	Aug 18	Perseids	5	06	80†
Aug 19	Aug 21	Aug 22	Chi Cygnids	1	21	90
Oct 16	Oct 21	Oct 26	Orionids	4	04	50
Oct 20	Nov 08	Nov 30	Taurids	3	01	60
Nov 07	Nov 09	Nov 11	Cepheids	2	20	80†
Nov 15	Nov 17	Nov 19	Leonids	2	06	60
Dec 07	Dec 14	Dec 15	Geminids	5	02	70
Dec 17	Dec 22	Dec 24	Ursids	1	08	60†

* Each step on the comparative rate scale represents a factor of 2.
† Above northern horizon.

returned signal, with deep and rapid fading when the satellite was just above the horizon. Brief and rapid 'pop-ups' of signal before and after predicted times of access were caused by scintillation of 144MHz uplink as it passed through tropospheric ducts. Sporadic-E effects were similar, but they may have taken place at quite high elevations. Fading suggested the presence of multiple diffraction paths in the ionosphere. Aurora caused marked degradation of tone on returned signals from some of the northern stations, often specific to small areas.

This topic has been brought back into prominence recently in connection with 136MHz aircraft-to-satellite-to-ground communications, 250MHz military satellite applications and global positioning systems working around 1.2 to 1.6MHz. (See, for example a paper by Dr J Aarons in *IEE Conference Publication* No 411.)

John Branegan, GM4IHJ, kept a regular check on where satellite scintillation occurred on polar paths and used it to define the instantaneous location of the auroral oval to provide an estimate of the total electron content along the satellite line of sight. In this way, very high electron densities were observed at heights above the normal 110km auroral reflection zone, considered capable of scattering frequencies of up to at least 250MHz. GM4IHJ, G3IOR and several Alaskan stations all heard double signals on 144MHz satellite transmissions, the second signal some 750Hz from the nominal frequency, sometimes with an auroral tone.

G4DGU and SM6CKU investigated the idea of using a large low-orbit satellite as a passive reflector. The relevant theory suggested that the total path length loss might be about 10dB better than moonbounce. There were problems – low-orbit satellite orbits soon decay and they could not be predicted with accuracy. There were very high Doppler shifts involved. At 432MHz signals are likely to appear first 12kHz high, shifting down at about 1kHz per second to become 12kHz low, six to eight seconds later, when signals disappear. Despite these difficulties, however, a four-second burst of SSB from G4DGU had been received by SM6CKU, 10dB over noise, off a Cosmos third-stage launcher. The transmitter used was of 400W PEP into an array of eight 17-element Yagis. The receiver had been coupled to an 8m dish.

SOLAR/GEOPHYSICAL CONNECTIONS

Precursors

The relationship between events on the Sun and the associated effects in the Earth's ionosphere has been dealt with at length in the propagation chapter of the *Radio Communication Handbook*. Suffice it to record here that the main solar events of interest to VHF and UHF operators are solar flares, radio bursts and emissions of high-energy protons. Solar flare effects appear in three time scales. Within about eight minutes of a major flare suitably placed on the Sun, electromagnetic radiation brings about increased D-region ionisation, sudden cosmic noise absorption (SCNA), short-wave fadeout phenomena (SWF), sudden phase anomaly of ELF signals (SPA), sudden enhancement of atmospherics at LF (SEA), sudden frequency deviation at LF (SFD), and noise bursts (not all of these will be detected in one particular event). Within an hour of the appearance of the flare, high-energy corpuscular radiation is likely to cause polar cap absorption (PCA) of anything from 20 minutes to 20 hours duration, which may extend to VHF. Some 20–40 hours after the flare, low-energy corpuscular radiation brings about magnetic storms, ionospheric storms and auroras, which may persist for more than a day.

The probability that this chain of events will lead to a magnetic storm and an aurora is greatest when the solar activity occurs when the region concerned is near, but slightly beyond, the central meridian.

There is a tendency for solar events to recur after periods of approximately 27 days, as has been discussed in connection with auroral propagation. Table 3.19 shows the starting dates of all solar rotations between 1991 and 2002 calculated in continuation of Carrington's photo-heliographic series, using a method described by the Belgian amateur astronomer Jean Meeus in the journal *Ciel et Terre*.

Table 3.19. Solar rotation calendar, 1991–2002

1991		1992		1993		1994		1995		1996	
1838	JA16	1851	JA05	1865	JA21	1878	JA11	1892	JA28	1905	JA17
1839	FE12	1852	FE02	1866	FE17	1879	FE07	1893	FE24	1906	FE13
1840	MR11	1853	FE29	1867	MR17	1880	MR06	1894	MR23	1907	MR12
1841	AP08	1854	MR27	1868	AP13	1881	AP03	1895	AP19	1908	AP08
1842	MY05	1855	AP23	1869	MY10	1882	AP30	1896	MY17	1909	MY05
1843	JE01	1856	MY21	1870	JE06	1883	MY27	1897	JE13	1910	JE01
1844	JE28	1857	JE17	1871	JL04	1884	JE23	1898	JL10	1911	JE29
1845	JL25	1858	JL14	1872	JL31	1885	JL20	1899	AU06	1912	JL26
1846	AU22	1859	AU10	1873	AU27	1886	AU17	1900	SE03	1913	AU22
1847	SE18	1860	SE07	1874	SE23	1887	SE13	1901	SE30	1914	SE18
1848	OC15	1861	OC04	1875	OC21	1888	OC10	1902	OC27	1915	OC16
1849	NV12	1862	OC31	1876	NV17	1889	NV07	1903	NV23	1916	NV12
1850	DE09	1863	NV27	1877	DE14	1890	DE04	1904	DE21	1917	DE09
		1864	DE25			1891	DE31				
1997		**1998**		**1999**		**2000**		**2001**		**2002**	
1918	JA06	1932	JA23	1945	JA12	1958	JA02	1972	JA17	1985	JA07
1919	FE02	1933	FE19	1946	FE08	1959	JA29	1973	FE14	1986	FE03
1920	MR01	1934	MR18	1947	MR08	1960	FE25	1974	MR13	1987	MR03
1921	MR29	1935	AP14	1948	AP04	1961	MR24	1975	APO9	1988	MR30
1922	AP25	1936	MY12	1949	MY01	1962	AP20	1976	MY07	1989	AP26
1923	MY22	1937	JE08	1950	MY29	1963	MY17	1977	JE03	1990	MY24
1924	JE18	1938	JL05	1951	JB25	1964	JE13	1978	JE30	1991	JE20
1925	JL16	1939	AU01	1952	JL22	1965	JL11	1979	JL27	1992	JL17
1926	AU12	1940	AU29	1953	AU18	1966	AU07	1980	AU24	1993	AU13
1927	SE08	1941	SE25	1954	SE14	1967	SE03	1981	SE20	1994	SE09
1928	OC05	1942	OC22	1955	OC12	1968	SE30	1982	OC17	1995	OC07
1929	NV02	1943	NV18	1956	NV08	1969	OC28	1983	NV13	1996	NV03
1930	NV29	1944	DE16	1957	DE05	1970	NV24	1984	DE11	1997	NV30
1931	DE26					1971	DE21				

The rotation numbers follow Carrington's series. The quoted date shows the first day on which a new rotation value of L_0 appears at 12UT.

A solar rotation diagram intended for use in connection with radio propagation studies appears each year in the *RSGB Yearbook*. It covers an 18-month period covering the year in which the *Yearbook* is current and the three-month periods immediately before and after.

Solar and geophysical data

There still remains much to be learned about the relationship between events on the Sun, their corresponding effect on the Earth's magnetic field and VHF/UHF radio propagation via the ionosphere as in sporadic-E, auroral-E and other similar 'openings' on the amateur bands. This is an interesting field for individual research but it requires access to basic solar and geophysical data. Four useful sources are suggested here.

The first, the easiest to obtain, is supplied by the brief summary of solar and magnetic trends over the previous week, together with a forecast covering the week to come, which is compiled from authoritative sources and included each week in the RSGB GB2RS news bulletin.

These are transmitted according to a schedule published regularly in *Radio Communication*. The summaries are compiled by Neil Clarke, G0CAS, who also edits a monthly 16-page booklet, which he calls *SunMag*, containing tables of relevant data, together with graphs and diagrams, as appropriate. For subscription details write to Neil at 39 Acacia Road, Cantley, Doncaster, DN4 6UR (Tel: 01302 531925). An explanation of the terms used may be obtained by post from RSGB Headquarters or through the RSGB Web Page at http://www.rsgb.org.

The second source of information is by far the most detailed, but it is subject to three to seven months' delay, which can seem a long time when waiting to analyse a particular period of observation. The material is to be found in the publication *Solar Geophysical Data*, produced in the USA by the Environmental Data Service of the National Oceanic and Atmospheric Administration (NOAA), of Ashville, NC. This appears in two parts each month – Part 1 (Prompt Reports) containing preliminary data covering the two months prior to the date of issue, and Part 2 (Comprehensive Reports) containing more-detailed data centred about six months prior to the date of issue. In this country the monthly parts arrive by post about the middle of the month following that shown on the cover.

For a trial inspection it is suggested that the most recent February issue (consisting of Part 1, Part 2 and a very detailed explanation booklet distributed annually at that time) should be requested through a library. In case of difficulty an approach should be made to a library specialising in science subjects.

Daily values of K_p, C_i, C_p, A_i, and other magnetic data, including an inferred interplanetary magnetic field indication derived from satellite observations, appear after a delay of about four months in the American publication *Journal of Geophysical Research (Space Physics)*.

Amateur radio-astronomical observations are co-ordinated by the British Astronomical Association and an edited summary appears as a regular feature in their *Journal*.

The third source ought to have been the data broadcaster GAM1, which was to have been operated by the Amateur Service to provide daily messages containing various solar and geophysical information obtained from 'official' sources, such as Meudon Observatory, Paris, and Boulder, Colorado. Table 3.20 shows the format of summary messages that used to be available 'over the air' (but no longer, unfortunately). It will serve here to indicate the type of information it was hoped to cover. After years of trying it now seems that we cannot get the authorities to allow the service to go ahead in the way we had hoped.

All is not lost, however. At the time of writing tests are being carried out by G4FKH which it is hoped will lead to an alternative approach. If the outcome is favourable no doubt full details will be printed in *Radio Communication*, but it is too soon to be able to give any reliable indications here.

The fourth source is the Internet/World Wide Web, but here again it may be unwise to attempt to be specific. However, it is clear that a great deal of information may be obtained from official sources if you know how to reach them. Neil Clarke, G0CAS, may be able to help you there. His e-mail address is

Table 3.20. Format of a typical GEOALERT message

	GEOALERT CCCNN DDHHMMZ 9HHDD 1SSSG 2FFFB 3AAAE 4//// 5MMXX QXXYY nnijk (QXXYY nnijk . . .) (Plain-language details of major optical flares and tenflares) 8hhdd 7777C QXXYY degree of activity . . . (Plain-language forecast of activity) SOLALERT JJ/KK MAGALERT JJ/KK
	Note: There is no fixed length to a GEOALERT message, groups being repeated as often as necessary.
	Key
CCCNN	Originating centre (MEU = Meudon, WWA = Boulder); serial number of message
DDHHMMZ	Date and time of origin of message; Z = GMT.
9HHDD	Indicates that various daily indices follow, for 24 hours ending at HH hours on DD day of month.
1 SSSG	Indicates sunspot number, SSS, and number of new groups observed, G.
2FFFB	Indicates 2800MHz solar flux value FFF, and number of important bursts.
3AAAE	Indicates geomagnetic activity, AAA = A_k value; E = events (0 = no events, 1 = end of magnetic storm, 2 = storm in progress, 6 = gradual storm commencement, 7 = sudden storm commencement, 8 = very pronounced sudden storm commencement).
4////	Indicates cosmic ray data (not used on Meudon message).
5MMXX	Indicates flare counts: MM = daily total of M flares, XX = daily total of X flares.
Then follow groups identifying active regions on the Sun:	
QXXYY	Q = Quadrant of the Sun (1 = NE, 2 = SE, 3 = SW, 4 = NW), XX = degrees of longitude, YY = degrees of latitude, relative to the centre of the Sun's visible disc.
nnijk	nn = total number of flares in active region indicated; i = number of flares greater than importance 1; j = number of M flares; k = number of X flares (in region QXXYY).
8hhdd	Indicates 24-hour forecast follows, starting at hh hours on dd day of month.
7777C	Indicator, C = types of observation used in forecast (1 = solar radio, 2 = partial solar optical, 3 = optical and radio, 4 = all, plus solar magnetic measurements).
QXXYY	Positions on the Sun, coded as before.
JJ/KK	Days of month between which the solar or magnetic alerts apply

neil@g0cas.demon.co.uk ; his postal address and telephone number have been given earlier.

The June 1997 issue of *Radio Communication* carried a reference to Solar Warning and Real-time Monitor (SWARM) software which may be downloaded from http://solar.uleth.ca/solar/www/swarm.html and which is said to provide current real-time solar and geomagnetic data.

Geomagnetic data

Solar events and associated ionospheric disturbances are of less direct interest to the VHF/UHF operator than are geomagnetic variations, which correlate well with auroral events and, though to a lesser extent, with E_s and TE.

These are comparisons which fall within the scope and capabilities of interested radio amateurs, but many find it difficult to extract the necessary, elementary information from conventional manuals on the subject.

In the studies carried out on auroral propagation by the RSGB the region of interest is admirably represented by a knowledge of the performance of the magnetometers at Lerwick Observatory, Shetland, which location, at over 60° north latitude, often experiences spectacular displays of visual aurora. Valuable experience was gained there during the International Quiet Sun Years, 1963–4, when the GB3LER beacon transmitters were set up nearby and direct comparison of results against the magnetometer records became possible. The remainder of this section is intended to provide an introduction to the subject to those who take part in the Society's auroral observation work, which forms an important part of its Propagation Studies programme.

The terrestrial magnetic field at the Earth's surface is not constant, but is subject to both long-term and short-term variations in intensity ranging from periods of centuries or more to hours, minutes or even less. The transient variations are small in comparison to the total field; they are measured in gammas, which are equal to 10^{-5} gauss, the force being determined in terms of three components mutually at right-angles, either in the directions X (geographic north), Y (east) and Z (vertically downward), or H (horizontal intensity), D (declination) and Z.

The main variometers at Lerwick record H, D and Z photographically on a sheet of sensitised paper approximately 40cm by 30cm, on which all three traces appear side-by-side, together with timing marks every five minutes, and suitable baselines. The present sensitivities are such that H changes of 3.45 gamma, D changes of 0.94 minutes of arc (corresponding to 4 gamma at right-angles to the meridian) and Z changes of 4.35 gamma move their respective traces by 1mm. A system of prisms ensures that any trace which approaches the edge of its section of the chart appears again from the other side, thereby extending the effective width so as to be able to handle the widest excursions.

The traces exhibit two features, one a fairly regular diurnal 'background' change due to solar and lunar effects, the other a superimposed irregular, and often violent, variation, the extent of which depends on particle radiation from the Sun. It is necessary to examine initially a large number of traces obtained during magnetically quiet periods in order to be able to assess the appearance of the 'normal' diurnal curve, which has the form of a shallow letter 'S' on its side, and allowance has to be made for season, solar flare effects and certain decreases which follow a magnetic storm. The sum of the highest positive and negative departures from the 'normal' curve are converted into a quasi-logarithmic scale of *K*-figures, where the actual values for the lower limit of each number vary from one observatory to another depending on the magnetic latitude. For Lerwick the scale is given in Table 3.21.

Table 3.21. *K*-figures for Lerwick

0	0 to 10	5	140 to 240
1	0 to 20	6	240 to 400
2	20 to 40	7	400 to 640
3	40 to 80	8	640 to 1000
4	80 to 140	9	1000 or more

At other observatories the ranges are proportional, and may be found from the value assigned for the lower limit of $K = 9$, which will be quoted as being either 300, 350, 500, 600, 750, 1000, 1200, 1500 or 2000 gamma.

Using this K-scale, the degree of activity is described for each directional component of the force during eight three-hourly periods of each day, and the highest of the three numbers for each period are grouped in two sets of four digits, separated by Greenwich noon, beginning with the period 0000 to 0300GMT.

A combination of K-figures from 12 widely spaced observatories (of which Lerwick is one) results in the planetary K-index, K_p, prepared monthly by the Committee on Characterisation of Magnetic Disturbances at the University of Göttingen, Germany, which is often preferred to single-station data in analytical work, particularly in connection with the ionosphere or when purely local effects are unwanted. The normal 0 to 9 scale is expanded into one of thirds by the addition of suffixes, eg 2–, 2o, 2+, 3–, 3o, 3+ . . .

A daily magnetic character figure, C, has been in use for over half a century, each observatory subscribing a figure descriptive of their assessment of the day's activity: 0 if it was judged quiet, 1 if it was moderately disturbed or 2 if it was very disturbed. The individual figures are rarely used, but an index C_i, the average to one decimal place of C-figures from a worldwide network of collaborating observatories, provides a convenient classification of daily activity. Another, apparently similar, character figure, C_p, is prepared directly from the K_p indices but, although its derivation is so different, it rarely differs from C_i by more than 0.2. To simplify machine tabulation the scales are sometimes expressed in terms of yet another, known as C_9, which uses whole numbers from 0 to 9 in place of the decimal range from 0.0 to 2.5.

The sum and arithmetic mean of the eight three-hourly K-figures provide further expressions of daily activity which are simple and convenient to obtain. They are not ideal, however, because the K-scale is a logarithmic one (as is the decibel scale), and the arithmetic average gives the logarithm of the geometric mean and not the logarithm of the arithmetic mean. To take an extreme example, consider the two series 1111 1111 and 0000 0008, both of which give a sum of 8 and a mean of 1; the first would be representative of a quiet day, whereas the second would be considered a highly disturbed day. For this reason it is preferable to turn each K-index back into an equivalent range, a_k, on a linear scale, by using the corresponding values:

K	0	1	2	3	4	5	6	7	8	9
a_k	0	3	7	15	27	48	80	140	240	400

which may be summed and meaned arithmetically to represent activity over a period. It should be noted that the same table used for all observatories, irrespective of their actual K-scales so that the resulting standardised figures are not the true gamma ranges, although those may be approximated from them, if required, by the use of a factor (which in the case of Lerwick is 4). The *daily amplitude* A_k is the average of the eight values of a_k for the day. In the case of the two examples cited above as leading to the same K-figure sum, the first, 1111 1111, gives an A_k of 3, whereas the second, 0000 0008, produces an A_k of 30, thereby reflecting the vastly differing states of activity.

A similar, but expanded, scale relates the planetary three-hour index K_p to the three-hourly *equivalent planetary amplitude*, a_p:

K_p	0o	0+	1–	1o	1+	2–	2o	2+	3–	3o	3+	4–	4o	4+	5–
a_p	0	2	3	4	5	6	7	9	12	15	18	22	27	32	39

K_p	5o	5+	6–	6o	6+	7–	7o	7+	8–	8o	8+	9–	9o
a_p	48	56	67	80	94	111	132	154	179	207	236	300	400

The *daily equivalent planetary amplitude,* A_p, *is* the average of the eight values of a_p for the day.

ACCURATE TIME RECORDING

When selected frequencies need to be monitored for propagation studies, some form of recording is essential so that the operator can attend to more productive things. It also results in a semi-permanent set of observations which can be transcribed and analysed as and when convenient.

Stereo cassette recorders provide a useful method of recording. The machines are relatively cheap because they are mass-produced and, as the tapes may be used again and again, running costs are low, particularly when compared with pen recorders using paper charts. The stereo facility provides the user with two synchronised (but entirely independent) channels, one of which may be used for signal data, the other for timing signals originating from one of the standard time and frequency radio transmissions.

For some purposes it is sufficient to record no more than the presence of the selected signal at a known time and for this the recorder may be connected to the output of the receiver in the conventional way. When an indication of the signal strength is required it is advisable to use some form of voltage-to-frequency or analogue-to-digital conversion before recording from either the AGC line or a special detector giving a suitable time-constant. This avoids the effects of differences between recording and playback, including the quality of the tape, from affecting the measurements.

Sometimes, to make more effective use of tape and to simplify the task of data reduction, discontinuous recording may be adopted. This can take the form of sampling, where, say, five minutes in every hour are transferred to tape and the transport mechanism is halted between-times, or by causing the tape to stop automatically in the absence of signals for more than a selected period. Both of these practices complicate the provision of radio time signals from the best-known sources, such as those which share 2.5, 5, 10 and 15MHz, because only minute and second intervals are available with no indication of absolute time.

MSF, Rugby, on 60kHz, carries time every minute in the form of a 0.5-second burst of pulse code modulation immediately before the minute indicator to which it refers. This could be recorded, together with successive one-second 'ticks' on one channel of a stereo recorder.

Nowadays very accurate time is a very cheap commodity. An analogue clock with a built-in decoder locked by radio to an atomic standard may be bought for under £25 from one of the big radio and electrical component chain stores.

For rapid access to a reliable time source try any teletext page from a terrestrial television station.

Finally, if you are using a computer as a real-time logging device, or to sample signal strengths through a suitable interface, then the internal clock is, or should be, a useful source

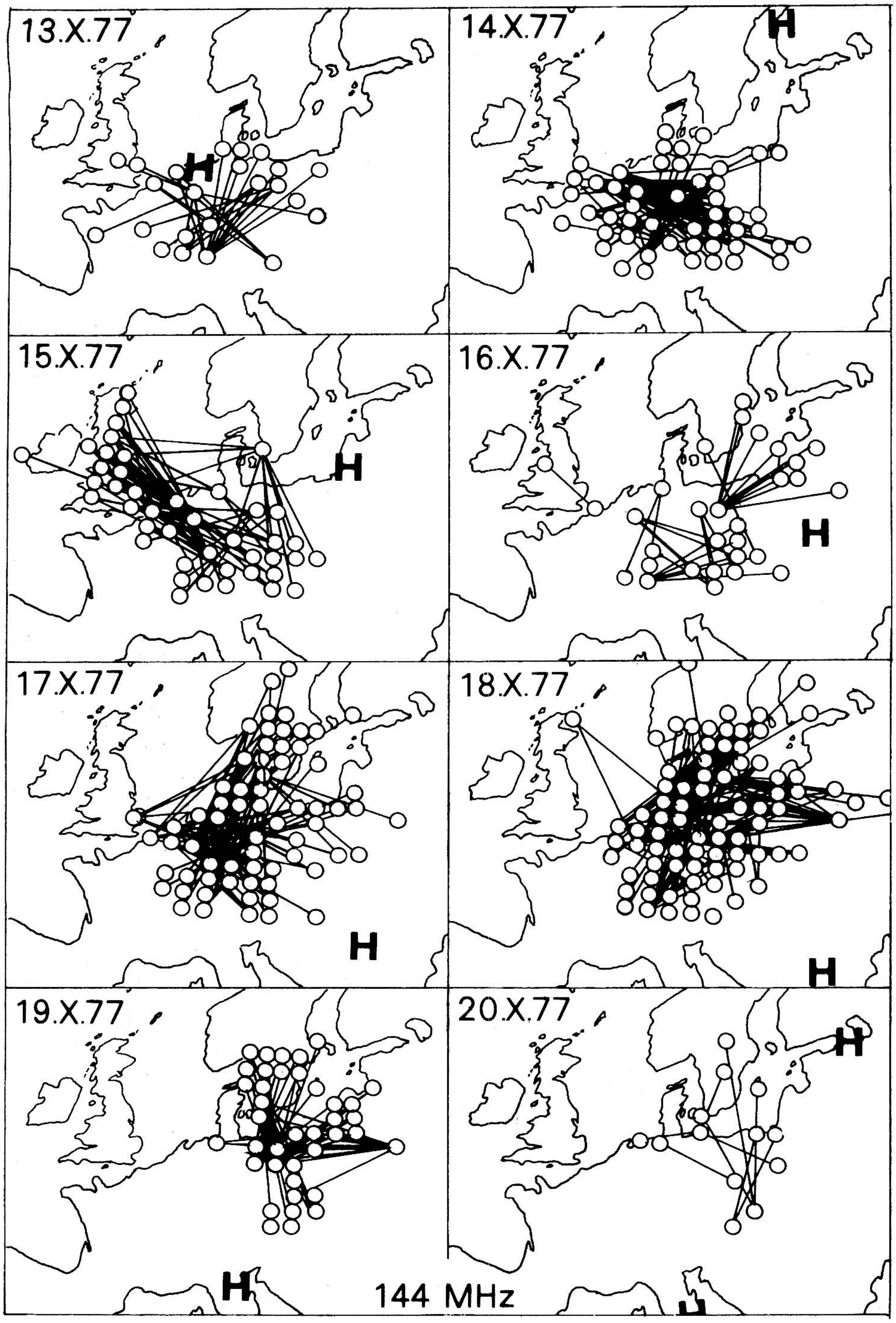


Fig 3.26(a). Analysis of 144MHz reports

organised groups of amateurs are particularly well placed to further our understanding of the physical processes involved in almost every mode other than direct line of sight.

There may never have been such an opportune time to put forward fresh evidence of the extent to which guidance given by such authorities as ITU and URSI to intending users of frequencies above 30MHz seriously underestimates the range and frequency coverage of the various modes in this part of the spectrum. The professionals acknowledge that such discrepancies exist and are looking to the Amateur Service for raw material that will help to bring present theory more into line with what is happening in the world outside.

There is a continuing need for long-term studies of over-the-horizon transmission paths, particularly those involving beacons. This type of work demands a degree of involvement that can be expected from only a very few individual amateurs, although it provides a very interesting project for a club, particularly one associated with an educational establishment, provided that continuity may be assured over long holiday periods and there is sufficient overall supervision to maintain the stability and calibration of the equipment.

Enough has been written already to show the value of studies which make use of a very high volume of simple reports from stations covering a large area. The mere fact that contact was made on a specified band, at a certain time, between two stations identified by callsign and QTH locator, is of little more than passing interest to either of the operators concerned. But when hundreds of such reports are collected and processed a powerful research tool emerges, and one, moreover, that is peculiar to the Amateur Service.

As an example of the capabilities of this technique, consider Figs 3.26(a) and (b), which were included in the IEE Conference paper referred to earlier in the section of signal strengths

that may be accessed along with the incoming data. But check it against a source known to be accurate before you rely on it.

AMATEUR NETWORKS AND FURTHER RESEARCH

It is hoped that the preceding pages of this chapter will have left the reader with the feeling that there is still much to be learned about radio propagation at VHF and UHF and that and ranges attainable using tropospheric propagation. The two sets of maps resulted from an analysis of many hundreds of individual reports, most of them collected in Germany by the DUBUS organisation. It is known that reporting amateurs were active in most parts of Europe during the whole of the period shown, and nearly all of them were within the boundaries of a very large anticyclone for a large proportion of the time, yet only a certain well-defined area was experiencing long-range anomalous propagation at any one time. That area is shown in the original paper to be where the steep-gradient boundary has formed between air of low refractive index, over air of high refractive index. Although subsidence is present elsewhere the sharp boundary is not present either because the turbulence in the lower atmosphere is too weak, allowing the low-refractive index air to reach ground level, or too strong, causing mixing at the interface and a consequent weakening of the gradient.

Observe how, in the course of the event, the centre of the anticyclone, marked with a letter 'H', moves from day to day, and how the axis of the really long-range paths rotates so as to maintain a broadside-on aspect relative to it. These paths form a chord across the curvature of the isobars (which had to be omitted from the diagrams in order to simplify them). In the last two maps the change in direction due to the approach of a fresh centre is of interest.

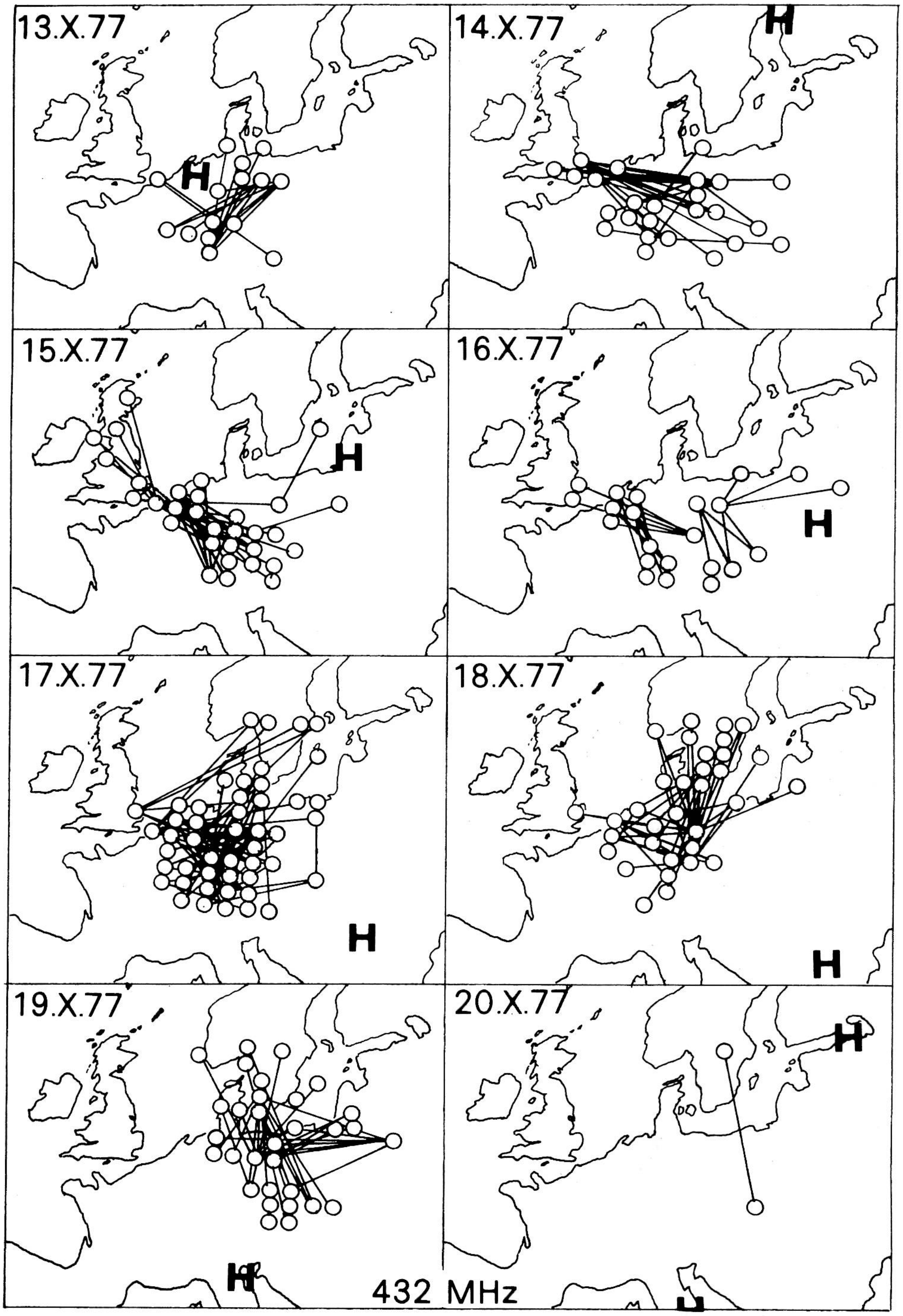


Fig 3.26(b). Analysis of 432MHz reports

Another conclusion to be drawn from the maps is that the area of enhancement is approximately the same at both frequencies, although the two sets of reports used are entirely independent of one another. Individual path lengths differed within the area, however, being roughly half as far again at 144MHz as compared to 432MHz.

It should be clear that these techniques, which require no more from operators than the reporting of contacts that may well have had nothing to do with propagation research in themselves, have enormous potential, not only in tropospheric propagation studies, but in VHF sporadic-E and auroral-E studies as well. It should be noted, however, that reports of bearings from both ends of the path are an important requisite of the latter.

It is only because of the truly international nature of amateur radio and the seemingly tireless enthusiasm of so many people, not the least those who collect the observations and make them available for studies such as the one described, that this technique can be employed to the full. Your reports may be the ones needed to complete the task.

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4 Receivers, transmitters and transceivers

THIS chapter is designed to give the reader a clear outline of the practical needs for receivers, transmitters etc in the VHF/UHF bands. As will be seen, these are different from those at HF and these differences will be identified. In receivers, the main difference is that, since external noise levels are much lower at VHF/UHF than at HF, the noise performance of the receiver must be that much better. The effect of stray capacitance and inductance on semiconductors is also important.

Note that devices such as ICs are mentioned only as illustrations and not as recommendations.

RECEIVER PARAMETERS [1, 2]

An ideal receiver would be stable, selective enough for the mode to be received and would convert an incoming VHF or UHF signal into an audio signal regardless of its strength or that of other signals whether or not they were close by in frequency. To do this it needs gain, stability, selectivity, a large dynamic range (see later) and detection. Needless to say, the ideal does not exist in practice but to approach it as closely as possible should be the goal of all amateurs, especially those wishing to read weak signals such as those from DX stations.

The factors causing deviation from the ideal are *noise, non-linearity, instability* and *gain compression.*

Noise

Noise is always present, and can come from within the receiver or from outside. The latter can be natural or man made and there is very little that can be done about it (but see below). Considering firstly the noise from within the receiver, a resistor R at any temperature above absolute zero (0K or –273°C) will generate a noise power w given by:

$$w = k.T.B$$

where w is in watts, k is Boltzmann's constant (a fundamental constant in physics/electronics) and B is the bandwidth in hertz. This is true for a resistor not carrying a current. Carrying a current will increase the noise and it will be worse for some types of resistor (eg carbon composition) than for others (eg metal film types).

Note that w is independent of the resistance of R. k is 1.38×10^{-23} watts per degree per hertz. Since k is very small, w is also small, eg at room temperature of 290K (17°C or 62.6°F) and a bandwidth of 2kHz, w is 8×10^{-18} watts or 8 attowatts. If the resistor has a value of 50Ω, the corresponding voltage is 0.02µV or 20nV.

Noise figure

This is often quoted along with the noise factor. Taking the latter first, it is defined as the signal-to-noise ratio at the input of a device divided by the signal-to-noise ratio at the output, ie:

$$\text{Noise factor (NF)} = \frac{S_{in} / N_{in}}{S_{out} / N_{out}}$$

and noise figure is simply the NF expressed as decibels, ie:

$$\text{Noise figure} = 10 \log \text{NF}$$

Noise is also generated by any device that dissipates energy. Pure capacitors and inductors do not generate noise but resistors, amplifying devices, semiconductors or valves do. If it is assumed that the noise due to the input resistance is doubled by the input amplifying device and added to further by noise from outside, a figure of about 0.1µV arises for 2kHz bandwidth. This sets a limit to the overall gain needed in the receiver. 0.1µV corresponds to 2×10^{-16}W. If an output of 1W is required the overall gain needs to be 5×10^{15} or 157dB. Less than this will be adequate since the full gain will produce 1W of noise. Narrow-band modes such as Morse code will allow less noise into the system and will therefore work with a smaller signal. However, they put a greater demand on frequency stability and selectivity.

Another way of looking at this is to consider a bandwidth of 1Hz at room temperature (290K) – the noise level will be –174dBm, ie –174dB relative to 1mW. This allows a very easy determination of signal-to-noise ratio under any circumstances when the various parameters are known. In a linear system such as for SSB, the signal-to-noise ratio is given by:

$$\text{S/N} = P_{in} + 174 - \text{NF} - 10 \log B \quad \text{dB}$$

where P_{in} is the input power in dBm, NF is the noise figure, and B is the bandwidth in hertz.

For example, a receiver with a 3dB noise figure with an input of 100nV (–127dBm assuming a 50Ω input impedance) will provide in an SSB bandwidth of 3kHz a S/N ratio of:

$$-127 + 174 - 3 - 35 = 9\text{dB}$$

This shows how important a narrow bandwidth is for ultimate sensitivity. However, since the effective noise bandwidth is only affected by the filtering to some 10 or 20dB down, the ultimate selectivity of high-performance crystal filters (ie the *stop-band* width, usually –60dB bandwidth) has little effect and simply from the requirements of detecting the smallest signal, stop-bands of 20dB are adequate.

Digital signal processing (DSP) at audio can provide an effective reduction of noise bandwidth greater than the actual bandwidth because DSP looks for correlation of the signal from sample to sample and so builds up the signal over time. This is because noise is essentially a random signal which tends to cancel over time while the wanted signal has some

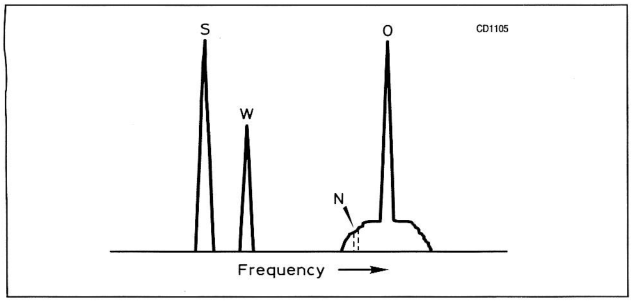

Fig 4.1. Phase noise. W is the wanted signal, S is a strong unwanted signal, O is the oscillator and N is part of the noise sideband of the oscillator. O – W = IF and N – S = IF noise

regularity. The details of DSP are rather complicated and will not be dealt with here.

In the case of FM, the output S/N ratio depends on the demodulator and the modulation index. As a guide, narrowband FM (NBFM) analogue voice systems can provide an output S/N ratio of 12dB (usually considered the minimum acceptable) with an input carrier S/N ratio of as little as 3dB.

Sometimes the term *MDS* (minimum discernible signal) is used. This is the input level of a signal 3dB above the noise floor of the receiver.

As will be described later, optimising the receiver including minimising noise is achieved by using the best devices, proper matching and the correct distribution of gain within the receiver.

All receivers in current use are superhets and may have more than one frequency conversion. (Straight or TRF, direct-conversion and super-regenerative receivers are possible but are not used in practice at VHF/UHF and will not be considered further.) Frequency converters (also known as *mixers*) contribute more noise than amplifiers.

Phase noise is a special case and results from use of a noisy oscillator, usually in the first frequency conversion in the receiver. It is caused by the noise side-band on the oscillator mixing with a strong, unwanted signal or signals close to the desired one and producing a noise signal or signals within the pass-band of the receiver. See Fig 4.1.

External noise

This comes from a multitude of sources. Natural noise comes from the heat of the landscape and from the stars. In particular, the Sun is a potent generator of noise. Whether or not the noise gets into the receiver depends on the direction that the beam antenna is pointing (see Chapter 5) and on the narrowness of its beam (its *directivity*). Since directivity and beamwidth are directly related, a high-gain antenna, besides producing a stronger signal, will produce less noise than a lower-gain one.

The same applies to man-made noise unless the noise emanates from the same direction as the wanted signal.

Non-linearity and intermodulation

The hypothetical ideal receiver would have a linear relationship between the input of any stage and its output. The overall gain would be controlled but the linear relationship would still exist. In reality, amplifiers are non-linear. That is, if two signals are applied to the input, the output consists of the two, amplified, signals and products (not in the mathematical sense

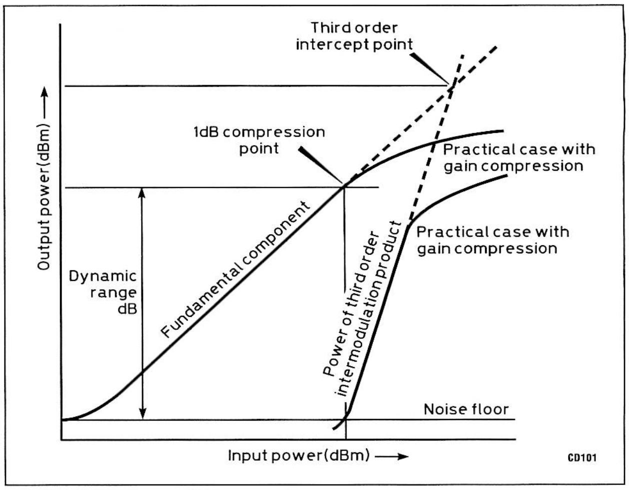

Fig 4.2. Noise floor, 1dB compression point, dynamic range and third-order intercept indicated on a mixer output versus input plot *(GEC-Plessey Professional Products Handbook)*

but harmonics, sums, differences and differences of products) of them. This is called *intermodulation* [1–3]. If the input frequencies are x and y hertz, then the output will contain:

$$x,\ y,\ 2x,\ 2y \text{ etc and } (x + y),\ (x - y),\ (2x - y),\ (2y - x) \text{ etc}$$

or, to generalise:

$$(mx \pm ny)$$

where m and n are integers (whole numbers), theoretically from one to infinity! Provided non-linearity is not too large, the higher-order products, where m and n are large, will be small. If x and y are close together, then $(2x - y)$ and $(2y - x)$ will be close to both of them. These are called *third-order intermodulation products* (3IPs).

These third-order intermodulation products are of most interest. It can be shown [1] that they grow at three times the rate of the wanted signal in an amplifier. If they are plotted on logarithmic paper (logarithmic in signal is also linear in decibels) the generalised result is shown in Fig 4.2. Third-order IPs in a receiver can produce unwanted signals in the pass-band of the receiver and need to be minimised.

Mixers are, by definition, non-linear since the required output is one intermodulation product, $y - x$, where x is the signal frequency and y is the local oscillator frequency. This assumes that the local oscillator frequency is higher than the signal frequency, a normal state of affairs. A simple mixer will have outputs of:

$$x,\ y,\ (x + y),\ (y - x),\ 2x,\ 2y,\ (2x \pm y) \text{ etc}$$

Only $(y - x)$, the intermediate frequency (IF), is wanted. All the others are eliminated by the selectivity of the IF amplifier. If the mixer is 'balanced' with respect to the input signal of the local oscillator (singly balanced) or to both (doubly balanced), a number of these output frequencies are balanced out.

For an unbalanced mixer Table 4.1 applies. In the case of a singly balanced mixer, there is half the number of mixer products (see Table 4.2). All the even 'y's are balanced out.

In the case of a doubly balanced mixer (Table 4.3), there is one quarter of the mixer products. All the even multiples of x

Table 4.1. Unbalanced mixer

	x	$2x$	$3x$	$4x$	$5x$
y	$x \pm y$	$2x \pm y$	$3x \pm y$	$4x \pm y$	$5x \pm y$
$2y$	$x \pm 2y$	$2x \pm 2y$	$3x \pm 2y$	$4x \pm 2y$	$5x \pm 2y$
$3y$	$x \pm 3y$	$2x \pm 3y$	$3x \pm 3y$	$4x \pm 3y$	$5x \pm 3y$
$4y$	$x \pm 4y$	$2x \pm 4y$	$3x \pm 4y$	$4x \pm 4y$	$5x \pm 4y$
$5y$	$x \pm 5y$	$2x \pm 5y$	$3x \pm 5y$	$4x \pm 5y$	$5x \pm 5y$

Table 4.2. Singly balanced mixer

	x	$2x$	$3x$	$4x$	$5x$
y	$x \pm y$	$2x \pm y$	$3x \pm y$	$4x \pm y$	$5x \pm y$
$2y$	*	*	*	*	*
$3y$	$x \pm 3y$	$2x \pm 3y$	$3x \pm 3y$	$4x \pm 3y$	$5x \pm 3y$
$4y$	*	*	*	*	*
$5y$	$x \pm 5y$	$2x \pm 5y$	$3x \pm 5y$	$4x \pm 5y$	$5x \pm 5y$

* This product is balanced out – see text.

Table 4.3. Doubly balanced mixer

	x	$2x$	$3x$	$4x$	$5x$
y	$x \pm y$	*	$3x \pm y$	*	$5x \pm y$
$2y$	*	*	*	*	*
$3y$	$x \pm 3y$	*	$3x \pm 3y$	*	$5x \pm 3y$
$4y$	*	*	*	*	*
$5y$	$x \pm 5y$	*	$3x \pm 5y$	*	$5x \pm 5y$

* This product is balanced out – see text.

and y are balanced out. This is relatively speaking – it is normal for the * terms to be 20 to 30dB below the other frequencies.

Note that harmonics up to five have been considered. Higher ones are possible but normally the fifth and higher ones are so small as to be negligible.

Other spurious responses in receivers

These can be classified into internal and external spurii. The former shows up with 'signals' appearing at various places in the tuning range(s) when there is no antenna attached to the receiver, ie no signal input. They arise from improper screening inside the receiver, allowing frequencies and their harmonics from intermediate stages to get to the input. For example, these could be harmonics of the beat frequency oscillator (BFO) or, in the case of double-superhet receivers, harmonics of the second oscillator. External spurii arise from *images* (the IF can be generated by subtracting the signal frequency from the local oscillator frequency, ie the frequency is as far above that of the local oscillator as the wanted one is below).

Gain compression

The output of, say, the RF amplifier (the 'front end' of a receiver) should be x times the input where x is the stage gain. This is always true up to a point but the stage can only give a certain output before it *saturates,* ie further increases of input cause no further change in the output. With most input devices, transistors or valves, this will be of the order of millivolts. Where inputs are made large, the graph of output versus input curves over (see Fig 4.2). Where there is 1dB difference between the actual line and the extrapolated (extended, dotted in the figure) line, this is called the *1dB compression point* and is often quoted in equipment reviews.

Blocking

One consequence of gain compression is *blocking* the receiver, ie reducing the output of the wanted signal to zero by the presence of a very strong signal close by in frequency. This will happen in all receivers to some extent. The better the receiver, the larger and closer can be the unwanted signal before it causes trouble.

Dynamic range

This depends on noise generated within the receiver which determines the *minimum discernible signal* (MDS). This is arbitrarily defined as 3dB above the noise level, ie twice as strong as the noise. The noise level referred to the input, ie that noise level at the input which would give the same noise output with a noiseless amplifier, is called the *noise floor*. The difference between this and the 1dB compression point is called the *dynamic range* and is often quoted in equipment reviews. It defines the maximum range of signal strengths that may be present at the input which do not cause unwanted responses.

There are, in fact, two dynamic ranges to consider:

1. that in which the limit is set by the onset of intermodulation;
2. that which is set by the effect of phase noise (see above).

The intermodulation-limited dynamic range, IMD, is also known as the *spurious-free dynamic range* (SFDR) and is given by:

$$\text{SFDR} = 2/3(I_3 - \text{NF})$$

where I_3 is the *third-order intercept point* (see Fig 4.2) and NF is the noise floor in decibels referred to the input, ie that noise level which, at the input, would give the same output with a noiseless receiver.

RECEIVER BUILDING BLOCKS

This section describes the various sections of a receiver with a few details at the end on transceivers and transverters.

Before discussing actual circuits, it is worthwhile to look at basic components and their behaviour at these frequencies. All have associated with them a 'parasitic' component or components and these modify their behaviour from that at LF or HF. For example, the leads of a capacitor have inductance, and inductors have capacitance and resistance associated with them. A simple resistor will have inductance (due to the leads) and capacitance (due to the proximity of the ends). To quantify this, the inductance of a straight piece of round wire is given by:

$$L\,(\text{nH}) = 0.461 \times b \times (\log_{10} 4b/a - 0.326)$$

where b is the length in millimetres and a is the diameter in millimetres. 'nH' is nanohenrys, ie 10^{-9} henrys. This implies that the larger the surface area, the smaller the inductance for a given length, hence the use of foil for UHF connections to reduce parasitic inductance.

Minimising the effects of these is done by designing the layout with the shortest possible leads and making those of foil rather than wire.

Individual components

The use of 'chip', ceramic dielectric, capacitors (with no leads) which either fit into a slot in the PC board or are soldered

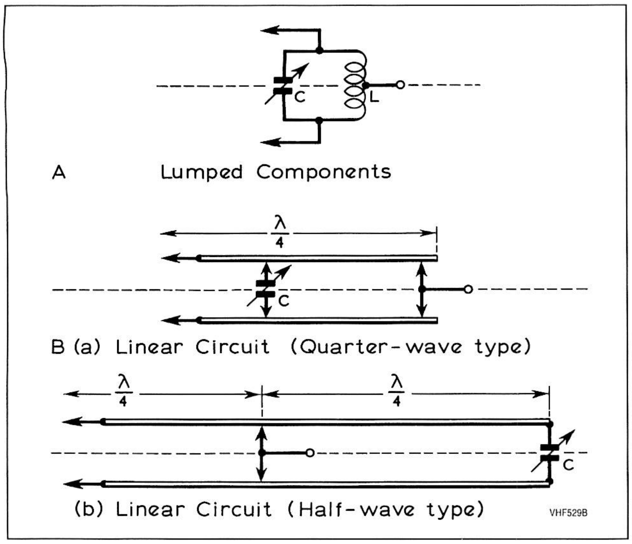

Fig 4.3. Three examples of tuned circuits

directly to its surface, gives the lowest possible inductance. Beware, however, of some multilayer surface-mounting capacitors which have internal inductance and sometimes a large loss. High-value components, eg above about 50pF, may have a large negative temperature coefficient.

Tuned circuits

As frequencies get higher, the conventional LC tuned circuit becomes smaller and smaller. It may be replaced, especially at the higher frequencies, by resonant lines or cavities as shown.

Fig 4.3 shows the conventional 'lumped' LC circuit (so called because the parts are 'lumps' of capacitance and inductance) and its equivalents in resonant quarter-wave and half-wave lines. Fig 4.4 shows the progression from a lumped circuit to a cavity resonator. The latter is normally used at 432MHz and above. The tuned line can be symmetrical, eg for a push-pull circuit, or an asymmetrical one such as a coaxial line. The latter is used in single-ended circuits both for

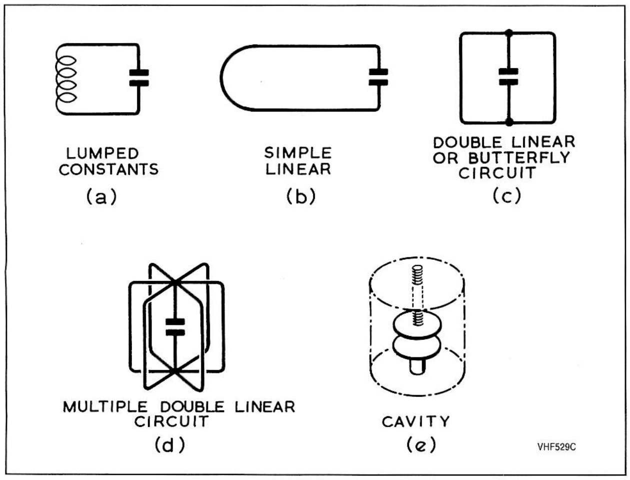

Fig 4.4. The development of the cavity from the original lumped-constant circuit

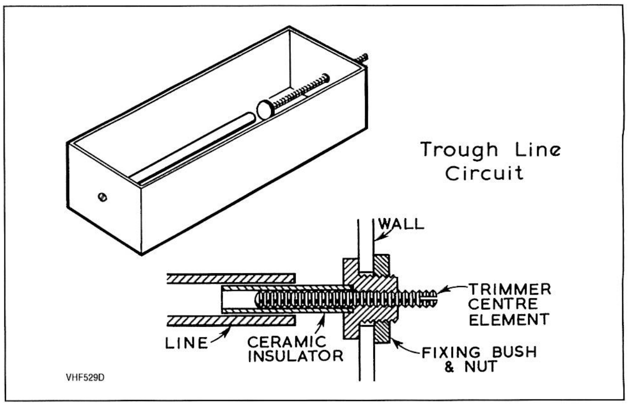

Fig 4.5. Trough-line circuits. The tuning capacitor may be either in line with the inner conductor (resonant line) or attached to the side wall of the trough. Another convenient method is to use a tube for the line and to fit into the end of it a ceramic trimmer centre element. In this case the outer element can be removed or connected to the line as required

transmitting and for receiving as a *trough-line* (Fig 4.5). Tuning for all types may be by conventional variable capacitors, by discs mounted on threaded rods (studding) or by *flappers* which are metal plates mounted with one end hinged by a flexible section and moved to or from the tuned line by means of an insulated cam or other insulated system. Occasionally, two plates may form a tuning capacitor and its capacitance is varied by moving a dielectric in or out of the gap between the plates.

If discs are used, the capacitance varies in a very non-linear manner with the distance between them. It is:

$$C\text{ (pF)} = \frac{0.00885 \times \text{ Area (sq mm)}}{\text{Spacing (mm)}}$$

DESIGNING RESONANT LINES

If this is for one frequency (eg for the output of a fixed-frequency oscillator), they can be cut to a quarter or half a wavelength and trimmed carefully to resonance. Otherwise, they should be cut shorter and have a tuning or loading capacitor which can be calculated from:

$$\frac{1}{2\pi fC} = \frac{Z_0 \tan 2\pi L}{\lambda}$$

where f is the frequency, C is the tuning capacitance, λ is the wavelength, L is the length and Z_0 is the characteristic impedance, calculated as follows. For a coaxial line:

$$Z_0 = 138 \log_{10}\left(\frac{D}{d}\right)$$

where D is the inside diameter of the outer tube and d is the outside diameter of the inner tube or wire. For parallel lines:

$$Z_0 = 276 \log_{10}\left(\frac{2D}{d}\right)$$

where D is the interline spacing and d is the line diameter. See Figs 4.6–4.9.

Fig 4.10 gives a series of graphs for coaxial and parallel lines where $f \times L$ (in megahertz and centimetres) is plotted against $f \times C$ (in megahertz and picofarads) for a number of different values of r or D/d.

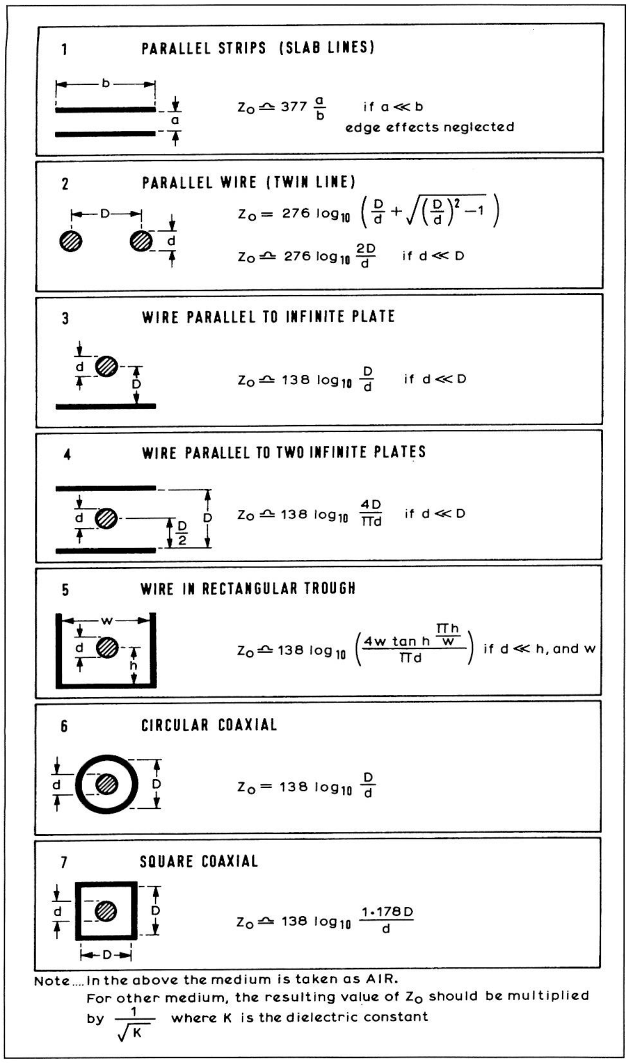

Fig 4.6. Various forms of transmission line

Coupling to tuned circuits

Coupling is used to take energy out of a circuit or to put it in. The nature of the coupling depends on the impedance of the input or output.

With a lumped circuit, the coupling may be by means of another coil close to the tuned circuit or by means of a tapping on it. Fig 4.11A shows this. The lower the impedance of the coupling, the smaller is the coupling coil or the closer is the tapping to the 'earthy' end of the tuned circuit.

Fig 4.11B shows various ways of coupling to tuned lines and Fig 4.12 to trough lines, coaxial resonators or cavities.

Helical resonators

These form high-Q circuits and are roughly equivalent to quarter-wave coaxial lines compressed. Fig 4.13 shows the basic idea. They are used at VHF and UHF and can be tuned by a variable capacitor between the case and the upper end. They can also be cascaded by making slots in the side of the cases and bonding them together so that the resonators can 'see' one another.

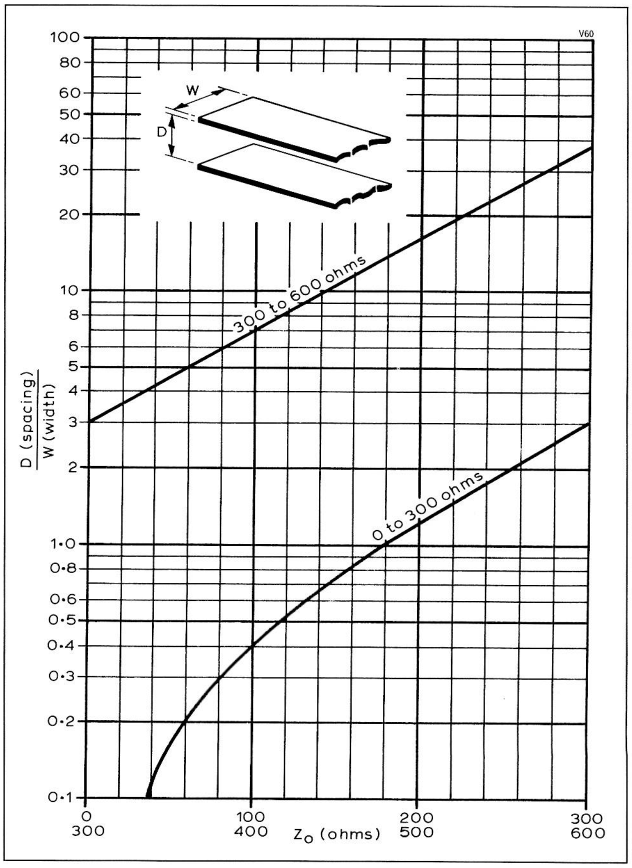

Fig 4.7. Characteristic impedance of balanced strip transmission line

The unloaded Q is $50Df^{1/2}$, where D is the internal diameter of the case and f is the frequency. If the case is of square section, this must be multiplied by a correction factor of 1.2.

Fig 4.14 shows a two-component filter using helical resonators. To design one, the following formulae are used:

$$\text{Pitch } (p \text{ in Fig 4.13}) = \frac{D^2 f}{90.6}$$

where p and D are in millimetres.

$$\text{Characteristic impedance} Z_0 = \frac{386}{fD}$$

These assume that $d/D = 0.55$ and $L/D = 1.5$.

Alternatively, the nomogram (Fig 4.16) will give an accurate enough answer for all practical purposes. Note that the dimensions are given in *inches*.

In order to maintain a high Q, it is desirable to plate the inside of the cavity and the outside of the wire with silver.

Microstrip circuits

At frequencies above 300MHz, microstrip circuits, which are formed directly on one side of double-sided printed circuit board, are useful. Figs 4.17 and 4.18 show quarter-wave and half-wave microstrip lines with the position of the tuning capacitor. It is essential that the other side or *ground plane* is bonded to the 'earthy' parts of the tuned circuit.

The dielectric should be as thin and of as low loss as

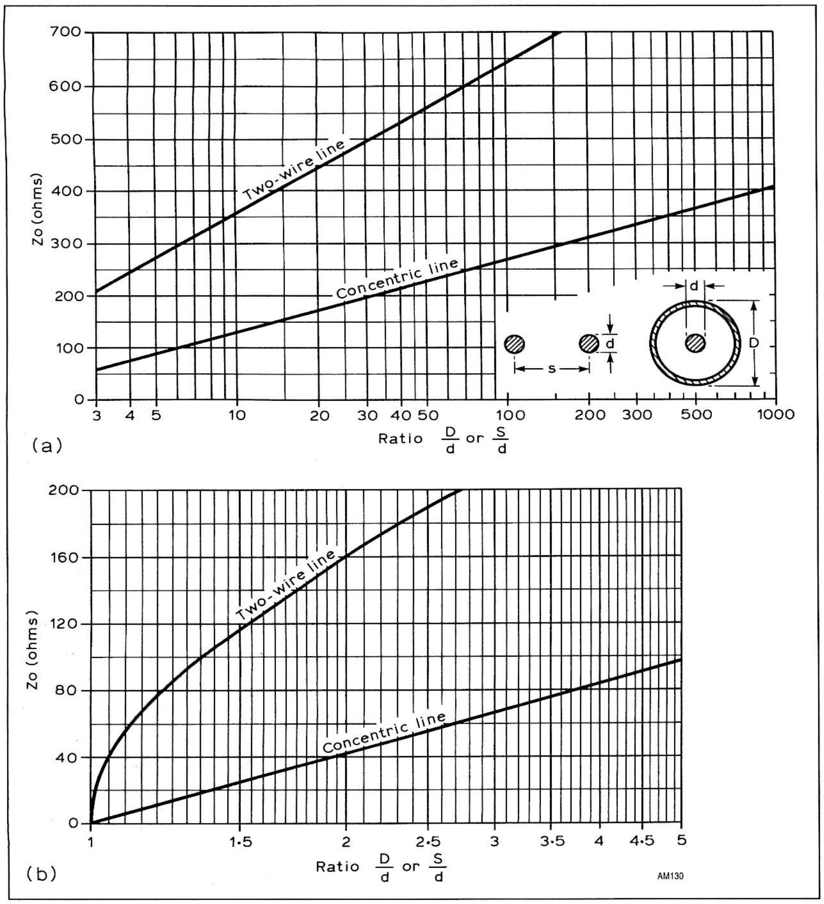

Fig 4.8. Chart giving characteristic impedances of concentric (coaxial) and two-wire lines in terms of their dimensional ratios, assuming air insulation. When the space around the wires is filled with insulation, the impedance given by the chart must be divided by the square root of its dielectric constant (permittivity). This ratio is called the *velocity factor* because the wave velocity is reduced in the same proportion

possible. PTFE (Teflon™ or Fluon™) is ideal but expensive. Glassfibre (epoxy resin reinforced with E-glass) is the next best material. The length is calculated in the same way as for a coaxial or trough line with modifications to take into account the different shape and the fact that the velocity of radio waves is lower where there is a dielectric other than air, just as the capacitance between two plates is greater when there is a solid dielectric present. The latter is greater by the dielectric constant which is about 2.1 for PTFE and about 5 for glassfibre-reinforced epoxy resin.

The exact value, which depends on precisely which epoxy is used, can be found by measuring the capacitance of a known area of double-sided material:

$$e = 113 \times C \times t/A$$

where e is the dielectric constant, t is the thickness in millimetres, A is the area in square millimetres and C is the capacitance in picofarads.

The width of the element depends on the design impedance Z and is given by:

$$\log_{10} W = 0.874 + \log_{10} t - 0.005 \times Z \times \sqrt{e} + 1.14$$

where t is the thickness in millimetres, e is the dielectric constant and W is the width in millimetres.

Having found this, the length is calculated from:

$$l = l_0 \times V_f$$

where l is the element length, l_0 is the length

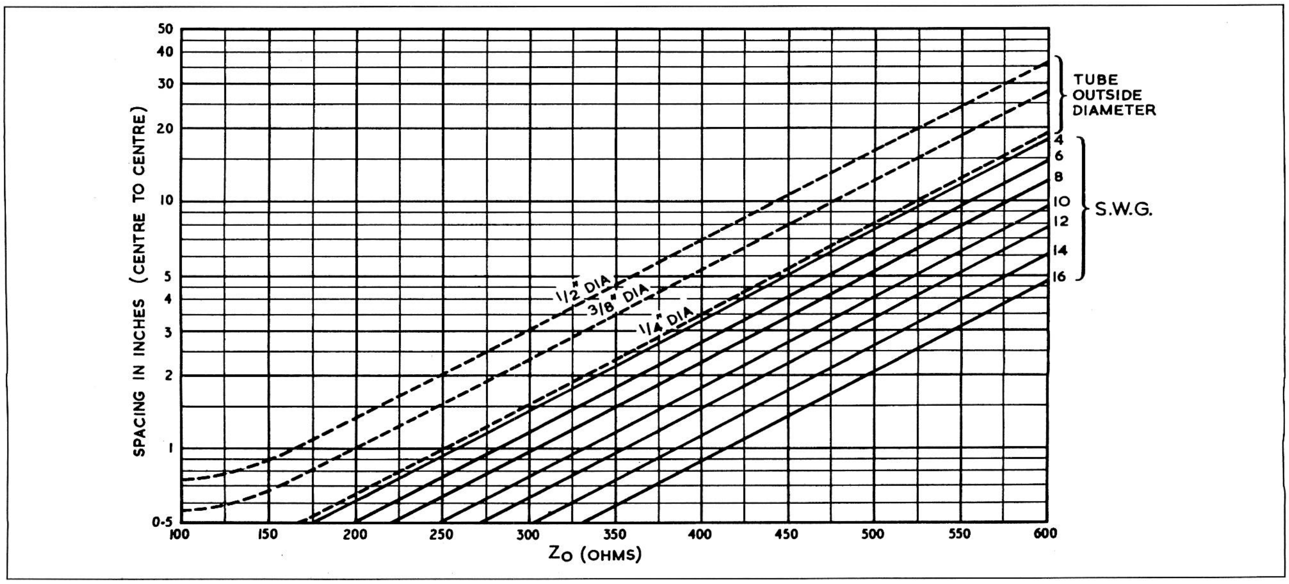

Fig 4.9. Characteristic impedance of balanced lines for different wire and tube sizes and spacings, for the range 200–600Ω. The curves for tubes are extended down to 100Ω to cover the design of Q-bar transformers. Air spacing is assumed

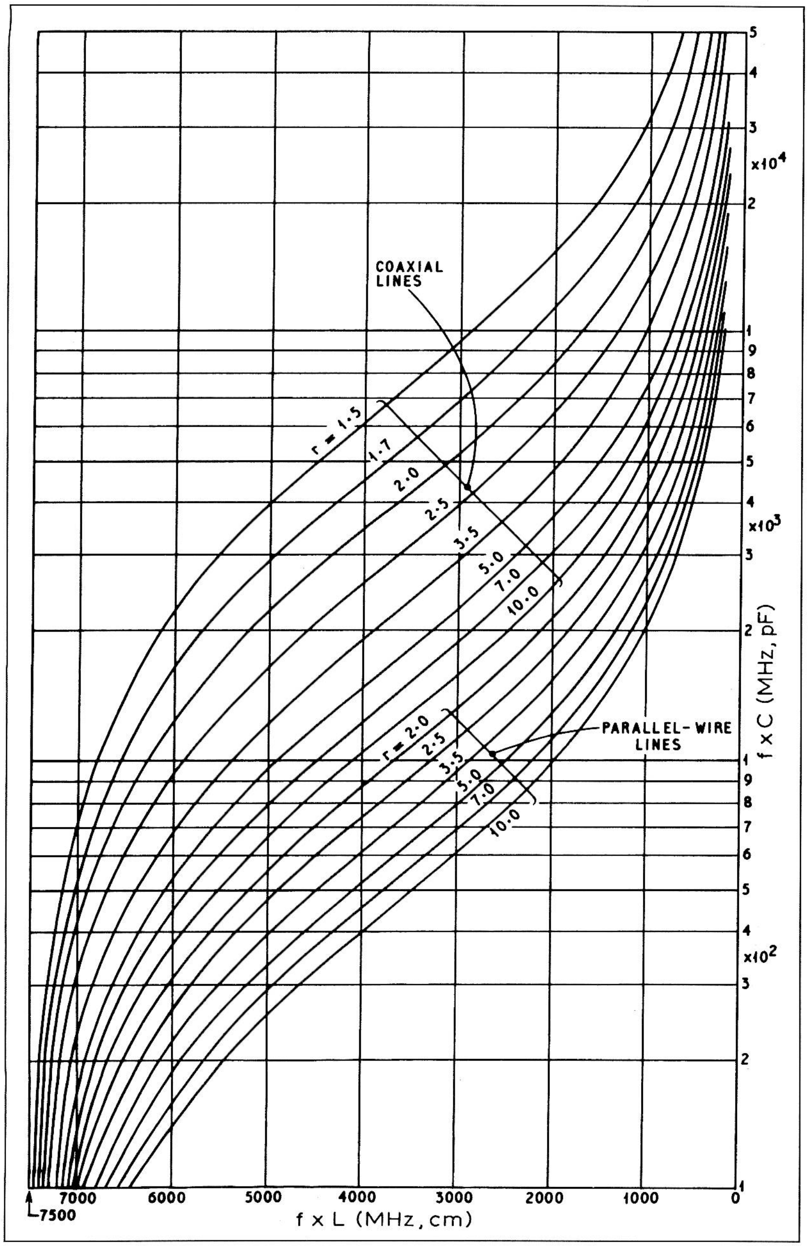

Fig 4.10. Resonance curves for capacitively loaded transmission-line resonators

(eg that of a half wavelength) in free space and V_f is the velocity factor which is related to the dielectric constant by:

$$V_f = \frac{1}{\sqrt{0.475 \times (e + 0.67)}}$$

Active components – handling precautions

Many active devices are sensitive to static discharge and precautions must be taken when handling them. FETs, both silicon MOSFETs and GaAsFETs, are especially sensitive and should be handled with extreme care. If particular, avoid wearing synthetic fibre clothing, walking on synthetic fibre carpets or wearing synthetic soled shoes. In winter, when the relative humidity is low, a static charge of up to 30kV (yes, 30,000V!) can build up on you just by walking across a room. 10–25V can destroy a MOSFET or a GaAsFET.

Other active devices are less sensitive but should still be treated with care. Device holders should not be used at VHF or UHF because they add to lead inductance and, at least, they can cause mismatching or, at worst, instability.

RECEIVER CIRCUITS

This section presents circuits for each section of the receiver. They may be combined to form a complete receiver or converter or, with the later section on transmitters, to form a complete transceiver or transverter. Designs for individual frequency bands follow the general introduction.

As mentioned above, these are always superhets and they may have more than one frequency conversion. Fig 4.19(a) shows a block diagram for a single-conversion superhet while Fig 4.19(b) shows the variation for a double-conversion superhet and Fig 4.19(c) for a double conversion superhet where the second local oscillator is the variable one, the so-called *tuneable IF* system. These are diagrams of multi-mode receivers. For dedicated single-mode receivers, FM or SSB/Morse, the switches S1 and S2 are omitted.

The distribution of gain between the RF, IF and audio stages is very important from the standpoint of strong-signal performance. Briefly, the gain between the antenna and the first sharp (IF) filter should be the minimum needed to overcome the noise created by the mixer(s).

In a single superhet (Fig 4.19(a)), this means an RF gain of about 20dB followed by a mixer loss of 6dB, giving an overall gain before the sharp filter of about 14dB. Methods for calculating exact values are given by G3SEK [1, 2]. Note, too, that a sharp crystal filter can behave in a non-linear manner if presented with a strong signal just out-of-band. In double superhets, there may be a need for gain between the two mixers before any sharp filter. In this case an amplifier with a very good strong-signal performance must be used.

There is a natural noise input from the antenna and all stages thereafter generate some noise. Mixers generate more noise than straight amplifiers so an RF amplifier must have sufficient gain to overcome the noise of the first mixer without contributing significant noise itself. It must also have a good large-signal handling capacity. This means good linearity.

Input circuits

A band-pass or low-pass filter is often placed between the antenna and the RF amplifier. In transceivers, this may be common to both receive and transmit circuits – in this case, a

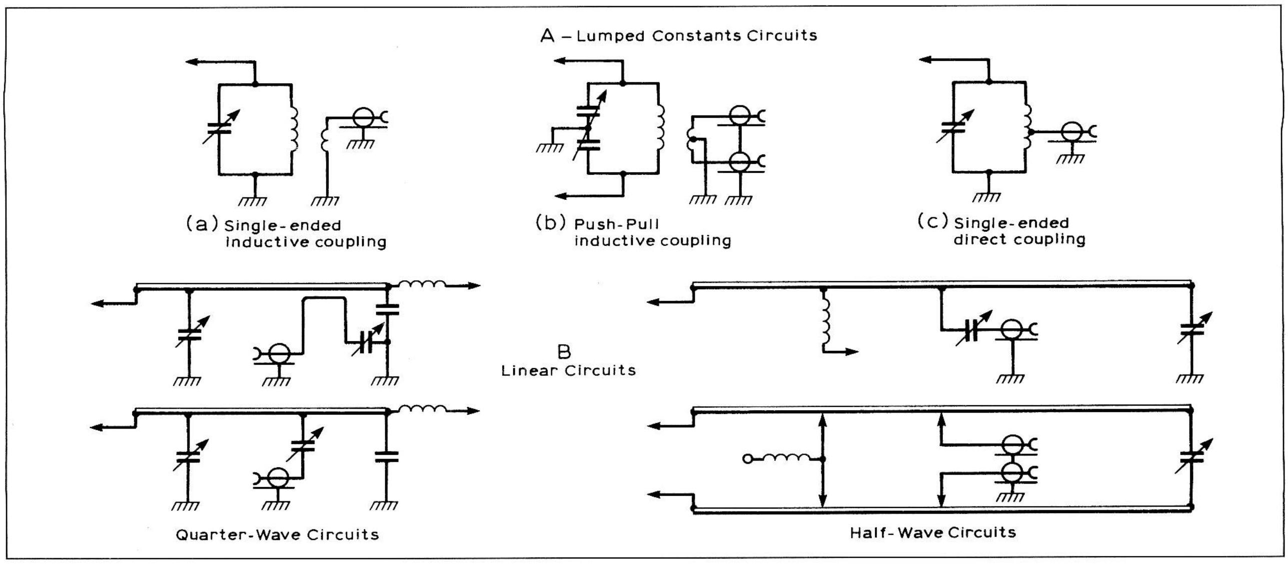

Fig 4.11. Methods of coupling to lumped constant and linear circuits

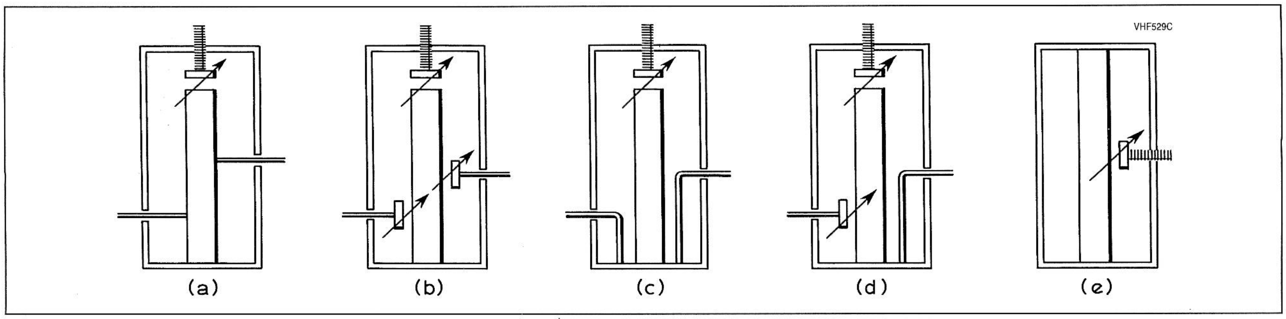

Fig 4.12. Methods of coupling to a trough cavity

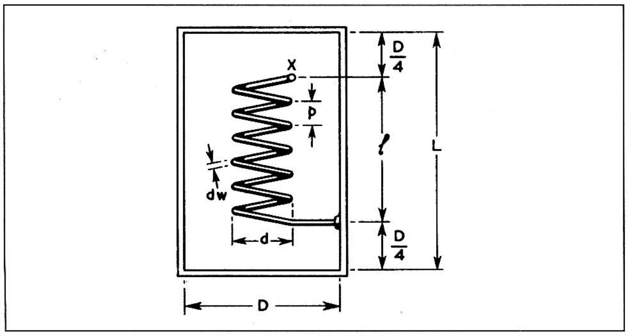

Fig 4.13. A helical resonator which can be used at VHF and UHF

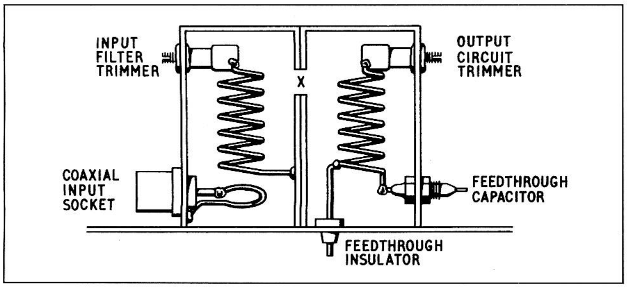

Fig 4.14. A typical arrangement of helical resonators

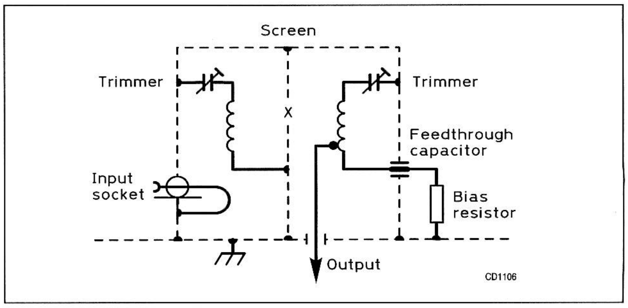

Fig 4.15. The equivalent circuit. This arrangement shows two tuned circuits as the input filter and input circuit to a grounded-gate amplifier. Coupling between the filter and input circuits is provided by an aperture in the screen (X). Alternatively, a normal coupling method can be used, such as a link, taps or probe. Note: when no bias is required, the feedthrough capacitor is omitted and the end of the helix is taken straight to earth

low-pass filter must be used. The alternative of a high-Q tuned circuit with a panel mounted 'antenna trimmer' (input circuit tuning control) is possible but not often used.

Fig 4.20 [8] shows a typical low-pass filter for 50MHz which has a cut-off (3dB down) frequency of 62MHz and an insertion loss of 0.2dB. Fig 4.21 [8] shows a typical band-pass filter, also for 50MHz, which has a pass-band of 3MHz centred on 51MHz.

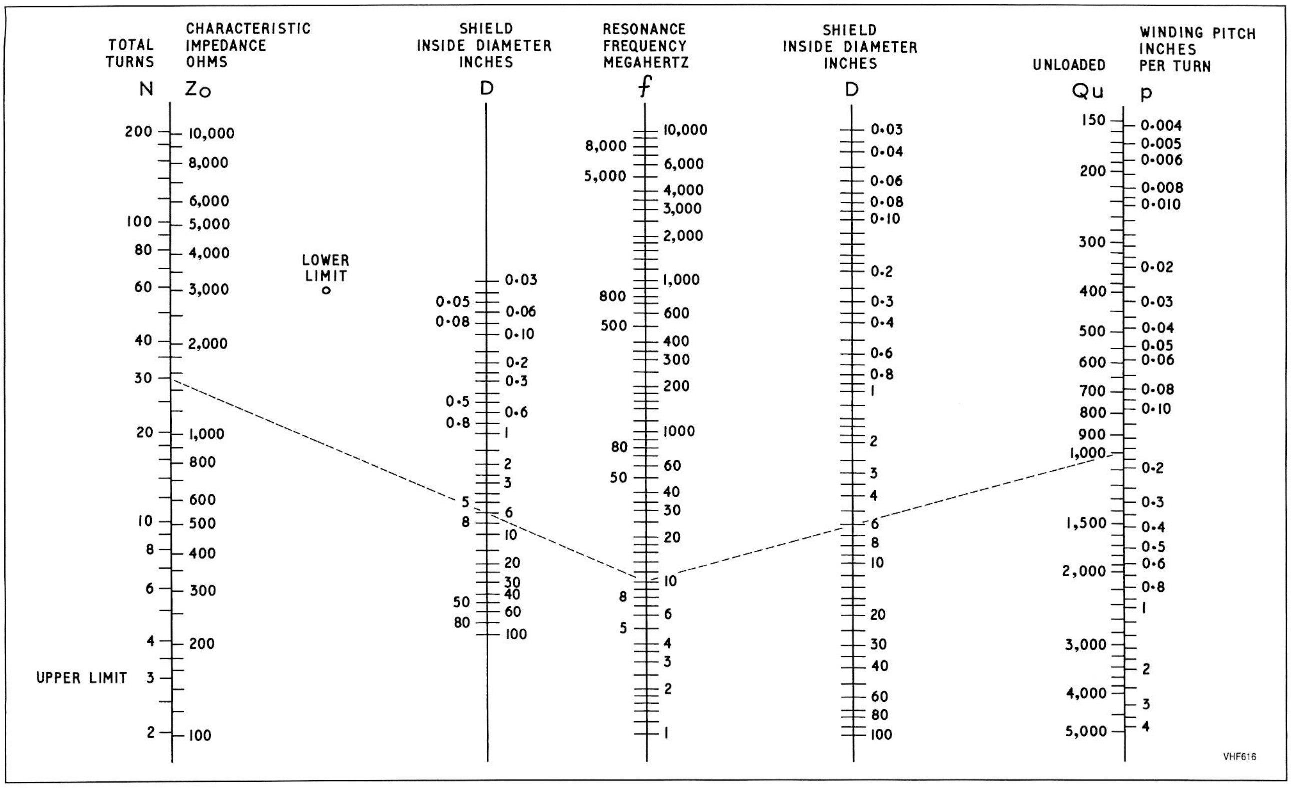

Fig 4.16. Design chart for λ/4 helical resonators. Lines indicate example

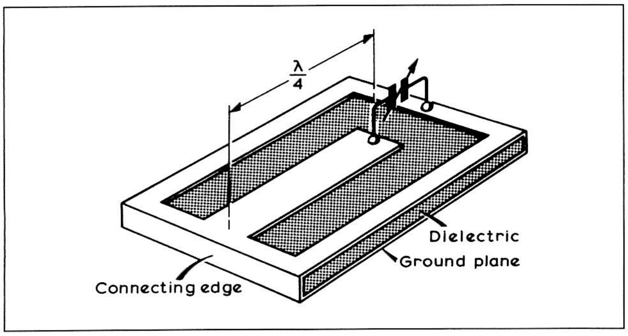

Fig 4.17. General arrangement of a λ/4 microstripline circuit

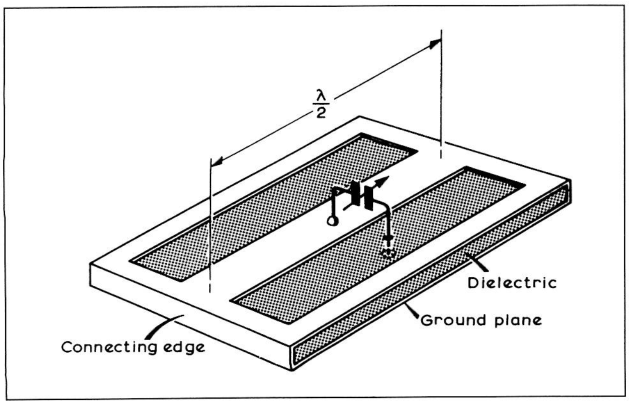

Fig 4.18. General arrangement of a λ/2 microstripline circuit

RF amplifiers

All modern RF stages are solid state. The best performance in terms of noise and linearity is from FETs, especially power devices such as the MGF1801 which can have a 3IP (third-order intercept point) of greater than +10dB. The snag is that they take a considerable drain current (about 60mA!) and are thus less suitable than ordinary MOSFETs for portable receivers. They are also expensive. A typical circuit of a slightly lower-grade design for a 50MHz RF amplifier using a BF981 is shown in Fig 4.22 [9]. C1, C2 and L1 form the input pi-circuit to match the antenna to gate 1 of the BF981 (C1 is the 45pF trimmer and the 33pF fixed capacitor in parallel). L2, L3 and C3 with their tuning capacitors form a band-pass output filter.

RF filters

These are placed between the RF amplifier and the mixer. Their purpose is to further reduce the level of out-of-band signals. They can have a higher insertion loss than the input filter since that can be compensated for by the gain of the RF stage.

At 50 and 70MHz, they are usually conventional coupled tuned circuits of the type shown in Fig 4.23. The coupling capacitor, C3, is chosen [6] to give a flat response over the band. It will be very small, of the order of 2pF. The position of the tappings on the coils will depend on the output and input impedances.

At 144, 430 and 1300MHz, helical filters (see above) come into their own. They generally have an insertion loss of 2–4dB and an input and output impedance of 50Ω.

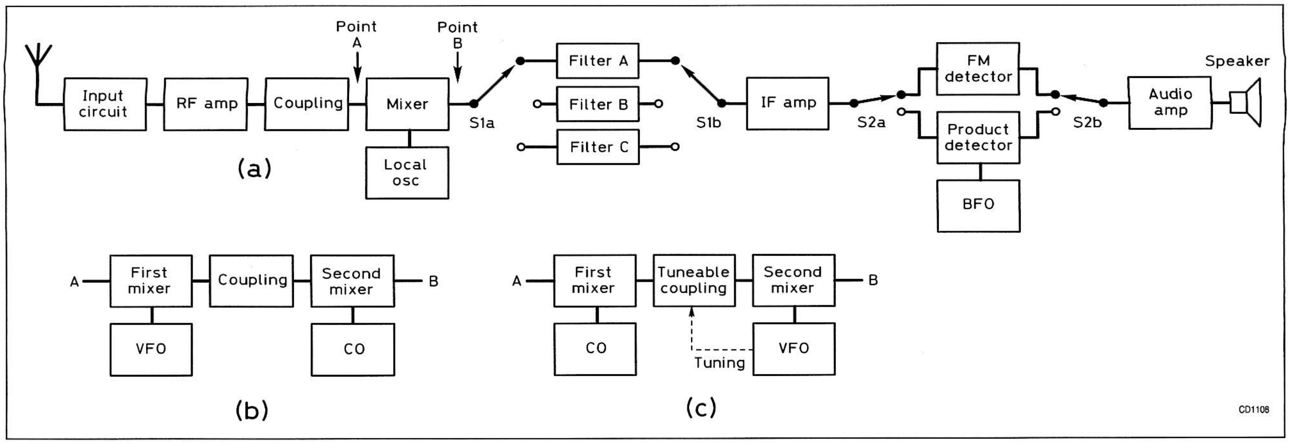

Fig 4.19. Block diagram of a multimode receiver

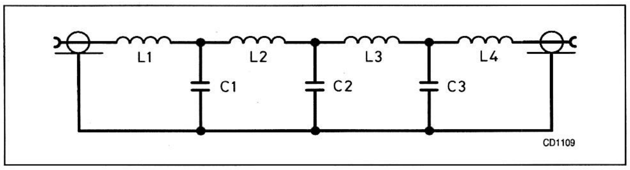

Fig 4.20. 50MHz low-pass filter with cut-off at 62MHz. L1, L4: 90nH, 5t 0.8mm wire close wound 4.8mm dia. L2, L3: 0.22µH, 10t 0.8mm wire close wound 4.8mm dia. C1, C3: 82pF silver mica. C2: 100pF silver mica *(QEX)*

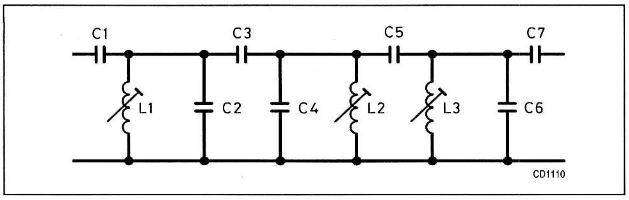

Fig 4.21. L1, L3: 293nH nominal; L2: 284nH nominal. All TOKO type MC130 (0.3µH adjustable). C1, C2, C6, C7: 15pF. C3, C5: 3pF. C4: 27pF. All capacitors silver mica

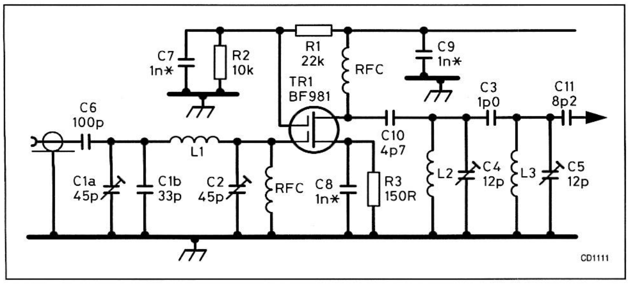

Fig 4.22. L1–L3: 0.87µH. RFC: 4.7µH. Asterisk (*) denotes ceramic chip capacitors *(VHF Communications)*

Mixers

Any non-linear device can in theory be used as a mixer. However, those commonly used are diodes, FETs and MOSFETs, ICs and occasionally bipolar transistors.

Diode ring mixers, which are doubly balanced (see above) are widely used because they contribute less noise than other types and have a better large-signal handling capacity.

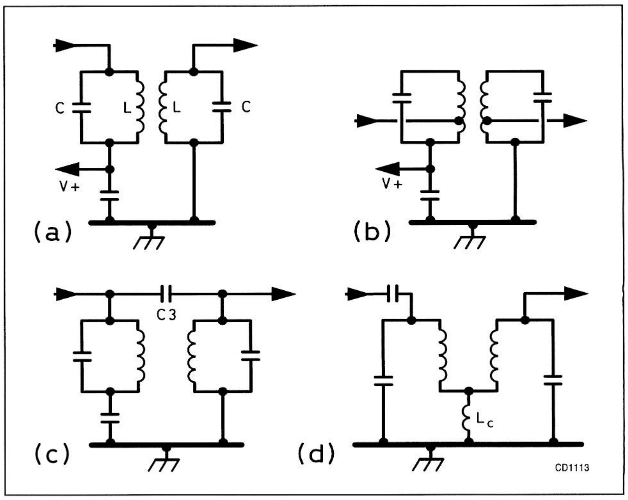

Fig 4.23. Coupled (band-pass) tuned circuits between stages

However, they do need a large oscillator injection and accurate matching at all ports. Also, the best ones are expensive. They have a conversion loss of between 6 and 10dB depending on type. They need a local oscillator input of 7 to 20dBm (decibels relative to 1mW, ie 0.5–2.5V into 50Ω). They can also be used in transceivers as mixers in the transmit chain.

A typical circuit for a 50 to 28MHz converter [9] using the Minicircuits TAK 11 is shown in Fig 4.24. The local oscillator level is +17.5dBm which corresponds to 1.7V into a 50Ω load. A similar circuit can be used at 144 and 430MHz.

Other forms of mixer are used. For example, in simple equipment, the dual-gate MOSFET is used with the signal into gate 1 and the local oscillator into gate 2. JFETs can also be used in a balanced circuit with excellent results.

There are ICs which work at these frequencies. One such is the NE600 which is the UHF version of the NE602, well known at HF. It has the advantage of being relatively cheap and of having a large dynamic range. It is noisy, having a noise figure of about 12dB at 1.3GHz, but this is overcome by a suitable RF amplifier. Also, it is only rated by its makers to 1.2GHz.

Finally, particularly at 1.3GHz, there are single diode

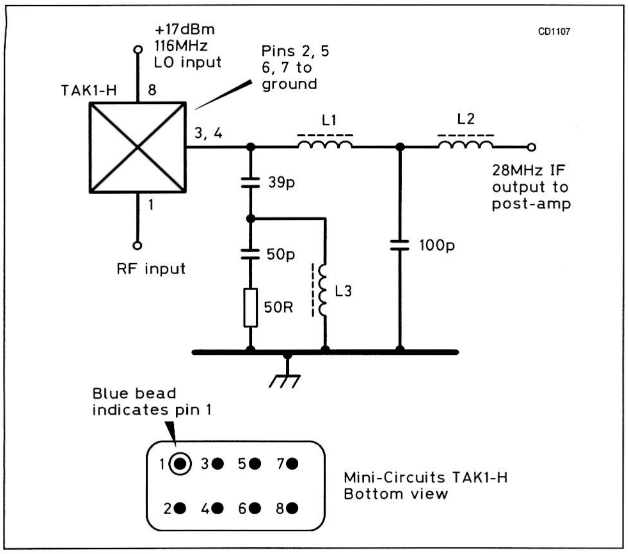

Fig 4.24. A 144MHz receive mixer and diplexer filter using a Mini-Circuits TAK-1H high-level mixer. The filter provides a 50Ω broad-band resistive match for improved third-order distortion characteristics. L1: 0.35μH, 5t 24 AWG enam on T50-6 toroid. L2: 0.21μH, 4t 24 AWG enam on T50-6 toroid. L3: 0.18μH, 4t 24 AWG on T37-6 toroid

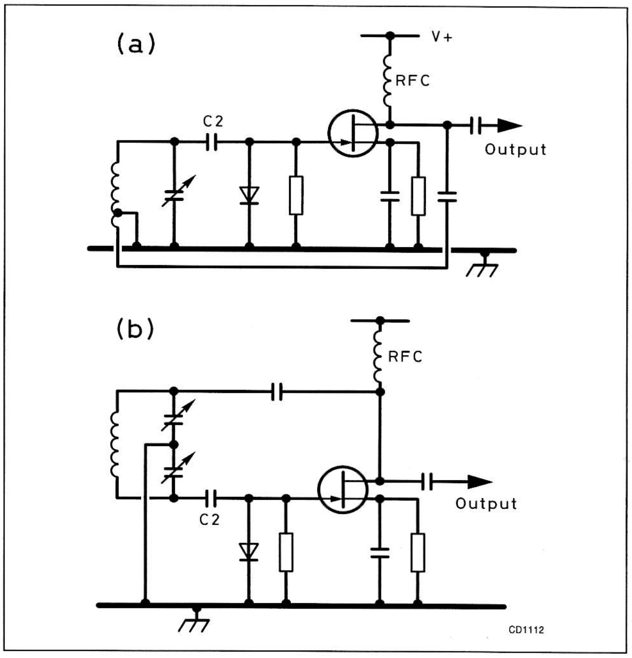

Fig 4.26. (a) Hartley and (b) Colpitts oscillators

mixers with either microstrip or solid 'interdigital' tuned circuits. One of the latter is shown in Fig 4.25 where the input is fed to the first resonator ('digit') followed by a tuned resonator and the diode mixer. The local oscillator is supplied at a quarter of the required frequency and harmonics generated by another diode. The correct harmonic is selected by a tuned resonator which is coupled to the mixer diode.

Local oscillators [4]

All oscillators except DDS [10] (see below) consist of a tuned circuit to set the frequency and a 'gain block' to make it oscillate. They start by small noise pulses being amplified and modified by the tuned circuit and increase until limiting in one form or another occurs. Some oscillators have a negative feedback system to limit the oscillations to sine waves [11]. This may not be a good thing as some mixers, especially diode rings, need hard switching by a square-wave drive, The gain is provided by a bipolar transistor, a JFET, a MOSFET or an IC. Best performance regarding noise is obtained from a JFET.

The two most important characteristics of a local oscillator are *stability* and *low noise*. In the case of single-conversion superhets, the oscillator must be tuneable over the width of the band for which reception is needed. This is 2MHz for 50 and 144MHz, only 0.5MHz for 70MHz, 10MHz for 430MHz and 85MHz for 1.3GHz. The latter bands would normally either be partially covered or would be covered in segments.

All oscillators drift in frequency and this is minimised by mechanical, electrical and temperature stability.

Frequency drift is of two kinds, short-term drift which, if very short term, is called *scintillation* or *jitter*, and is caused by mechanical instability, and long-term drift which is caused by thermal effects. This assumes that the power supply is stable. Thermal effects can be minimised by suitable choice of components in the tuned circuit and by operating the oscillator at the smallest power level possible.

Variable oscillators (VFOs)

There are four types of variable oscillator:

1. The *straight analogue oscillator* typified by the Colpitts, Hartley, Clapp, Vackar and Franklin (Figs 4.26–4.29) often followed by a frequency multiplier to get to the correct frequency. The various types differ only in the nature of the feedback circuits, eg the Hartley and Colpitts (Fig 4.26) use inductive and capacitive devices to provide the necessary phase shift. The source-follower Hartley and Colpitts (Fig 4.27) do not need the phase shift and a variant of the

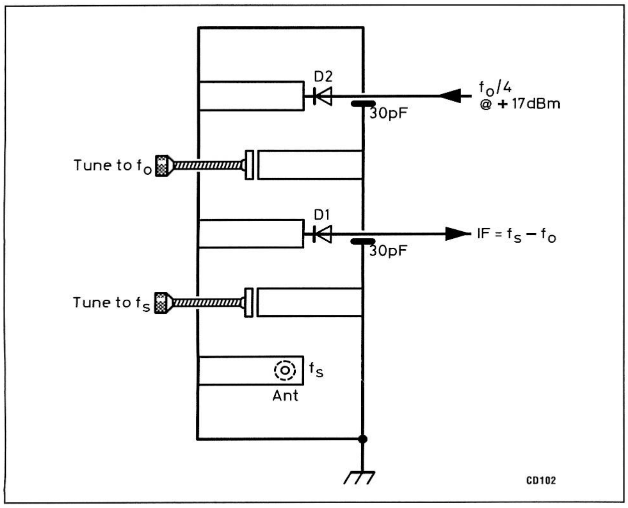

Fig 4.25. Single-diode mixer for 1296MHz. An interdigital filter provides isolation between ports. D1 (HP5082-2817) is the mixer and D2 (HP5082-2853) is the last multiplier in the LO chain *(QST)*

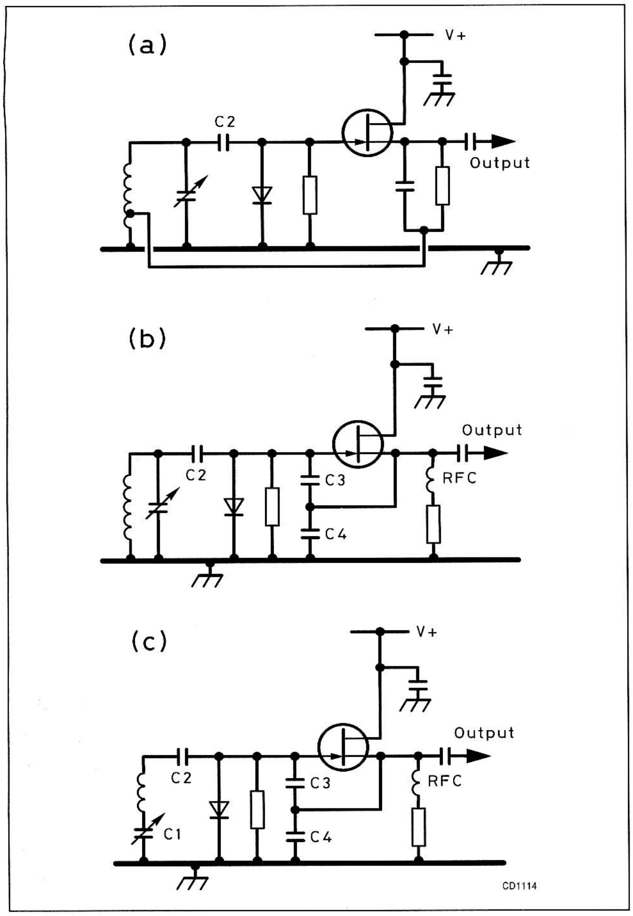

Fig 4.27. Source-coupled oscillators: (a) Hartley, (b) Colpitts and (c) Clapp

latter, the Clapp, uses a 'series tuned' circuit but this is directly equivalent to a parallel-tuned oscillator with C1, C2, C3 and C4 in series. Since C3 and C4 can be large without affecting the oscillation, they effectively swamp the interelectrode capacitance of the active device. The

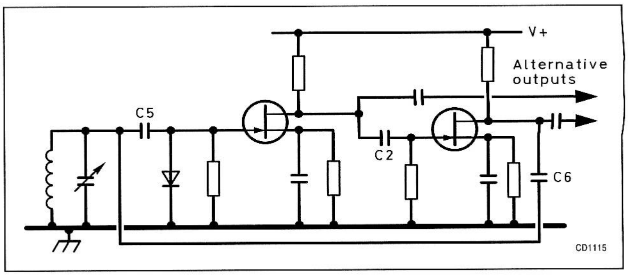

Fig 4.29. The Franklin oscillator

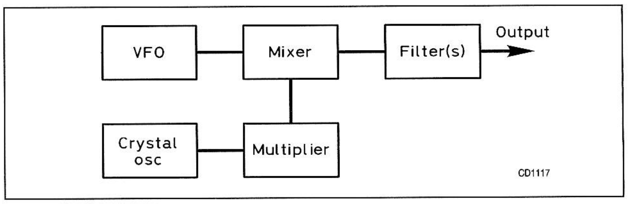

Fig 4.30. Block diagram of simple mixer-oscillator

presence of the diode does not seem to be essential. A variant of the Colpitts is the Vackar (Fig 4.28) where the large drain-to-earth capacitance also swamps interelectrode capacitances.

Decoupled source resistors (for bias) are not always needed but if used should be 150–1000Ω – the larger, the better from a stability view so long as oscillations start easily and are of sufficient amplitude. Gate resistors should be 10kΩ–100kΩ, gate capacitors (C2) should be 47–100pF and the source follower capacitors (C3 and C4) should be 200–470pF with C4 being larger but still allowing easy starting and enough output.

There are some advantages to a two-stage oscillator such as the Franklin (Fig 4.29) because the two stages provide the necessary phase shift and the gain is so high that the coupling capacitors, C5 and C6, can be very small, so reducing the effect of the active device on the tuned circuit. Also, the output can be taken from the 'mid-point' of the amplifier. C5 and C6 should be of the order of 1–2pF and are determined by making them as small as possible while still getting easy starting. The Franklin has another advantage, that of needing only one switch wafer if band changing is necessary.

Every VFO should be followed by a low-gain buffer amplifier to avoid subsequent stages affecting the frequency (*pulling*).

2. The *mixer oscillator* where the variable oscillator is mixed with a fixed frequency (a crystal oscillator – see below) to give the desired frequency. Fig 4.30 shows a block diagram of a simple mixer-oscillator where the VFO and a crystal oscillator (CO) with or without a multiplier are fed to a mixer. At this level, a double-balanced IC mixer such

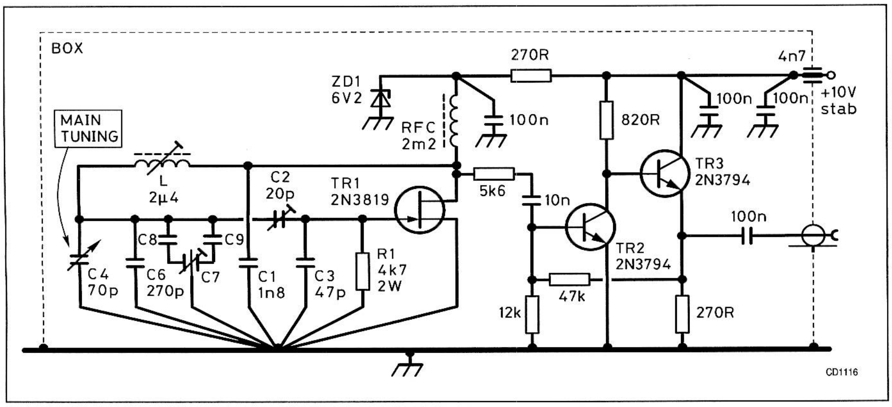

Fig 4.28. The high-stability FET Vackar oscillator developed by G3PDM to cover 5.88 to 6.93MHz for the Mk2 version of his *Radio Communication Handbook* (5th edn) receiver. C6 is silver mica. The replacements for the obsolete Tempatrimmer featured in the original design are: C7 (100pF + 100pF differential), C8 (100pF negative temperature coefficient) and C9 (100pF positive temperature coefficient)

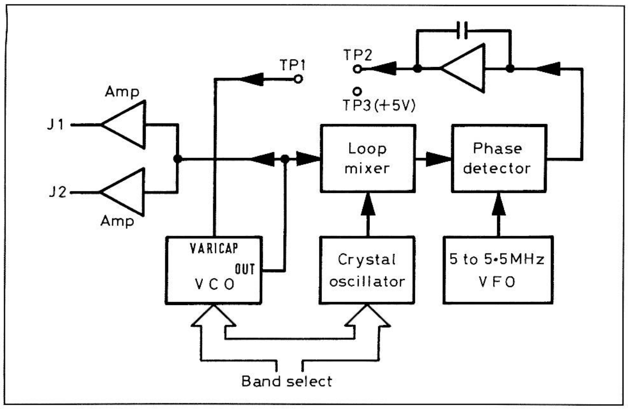

Fig 4.31. Block diagram of simple synthesised oscillator

as the MC1496 is suitable. Its output must go through a suitable filter, normally a band-pass filter. This design is useful for transceivers because the CO can be changed to give the transmit frequency, leaving the VFO running continuously. This leads to low drift. In this case, a low-pass filter should be followed by a high-pass filter to include both frequencies.

3. The *synthesised oscillator* using digital systems and a phase-locked loop (PLL) comparing an adjustable fraction of the oscillator frequency with another frequency derived from a stable, crystal, oscillator. It uses a voltage corresponding to the difference in phase to keep the variable oscillator constant – a negative feedback system. Most designs for PLLs have most of the components within one IC. The IC manufacturers supply data on the operation of the IC and the realisation of the PLL circuit.

   A simple synthesised oscillator has been described [12] for a HF receiver and can easily be adapted for VHF or UHF use by making it for one band only and either multiplying or by mixing with a suitable fixed frequency to get to the wanted frequency. Fig 4.31 shows a block diagram.

   There is a halfway house between straight or free-running oscillators and the fully fledged synthesiser. This is called the *Huff and Puff* and was invented by PA0KSB. Its most recent development [13] is shown in the block diagram, Fig 4.32. The crystal oscillator has an overtone mode 50MHz crystal and the 5–5.5MHz VFO frequency is divided by M, say 50,000, to give an output of around 100Hz which is applied to the clock input of a high-speed D-type flip-flop (labelled 'digital mixer') so that the maximum frequency output is about 50Hz. If N is the ratio of the

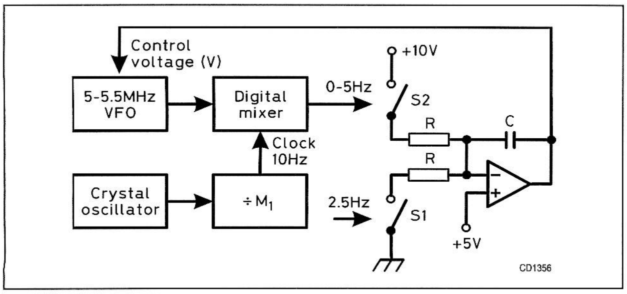

Fig 4.32. 'Huff and puff' oscillator block diagram *(QEX)*

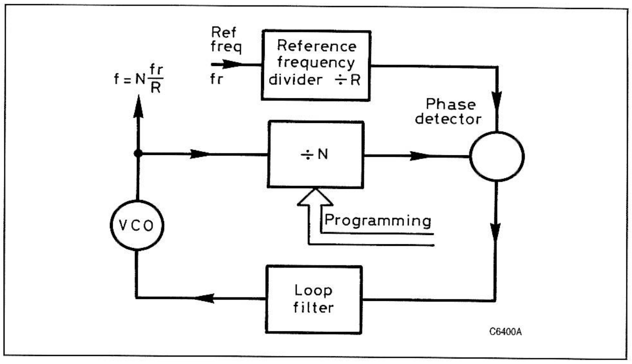

Fig 4.33. Basic PLL synthesiser

crystal oscillator to the output (100Hz) of the divider, in this case N = 500,000, then the system will be stable and:

$$F_{VFO}/M = F_{xtal}/N$$

which rearranges to:

$$F_{VFO} = F_{xtal} \times M/N$$

With the above values of M and N and the crystal oscillator, the VFO frequency will be 5,000,000Hz (5MHz). The next stable point will be where N goes to $N + 1$ which results in a VFO frequency of 4.999990MHz, ie 10Hz lower, and so on. The steps are 10Hz only with these parameters and are perfectly suitable for reception of Morse or SSB.

   On manual tuning the frequency 'creeps' to the nearest lock point and stays there. Bear in mind it is only as stable as the crystal oscillator which should be the best possible.

4. *Direct digital synthesis* (DDS) [10] where a digital signal several times higher in frequency than the desired frequency is applied to a digital device (a high-speed ROM – read only memory) which synthesises an approximate sine wave from which a pure sine wave is extracted using a low-pass filter.

All have their advantages and disadvantages:

1. Straight oscillators are generally unsatisfactory on stability grounds although, with great care, they could be used for a 50MHz receiver. They do have the lowest noise, ie the smallest noise side-bands.
2. Mixer oscillators also have a low noise and their stability is better than straight analogue oscillators because a small drift in frequency applies directly to the output while in the latter it is multiplied by whatever the original frequency was multiplied. Care must be taken with the two frequencies, the variable and the fixed, to ensure that harmonics and other mixer products don't fall within the receiver band and cause interference (*birdies*).
3. There are several forms of synthesiser. The basic PLL (Figs 4.33 and 4.34) uses a programmable divider for the voltage-controlled oscillator (VCO), the output of which is fed to a phase detector and compared with the fixed frequency divided down from a crystal oscillator. The output of the phase detector is a direct voltage proportional to the phase difference and is fed back to the VCO to counteract the drift. PLL synthesisers tend to have a higher phase noise than other types. The 'Huff and Puff' has a lower phase

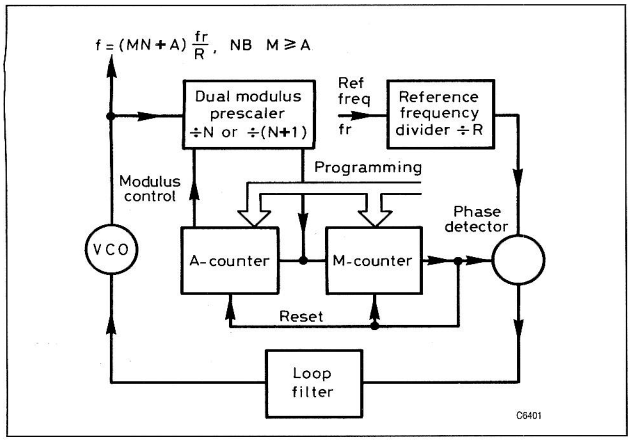

Fig 4.34. Dual modulus prescaler

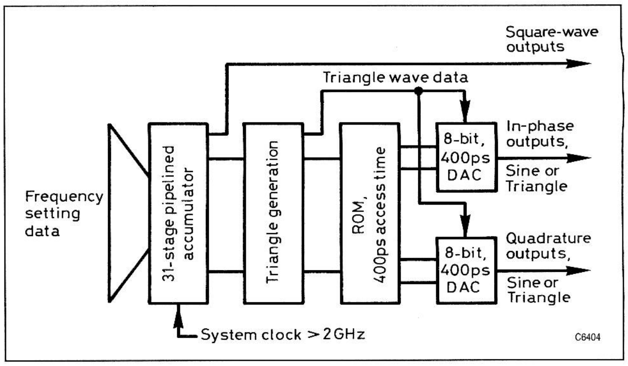

Fig 4.36. More complex DDS

noise than a regular PLL synthesiser but does not have a direct digital frequency readout and its frequency moves in steps.

4. DDS systems (Figs 4.35 and 4.36) need pulses generated at a very high frequency, at least three times the maximum frequency needed at the output, and consequently need very-high-frequency devices.

Both PLL and DDS methods depend on fast pulses which, in turn, generate many harmonics of the pulse repetition frequency (PRF). If not kept well screened, these can cause interference.

Fixed-frequency oscillators

These usually use quartz crystals as the frequency determining element, equivalent to the tuned circuit in the VFO. The crystal has a very large *Q* (values of 50,000–1,000,000 are possible) and produces a very stable frequency. Quartz is cut in different angles for this purpose and these cuts have slightly different properties, in particular, their temperature coefficient [14]. For frequencies between 5 and 20MHz, the usual cut is 'AT' with a temperature coefficient of ±50ppm (parts per million) between –10 and +60°C.

At higher frequencies, *overtone* crystals are used which vibrate at a mechanical multiple of the fundamental frequency (rather as a violin string can be made to vibrate at its third or fifth harmonic) which is not an exact multiple of its fundamental frequency although they are very close. For example, a 8.2013MHz crystal had a third overtone of 24.5773MHz while three times the fundamental is 24.6039MHz, a difference of 26.6kHz. All overtones are odd numbers and up to the seventh are available. Overtone crystals should be bought as such since they are finished in a different manner from fundamental crystals. See Fig 4.37 for a typical circuit.

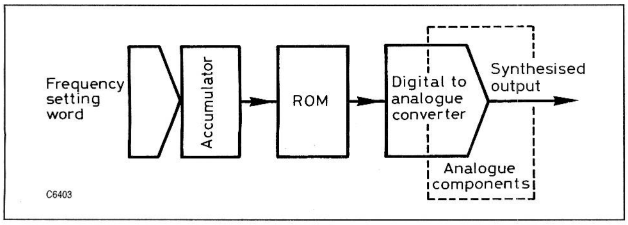

Fig 4.35. DDS concept

Crystal oscillators may be varied in frequency a little by adding an inductor and a capacitor to the circuit (see Fig 4.38(a)). This is the so-called *VXO* (variable crystal oscillator). Only a small deviation from the basic frequency is possible. Fig 4.38(b) shows a varicap-controlled VXO which can be used to cover upper and lower sidebands in a SSB receiver with one crystal.

Frequency measurement

Since the frequency to which the receiver is tuned is related to the oscillator frequency ($f_{rec} = f_{osc}$ ± IF for a single conversion superhet), measurement of the oscillator frequency with an offset of the IF gives the frequency to which the receiver is tuned. Digital frequency meters also depend on fast pulses and should be well screened from the rest of the receiver circuitry.

Intermediate frequency

Choice

IFs are chosen with three considerations in mind:

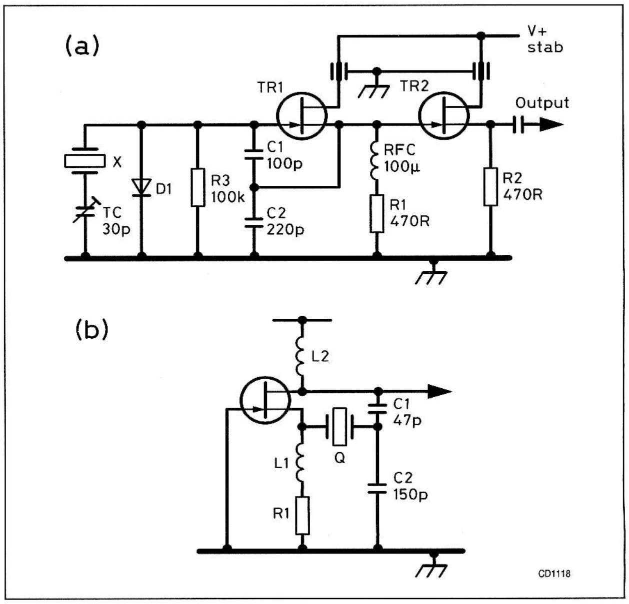

Fig 4.37. (a) Fundamental-mode crystal oscillator. The values shown are for a 10MHz crystal X. (b) Overtone-mode crystal oscillator. L1: 1.0µH. L2: 0.3µH. Q: 36MHz

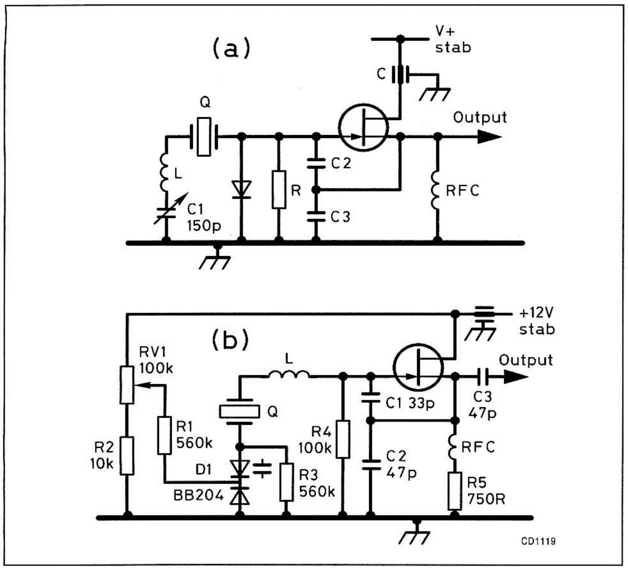

Fig 4.38. (a) L-C VXO. Values are for 10MHz crystal. (b) Varicap VXO with range 10.6990 to 10.7013MHz. L: 7μH. Q: 10.7MHz

1. Recognising the possibility of second-channel or image response. This means that the IF should be 7–10% of the operating frequency to make sure that the image is fully attenuated by the input circuits.
2. At the chosen IF there should be available relatively cheap crystal filters. This means that 10.7, 21.4 and 45MHz are preferred frequencies.
3. If dual conversion is used, the second local oscillator should not have harmonics within the band in use. For example, a 10.7MHz first IF and a 455kHz second IF needs a local oscillator at 11.155 or 10.245MHz. The former has a harmonic at 145.015MHz so the latter would be chosen for a 145MHz receiver.

If gain is needed between the first and second mixers in a double superhet, a highly linear amplifier with a large dynamic range must be used. A good example of this is a design by G3SBI [15] shown in Fig 4.39 with details of its performance in Table 4.4.

Table 4.4. Performance of 9MHz AGC-controlled amplifier

Noise figure (dB)		0.6	0.6
Input impedance (Ω)		50	50
Output impedance (Ω)		50	50
Third-order intercept (dBm)			
Vs = +12V	(max gain)	23	26
Vs = +20V	(max gain)	28	30
Input for 1dB compression (dBm)			
Vs = +12V	(max gain)	0	+3
	(max AGC)	+7	+11
Vs = +20V	(max gain)	+5	+8
	(max AGC)	+11	+14

Amplifier input and output impedances are 50Ω regardless of AGC-controlled gain. Gain range is 45dB.

Selectivity

This is provided at the 'front end' of the IF strip by crystal filters. The bandwidth of these depends on the mode to be used. For Morse, a bandwidth of 500Hz is normal, while 2.5–3kHz is used for SSB. FM requires about 7.5kHz for the present system of 5kHz maximum deviation and 25kHz between channels. When this is changed to 12.5kHz between channels, a lower maximum deviation will be used and the bandwidth will be narrowed. It is of the utmost importance that the filters are correctly terminated. The manufacturer will give details of the proper resistance and capacitance for this. It should be noted that sharp filters are subject to overloading and this can cause non-linearity.

Further down the IF amplifier chain, so-called *roofing filters* are used to prevent noise generated by the early IF amplifiers and outside the pass-band from reaching the detector. These can be simple ceramic resonators or further crystal filters.

A wide-band IF chain with 100kHz (or so) bandwidth may be used to amplify sharp noise pulses which can then be used for noise blanking by cutting off the input to the main chain be means of a semiconductor switch.

Gain

Gain can be provided by any suitable devices – bipolar transistors, FETs or ICs. At present, ICs seem the most used and many combine 'gain blocks' with various types of detector to produce an audio output. The overall gain needed is about 100dB and this should be controllable by an automatic gain control for Morse and SSB reception. FM reception requires that the last IF amplifier saturates and limits the signal, so removing noise 'spikes'. This feature is built into ICs that are specific to FM – they also have a *squelch* facility which shuts off the audio amplifier in the absence of a signal to prevent noise getting through to the audio stages.

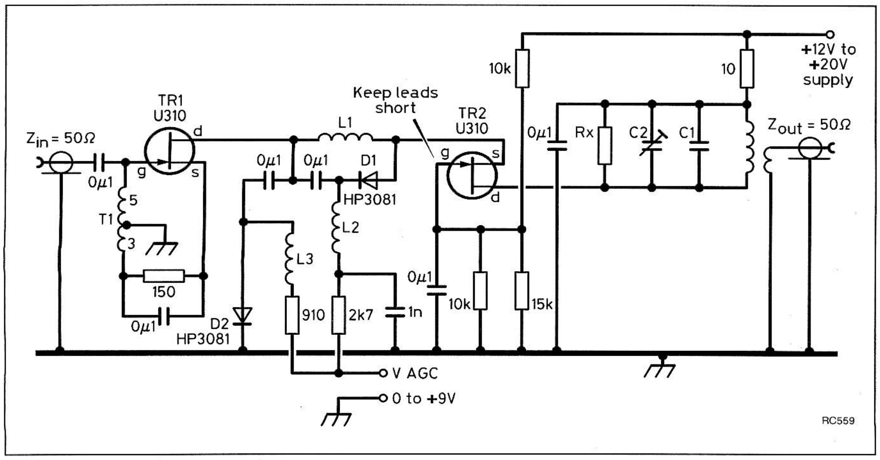

Fig 4.39. G3SBI's low-noise, AGC-controlled cascode IF amplifier

Demodulators or detectors

SSB/morse code

Here a product detector is the norm and it produces an output which is the

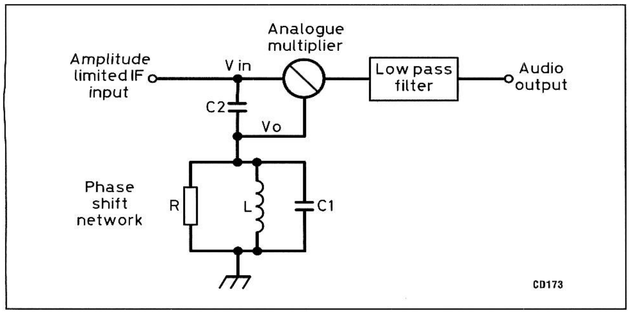

Fig 4.40(a). Block diagram of quadrature detector

product of the signal and local oscillator (or BFO, beat frequency oscillator) frequencies. This results in an audio signal, either speech from SSB or a tone from Morse code. A product detector is exactly the same as a mixer and all the mixer circuits can be used. Of course, at the lower frequency, other, simpler, product detector ICs are available such as the MC1496 which is a double-balanced mixer. At this signal level, exact balance is not important.

FM

ICs for FM demodulation are of three main types – the quadrature, the phase-locked-loop and the pulse counting. The *quadrature* or *co-incidence* detector (Fig 4.40(a)), as its name implies, splits the incoming signal into two parts, one of which is at 90° different in phase (quadrature) from the other and feeds both into an analogue multiplier. This results in a number of output frequencies including the original audio which is recovered through a low-pass filter. In a practical circuit (Fig 4.40(b)), the MC3361 carries out a second conversion to, say, 455kHz and generates the audio from that. It also contains a limiting amplifier and a crystal oscillator.

The PLL detector is similar to a PLL frequency synthesiser except that the voltage needed to lock the oscillator to the incoming frequency is the audio signal.

In the pulse-counting demodulator, a second conversion to a low IF of 100kHz or less is made and converted into pulses. The pulse rate is measured and corresponds to the modulating frequency. An IC such as the SL6601 has a limiting amplifier, a pulse generator, a counting stage and a crystal oscillator for the frequency conversion all in one package.

A great deal of emphasis has been placed on the ability of a receiver's front end to cope with multiple strong signals. Less has been published on the IF/AGC system of the receiver, despite the fact it determines how every signal sounds. This design by Bill Carver, K6OLG [16] covers that important territory.

The IF/AGC subsystem shown in Fig 4.41 has a minimum discernible signal (MDS) level of 0.03μV in a 2.5kHz bandwidth, and its AGC can be set to have a few decibels rise in audio with a range of signal amplitudes of less than 0.1μV to over 0.2V. The input-intercept point is about 20dBm, so in-band intermodulation distortion (IMD) is 40dB down even for S9 + 70dB signals. The constant-gain time interval of the 'hang' AGC circuit is smoothly and continuously varied from 100ms to 2s using a panel-mounted control. This IF/AGC system is intended for use with a front end having a net gain of +3dB and a 6dB noise figure (NF), and *directly* drives a +7dBm diode-mixer product detector.

AGC basics. An amplifier with feedback to control its gain can be resolved into two components: the amplifier itself and the detector and processing circuits that develop the gain-control feedback voltage.

The Analog Devices AD600 is a dual, low-noise, wide-band, variable-gain amplifier IC. The gains of the two internal 40dB amplifiers are controlled by the potential difference between the CxLO (that is, C1LO or C2LO) and CxHI (C1HI or C2HI) pins. This IC has about a 40dB gain control range with a constant-gain control scale factor of 32dB/V. CxLO voltage above or below the 0.625V difference leaves the gain at 0 or 40dB respectively. With a 2dB NF and 30MHz bandwidth, it's ideal for the gain portion of an IF system.

The logarithmic envelope detectors produce a 3V output change from an 8dB input signal change, a scale factor of 0.375V/dB. Coupled with the AD600, this produces a loop gain of 12, which means an input-signal

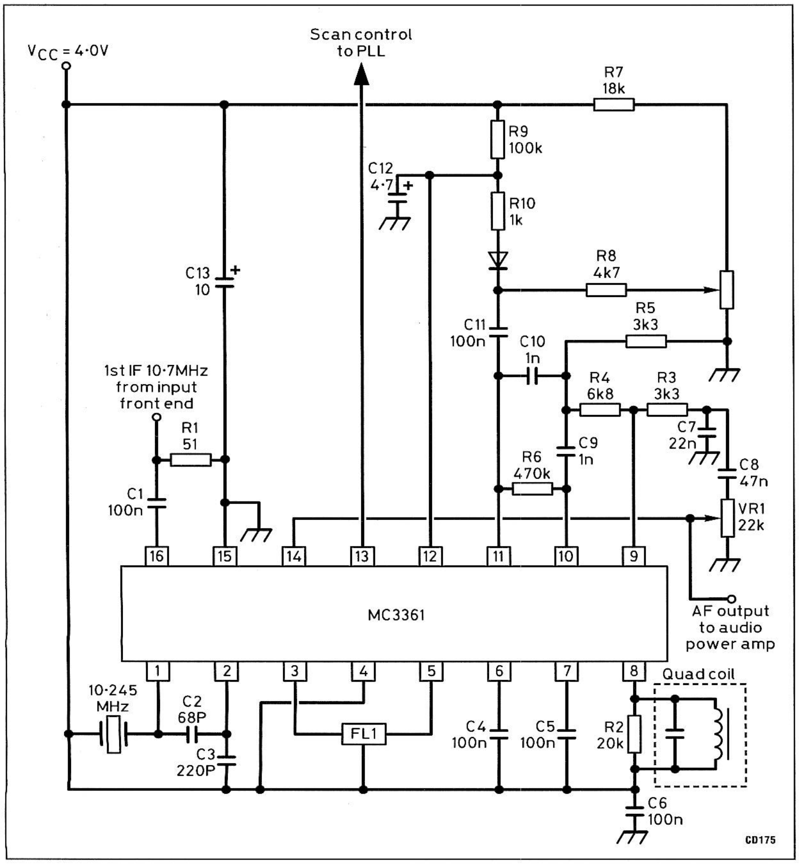

Fig 4.40(b). Practical circuit using the Motorola MC3361 (Motorola)

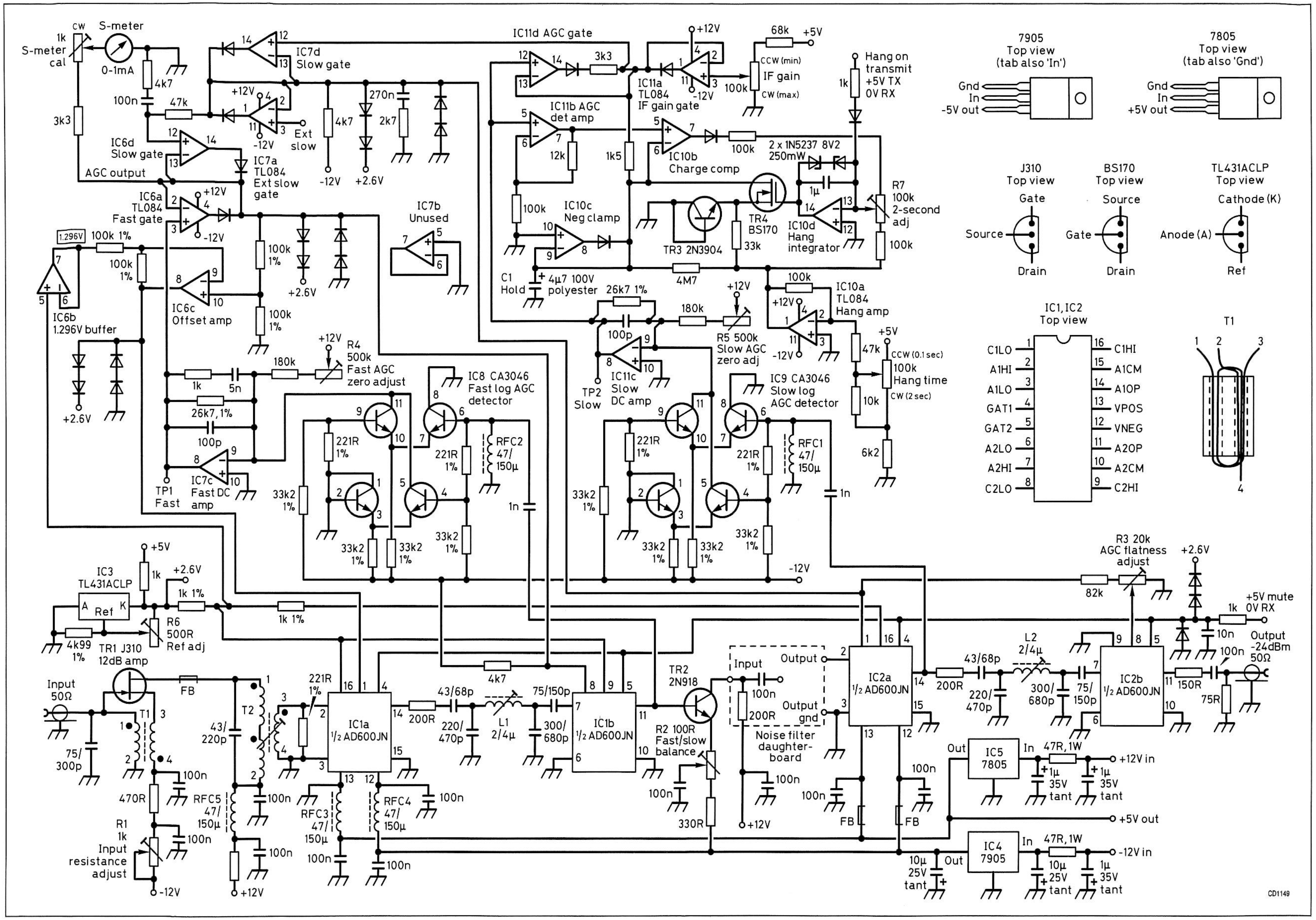

Fig 4.41. The IF/AGC circuit. Where two values for a component are shown (one before and one after an oblique line) the upper value is for use at an IF of 9MHz, the lower one for a 4.434MHz IF. IC8 and IC9 are shown uncharacteristically here so as to provide a better understanding of circuit operation. All diodes 1N4148 or equiv except as noted. FB = ferrite bead, -43 material (FB-43-101 ferrite beads can be used). L1, L2: (4.434MHz) 26t 30 AWG Amidon L45-7; (9MHz) 8t trifilar 30 AWG Amidon L43-6t. T1: (4.434 and 9MHz) constructed as shown on BN-61-202 ferrite form (do not substitute ferrite type), 1-2 = 4t 28 AWG, 3-4 = 3½t 28 AWG. T2: (4.434MHz) 15t trifilar 30 AWG on L45-7; (9MHz) 8t trifilar 30 AWG on L43-6 *(QST)*

change causes an output-signal change equal to the input signal change divided by 13. For example, a 65dB input signal change results in a 5dB output-signal change.

Two AGC detectors? Wide-band noise at the AGC detector would preclude using AGC for low signal levels; the amplifier's noise bandwidth must be limited. But the time delay of a filter placed between controlled stages and the AGC detector destabilises an AGC system, so AGC voltage to the gain-controlled stages ahead of the filter must be delayed. This permits brief clipping of large signals and envelope distortion at the leading edge of a signal.

This amplifier system uses two AGC detectors. The first one, called 'FAST', reduces the gain of the stages ahead of the noise filter, preventing them from clipping until the 'SLOW' detector – placed after the filter – can catch up. This prevents overloading IC1a and IC1b for signal levels above 10μV for a few milliseconds until the SLOW AGC responds. This 'minor' detail makes a big difference in how a receiver sounds even though most ears can't identify why.

Identical FAST and SLOW detectors operating at the same signal level – combined with the well-defined gain control characteristics of each AD600 – permit seamless combination of the detector outputs. TR2's gain is adjusted so both detectors have the same signal voltage when IC2a's gain is at minimum.

The JFET and first three AD600 amplifier stages. The AGC threshold occurs when a signal level of –32dBm appears at the output of IC2b. This results from an input-signal level of about –128dBm (0.09μV) when the signal-to-noise ratio is approaching 10dB. Maximum signal is reached with –20dBm at the output of IC2b, corresponding to an input-signal level of 0dBm (0.23V).

TR1 is a J310 FET with gate-source transformer feedback. Its theory of operation is covered in [17]; an almost identical implementation to this one was described by Colin Horrabin, G3SBI [15].

TR1's gain is 12dB and is set by the 2:1 transformer (T2), the load resistance of the following stage and the 221Ω metal-film resistor. The resistor value is optimised for the lowest AD600 noise figure. To produce a precise 50Ω input resistance, TR1's source current is adjustable and a capacitor across the input terminals cancels the leakage reactance of the feedback transformer T1. Minimum transformer loss is necessary to achieve a low noise figure, so the input transformer (T1) must be wound exactly as shown on a BN-61-202 balun core.

There are four amplifier stages in two AD600 packages. The first 40dB of gain reduction is done only by the third stage; the next 40dB in the second stage, and only the last 40dB reduction applied causes any gain reduction in the first stage. As AGC voltage passes halfway between the threshold of adjacent stages, gain control is being handed off from one IC to the next. This sequential gain reduction maximises signal-to-noise ratio.

To fully understand the AD600, you'll need a data sheet. Briefly, as the AD600's CxLO control voltage (either C1LO or C2LO) changes from –0.625 to +0.625V with respect to its CxHI, its gain changes linearly from +40dB to 0dB (unity gain). Because C1HI of IC2a is biased to 0.649V, its gain will change by 40dB as the AGC voltage at IC7 pins 2 and 13 changes from 0.024 to 1.274V. C2HI of IC1b is at 1.947V, so

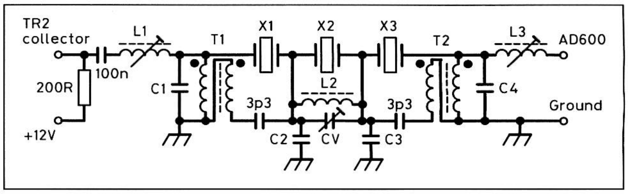

Fig 4.42. The ladder filter. CV: 1.5–7p. For X1–X3 = 4.434MHz, C1: 47p; C2, C3: 12p; C4: 53p (47p + 5p); L1: 11.7mH (27t 30 AWG on L-45-7 core in can); L2: 160mH (60t 34 AWG on FT-37-61); L3: 6.4mH (23t 30 AWG on L-45-7 core in can); T1, T2: 11.5t bifilar 30 AWG on BN-61-203 core

its gain will change by 40dB as the AGC voltage at IC6 pins 2 and 13 changes from 1.322 to 2.572V. The small gap between 1.274 and 1.322V provides a smooth transition of gain control from IC2a to IC1b, as explained in the AD600 data sheet.

The threshold of IC1a is also 1.947V, but 1.298V is subtracted from the AGC voltage by IC6c. Thus, the 40dB gain change of IC1a occurs as the AGC voltage changes from 2.620 to 3.870V. Taken together, as the AGC voltage at TP2 varies from 0 to 3.870V, the gain of the three cascaded stages changes by 120dB, with gain reduction starting at the last amplifier stage.

Noise filter. The home-brew crystal-ladder filter between the second and third AD600 stages is not intended for selectivity. This filter removes the broad-band noise generated by TR1 and IC1, permitting a lower AGC threshold. The filter is 2500Hz wide at the –3dB points, and 5500Hz wide at the –20dB points, removing much of the noise from the opposite side of zero beat, improving the IF NF and thus the overall receiver NF. Selectivity should be accomplished ahead of this IF strip.

After measuring the series resistance and motional inductance of your crystals, ladder-filter design becomes essentially a cookbook process using the X software bound with Wes Hayward's *Introduction to RF Design*. At 4.43MHz. the usual crystal-ladder filter cannot be made 3kHz wide. Hayward's article 'Refinements in Crystal Ladder Filter Design' in the June 1995 issue of *QEX* shows how the holder capacitance of each crystal can be parallel-tuned to permit construction of SSB-width filters.

Fig 4.42 shows the ladder filter. The middle crystal is parallel tuned as Hayward suggests. However K6OLG 'neutralised' the capacitance of the input and output crystals with a phase-inverting transformer, self-resonant at the IF. Winding information is given for 4.434MHz. Small trimmer capacitors could have been used to precisely adjust the neutralisation, but fixed-value 3.3pF capacitors work well and eliminate some adjustments.

At the 2500Hz bandwidth, the filter impedance is 2.9kHz. L-networks are used at each end to match the filter to the 100Ω input impedance of IC2a at one end, and the 200Ω collector resistor of TR2 at the other end. Every set of crystals will be different; exact values for the L-networks must be determined after the filter impedance has been computed by Hayward's X program. Once the filter is designed, breadboard it and verify proper operation before committing parts to a PCB.

At 9MHz, neither parallel tuning nor neutralising is necessary to get the SSB bandwidth and, if used only for receiving digital modes, you can even omit the neutralisation at

4.434MHz. Making a three-pole filter sounds more difficult than it is: with the X program and one of the many tutorial articles on ladder filters, it's a tedious process.

It is anticipated that many will choose to use a commercial filter instead of home-brewing a ladder filter. There's enough gain in TR2 to absorb a filter loss of up to 10dB. TR2's 200Ω collector resistor value can be increased to about 330Ω to match a commercial filter's impedance and eliminate one L-network. Using more filter poles and/or narrower filters may require increasing the value of C2 to delay the SLOW AGC more than the filter delay, although Harold Johnson, W4ZCB, didn't find this necessary with an eight-pole SSB-width filter he used.

Forward gain correction of the output. The fourth AD600 stage, IC2b, is outside the AGC loop. A signal 130dB above threshold causes the OFFSET SLOW AGC voltage at TP4 to increase from its zero-signal value of 0.625V to 3.125V, and with a loop gain of 13, the output of IC2a has increased by about 10dB. When about 10% of the OFFSET SLOW AGC voltage is applied to IC2b's C2LO pin, its gain decreases by 10dB, resulting in no signal change at its output.

R3 can be adjusted so the output is flat within a few decibels for signal levels of 0.23μV to 0.23V (−120dBm to 0dBm). Because of finite signal-to-noise ratio for small signals, it is not possible to compensate perfectly, but the resulting flatness is remarkable and addictive. Some may prefer the audio level to rise somewhat with increasing signal. R3 allows adjustment of IC2b's contribution to gain control to suit each builder's taste.

The output of IC2b is attenuated by a 150/75Ω resistor pair to produce a −26dBm signal level and 50Ω output resistance perfect for the +7dBm diode-mixer product detector. IC2b and the resistors produce low IMD and a high S/N ratio from the product detector while preventing BFO signal leakage to the SLOW detector.

Reference-voltage details. IC3 is a TL431 shunt regulator whose 2.6V output is divided by four identical 1% tolerance resistors to produce three close-tolerance voltages with only one adjustment: 1.947V for the C1HI threshold for IC1a and IC2b; 1.298V to offset the AGC voltage for IC1a and 0.649V for the threshold of IC2b. R6 is adjusted to produce 1.947V at TP3. The 1% tolerance resistors need only have the same value – between 1kΩ and 5kΩ. DigiKey and other suppliers offer suitable resistors at a modest price.

The AD600 CxLO and MUTE pins have 15 small-signal silicon diodes clamping the pins to ground or to the 2.6V potential at IC3. Normally, the diodes won't conduct. They're there to prevent damage to the expensive AD600s in case of misconnection, loss of a power supply or op-amp failure. Being a shunt regulator rather than a three-terminal regulator, IC3 is able to sink diode current should a fault occur.

The AGC detectors. The two AGC detectors, IC8/IC7c and IC9/IC11c, are not rectifiers but an interconnection of matched transistors that produce an average current equal to the logarithm of the applied IF signal over a range of signals. These detectors produce several DC volts from only 10mV of signal, and their logarithmic characteristic complements the AD600 scale factor almost perfectly.

Processing the AGC voltages. The large number of op-amps makes this circuit look complex. But the ideal performance of op-amps isolates each component's contribution to the circuit, making it easier to understand and troubleshoot than it may appear.

The SLOW detector output at TP2 and the IF GAIN potentiometer voltage at IC11a pin 3 are gated by IC11a and IC11d. Whichever one has the higher voltage charges the HOLD capacitor, C1, through its diode and a 4.7kΩ resistor. The output pin of the amplifier whose input is lower (not in control of the output) will swing to about −11V. This bizarre action is perfectly normal and logical, but takes some getting used to when troubleshooting. Similar action occurs at IC6a–IC6d and IC7a–IC7d.

The output from IC11a–IC11d is combined with an external AGC signal (if present) in IC7a–IC7d. Whichever signal is higher appears at IC7 pins 2 and 13, and controls the gain of IC2a. The output of IC7a–IC7d is combined with the FAST AGC detector output in IC6a and IC6d to control the gain of IC1b and drive the S-meter. Buffered by IC6b, 1.298V is subtracted from the combined detector outputs in IC6c to become the OFFSET AGC voltage with a total swing of −1.2 to 2.6V appearing at IC1a pin 1 to control its gain.

Hanging AGC. When the SLOW detector output is more than 90% of the voltage on C1, IC10b's output (pin 7) will be about 11V. Current through the 47kΩ resistor and diode to IC10D pin 13 charges the 1μF integration capacitor C1, its output moving toward the negative supply until the 1N5237 zener diode starts to conduct with about −8V at pin 14 of IC10d. TR4 is cut off by the negative base voltage and no drain current flows to discharge C1. The 4.7MΩ resistor slowly discharges C1, permitting the AGC to track modest amounts of fading with TR4 cut off. Clamp IC10c provides current as necessary to keep TR4 from developing a negative voltage on C1.

When the SLOW AGC detector voltage is not 90% of the voltage on C1, IC10b's output is negative and the integrator output swings in the positive direction at a rate determined by the setting of the HANG TIME potentiometer. It stops charging when C1 is discharged by TR4 down to 90% of signals-plus-background noise, or the other 1N5237 starts to conduct with 8V on pin 14 of IC10d. When the HANG TIME pot is fully clockwise, it not only takes a longer time for receiver gain to return, it returns at a slower rate.

S-meter. The linear-in-decibel nature of the AD600 AGC voltage means the 0 to 3.87V swing at TP2 corresponds to a 120dB gain change, plus the signal rise at the output of IC2a, a total of about 130dB. The AGC voltage is almost perfectly logarithmic or linear in S-units over this range.

Multiplier resistors develop a 1mA meter current for 3.75V at TP2. With appropriate changes to the resistors, more sensitive meter movements can be used. It's convenient to have a portion of the multiplier adjustable; K6OLG set his meter so the needle was vertical for an S9 signal. The linearity of the AD600 permits the S-meter reading to be corrected for the gain of a preamp or loss in a switched attenuator. For example, the voltage across the coil of a relay used to switch an attenuator can be fed to the S-meter through a variable resistor. Hence, the reduction in AGC voltage caused by the attenuator is perfectly compensated for by current developed from the coil voltage.

Adjustment. Seven potentiometers need adjustment. These adjustments do not interact and six of them are quick and easy to make.

1. Using a noise bridge (or other suitable impedance-measuring instrument), set the input resistance to 50Ω with R1. A ground connection is provided at the top side of T1 so that an additional half-turn can be added to the 1-2 winding of T1 if varying R1 does not produce the 50Ω input resistance. The input reactance should be very close to zero. At frequencies other than 4.434 or 9MHz, scale the input capacitor's value and verify zero reactance using a noise bridge.
2. Set R6 for 1.947V at TP3. The potential at IC6b pin 7 should be very close to 1.298V, (1.272min, 1.324 max). The voltage at IC2 pin 16 should be close to 0.649V (0.636min, 0.662 max).
3. Turn the IF GAIN pot fully counter-clockwise, to minimum gain. Adjust R5 for 0V at TP2, then adjust R4 to 0V at TP1.
4. Turn the IF GAIN control clockwise, and connect a signal generator to the input. Peak T2, L1 and L2 for maximum gain. These are broad-tuning, low-*Q* circuits. Adjust the noise filter L-networks and verify the expected filter bandwidth.
5. Apply sufficient signal to produce more than 1.6V at TP1. Adjust R2 so that the voltage at TP1 decreases slightly. This is the only tricky adjustment; don't be surprised if you have to repeat it several times before you get it right. Mid-scale S-meter linearity depends on proper balance of the two detectors. Large changes in the S-meter reading for small signal changes of around 10μV is a sure sign that R2 is not adjusted properly.
6. Connect a pulsed test signal to the input (receiving the signal from a keyed transmitter connected to a dummy load will do). Turn the HANG TIME potentiometer clockwise to its two-second position. Adjust R7 for a two-second delay between the end of the signal and the beginning of the voltage drop at AGC OUTPUT.
7. Adjust R3 (FLATNESS) to suit your taste. This can be done by ear – while listening to an on-the-air roundtable for example – or by using a signal generator and test instruments.

Additional points. This amplifier is insensitive to exact supply voltage, but the supply voltage needs to be stable so that the zero adjustments R4 and R5 will be stable. The +12 and −12V potentials are supplied by off-board 7812 and 7912 regulators respectively.

The 10kΩ resistor between the HANG TIME pot's wiper and the pot's low end simulates an audio-taper pot (these are hard to find!), providing the proper feel with a common linear-taper control. The IF GAIN control is also a linear-taper control. For computer control, both HANG and IF GAIN can be controlled by a 0 to 5V DC signal from a digital-to-analogue converter.

The 150μH RF chokes are parallel-resonant near the IF. RF choke values of 100 to 200μH will have virtually identical performance at 4.434MHz. At 9MHz, values between 27 and 47μH are suitable. There need be no fear of substitution.

Because of fading, on-the-air signals are not an ideal source of test signals. A keyed transmitter, known to have a good keying envelope, can be used with an oscilloscope to confirm proper circuit operation. Alternatively, you can use a diode mixer with a signal generator tied to the RF port, drive the DC-coupled IF port with a pulse generator and take the output from the LO port.

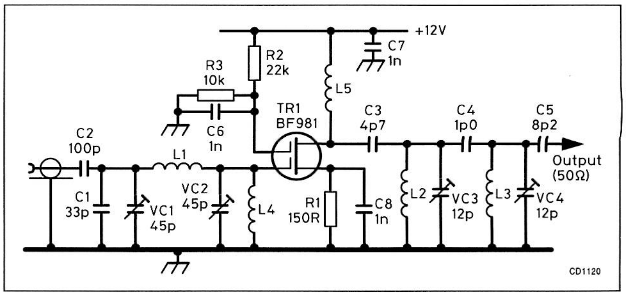

Fig 4.43. 50MHz RF receive amplifier. C6, C7: 1nF chip. L1: 0.13μH. L2, L3: 0.56μH. L4, L5: 4.7μH

Transmit hang, diversity reception and external AGC. A 5V logic potential during transmit forces the hang integrator output to −10V and C1 discharges slowly through the 4.7MΩ resistor during transmit intervals. In fast-break situations (full QSK CW, AMTOR and sometimes even SSB) this causes the receiver gain to return to nearly its previous level rather than full gain. C1 discharges to 0V during longer transmissions.

With two IF strips and two mixers, connecting both SLOW to the EXT SLOW of the other gives gain control to the stronger signal, providing diversity reception to combat multipath reception on the digital modes. Some older, but very good Fredrick modems provide an AGC output that may benefit from the EXT SLOW connection.

Construction. K6OLG successfully used IC sockets on an 'ugly' groundplane breadboard, and with several iterations of PC board layout. On the other hand, an etched and drilled PC board saves a lot of time. Like the hand-wired board described earlier, the PC board is double-sided with a ground plane and the components on top, traces on the bottom. To minimise cost, the board doesn't have plated-through holes; component leads and grounded socket pins are soldered directly to the ground plane.

Although this amplifier is stable even as a breadboard, remember that it has a lot of gain. Its stability can become marginal with scope probes radiating signals at amplifier outputs. Given the opportunity, the amplifier can pick up the BFO signal. K6OLG eliminated this possibility by placing the BFO and product detector in a shielded box. In the final package, each circuit should be in its own shielded box and no BFO signal should be detectable in the IF output.

Designs for specific bands

RF amplifiers

With suitable switching, an RF amplifier can with advantage be placed at the feed point of the antenna – its gain will offset the loss of the feeder cable. Gains of 15 to 25dB will normally be sufficient to overcome that loss and the noise of the first mixer.

50MHz. Circuits based on the BF981 dual-gate MOSFET have been widely used. In one typical circuit (Fig 4.43), an input pi-network is used to match the antenna to the gate circuit of

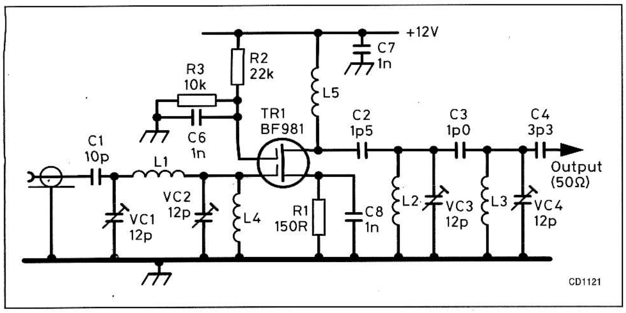

Fig 4.44. 145MHz RF receive amplifier. C6, C7: 1nF chip. L1: 0.08µH, L2, L3: 0.1µH

the device and the output is taken to a band-pass filter and thence to the mixer. Tuning is fixed at the centre of the band so only the local oscillator needs to be tuned.

70MHz. There are few published circuits for 70MHz equipment because the band is of such limited appeal (it is only available to the UK, Gibraltar and to parts of Cyprus). Most circuits that work at 50MHz will also do so at 70MHz with suitable reduction (about 16%) in the values of the tuning components, L's and C's.

144MHz. Here again, circuits based on the BF981 are widely used and Fig 4.44 shows a typical one. Apart from the tuning components, it is very similar to Fig 4.43 above.

For more advanced work where absolute minimum noise, maximum immunity to nearby strong signals and unconditional stability is necessary, there is a modern circuit (by G4SWX) as in Fig 4.45. This has a fixed tuned input circuit in the form of a quarter-wave line and an aperiodic 50Ω output. It has a gain of 18dB, a noise figure of 0.4dB and its 3IP is +8dB. Its output is of the correct impedance to feed a multiple helical filter and then a double balanced mixer of the diode ring type (see above).

It could with care be used at lower and higher frequencies. At lower frequencies, it will probably be necessary to replace the input circuit by a 'lumped' circuit of a conventional coil and capacitor.

432MHz. FETs, especially dual-gate MOSFETs, are the preferred RF amplifiers at this frequency. While silicon devices will give good results, the lowest noise comes from using

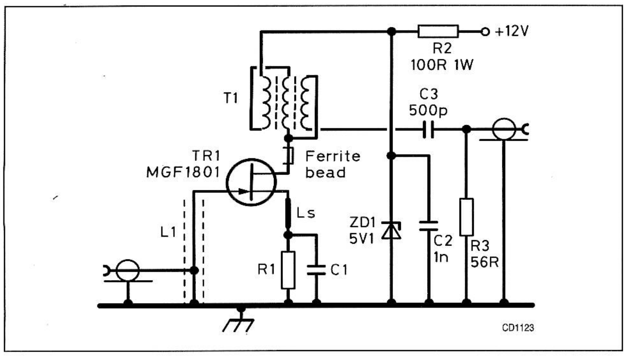

Fig 4.45. 144MHz MGF1801 preamp. R1: 2 × 47Ω chip. C1: 2 × 500pF chip. L1: 364mm long RG401, tap at 68mm from ground. Ls: 2 × 5mm. T1: 6t 30 SWG trifilar on T25-12 core connected as 1:2 transformer *(VHF/UHF DXer)*

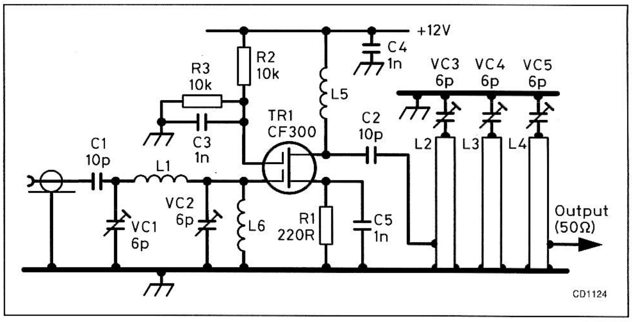

Fig 4.46. 432MHz receive amplifier. C3–C5: 1nF chip. L1: 25nH, L2–L4: stripline on 1.6mm thick epoxy glassfibre board, 70mm long, 2.8mm wide, spaced 2.0mm, L5: 0.1µH

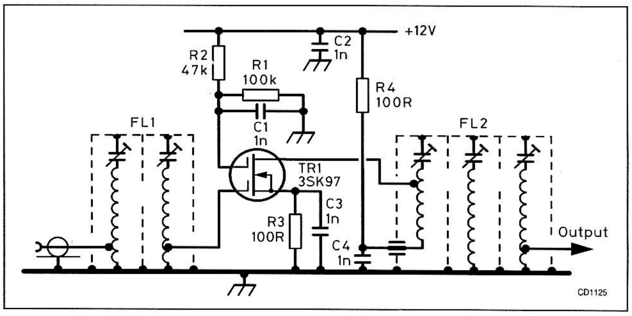

Fig 4.47. Wide-band 432MHz receive amplifier. FL1: 2-stage helical filter, FL2: 3-stage helical filter

GaAsFETs such as the 3SK97 or CF300. (*Note*: there are many others; these have been listed because they are most often mentioned.) The antenna input circuit can be 'lumped' but is usually resonant lines, either as wire/rod or as etched micro-circuits (see above).

A typical RF amplifier is shown in Fig 4.46 and consists of a tuned line input and output and a CF300 amplifier. Note the essential screen between the input and the output sides of the circuit. Setting up is done simply by tuning the input and output for maximum signal at the desired frequency then retuning the input slightly for minimum noise factor. As it stands, the selectivity is high, being 6dB down at ±5MHz off tune and 20dB down at ±12MHz so it needs to be tuned to the part of the band of interest.

A wider-band device is shown in Fig 4.47. This uses a GaAsFET, the 3SK97.

1.3GHz. Power GaAsFETs provide the best gain with the least noise and a good large-signal handling capacity. Fig 4.48 shows a typical design (due to WA7CJO [18]) which uses a half-wave coaxial line for its input circuit, an MGF1402 power GaAsFET and an aperiodic (balun) output circuit. Its gain is 15dB, noise factor about 0.4dB and its output impedance is 50Ω.

The tuned circuit is the most difficult part to make but it is well within the capability of an amateur with only a modest workshop. The tuned line consists of 25mm diameter copper tube, 87.6mm long with the centre part 9.5mm dia. The plain ends are soldered in place. The input capacitor (C1, antenna coupling) is a 18mm dia copper disc soldered directly onto a SMA connector and screwed into the cavity. Because thin copper tube will not take a thread, a nut of the right size is

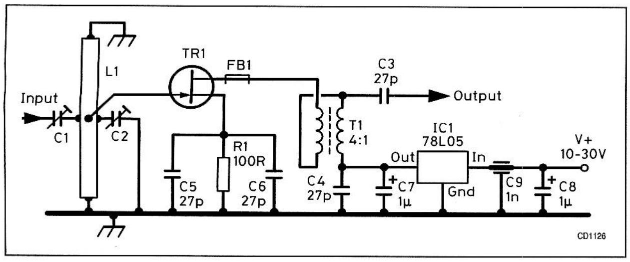

Fig 4.48. The WA7CJO preamplifier. See Table 4.5 for further details *(QST)*

Table 4.5. Components list for the WA7CJO preamplifier

R1	100R nominal, adjust to get 8–15mA drain current
C1	0.7in dia, 20-mil thick copper disk attached to the SMA connector
C2	0.7in dia, 20 mil thick copper disc attached to a 10-32 brass flat-head screw
C3–C6	0.05 × 0.05in chip capacitors
FB1	Ferrite bead. Optional but may reduce the tendency of the circuit to oscillate
T1	4:1 balun, 3t 32 AWG enam wire, bifilar wound on a Siemens B62152-A008-X-060 double-aperture core
TR1	MGF1802, MGF1412, ATF10135 or similar GaAsFET

soldered onto the tube and the connector locked in the optimum position by a lock-nut. The tuning capacitor C2 is made in the same way but the disc is soldered onto a piece of studding. The connection to TR1 is made from 1.5mm dia wire. TR1 and the output circuit is mounted saddlewise in a box on top of the resonator.

Another method of overcoming the noise, stability and large-signal handling capabilities of GaAsFETs is to use multiple low-gain stages. The following design by Matjaz Vidmar, S53MV (formerly YT3MV), [19] shown in Fig 4.49 uses two stages of amplification to provide an overall gain of approximately 25dB, with very low noise figures attainable.

The L-band low-noise amplifier is housed in a small case made from 0.3mm thick brass plate, which is 50mm long, 20mm wide and 15mm high (Fig 4.50). It is thus small enough to permit no resonance below 7GHz, so that no absorber material is required for the damping of parasitic vibrations at these frequencies. The BNC sockets (UG 1094, without nuts or washers) must be soldered on, as shown in Fig 4.50.

Next, the six 470pF DC blocking capacitors are soldered into the tin case with sockets. They must be ceramic disc or trapezoidal capacitors, not wire ended, and their value can be higher. When soldering them in, make sure that both the metal coating of the disc capacitor and the brass plate surface are well pre-tinned, so that the fragile capacitors are not destroyed by direct contact with the base plate bending in the heat.

Warning – in no case use multi-layer capacitors, as used in surface-mounted device technology! These capacitors display high levels of internal parasitic inductance and loss resistance, with natural resonances down to under 1GHz. In spite of their small dimensions, surface-mounted device capacitors and other multi-layer capacitors are completely unsuitable for microwave applications!

All the resistors used in the low-noise amplifier are wired miniature types rated at 1/8W (0204). The source resistances marked with an asterisk in Fig 4.49 are not soldered in immediately, but are required only for calibration of the amplifier, and the final values are dependent on the ID tolerances of the GaAsFETs used.

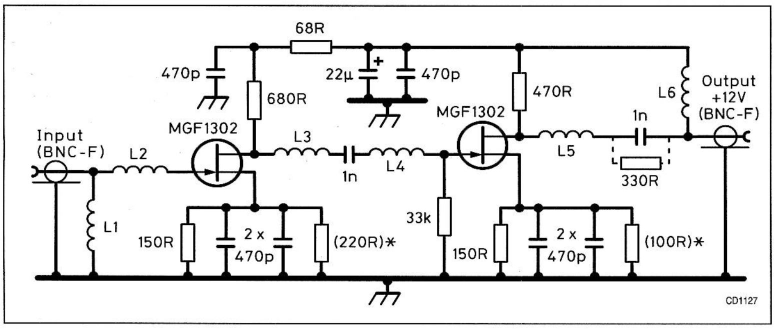

Fig 4.49. A two-stage GaAs FET antenna amplifier for L-band *(VHF Communications)*

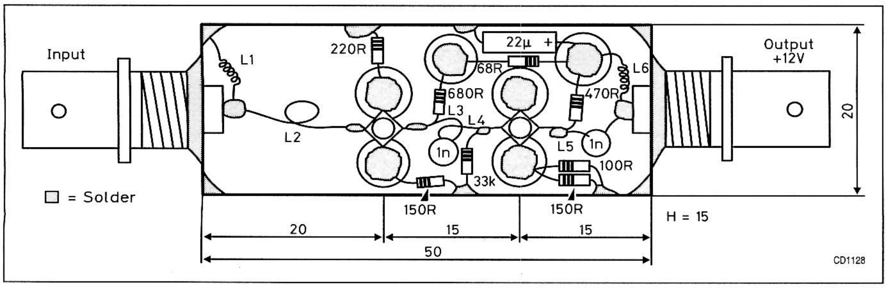

Fig 4.50. The low-loss structure with air as the dielectric makes an extremely low noise factor possible *(VHF Communications)*

The λ/4 chokes, L1 and L6, are each manufactured from a 6cm long piece of 0.15mm thick enamelled copper wire. The pieces of wire are first tinned over about 5mm at each end, and then the enamelled wire is wound around the shaft of a 1mm drill to form a coil. How many windings finally result from this is unimportant.

L2 is manufactured from a 0.6mm thick piece of silver-plated copper wire. For example, the internal conductor of an RG-214 unit can be used for this. For the frequency range from 1.5 to 1.7GHz, L2 has a single winding with an internal diameter of 3.5mm.

L3 and L4 simply represent the connection wires of the 1nF disc capacitor between the two transistors (Fig 4.50), each bent into a half-winding. Similarly, L5 is just a slightly longer connection wire of the output coupling capacitor. The inclination of the loop of L2 and the distance between L3 and L4 are adjusted last, on the basis of select-on-test methods.

The GaAsFETs are fitted last. After this operation, you should set your DC working points. In order to avoid any wild oscillations here, the input and output should be terminated with 50Ω.

The amplifier is now linked to an adjustable voltage source, which to start with should be set to about 7V. The DC voltage between drain and source is measured for both FETs. The thing to do now is to keep both drain-source voltages between 3 and 4V. To this end, you slowly increase the operating voltage and switch the initial source resistances of 150Ω in parallel again, until the final operating voltage of 12V is reached. Naturally, the parallel resistances should always be soldered on only when the operating voltage is switched off. The final voltage drop across the source resistances is typically 1 to 1.5V.

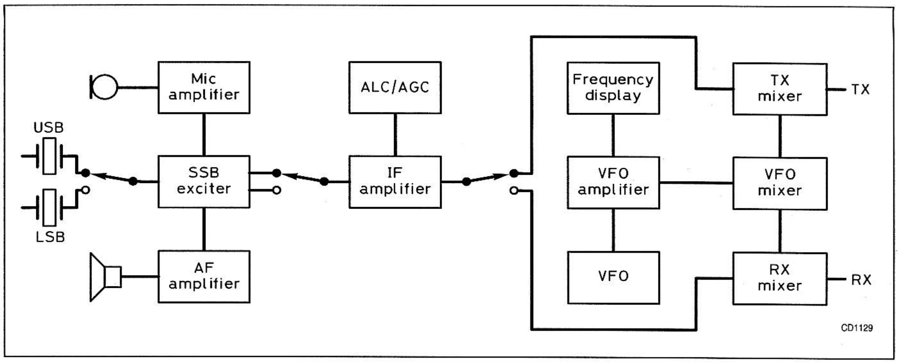

Fig 4.51. Block diagram of the 50MHz transceiver *(VHF Communications)*

Now the amplifier must be put into a test rig to check amplification and noise factor. For this, you need a noise generator suitable for this frequency range (which does not need to be standard for calibration), a receive converter and an SSB receiver with a large S-meter (if necessary, display the AGC voltage externally), or with a low-frequency voltmeter connected to the audio frequency output and with the AGC switched off.

The noise factor is predominantly influenced by L2. L3 and L4 are to be set to maximum amplification, as is L5, even if its influence is much less than that of L3 and L4.

L3 and L4 are normally each up to 10mm long. The distance between the two can be precisely adjusted. L2 can also be precisely adjusted in the same way if you compress or expand the loop slightly.

If a FET with lower amplification is used in the second stage, such as the MWT11, or an older type from the CFY range, and/or more amplification is desired, then the output network can be modified. The aim is to increase the DC through the second FET. To this end, you remove the 470Ω resistor and in its place solder in a 330Ω resistor parallel to the output coupling capacitor. This is shown as a dotted line in Fig 4.49.

COMPLETE RECEIVER CIRCUITS

It would be a pointless exercise in reinventing the wheel to describe a complete receiver for any particular band – it is normally assumed that a short-wave receiver or, in the case of 23cm, a 2m receiver, is available. Thus what is required are receive down-converters for each band in question.

Whilst receive down-converters do exist, it is more general to find receive converters combined with transmit converters in single units, each sharing common circuits. These units are transmit/receive converters or, as they are normally called, *transverters*. This now means that we shall take our presumption a stage further, and assume that either a 28MHz or 144MHz transceiver is available.

Finally, because we shall now place the emphasis on transverter designs this section shall be dealing not only with receivers, but transmitters as well.

50MHz

The design featured below by Wolfgang Schneider, DJ8ES, [20] is the exception to the 'rule' set out above and is for a full transceiver for the 6m band. The concept of the design is that each sub-assembly is a stand-alone 50Ω module, thus allowing for simple interconnection, switching etc. A block diagram of the transceiver is shown in Fig 4.51.

The received signal arriving from the antenna at the RX input is amplified in the RX mixer, passed through a filter and then transposed to the IF level (9MHz). For that the module requires an oscillator signal tuneable between 41 and 42MHz. In this way the range 51 to 52MHz is covered. The oscillator signal needed is provided by the VFO unit. The tuneable oscillator works in the region from 5 to 6MHz. The amplifier following operates as a buffer and at the same time decouples the frequency counter and VFO mixer from one another. Mixed with a 36MHz crystal oscillator, it produces the desired output frequency of 41 to 42MHz.

The digital frequency display has eight digits, so the last digit shows hundred hertz units. This counter evaluates the VFO signal directly. To achieve an accurate frequency display the difference from the wanted frequency (ie 45.000MHz) must be programmed. The received signal now reaches the IF amplifier. Here the crystal filter limits the bandwidth to the 2.3kHz necessary for SSB. For the signal-strength display (S-meter) the received signal from the AGC/ALC module is processed. At the same time this is used to set the regulator voltage for controlling the IF amplifier. For demodulation the SSB exciter is employed. In this process, by activating the corresponding crystal oscillator, the desired sideband is selected. At the output of the ring demodulator the audio achieved in this way is passed to a low-pass filter and a two-stage amplifier. The power amplification needed for loudspeaker operation can be undertaken in a separate module.

During transmit, speech from the microphone is amplified and mixed in the SSB exciter with the desired sideband oscillator. At the output we get a double sideband (DSB) signal. The switchable IF amplifier is used in the transmit path. The crystal filter selects the desired sideband. By using an AGC/ALC circuit we can avoid over-driving the following stages.

The transmit mixer converts the SSB signal already produced up to 50MHz. Like the receive mixer, this contains filters and amplifier stages. At the output of the TX mixer we have around 100mW available on the desired frequency.

The tuneable oscillator (VFO) (Fig 4.52)

In this transceiver an oscillator from a Collins receiver is used. VFOs of this kind or similar ones can be found at radio rallies and boot sales.

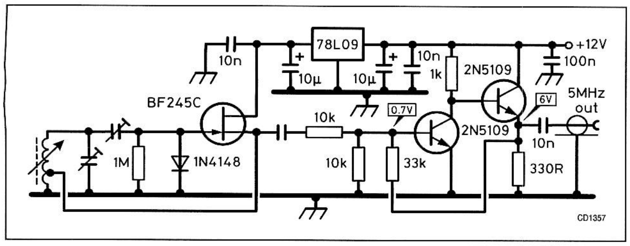

Fig 4.52. The tuneable oscillator (VFO)

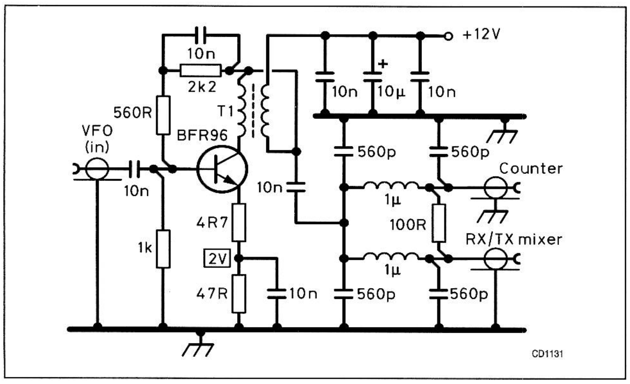

Fig 4.53. Broad-band amplifier

The VFO is designed around a Hartley circuit, which ensures good stability. The oscillator gets reverse feedback from a tap located about 10 to 25% away from the earthy end of the coil.

The higher the slope of the FET used, the less feedback is needed. This should only be enough to ensure foolproof operation of the oscillator – too much leads to instability.

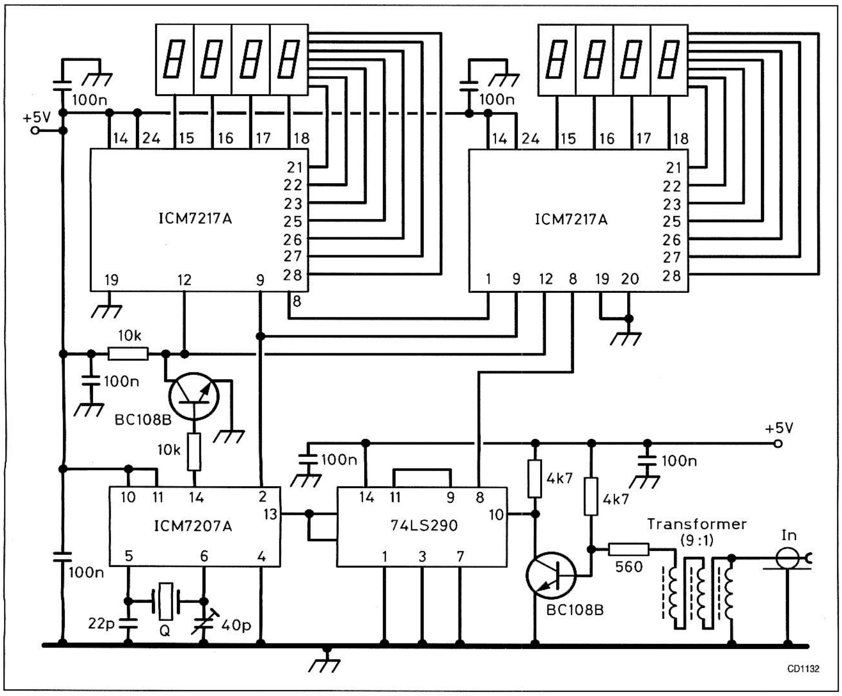

Fig 4.55. The programmable counter module

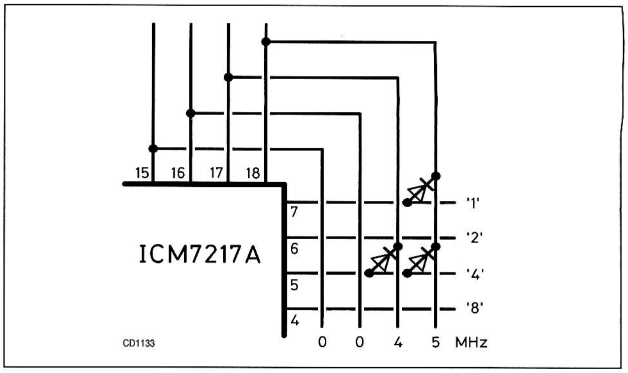

Fig 4.56. Sample programming arrangement

The oscillator is tuned using the core of the coil, giving a tuning range of around 5 to 6MHz. To decouple the VFO from the following stages a two-stage push-pull amplifier is used, and this ensures feedback-free operation of the VFO.

VFO broad-band amplifier and divider (Fig 4.53)

Following the VFO is a single-stage broad-band amplifier with an amplification of approximately 20dB. A Wilkinson divider is provided at the output to provide the two outputs, one for the frequency display and the other for the VFO mixer. A PCB design and component overlay are shown in Fig 4.54 in Appendix 1. The PCB measures 34 × 72mm.

Digital frequency display (Fig 4.55)

The digital frequency display is an eight-digit one and is constructed from two four-digit counters, type ICM7217A. These ICs are designed for seven-segment common-cathode LED displays and are programmable for display ranges. This feature is utilised here to programme a 45MHz offset in the display, required due to the fact that the counter is actually reading the 5 to 6MHz VFO. The method of pre-programming the display using diodes is shown in Fig 4.56.

The assembly is produced on two PCBs, one a standard display PCB for the LEDs, which can be remotely mounted from the driver board using ribbon cable. The driver PCB and component overlay is shown in Fig 4.57 (Appendix 1). Provision is made on the PCB to include the pre-progamming diodes for the 45MHz display offset (these are not shown in the circuit diagram).

VFO mixer (Fig 4.58)

In this 50MHz SSB transceiver a VFO signal in the range of 41 to 42MHz is required, taking into consideration the 9MHz IF being used. This will enable coverage of the 6m band from 50 to 51MHz.

The required VFO signal is produced by mixing the 5 to 6MHz VFO with a 36MHz fixed-frequency oscillator, thus producing an output of 41 to 42MHz. The

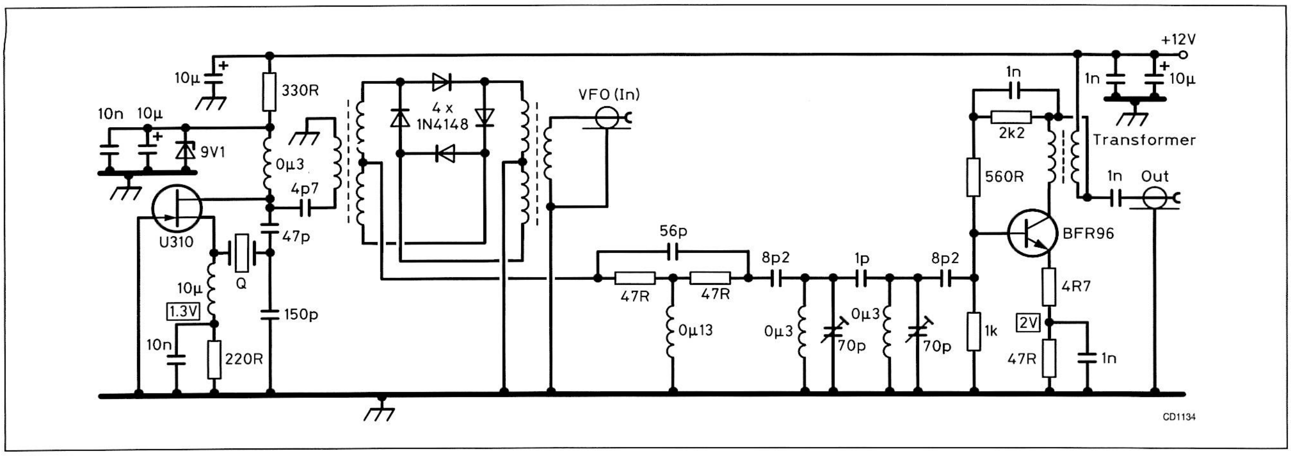

Fig 4.58. The VFO mixer

low-noise 36MHz oscillator is based around a U310 and produces of the order of 10mW into the ring mixer. The output from the ring mixer passes through a two-stage bandpass filter to a buffer amplifier, giving around 20mW tuneable in the range 41 to 42MHz.

In this and the previous broad-band amplifier, and elsewhere in this project, bifilar wound transformers are used. These are simply constructed as follows: take two equal lengths of 0.2mm diameter varnished copper wire and carefully twist around each other; six turns of this twisted wire are then wound onto a 5mm ferrite bead.

The transformer in the ring mixer is a trifilar one and is constructed exactly as the bifilar types but using three wires.

A PCB design and component overlay for the VFO mixer is shown in Fig 4.59 (Appendix 1). The PCB measures 53.5 × 72mm.

Receive mixer (Fig 4.60)

The receive mixer comprises a three-pole filter, first RF amplifier, ring mixer, second three-pole filter and a second RF amplifier. The PCB design is shown in Fig 4.61 (Appendix 1) and measures 72 × 72mm. The mixer is best aligned using a signal generator at the input set to 50.5MHz and a high-impedance detector at the output. The trimmer capacitors are then tuned for maximum output.

Transmit mixer (Fig 4.62)

The transmit mixer comprises a ring mixer, broad-band matching, three-pole filter and a broad-band two-stage amplifier.

With a drive level of 100µW (two-tone signal, each at –13dBm) there should be 55mW (+18dBm) per single tone at the output, corresponding to +21dBm PEP. Under these conditions the third-order intermodulation products (3IP) should be depressed by about 40dB and the 5IP products to 60dB. Although no harmonic filtering is included in this circuit, any harmonics produced should be below –40dBc.

A PCB design and component overlays for the transmit mixer are shown in Fig 4.63 (Appendix 1). The PCB measures 53.5 × 72mm.

IF amplifier with AGC/ALC (Fig 4.64)

The IF amplifier consists of a crystal filter, followed by an amplifier, a 20dB coupler for the AGC signal output, a pin-diode limiting stage and a final broad-band amplifier. A regulating signal from the AGC/ALC stage is fed into the pin-diode limiter.

The crystal filter used is an XF9B and is matched to 50Ω at its input and output by 9:1 transformers and 90pF trimmers. Special care should be taken when winding the transformers – any twist in the windings will have significant influence on the operation of the module. Matching of the

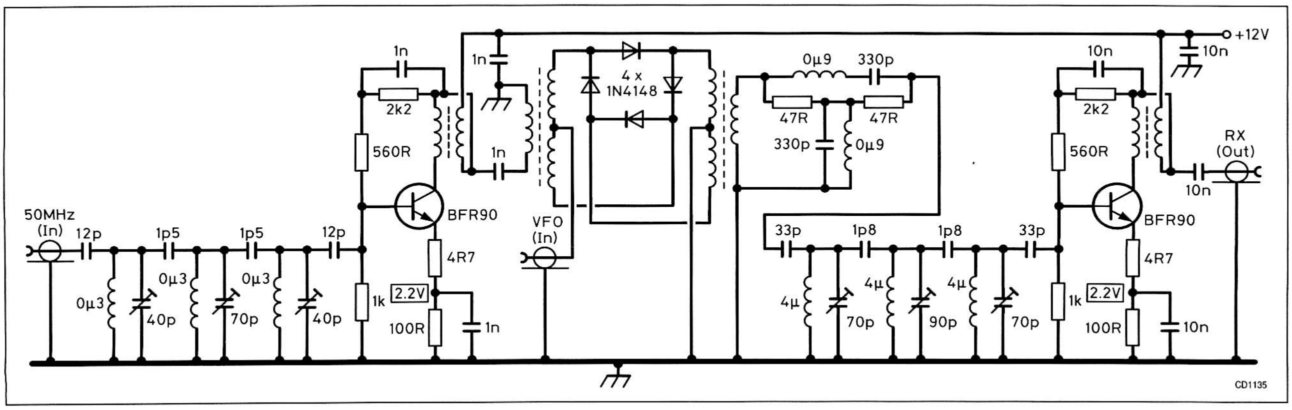

Fig 4.60. The receive mixer

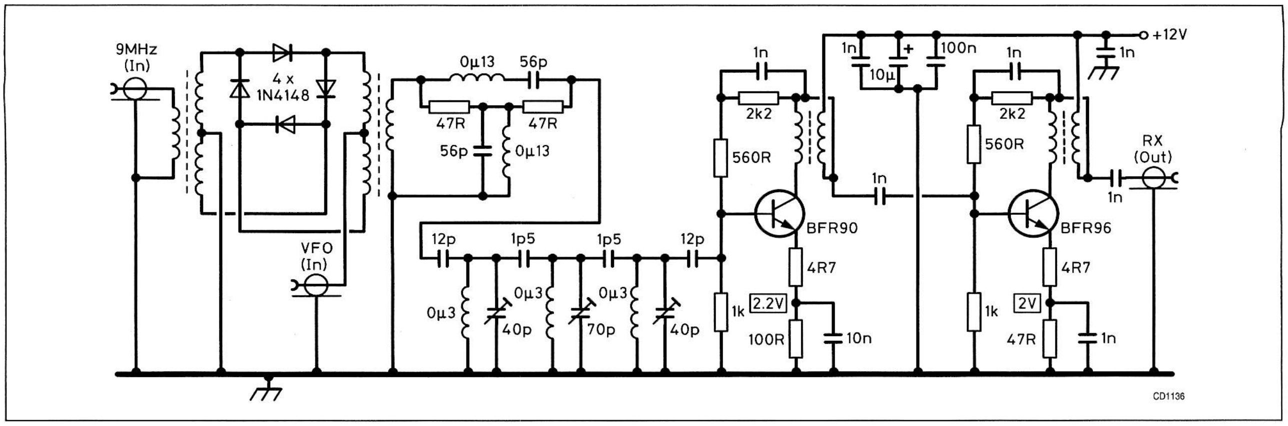

Fig 4.62. The transmit mixer

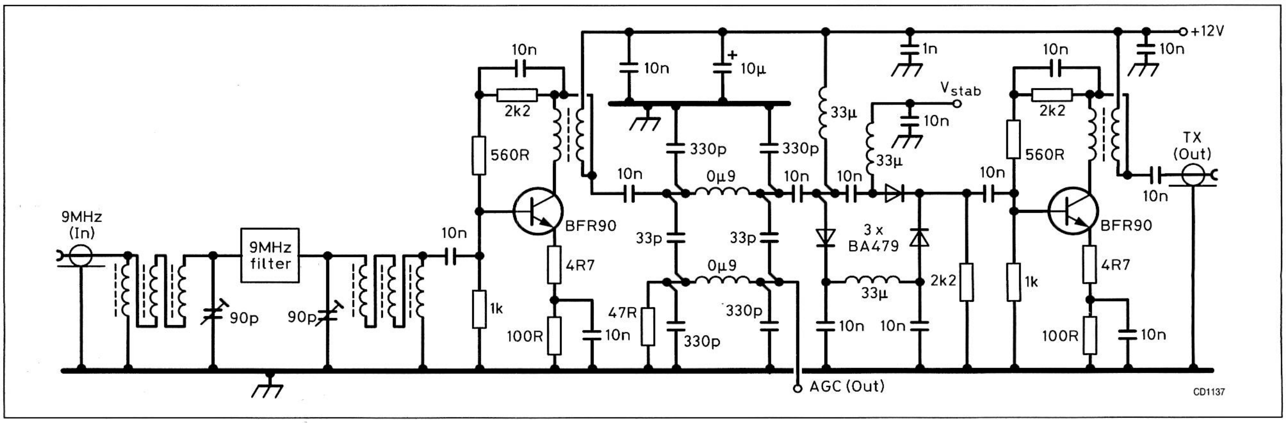

Fig 4.64. The IF amplifier

crystal filter is best achieved by feeding in a 9MHz DSB signal from the SSB exciter and tuning the trimmers for minimum residual carrier.

The PCB design and component overlay for the IF amplifier section is shown in Fig 4.65 (Appendix 1). The PCB measures 72 × 72mm.

The AGC/ALC module is shown in Fig 4.66 with the PCB design and overlay in Fig 4.67 (Appendix 1). The PCB measures 53.5 × 72mm. It is recommended that all the modules are built into tin-plate boxes. However, this is particularly important with this unit, as stray electrical fields may cause unwanted S-meter readings.

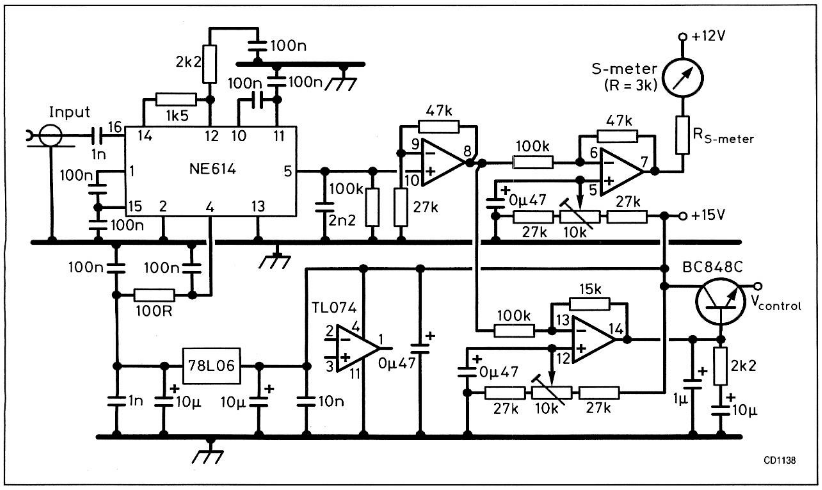

Fig 4.66. Complete circuit for the AGC/ALC module

The circuit is based on the NE614 IC and the signal-strength output is utilised here. This output is buffered by an op-amp and then split into two paths. One path provides the S-meter signal and the other path the regulating signal for the IF amplifier pin-diode limiter.

Calibration of the S-meter is carried out when the entire receive chain is complete and operational. A signal of approximately –73dBm is fed into the RF input. This signal is adjusted until the AGC/ALC output level from the IF amplifier is –29dBm. This then equates to an S-meter reading of 8, and the series resistor $R_{S\text{-meter}}$ should be selected accordingly to give this meter reading

SSB exciter (Fig 4.68)

The SSB exciter consists of an SSB modulator, demodulator, USB/LSB oscillator,

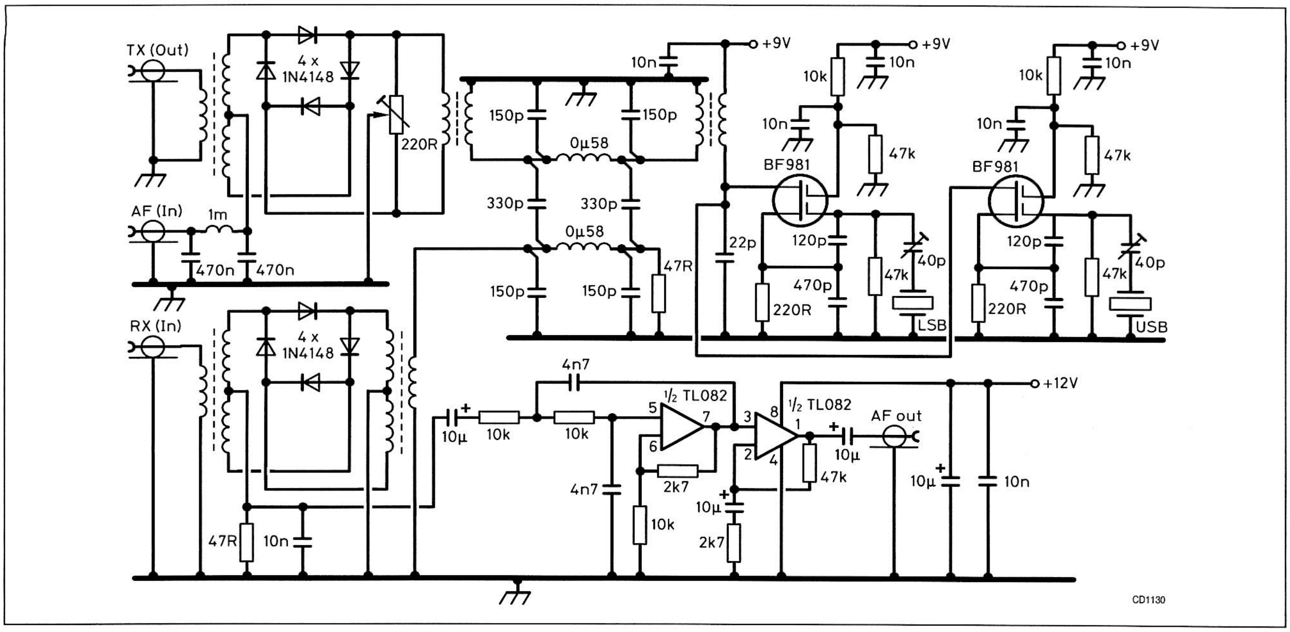

Fig 4.68. The SSB/CW exciter

microphone amplifier and a low-frequency amplifier. The sideband oscillator utilises the dual-gate MOSFET (BF981) in the oscillator stages. The two oscillators (LSB and USB) have common drain circuits and either is selected into operation by the application of the +9V supply for g2. The output from the oscillators is fed to a capacitively coupled hybrid, splitting the output to feed the TX modulator and the RX demodulator.

The amplified microphone signal is fed into the transmit modulator ring mixer via a low-pass filter. A 250Ω trimmer potentiometer is provided to allow for circuit balancing.

The audio output from the receive demodulator is fed to the low-frequency amplifier via a low-pass filter. The LF amplifier is configured using a TL082 operational amplifier.

The PCB layout and component overlay is shown in Fig 4.69 (Appendix 1). The PCB measures 72 × 72mm.

Microphone amplifier (Fig 4.70)

The microphone amplifier is constructed using the two halves of a TL082 op-amp. A standard 600Ω dynamic microphone is required and the output level required by the SSB modulator (approx 1V) is adjusted by means of the 10kΩ potentiometer.

The PCB layout and component overlay is shown in Fig 4.71 (Appendix 1). The PCB measures 34 × 34mm.

AF amplifier (Fig 4.72)

The heart of the low-frequency amplifier is an MC34119 integrated circuit. This IC is available in SMD format in an SO-8 package. The audio output is 200mW into 8Ω.

The PCB layout and component overlay is shown in Fig 4.73 (Appendix 1). The PCB measures 34 × 34mm.

Interconnection of modules

Once all the modules have been assembled and pre-tuned they are ready for interconnection. Various supply voltages are required in the transceiver:

+15V	AGC/ALC
+12V	SSB exciter, microphone amplifier, LF amplifier, receive mixer, transmit mixer, oscillator, VFO amplifier, VFO mixer, S-meter
+9V	SSB exciter
+5V	Frequency display

Not only the various voltages, but also operational considerations play a part in the supplies. For example, the VFO,

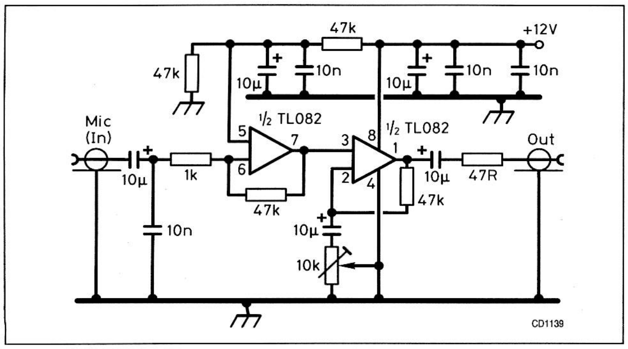

Fig 4.70. The microphone amplifier

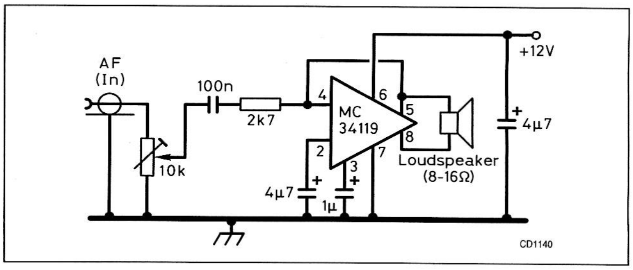

Fig 4.72. The AF amplifier

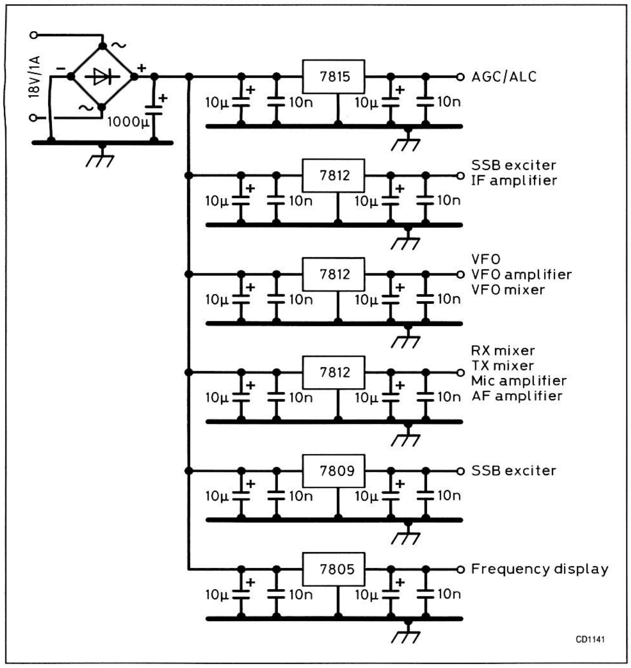

Fig 4.74. Separately stabilised voltages for the individual modules

audio amplifier and the mixer each require to have separate voltage stabilisers. This prevents interaction between the various units through a common supply. A suggested supply regulation circuit is shown in Fig 4.74.

The IF amplifier is used both in transmit mode and receive, thus switching must be provided to reverse the input and output connections. This is most simply achieved by using miniature relays operated by the PTT function. Similarly, the VFO and various supply voltages can be switched by relays. Suggested switching arrangements are shown in Fig 4.75.

Fig 4.76 shows the method for interconnecting the IF amplifier and the AGC/ALC unit.

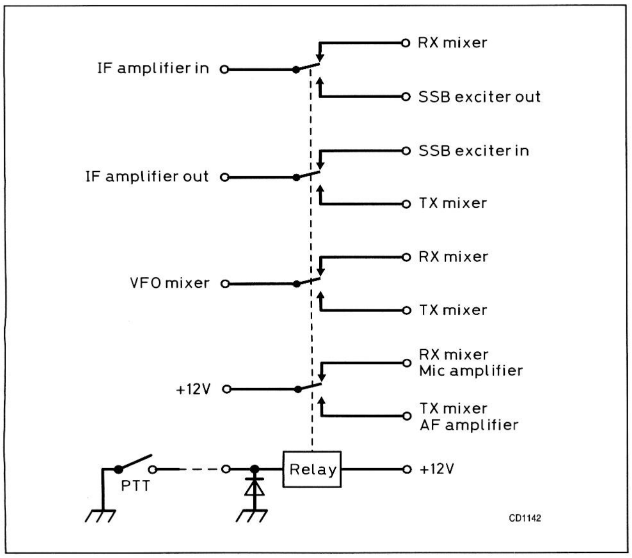

Fig 4.75. Transmit/receive switching

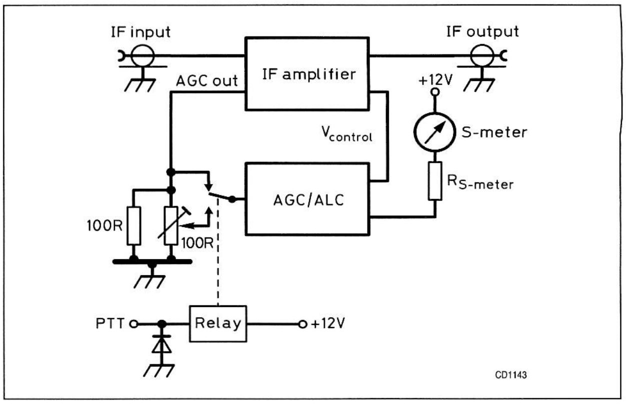

Fig 4.76. Interconnection of the IF amplifier and the AGC/ALC unit

70MHz

Again, due to the limited availability of the 4m band no specific designs are included here. It is conceivable that the 50MHz transceiver described above could be modified for use on 4m without much difficulty.

144MHz

The following 28MHz to 144MHz transverter was also designed by Wolfgang Schneider, DJ8ES [21]. This design incorporates features very important in equipment today, such as high signal strength handling, adjacent signal rejection and good spectral purity, which with increasingly occupied bands are perhaps more important than any other consideration.

The complete circuit of the transverter is shown in Fig 4.77. The design uses a fixed crystal oscillator running at 116MHz which is then amplified by a MMIC (microwave monolithic integrated circuit) device to give an level of 50mW to feed into the SRA1H mixer.

The input level from the 10m transmitter is fed into a pi-damping circuit comprised of R1, R2 and R3. Values for the three resistors (E12 or E24 range) for various input levels are given in Table 4.6.

In order to derive a clean transmit signal (intermodulation products < –50dBc) the mixer must be driven to full output with a maximum of 1mW at its input. The input pi-circuit also gives the mixer a good 50Ω match.

Parallel to the transmit input feed to the mixer is the take-off point for the received signal output. The received signal (144MHz now transformed to 28MHz) is amplified by TR3 (BF981) and fed to the RX output of the transverter.

In receive mode the +12V switched RX voltage, apart from supplying the receive chain, also biases D1 on, thus switching the received signal through to the mixer. The 144MHz

Table 4.6. Values for the three resistors (E12 or E24 range) for various input levels

P_{in} (mW)	P_{in} (dB)	R1(Ω)	R2 (Ω)	R3 (Ω)
1	0	-	0	51
2	3	300	18	300
5	7	120	47	120
10	10	100	68	100
20	13	82	100	82
50	17	68	180	68
100	20	62	240	62

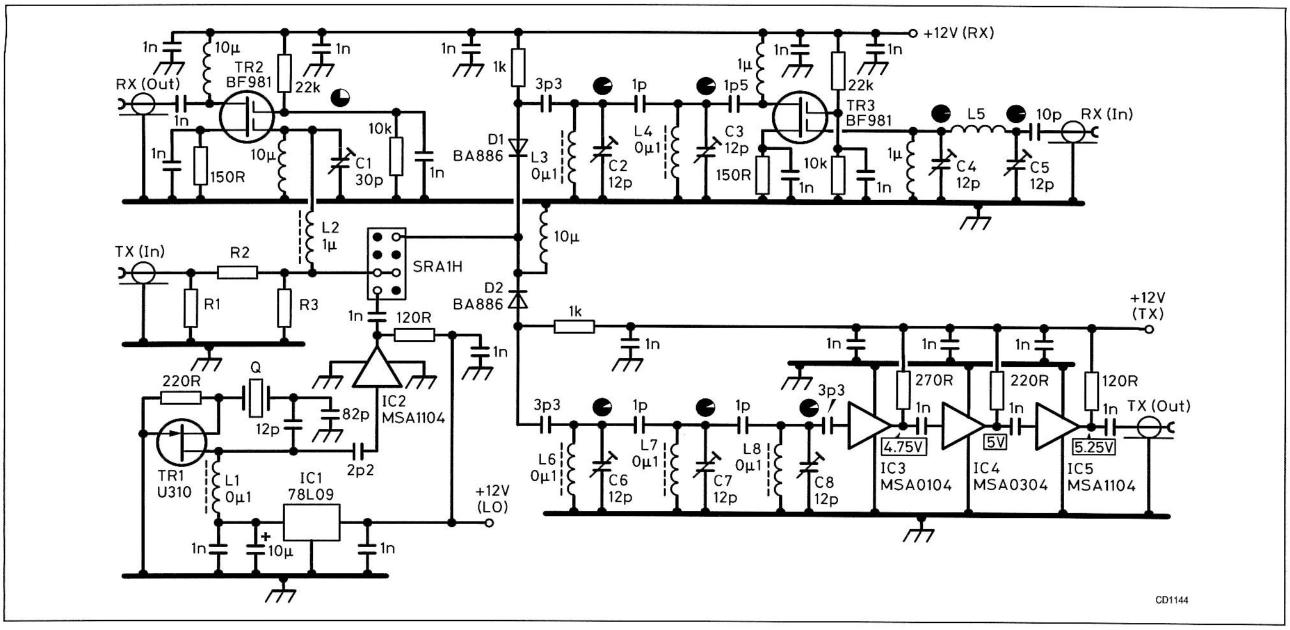

Fig 4.77. Circuit of the 28/144MHz transverter *(VHF Communications)*

signal from the antenna is filtered and then amplified by TR3 (BF981), and filtered again. The receive conversion gain of the unit is of the order of 20dB with a noise figure of the order of 2dB attainable.

In transmit mode the +12V TX supply line biases D2 on and feeds the transmit signal from the mixer to the MMIC amplifier chain via a three-pole filter network.

Note: although this may seem obvious it is not always appreciated – in a transverter such as this, the +12V RX and TX supplies should not be connected to the board simultaneously, but preferably relay switched by the action of the PTT from the controlling transceiver/transmitter.

The output level at the TX out should be of the order of 50mW with any harmonics present –55dBc or better.

Construction and setting up

The PCB layout and component overlays are shown in Fig 4.78 in Appendix 1. The PCB measures 54 × 108mm. Suitably sized holes should be drilled in the PCB in which to locate the MMIC amplifiers and the BF981 transistors, such that the connecting leads are flush with the circuit connections. Use a minimum of solder on all joints. The through connections required for the coils and the mixer are provided by 1.5mm copper rivets.

The unit is aligned as follows. First set all trimmers to the approximate positions as shown in Fig 4.77.

1. Tune the oscillator to 116.000MHz with L1.
2. Inject a 28MHz signal, after setting the values of the input pi-network according to Table 4.6. Adjust C6, C7 and C7 for maximum output (approx 50mW).
3. Using a strong 2m signal (eg a beacon) connected to the RX input adjust C2 and C3 for maximum signal strength on the 28MHz receiver.
4. Adjust C1 for maximum received signal on the 28MHz receiver.
5. Repeat steps 3 and 4 until no further increase can be obtained.
6. Adjust C4, C5 and L5 for best signal-to-noise ratio.

432MHz

Here is another transverter design by Wolfgang Schneider, DJ8ES [22], this time for 432MHz and again using 28MHz as the base frequency. The design of this transverter was based on the same premise and concepts as the 28/144MHz one, and in many ways the design is complementary.

The oscillator (Fig 4.79) is based around a well-known circuit using the U310 transistor. The 101MHz output from the oscillator is quadrupled by TR2 (BFR90a) and amplified by two MMIC stages to the required level of 50mW for the SRA1H ring mixer.

The circuit diagram of the transverter is shown in Fig 4.80. The input level from the 28MHz transmitter is again fed to

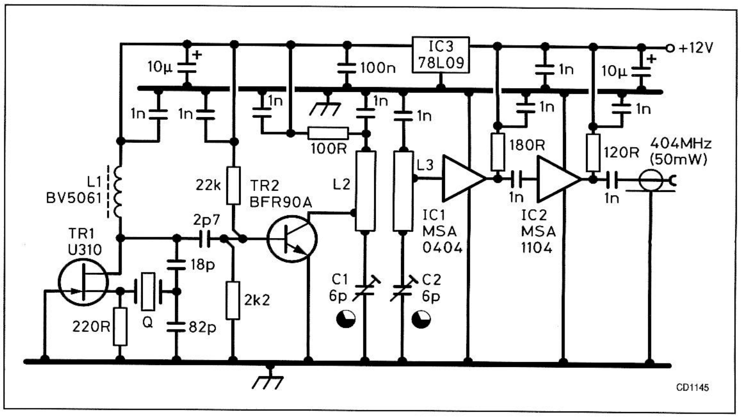

Fig 4.79. Crystal oscillator with quadrupler for 404MHz

the ring mixer via the same pi-damping circuit as used in the 28/144MHz transverter. Consequently, reference to Table 4.6 in the description of that unit will give the required values of the components in the input network, for various 28MHz transmit levels.

The 432MHz receive signal is routed to TR4, a dual-gate MOSFET preamplifier stage (CF300), via a pi-filter. The preamplifier is followed by a MMIC amplifier stage, which amplifies the signal sufficiently enough to overcome the losses in the following PCB etched three-pole filter (L5, L6 and L7). The receive signal from the ring mixer is fed through a filter network to TR3 and the subsequent amplified signal taken to the 28MHz receiver from RX OUT. Transmit/ receive switching is accomplished using PIN diodes D1 and D2, which are switched by the appropriate TX or RX +12V.

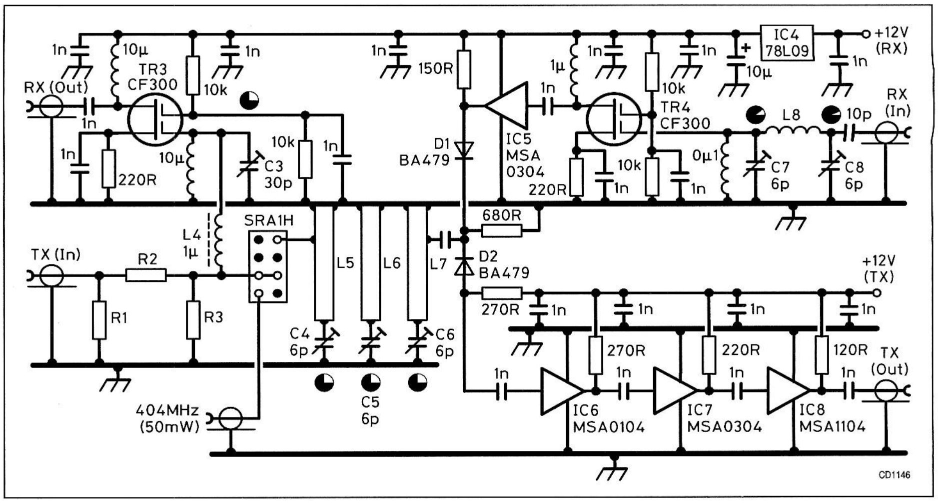

Fig 4.80. The 28/432MHz transverter *(VHF Communications)*

In transmit mode, the output from the ring mixer passes through the three-pole filter and via PIN diode D2 to the transmit MMIC amplifier chain. The 432MHz output level at TX OUT is of the order of 50mW.

Construction and setting up

The transverter is split into two assemblies: the oscillator and the transmit/receive converter. The oscillator PCB and layout is shown in Fig 4.81 in Appendix 1 and measures 54 × 72mm. The transmit/receive converter PCB is shown in Fig 4.82 (Appendix 1) and measures 54 × 108mm. Construction is much the same as for the 28/144MHz unit, with the MMICs and transistors being mounted in holes drilled in the PCBs.

Alignment of the oscillator is as follows. First set all trimmers to the approximate positions as shown in Fig 4.79.

1. Set the oscillator to 101.000MHz using L1.
2. Adjust C1 and C2 for maximum output, which must be a minimum of 50mW.

Alignment of the transmit/receive converter is as follows. First et all trimmers to the approximate positions as shown in Fig 4.80.

1. Having correctly dimensioned the pi input filter, inject a 28MHz signal at TX IN. Adjust C4, C5 and C6 for maximum 70cm output at TX OUT. The level should be at least 50mW, with all spurious signals and harmonics better than −50dBc.
2. Using a strong 70cm receive signal (eg a beacon) adjust C3 for maximum received signal on the 28MHz receiver.

   Note: C4, C5 and C6 should not be adjusted in receive mode. If the received signal level is lower than expected, careful retuning of C4, C5 and C6 should be carried out in transmit mode only.
3. Adjust C7, C8 and L8 for best signal-to-noise performance.

1.3GHz

This 144Hz to 1.3GHz transverter designed by Michael Kuhne, DB6NT [23] is compact (55 × 74 × 30mm), has 1.5W power output, 1.4dB noise figure and 70dB spurious rejection. This is achieved by modern SMD techniques utilising the latest MMICs and power modules, an SMD double-balanced mixer and commercial pre-tuned helical filters. The performance renders it useful for small portable set-ups, as well as for home stations. The circuit diagram is shown in Fig 4.83, a PCB design in Fig 4.84 (Appendix 1) and a component overlay in Fig 4.85.

Design

A 96MHz crystal oscillator works with a J310 junction FET in a source feedback circuit. A quadrupler with a BFR92 feeds a helical filter on 384MHz. A subsequent tripler with a BFG93 and a helical filter on 1152MHz provides a clean 7dBm output on 1152MHz. A double-balanced Schottky mixer SMD-C3 is used for both transmit and receive. Its IF port is terminated in a diplexer and has switched attenuators for RX and TX. These allow for independent adjustment of transmit and receive gain.

In transmit mode the 144MHz IF is mixed with the 1152MHz LO to an output on 1296MHz. The mixer output is filtered by a helical filter and routed through a PIN-diode switch to the transmit chain.

This chain uses a MMIC INA10386, another helical filter and the output power module M67715, which can supply a linear output power of 1.5W. For higher output power a second, external power module with the M67762 can provide 15W at a drive power level of 0.2W.

A control voltage of 12V/2A for external use is provided by the transverter.

The RX chain is comprised of an RF stage with a MGF1302, a helical filter and a second RF stage with a MAR6 MMIC. The output is switched to the common mixer circuit by means of a PIN diode switch. The noise matching in the first stage is made by series-L matching. The two helical filters provide optimum RF selectivity for image and spurious rejection. The overall noise figure is less than 2dB and typically 1.4dB.

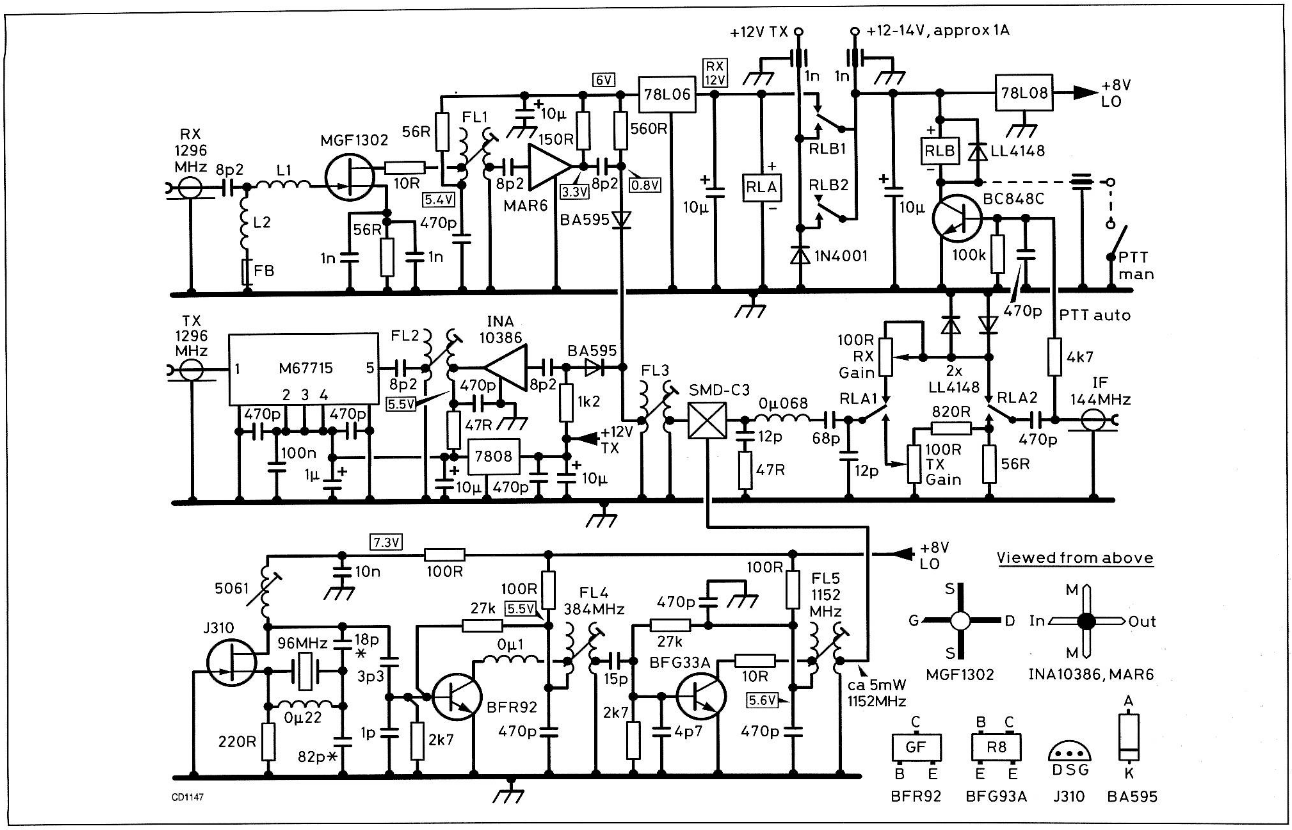

Fig 4.83. The 1296MHz transverter circuit diagram. Components marked with an asterisk (*) are temperature coefficient N750 violet (to suit quartz crystal). L1 and L2 are wound with 5t and 4t respectively of 0.22mm enam wire and have an inside diameter of 1mm. FL1–FL5 are helical filters. FL1–FL3: 367MN113F, FL4: 5HW36535A-385, FL5: 367MN110A. An MSA0685 can be substituted for the MAR6 MMIC *(DUBUS Technik IV)*

Construction

1. Cut PTFE board to dimension of the cabinet.
2. Drill PCB (0.8mm and 2 × 5.8mm for the two 1nF leadless disk caps).
3. Mount coaxial connectors.
4. Put PCB onto inner conductors of coaxial connectors and solder to the cabinet on both sides.
5. Solder feedthroughs.
6. Mount and solder all parts according to the pans layout except filters, MMICs, MGF1302 and SMD-C3.
7. The 1nF disk caps are soldered to the ground plane with small pieces of copper-foil.
8. Remove free leg from Neosid 5061 and also the ground lugs of the enclosure.
9. Mount all filters. Fully solder the enclosures to the ground plane.
10. Mount pots for TX GAIN (Cermet) and RX GAIN (SMD).
11. Solder 7808 to tin-plate enclosure.
12. Solder M67715 approximately 1mm above the PCB.
13. Mount the external cooler with heatsink and some thermal conductive compound.
14. L1 and L2 are made from 0.22mm copper enamelled wire wound closely on a 1mm form. Mount them with short legs.
15. Mount MGF1302 and solder L1 onto the free gate leg (bent through 90°).
16. Make contact troughs for the ground lugs of the MMICs.
17. Mount MMICs.
18. Mount BFR92 and BF093 'overhead'.
19. Solder 56Ω load resistor
20. Mixer C3 can be soldered after tuning of LO chain.

LO adjustment

Apply supply voltage and adjust core of oscillator coil 5061 until voltage at collector of quadrupler BFR92 drops to a minimum value. The 384MHz helical filter is tuned for minimum collector voltage at tripler transistor BFG93. Connect power meter via 50Ω coaxial cable to LO port (mixer SMD-C3 is not fitted yet) and adjust 1152MHz helical filter to maximum output (7dBm min). With the aid of a frequency counter the 96MHz crystal oscillator can be set to the exact LO frequency. Negative temperature coefficient (N750) components for the 10pF/82pF capacitors in the crystal oscillator circuit compensate for low-frequency drift, but may be changed in temperature coefficient in order to accommodate different types of crystals. An external TCXO or OCXO can be connected via a 47pF capacitor to the source of the J310. In this case the crystal and the 0.22µH choke must be omitted.

After finishing the tuning procedure mixer SMD-C3 should be fitted (see step 20).

RX adjustment

Adjust pot RX GAIN to minimum value. Connect antenna or 50Ω dummy load to RX input and a 144MHz RX to the IF

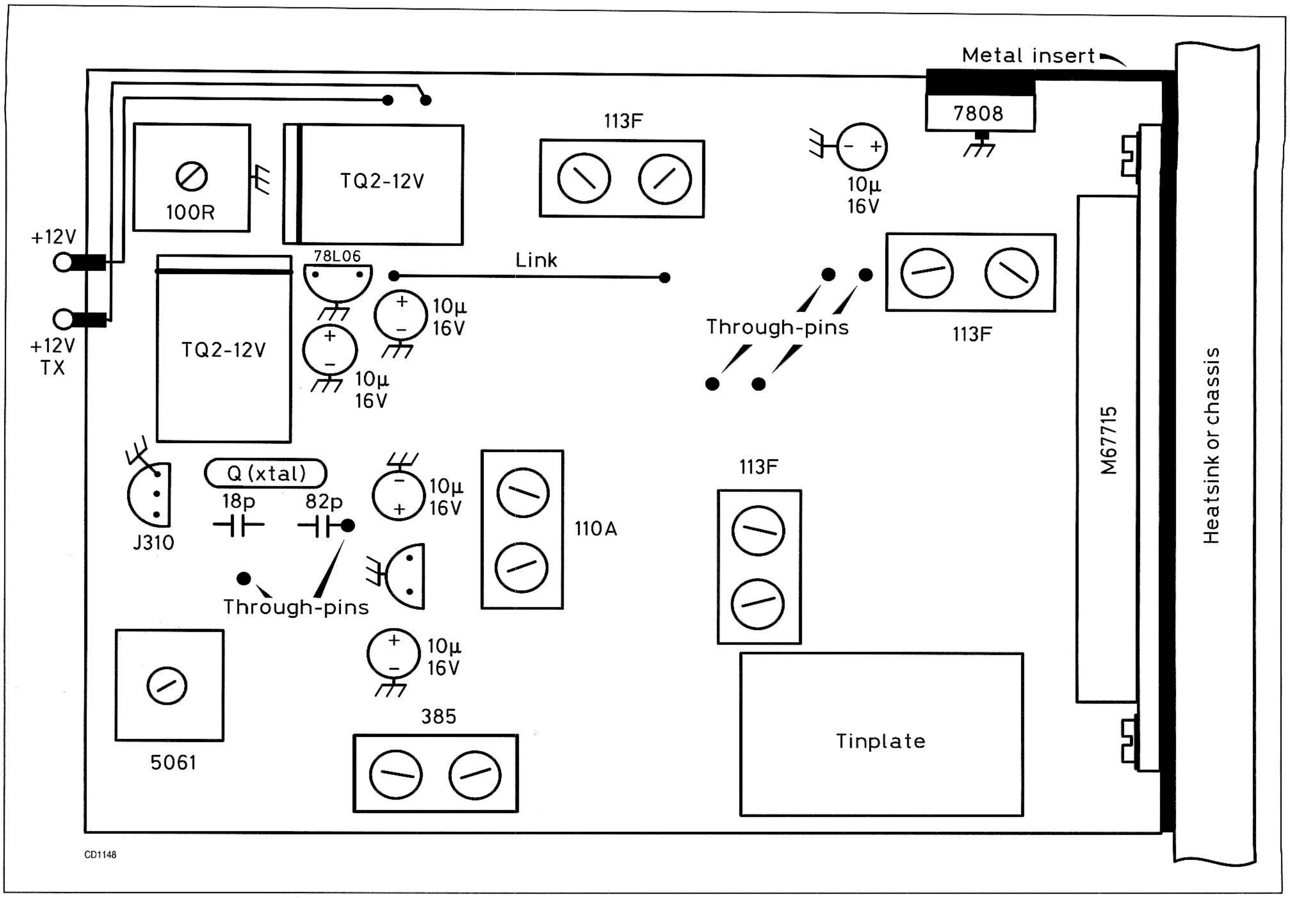

Fig 4.85. Component layout *(DUBUS Technik IV)*

output. Check operating voltages on the MGF1302 and the MAR6. Adjust helical filters in the RX chain to maximum noise level on 144MHz. Tweaking of the helical filters can be done with a signal generator or by listening to a beacon.

Normally tuning of L1 is not required because of its low *Q* characteristic, but if a noise figure meter is available, tuning can he done by squeezing coil L1 with a plastic tool.

Because maximum gain is quite large, an adjustment to the input characteristic of the 144MHz RX can be performed by adjusting the RX GAIN pot for an appropriate S-meter reading.

TX adjustment

Transmit/receive switching is initiated by a DC voltage on the IF line. Alternatively, an external facility could be fitted (see circuit diagram). Connect a power meter to TX output and switch to transmit. Check voltages on INA10386 and M67715. Key 144MHz TX and adjust TX GAIN pot for some readable output power. Adjust helical filters for maximum output power. Finally, the TX GAIN pot can be adjusted to a nominal output of 1.5W.

Measurement results

Power output = 1.5W
Spurious < –70dBc
Image rejection > 70dB
Harmonics <–40dB
RX gain = 16–18dB
Noise figure = 1.2–1.4dB

TRANSMITTERS

There are two main types of transmitter, one having a constant, if interrupted, output such as is used for Morse, data or FM, and the other for modes where the output must follow the input exactly such as for SSB. The latter are called *linear* circuits. A transmitter consists of several stages, some of which are common to receivers. For example, low-power amplifiers, mixers and some forms of modulator are common.

Fig 4.86 shows a block diagram of a transmitter suitable for 'on-off' modulation, ie for Morse code or data transmission. Fig 4.87 shows one for a FM transmitter and Fig 4.88 for a SSB system of the 'filter' type. Fig 4.89 shows a similar block diagram for a FM transceiver. A SSB transceiver is similar except that frequency translation is done by mixing rather than by multiplication.

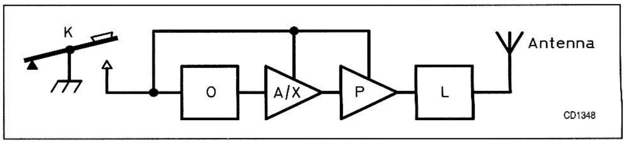

Fig 4.86. Morse code transmitter. ***Key:*** **K – key; O – oscillator; A/X – amplifier/frequency multiplier; P – power amplifier; L – low-pass filter**

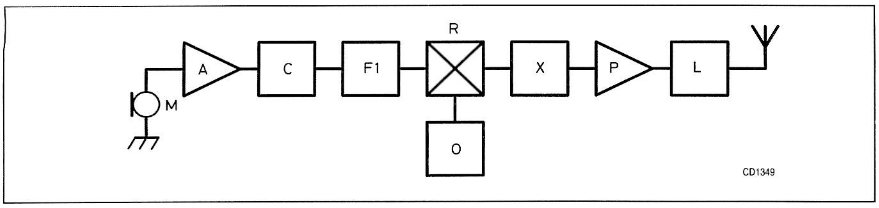

Fig 4.87. FM transmitter. *Key:* M – microphone; A – audio amplifier; C – clipper; F1 – low-pass AF filter; R – reactance modulator; O – oscillator; X – frequency multiplier; P – power amplifier; L – low-pass RF filter

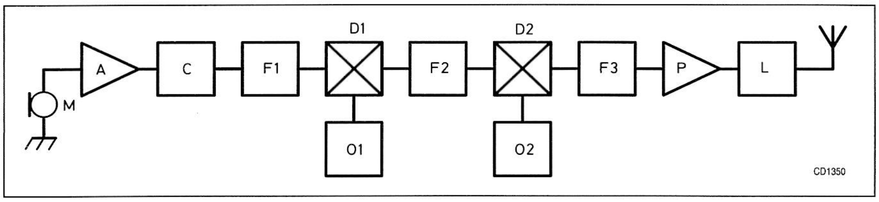

Fig 4.88. Filter-type SSB transmitter. *Key:* M – microphone; A – audio amplifier; C – clipper; F1 – low-pass audio filter; D1 – double-balanced modulator; F2 – sideband filter; O1 – fixed oscillator; D2 – double-balanced modulator; O2 – variable oscillator; F3 – filter; P – power amplifiier; L – low-pass filter

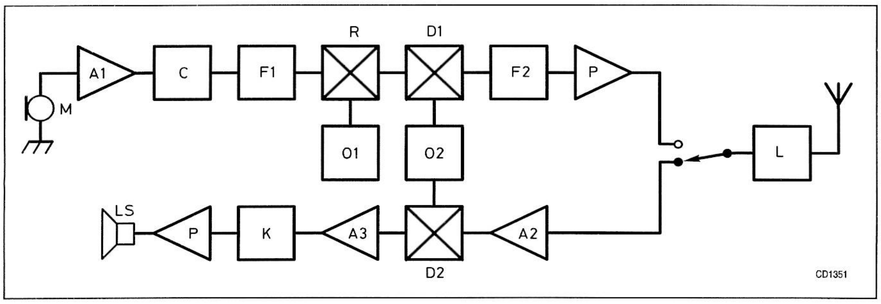

Fig 4.89. FM transceiver. FM is generated at O1 frequency which is equal to the IF. *Key:* M – microphone; A1 – audio amplifier; C – clipper or speech processor; F1 – low-pass audio filter; R – reactance or phase modulator; O1 – fixed frequency oscillator; D1 – double-balanced mixer; O2 – variable-frequency oscillator; F2 – RF filter; P – power amplifier; L – RF low-pass filter; A2 – receiver RF amplifier; D2 – double-balanced mixer; A3 – IF amplifier; K – FM demodulator; A4 – audio amplifier; LS – loudspeaker

The following paragraphs are a short introduction to the various blocks in the diagram. Note, however, that blocks which are common to receivers are dealt with in that section.

RF amplifiers

These follow exactly those in the receiver section – for further details see p4.21.

Frequency multipliers

These are RF amplifiers in which deliberate distortion of the waveform is carried out so that the output contains harmonics of the input. This is done by using very low bias in a bipolar transistor stage or high bias in a FET stage. In either case, the output is tuned to the desired harmonic. Single devices tend to produce better odd harmonic output while even harmonics can be better produced using a 'push-push' circuit in which the input is 'push-pull' and the output is parallel.

Fig 4.90 shows a circuit for an odd harmonic multiplier and Fig 4.91 one for even harmonics. Many devices will work in these circuits; those given are known to work but are not specifically recommended.

Interstage coupling

The coupling of signals between stages is important for one or two reasons:

1. It provides matching between the output of one stage to the input of the next and
2. It may provide selectivity, ie it may contain a tuned circuit. The second is not always the case since there are wide-band amplifiers which, for example, cover most of the HF part of the spectrum.

In general, in VHF and UHF transmitters, we are dealing with both reasons and, in the case of selectivity, this must be either low enough to cover the whole band, or tuneable with a front-panel control, or possibly cover only the section of the band of interest to the operator.

With tuned circuits, it is often possible to provide a low impedance output by means of a tapping or a link winding of relatively few turns. Fig 4.92 shows a typical interstage coupling between the collector of one transistor and the base of

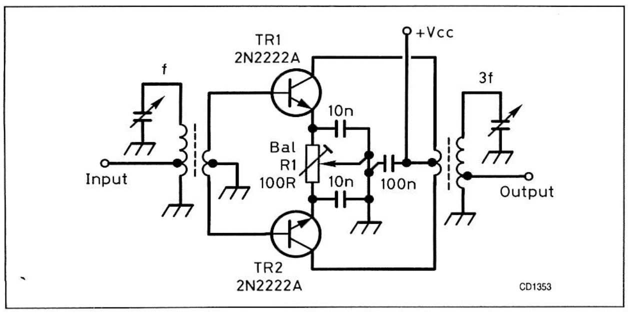

Fig 4.90. Odd-harmonic multiplier *(ARRL Handbook)*

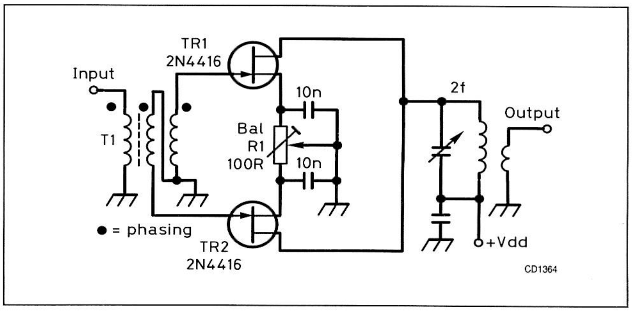

Fig 4.91. Even-harmonic multiplier *(ARRL Handbook)*

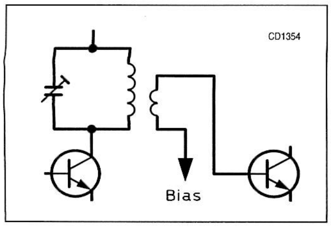

Fig 4.92. Tuned coupling between transistor amplifiers

the next. In general, FETs are easier to deal with because their input impedance is much higher than bipolar transistors. They do, however, have a large input capacitance which varies with bias and signal. Fig 4.93 shows a FET interstage coupling of the broad-band type.

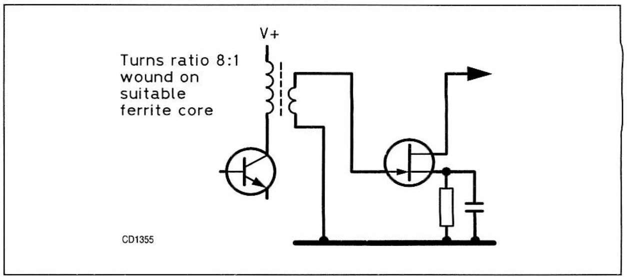

Fig 4.93. Wide-band coupling between transistor amplifiers

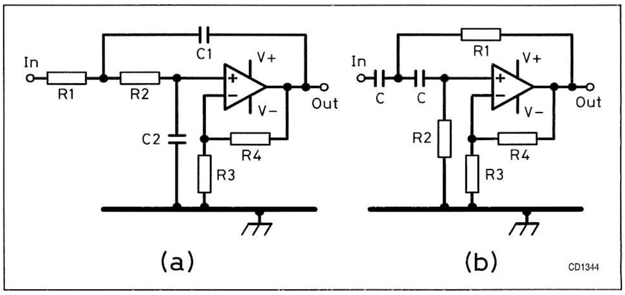

Fig 4.94. Low-pass (a) and high-pass (b) Sallen and Key type audio filters. V+ and V− are symmetrical about the earth line. Both filters have a Chebyshev response and a gain of 2

Broad-band couplings are useable up to a point. At high levels, which may only mean 1W or so, by-product signals can be formed which can, in non-linear circuits, form unwanted signals, some of which may end up in the band. It is therefore necessary to put in some selectivity to attenuate all but the wanted signals.

Spurious outputs

In all cases the ideal transmitter will have frequency stability and a 'clean' output, ie with no other RF signals present than those wanted. This means no unwanted outputs such as harmonics or other spurious frequencies. It will also have means for controlling the power output so that minimum power necessary for communication can be used.

These ideals are not attainable in practice so it is generally recognised that, if the spurious outputs are lower than −60dBc (decibels with respect to the carrier power) or −30dBm (with respect to 1 milliwatt), whichever is the lower, then the output is clean enough. The VHF Contest Committee of the RSGB recommended a lower level of −90dBc but this has proved too difficult to achieve.

Power levels

The transmitter starts with low-power stages and only in the last stages does the power increase significantly. These high-power stages are dealt with later. Arbitrarily, a RF power of 1W is considered the boundary. Below this, many of the circuits are identical to those used in receivers although there are functions performed in a transmitter which do not occur in receivers.

Audio inputs

The average normal microphone produces a few millivolts (mV) and the modulator (see below) needs of the order of volts so a gain of about 1000 is needed. It is also necessary to 'tailor' the frequency response of the microphone to about 300–2500 Hz to avoid broadening the bandwidth requirement for SSB. Both operations can be carried out using low-noise op-amps. A quad op-amp can provide all the necessary functions. The first stage provides most of the gain, the second stage is a high-pass filter with a cut-off of 300 Hz and the third stage is a low-pass filter with a cut-off of 2.5kHz. The filters are of the Sallen and Key type (Fig 4.94) using selective feedback around the op-amps and have a Chebyshev response. Fig 4.95 shows the practical circuit.

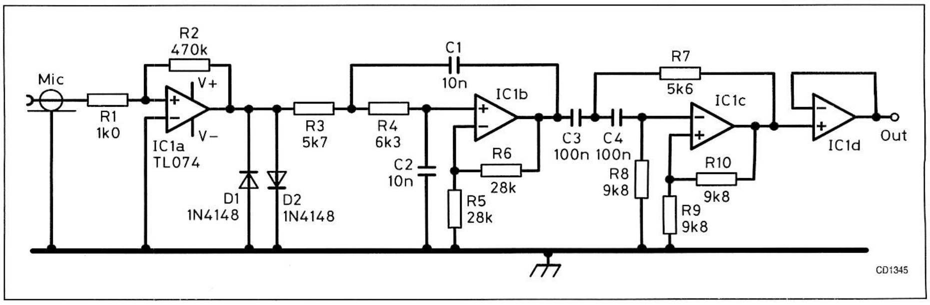

Fig 4.95. Practical filter circuit. 'Odd' values of resistors should be made up from series or parallel combinations. C1, C2 and C3, C4 should be matched if possible

In order to control the maximum output, it may be useful to put a simple clipper between the amplifier (IC1a) and the filters. In this way some of the distortion products made by the clipper are removed.

Modulation

On-off modulation is used for Morse code and for data modes, and the only essential is that the on-off transitions are gentle. If they are sharp, a large number of sidebands will be produced and these appear as *key clicks*. The aim should be to have a waveform which is rounded (see Fig 4.96). This is done by arranging the rise and fall times to be 0.1 of the time of the bit. In the case of Morse, the bit is the single dot and in digimode, the bit is the 'mark' signal.

Frequency modulation (FM)

There are two methods of achieving this: the direct method and the indirect or phase modulation (PM) method. The former

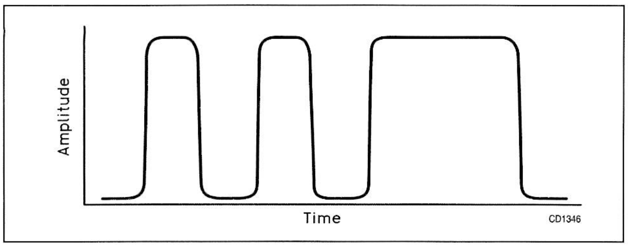

Fig 4.96. The letter 'U', showing gentle transitions from mark to space and vice versa

consists of a reactance modulation of an oscillator which, at its simplest, can be achieved by applying the audio signal to a suitable biased varicap diode (Fig 4.97) and the latter by attaching the same sort of varicap to one of the tuned circuits *after* the oscillator to alter the phase with the audio signal (Fig 4.98).

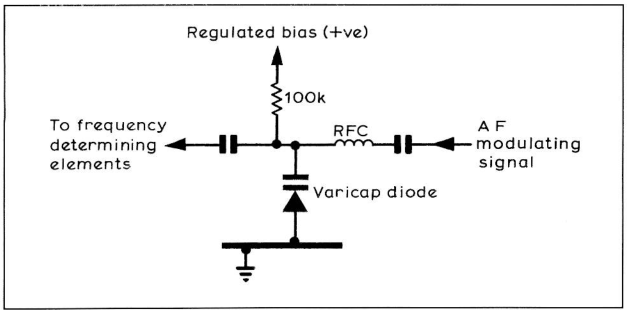

Fig 4.97. Basic varicap diode modulator

The two waveforms are shown in Fig 4.99 together with the audio signal (top). The FM is shown in the centre and the PM at the bottom.

It will be seen that there is little difference between the FM and the PM waveform. However, in PM the effective FM is proportional to the modulating frequency so to get true FM, in the PM system, the audio must be tailored to have an amplitude that falls with increasing frequency.

By whatever means FM is achieved, the signal is described mathematically by the Bessel function. It is not necessary to understand this, only to apply it. Table 4.7 and Fig 4.100 show the Bessel function as a function of modulation index (MI) which is defined as:

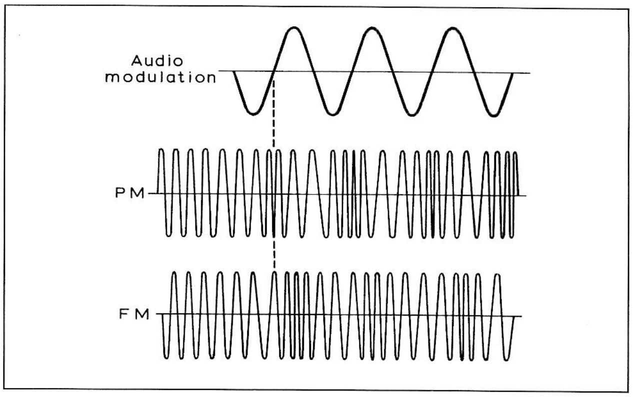

Fig 4.99. Angular modulation of an RF carrier

$$\text{MI} = \frac{\text{Deviation of the FM carrier}}{\text{Audio frequency producing that deviation}}$$

Fig 4.100 is easiest to understand as it shows the amplitude of the carrier and the various sidebands for various values of MI. It will be seen that the carrier *decreases* as MI increases and at a value of MI of about 2.4 (actually 2.405), the carrier disappears. When it reappears at higher MIs, it is negative which only means its has its phase shifted by 180°. Fig 4.101 shows the amplitude and distribution of sidebands at various MIs.

Maximum deviation needs to be set to suit the band and the system. At present, repeaters will not accept deviations greater than 5kHz and this is also used for simplex working at the present channel spacing of 25kHz. When the channel spacing is reduced to 12.5kHz, the maximum deviation must also be reduced to 2.5kHz.

The fact that the carrier disappears at a MI of 2.4 gives us a way of setting up an FM transmitter (or just an FM generator) to a specific deviation using an audio generator and a sharply tuned HF receiver. This assumes that the FM is generated on a carrier within the range of the HF receiver and multiplied up to the band in question. As an example, take a final frequency of 145MHz which is generated at 12MHz and multiplied by 12. If the final deviation is 5MHz, the deviation at 12MHz is 417Hz. If this is caused by an audio tone of 174Hz the carrier will disappear at a deviation of 417Hz (ie at a MI of 2.4). This disappearance is monitored on the HF receiver.

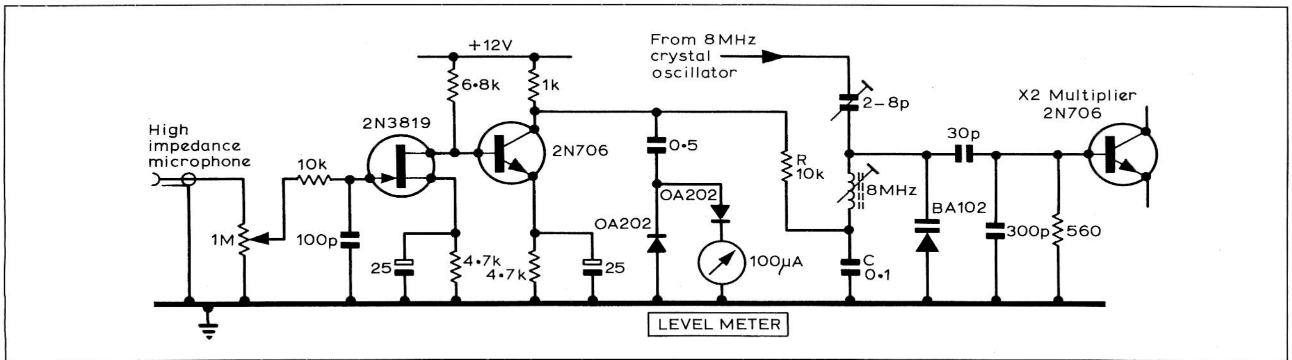

Fig 4.98. Practical circuit using a varicap diode

Table 4.7. Bessel function

Modulation index	Carrier value	1st set of side-bands	2nd set	3rd set	4th set	5th set	6th set	7th set	8th set	9th set	10th set	11th set	12th set	13th set	14th set
0.00	1.000	–	–	–	–	–	–	–	–	–	–	–	–	–	–
0.01	1.000	0.005	–	–	–	–	–	–	–	–	–	–	–	–	–
0.05	0.9994	0.025	–	–	–	–	–	–	–	–	–	–	–	–	–
0.02	0.9900	0.0995	–	–	–	–	–	–	–	–	–	–	–	–	–
1.00	0.7652	0.4401	0.1149	0.0020	–	–	–	–	–	–	–	–	–	–	–
2.00	0.2239	0.5767	0.3528	0.1289	0.0341	–	–	–	–	–	–	–	–	–	–
4.00	−0.3971	−0.0661	0.3641	0.4302	0.2811	0.1321	0.0491	0.0152	–	–	–	–	–	–	–
5.00	−0.1776	−0.3276	0.0466	0.3648	0.3912	0.2611	0.1310	0.0534	0.0184	–	–	–	–	–	–
7.00	0.3001	−0.0047	−0.3014	−0.1676	0.1578	0.3479	0.3392	0.2336	0.1280	0.0589	0.2035	–	–	–	–
10.00	−0.2459	0.0435	0.2546	0.0584	−0.2196	−0.2341	−0.0145	0.2167	0.3179	0.2919	0.2075	0.1231	0.0634	0.0290	0.0120

A negative sign indicates that the component is 180° out of phase with respect to the others. Blank spaces indicate that the values of the sidebands are insignificant

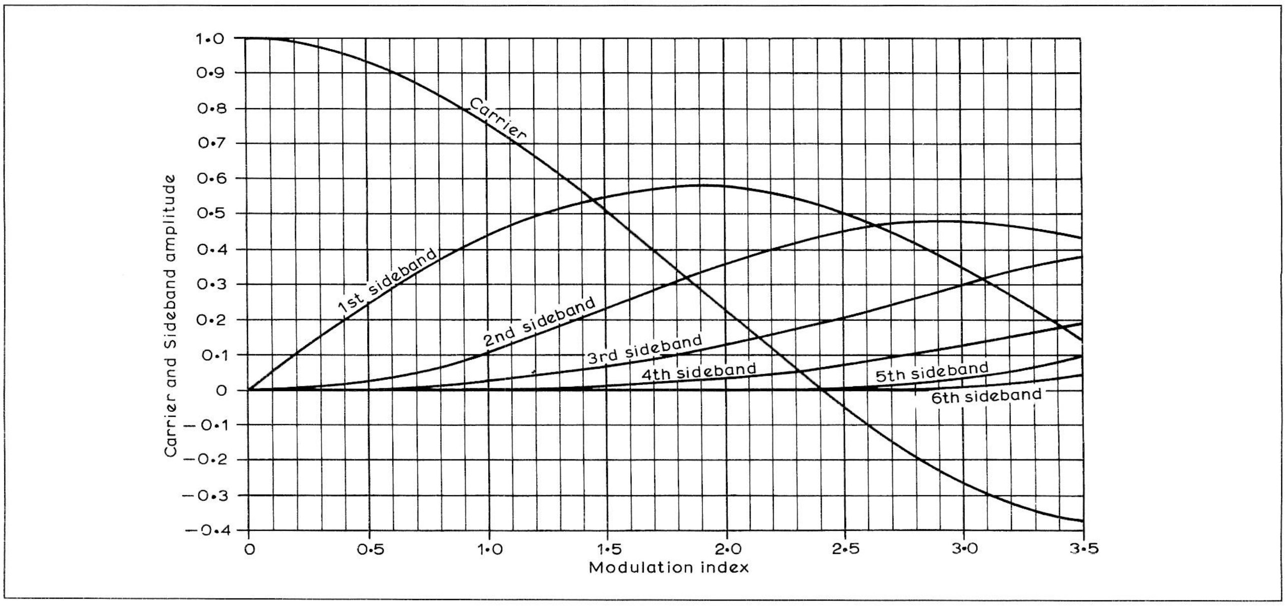

Fig 4.100. Bessel curves showing variation in carrier and sideband amplitude with modulation index

SSB

Fig 4.102 shows the relationship between the audio modulating frequency, the carrier frequency and the AM, DSBSC and SSB products. There are three methods for generating SSB.

The filter method

This is where a carrier is generated and fed to a balanced modulator (DBM, see above, the balanced mixer has exactly the same characteristics) which produces two sidebands with very little carrier (a perfect DBM produces zero carrier), ie a DBSC (double sideband suppressed carrier) signal. A sharp filter removes one of the sidebands and what is left of the carrier. This method is that used in the majority of commercial transceivers.

The mechanism can be expressed mathematically as follows. We assume that a simple tone is modulating a carrier:

$$\cos A \times \cos B = \tfrac{1}{2}\cos(A + B) + \tfrac{1}{2}\cos(A - B)$$

where A is the carrier frequency and B is that of the tone (cos or cosine functions are used because the mathematics are simpler than sine functions and the difference is only a 90° phase shift.) $A + B$ is the upper sideband and $A - B$ the lower. If B is biased so that A never goes negative, the carrier is not suppressed and the result is AM (amplitude modulation) – see Fig 4.102.

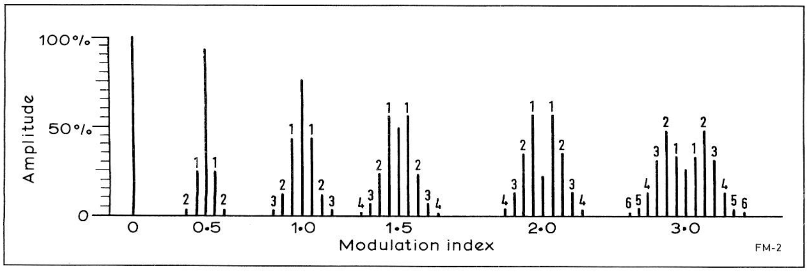

Fig 4.101. Relative carrier and sideband levels for various modulation indices

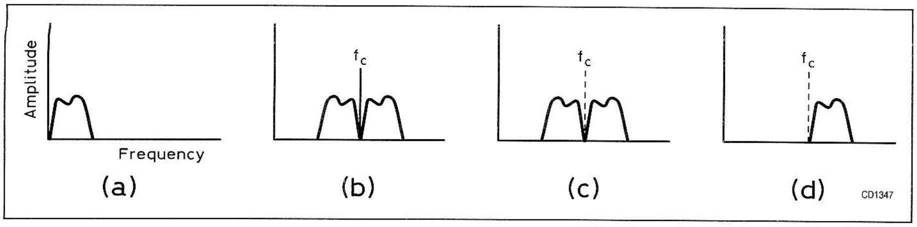

Fig 4.102. (a) The audio signal. (b) AM signal derived from it, f_c being the carrier frequency. (c) DSBSC signal, f_c being where the carrier would be. (d) SSB signal (upper sideband), f_c being where the carrier would be

The phasing method

In this method [24–26] the audio is used to generate two signals which are mutually at 90° phase difference to one another. These are used to modulate two carriers which are also at 90° phase difference and the resulting signals are combined. A SSB signal results and this can be shown mathematically.

If the two carrier signals are A and $A + 90$ and the two audio signals are B and $B + 90$, then:

$$\cos A \times \cos B = \tfrac{1}{2} \cos (A + B) + \tfrac{1}{2} \cos (A - B) \qquad \text{(i)}$$

and

$$\cos (A + 90) \times \cos (B + 90) = \tfrac{1}{2} \cos (A + B + 180) + \tfrac{1}{2} \cos (A - B) \qquad \text{(ii)}$$

now

$$\cos (A + B + 180) = - \cos (A + B)$$

so by adding (i) and (ii) we get $\cos (A - B)$ and by subtracting (ii) from (i),we get $\cos (A + B)$ which are the lower and upper sidebands.

Analogue methods can be used to produce the phase shifts in the audio signal and these are quite complex if an accurate 90° phase shift necessary for good carrier and sideband suppression is to be obtained.

The Third Method

This is a different phasing method [27] which does not need two audio channels accurately at 90° phase to one another. The audio signal is passed to two DBMs which are fed in quadrature (90° phase difference) by a pilot tone at, say, 1.7kHz. Each produces a DSBSC signal from which the upper sideband is stripped by two low-pass audio filters. The resulting signals are passed to two further DBMs fed in quadrature by the HF oscillator. Finally, the four sets of sidebands are combined whern some reinforce and the others cancel, resulting in a SSB signal. The mathematics of the process are rather complex. The originator of the Third Method was D K Weaver [28] who was not a radio amateur.

All SSB signals are generated at HF and transformed to VHF or UHF by mixing with a suitable frequency and filtering out the unwanted sideband. Frequency multiplication cannot be used because this would multiply the audio frequencies.

It should also be noted that while all the information is transmitted on each sideband, it is not possible to regenerate the audio signal *exactly*. The normal technique is to adjust the added carrier until speech sounds 'normal', ie not like Donald Duck! For this reason, SSB cannot be used to transmit music unless there is some method for reintroducing the carrier exactly, eg by transmitting a very much reduced or *pilot* carrier to which the receiver carrier can be synchronised.

SEMICONDUCTOR POWER AMPLIFIERS

Lower-power transistors are readily available from specialist suppliers and so can be used in designs intended for copying. Higher-power devices tend to be very expensive, the selection available to the amateur buyer is very variable, and it can prove difficult to duplicate some designs. Nevertheless, such designs are valuable in demonstrating the specialised design and construction techniques.

Cooling

Proper cooling is vital for high-power transistors as lot of heat has to be conducted through a small area with a low temperature rise. Philips' recommendation [29] for the heatsink is for a surface flatness of 0.02mm (0.001in) and a surface roughness of 0.5µm (almost mirror finish). These specifications are difficult to achieve without machining and probably unnecessary in most amateur applications, but most extruded heatsink shows a far from ideal surface. Sanding with fine wet/dry paper can make a worthwhile improvement in the flatness at the expense of roughness. It is worth asking local machine shops for a price for skimming a heatsink; some charge very reasonable prices. Figs 4.103 and 4.104 show enlarged views of the transistor/heatsink joint; if the surfaces are not flat there are few points of contact for heat flow. The heatsink is not the only source of imperfection; most transistor flanges are quite soft and can become distorted. If necessary, they can be sanded against fine wet/dry paper which is placed face upwards on a flat surface.

White heatsink compound should be used in preference to clear silicone grease and it is intended to fill the micro crevices in Fig 4.104, rather than the large gaps. Heatsink compound is a hundred times worse than metal at conducting heat (but slightly better than nothing at all), hence the emphasis on flatness and roughness to get maximum metal-to-metal contact at a microscopic level [30]. Before fitting the transistor, deburr the fixing holes well and clean all swarf, dust from sanding and solder/flux splashes from both heatsink and transistor using methylated spirits or similar; any debris whatsoever under the transistor will prevent the important metal-metal contact. Using a thin layer of thermal compound, just enough to obscure the metal, press the transistor into

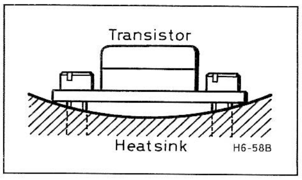

Fig 4.103. View of transistor on concave heatsink, showing gap

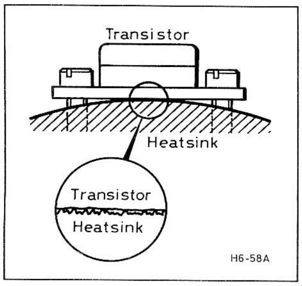

Fig 4.104. View of transistor on convex heatsink, showing gaps. Magnified view shows micro crevices in the transistor-heatsink interface

place as hard as possible with a finger, then remove it when any voids in contact will be obvious. It should be possible to achieve contact over the whole flange with this amount of thermal compound; if this is not possible, then either the transistor or heatsink is not flat enough. Be careful: thermal compound spreads a long way and does not wash out of clothes easily. Methylated spirits will remove and disperse it prior to washing.

Make sure that the fixing screws do not foul against the main body of the transistor or the BeO (beryllium oxide, or 'beryllia') between the flange and the leads might be damaged; particles of BeO are very toxic if inhaled. The top cap is made from alumina which is razor sharp if broken but non-toxic. Most transistors are designed to be used with American 4-40 UNC screws, but M2.5 or M3 can be used as an alternative. M2.5 is probably better as there is greater clearance for the head and less likelihood of overtightening (which actually reduces the thermal contact to the heatsink). Screws should be tightened to 0.7Nm, which is tight but well short of needing great effort. Always use a plain washer between the screw and the flange.

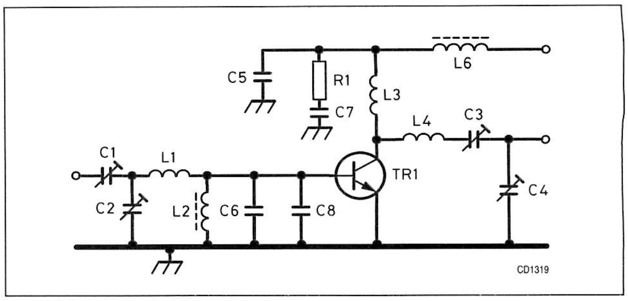

Fig 4.105(a). Typical VHF amplifier circuit

If the transistor is correctly fitted to the heatsink, a heatsink temperature of 70°C is a reasonable upper limit to apply. The

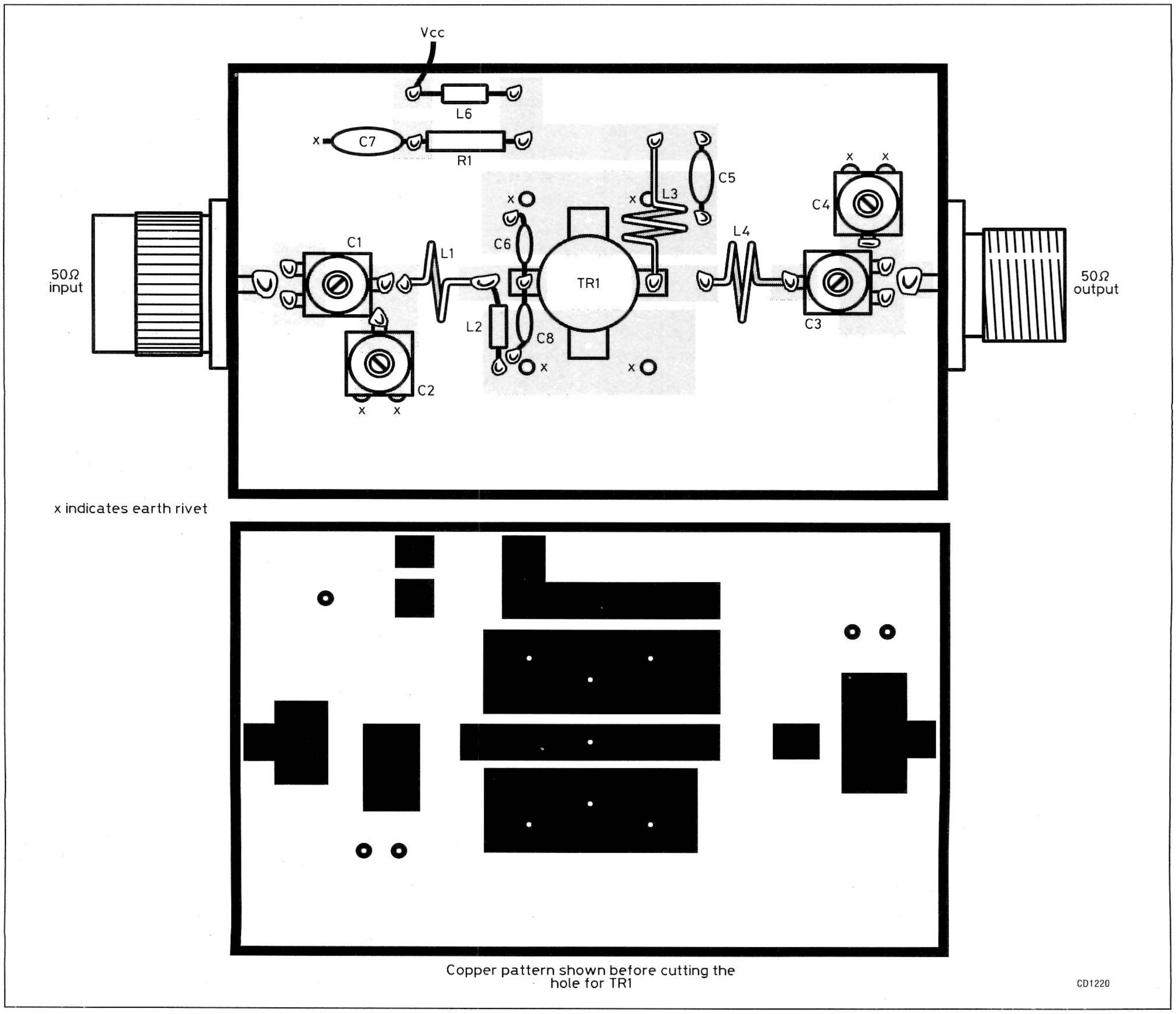

Fig 4.105(b). Component layout and component side copper (reverse side is continuous copper)

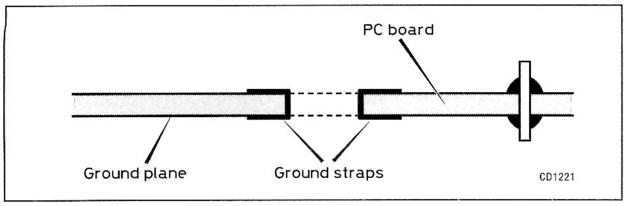

Fig 4.105(c). Earthing straps and wires

transistor will typically be operating with a junction temperature of around 120°C which will give very good reliability, and many of the other components in the circuit will have a temperature rating of 70 or 85°C. Remember that 70°C is hot enough to cause skin burns; using a fan or larger heatsink to reduce the heatsink temperature so that it is not uncomfortable to hold will provide extremely good reliability, provided that individual components are not over stressed.

It is also important to ensure correct mechanical alignment between the RF power transistor and the PCB. The transistor leads should ideally be level with the PCB; slight downwards tilt is acceptable, but avoid any upwards bend as this can result in the leads becoming detached or the top cap being dislodged. The PCB will normally need to be spaced away from the heatsink using, say, a stack of washers. For most flange and stud mounting transistors, a 4BA half nut (2.5mm thick) is about correct if the PCB is the most common 1.6mm thickness. The nut has a centre hole which gives clearance for a M3 screw.

Earthing and construction

An important aspect of all transistor amplifiers is earthing. The low impedances result in high RF currents and solid, low inductance earthing is key in ensuring circuits operate as expected. Normally, circuits will be built on double-sided PCB with the underside a continuous copper ground plane. Fig 4.105(a)–(c) and 4.106 show a typical amplifier circuit and a suitable way to construct it, including how the emitter leads should be connected to the ground plane with straps of copper or brass foil. Additionally, any areas of earth connection on the top surface should have pins or wires connecting the top and bottom together at regular intervals (20mm or less spacing).

Stability

VHF/UHF power transistors have increasing gain at lower frequencies which can easily bring about destructive oscillation. Many designs show no signs of instability until an 'equivalent' transistor, or even a different batch of the original, is tried. A precaution against this is to ensure that the transistor sees a resistive load at frequencies much below the operating frequency. In Fig 4.107 the RFC and RF decoupling are transparent at low frequencies and the transistor is loaded by the resistors through the LF decoupling to ground. The LF choke provides the DC path for the supply current. In the base circuit of a Class C amplifier, point X is grounded and C4/5 are omitted. Another technique which improves LF stability is resistive feedback, shown in Fig 4.108. The RF choke isolates the resistor at the operating frequency and the capacitor provides DC blocking.

Bipolar transistors in particular can also suffer from parametric instability. The collector base capacitance varies with

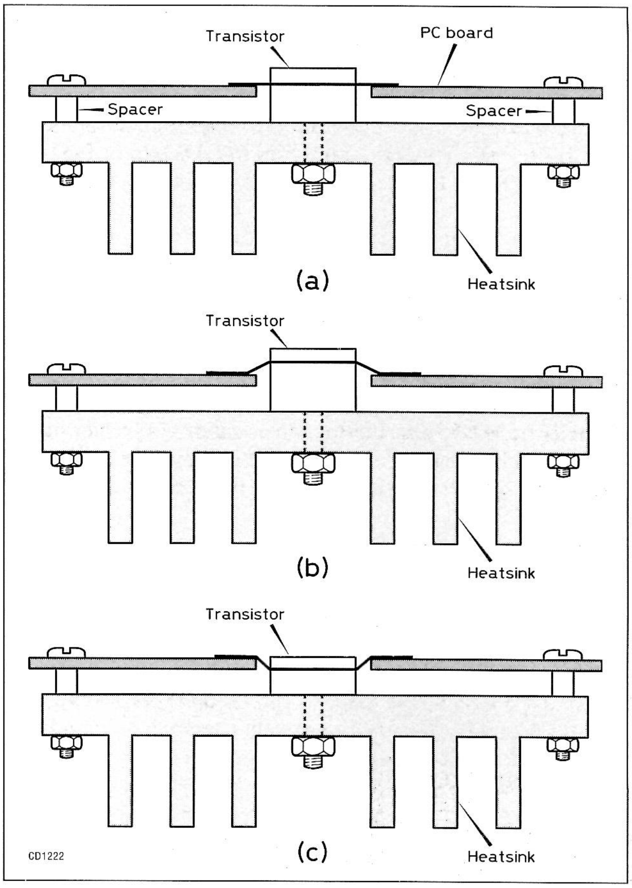

Fig 4.106. (a) Mounting arrangements for PCB, transistor and heatsink. (b) Acceptable lead forms (c) Incorrect lead form

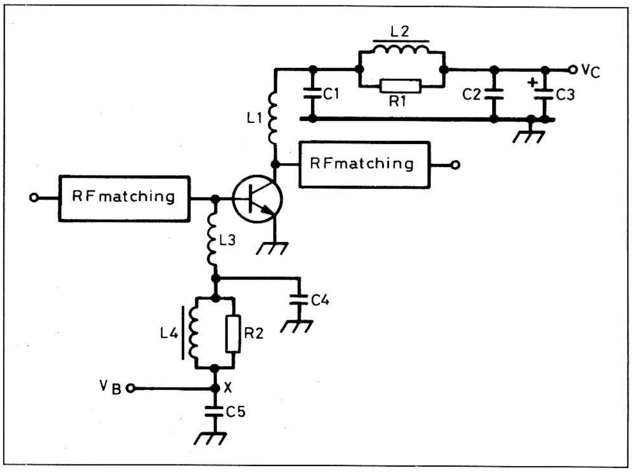

Fig 4.107. Transistor decoupling networks. L1, L3: RFC suitable for the operating frequency. L1 is usually an air-spaced inductor and L3 can be a moulded choke. L2, L4: ferrite-cored inductor, eg 2t on Siemens B62152 two-hole core, type A1X1 for L2, A4X1 for L4. C1, C4: decoupling at operating frequency; typically 1000pF. C2: 0.1µF ceramic. C3: 47µF. R1, R2: 10Ω 1W carbon or metal film

collector voltage throughout each RF cycle and thus provides a varying feedback path. Under some operating conditions spurious signals can be produced and, once present, the

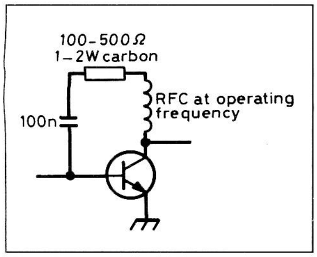

Fig 4.108. Transistor RLC feedback

non-linear feedback makes these self sustaining. The sustaining mechanism makes the effect prone to produce exact sub-harmonics (*f*/2, *f*/3 etc) of the input signal, although the signals will often interact with each other and produce a very wide band of spurii.

The effect is affected by the choice of matching circuits, supply voltage and operating power level. Avoiding heavy saturation is the single most effective preventative measure.

FETs are not totally immune to instability, but their internal feedback tends to be much lower than bipolar transistors and varies less with drain voltage, so the effects are not so severe.

Biasing

For linear operation transistors require some form of fixed bias. For FM only use, bipolars can be used without bias, although there is usually slight loss of gain. Power FETs always need fixed bias.

FETs are easy to bias as they only require a fixed voltage and, for normal amateur environments, temperature compensation is not needed. Biasing bipolars correctly is a little more complex. Bipolars are biased with current injected into the base. The bias source has to maintain a constant bias voltage as the base current varies over the RF and modulation cycles; the base current can be in the regions of 1A peak in high-power devices.

The base-emitter junction has the characteristics of a silicon diode, with a negative temperature coefficient of about 2mV/degree. As the transistor heats up, the base emitter voltage drops and, if the bias voltage is not reduced, more base current will flow, resulting in higher collector current, and more heating etc. The overall effect is termed *thermal runaway*. The solution is to use a diode, or another base-emitter junction, to sense the temperature of the power transistor (or the heatsink close by) and use this to adjust the bias voltage in line with the temperature.

Fig 4.109 shows a simple circuit using a power diode to set the bias voltage which is suitable for low-power amplifiers; the maximum base current of the transistor (I_{cmax}/H_{FE}) should not exceed about 10% of the current in the diode. For higher powers the bias circuit has to supply more current. Fig 4.110 shows the addition of an emitter follower to give higher current capacity; D1 is mounted by the RF transistor, D2 should be mounted close to TR1 and compensates for variations in V_{BE} of TR1. Other bias circuits are used in the amplifier circuit examples and these can be applied to most circuits without difficulty.

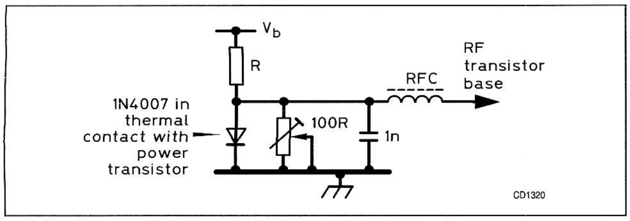

Fig 4.109. Basic diode bias circuit. Select R to give approx 250mA through the diode

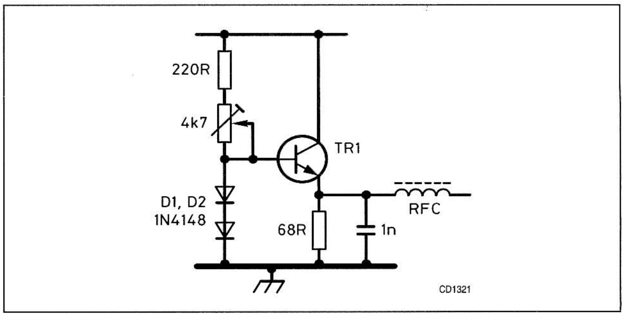

Fig 4.110. Emitter-follower bias circuit

Matching

The use of output impedance is inappropriate for RF power transistors; they are characterised in terms of load impedance. The external load impedance (often 50Ω) is modified by the matching circuits and presented to the transistor. The load impedance affects the balance between gain, linearity and efficiency at any given power level. Manufacturer's data will usually give figures for maximum output power and efficiency, and lower impedances might be needed for best linearity.

S parameters have very little value when applied to high-power amplifiers and transistors; by definition they are measured at signal levels where no non-linear effects occur and this is totally inappropriate for high-power situations. It is much more useful to define the load impedance required by the transistor for given operating conditions, and the input impedance of the transistor under those conditions.

The input matching circuitry converts the input impedance of the transistor to whatever impedance is needed to load the preceding stage. In the case of a stand-alone amplifier this will usually be 50Ω, but in an amplifier chain the conversion might be directly to the optimum load impedance of the previous transistor. This will simplify circuitry compared with matching each stage to 50Ω and then connecting them together, although it means that the output power of the driving transistor cannot be measured directly.

Modern simulation programs can use S parameters in conjunction with non-linear models of transistors to predict circuit behaviour but this is beyond the scope of this book.

$(V_{cc} - V_{cesat})^2/2P_o$ is widely used to establish a starting point for load impedance where this is not specified by the manufacturer. In practice, high-power transistors used near the upper limit of their frequency range will have an optimum load resistance which is one-half to one-third of the calculated value and which will also have a significant reactive component. The calculated value is still a reasonable starting point, but experimentation is likely to be needed to optimise performance in practice so there is benefit in using matching circuits which offer a wide tuning range.

At high powers, and especially at lower supply voltages (eg 12V), the load impedances can be very low with the resistive component being a fraction of an ohm. At 50W output, this corresponds to a RF current of 10A or more; components

Table 4.8. Capacitance of Philips 680 series miniature ceramic capacitors at 144MHz, lead length < 2mm

Marked value (pF)	Measured value (pF)
2	2.2
4.7	4.7
10	10
15	15.5
22	23
47	53
100	132
220	480
330	2760

carrying this level of current must be of low loss and suitable for the purpose. Layout and construction become critical; inductors typically become wide printed tracks or thick copper straps to give large surface area. Do not use tinned copper wire, which is lossy as skin effect means that the RF current flows in the surface plating and not in the copper wire. Capacitors capable of carrying high current are normally either metal clad mica, or porcelain ceramic chips (ATC or equivalent) and very few other types are suitable in these situations. Often several capacitors are connected in parallel make up the total value and share the current. Using higher supply voltages brings several benefits: higher impedances which makes matching easier and less lossy, higher gain and lower price/watt.

To some extent at 144MHz, and certainly at higher frequencies, the parasitic inductance of capacitors must be taken into account. For both chip and metal mica types, this is about 0.5nH. At 432MHz this is a reactance of +1.3Ω. A final reactance of, for example, −8Ω (46pF at 432MHz) actually requires a capacitor with a capacitive reactance of −9.3Ω. This is 40pF, 15% less than calculated. Table 4.8 shows another example, comparing the marked and measured values of leaded miniature ceramic capacitors at 144MHz. A leaded capacitor, even with minimum lead length, will have an unpredictable parasitic inductance of several nanohenrys. This makes it impossible to swap capacitors in matching networks with any certainty, so designs need to be reproduced accurately to have a good chance of success. Motorola application note EB46 [32] shows an excellent example; a VHF 80W amplifier is fixed-tuned in bands by component selection. The difference between 143–170MHz and 155–175MHz is that in the first, a 500pF capacitor is made from two 250pF components and in the second it is made from 200 and 300pF components.

The low impedances mean that there is a practical limit to the power level which can be obtained from a transistor. For example, trying to reach 150W by putting two 75W devices in parallel in one transistor package can reduce the impedance to a level which becomes unmanageable. One means of reaching the higher power level is to use two 75W devices in push-pull. Here the input and load impedances are in series, so the impedance is double that of a single transistor (see Fig 4.111), making matching much easier. Where the two transistors are mounted on a single package, further benefits arise. If the drive to the two halves of the device is truly antiphase, a RF ground point exists within the package and this means that there are no circulating ground currents in the external circuitry. If this circuit arrangement is used with two separate transistors, the physical separation introduces current in the path between the transistors which then demands the same low-inductance ground connections as are needed in single-ended circuits.

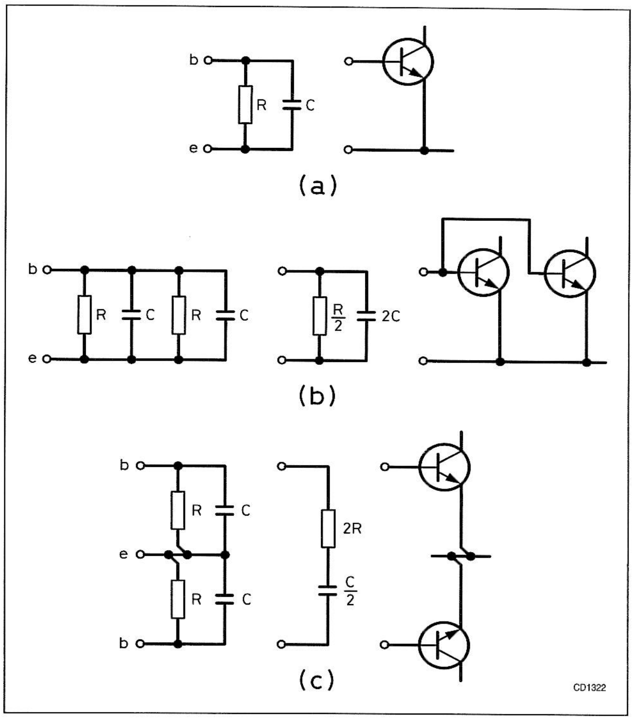

Fig 4.111. (a) Typical equivalent input circuit of high-power RF transistor. (b) Equivalent for two transistors in parallel. (c) Equivalent for two transistors in push-pull

At low frequencies the balanced drive signals are often generated with a conventional centre-tapped, wound transformer. Leakage inductance, stray capacitance and losses in ferrite make this unpredictable at VHF and transmission line transformers are widely used instead. The principle is shown in Fig 4.112. Basic laws of physics mean that if a current flows in the inner of a coaxial cable, exactly the opposite current flows in the outer. If the output end is not connected to ground, then the inner and outer are in antiphase or push-pull. Satisfactory operation depends on choosing the type and length of cable correctly. Lengths should not exceed 0.4λ and

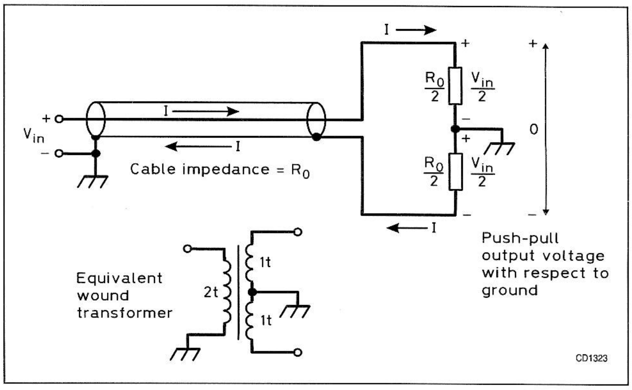

Fig 4.112. Operation of coaxial cable balun

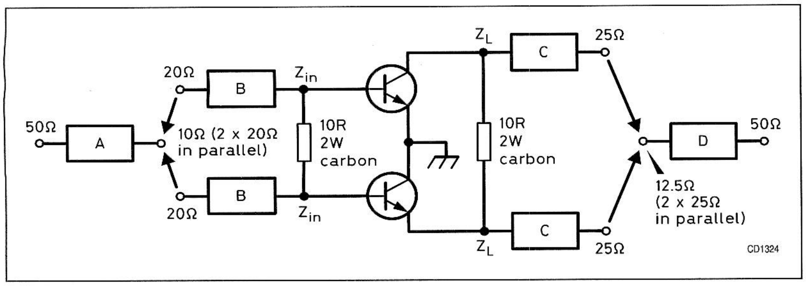

Fig 4.113. Matching for transistors in parallel (bias and power connections omitted). Impedance values shown are typical and need not be duplicated exactly

the minimum lengths needed depends on the cable size. As a rule of thumb, a length of λ/6 of UR43 type cable (5mm diameter) will have balanced outputs. If miniature cable is used (RG179 or smaller) then about half this length is adequate. As an alternative, a shorter length of cable can be loaded with low-loss ferrite material, sufficient to produce a few hundred ohms of reactance (calculated as if the cable was replaced by a single wire of the same diameter).

Another method for combining transistors in a circuit is to operate them in-phase [33]. Mechanically it is impractical to simply connect them in parallel, and Fig 4.113 shows how each transistor has some individual matching circuitry to a higher impedance and then the circuits are connected in parallel. There is a risk that the two-transistor circuit will form a push-pull oscillator, and resistors are often added to suppress this. If the two sides are in-phase and at equal amplitude, the resistors dissipate no power. In many instances where low-gain transistors are being used, these resistors can be omitted.

Matching circuits

At VHF, matching circuits are usually constructed from 'lumped', ie discrete, components and are usually designed to provide a useful amount of tuning range to allow for differences between devices, and as a result of variations in the construction of the circuit. The tuning range is also valuable where the impedances have been estimated and the actual values required can be significantly different to those calculated. At UHF and microwave frequencies, matching techniques tend to use printed striplines instead of wire inductors, and construction techniques become very important as even a millimetre of lead or track length represents a significant reactance. The formulae for estimating load impedance are highly inaccurate at these frequencies and designs need to be based on accurate data. Published designs need to be copied very accurately to ensure success.

Some circuit configurations widely used for narrow-band transistor amplifiers at VHF are shown in Fig 4.114 [34]. Power connections to the base and collector are not shown; these are assumed to be RF chokes so that they do not affect the operation of the matching circuit. The 'L' circuit of Fig 4.114(a) lacks flexibility in that the loaded *Q* is set by the impedance transformation ratio between R1 and R2, and there is limited tuning capability as the inductance is a fixed value; variable inductors are impractical in VHF power amplifiers.

The circuits of Figs 4.114(b) and 4.113(c) have greater flexibility through the use of two tuning capacitors. This allows the loaded *Q* to be chosen by the designer independently of the values of R1 and R2 and provides a wider range of impedance matching. A convenient by-product of the circuit is the DC blocking action of the series capacitor. A loaded *Q* of about 5–10 is usually chosen. Higher values give higher harmonic rejection, but at the cost of higher RF current and voltage in the network which leads to increased losses. In practice, a harmonic filter should be used at the output of every amplifier, so there is no benefit in choosing a high loaded *Q* in the matching network.

Where the impedance ratio to be matched is high, component values become impractical if the matching is carried out in a single stage. In these cases, it is common to use two circuits in series, each carrying out about half of the total transformation. The intermediate impedance can be chosen by:

$$\sqrt{R_1 + R_2}$$

The definition of 'high' varies with frequency and impedance, but as a rough guide, a single three-component section will match an impedance ratio of up to 15:1 at 50/70MHz and 10:1 at 144MHz.

Quarter-wave coaxial transformers provide another convenient method of transforming impedances. As the transistor impedances are <50Ω, cable impedances of <50Ω are needed. Such cables are rare and expensive, but readily available 50Ω cable can be used by connecting lengths in parallel,

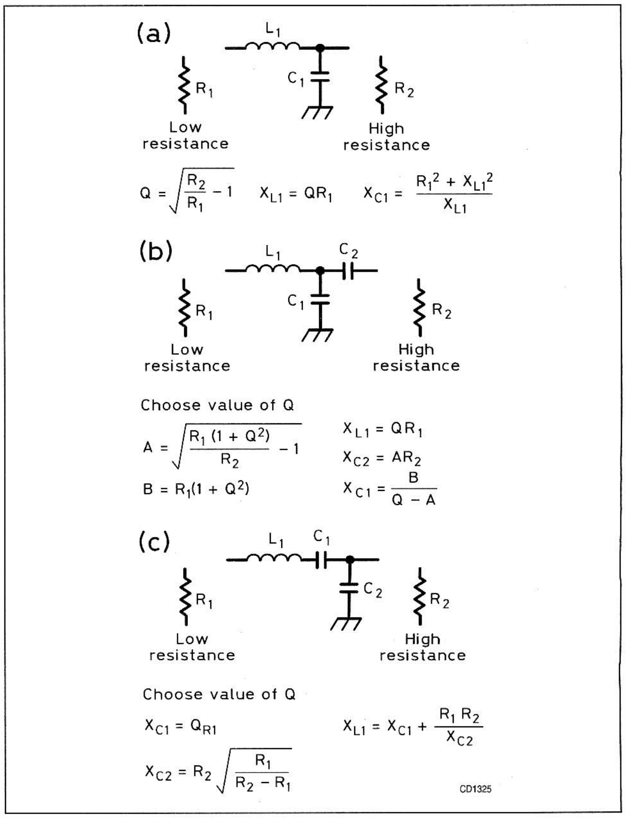

Fig 4.114. Transistor matching networks

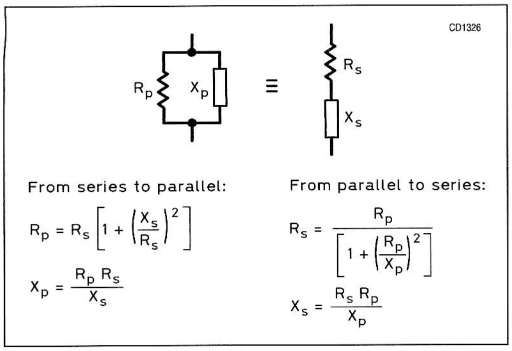

Fig 4.115. Conversion between parallel and series impedances

for example 2 × 50Ω behaves like 25Ω, and 3 × 50Ω like 16.7Ω.

All of the matching circuits discussed here are most easily designed when transforming between purely resistive values. In practice, transistor impedances normally include some reactive component. The design process involves cancelling ('tuning out') the reactive part of the impedance to leave a resistance and then designing the matching network to transform that resistance to the desired value. Impedances are normally given as series components; a resistance in series with a capacitive or inductive reactance. In designing matching circuits it is often useful to convert between series and parallel descriptions of the impedance; any given impedance can be defined in both ways, as shown in Fig 4.115.

General rules for designing matching networks

Input matching (Fig 4.116)

If the transistor impedance is inductive, convert it to the parallel form and add a parallel capacitor at the base lead to cancel the reactance of the inductive component, then match to the resistive value. If the transistor impedance is capacitive, add a series inductance to cancel the capacitive reactance.

Output matching (Fig 4.117)

If the manufacturer gives a load impedance for the desired operating conditions then proceed as follows. If the desired load impedance is inductive, transform the external load (50Ω) to give the correct resistive value and add a series inductor to give the required inductive reactance. Alternatively, the inductance can be provided as a parallel element, which doubles as the power supply connection. While this is often used it introduces high RF currents into the decoupling capacitors. For this reason, the technique is not recommended for those without experience or access to RF test equipment.

If the desired load is capacitive, convert it to the parallel form, make the resistive transformation from the external load to the required parallel

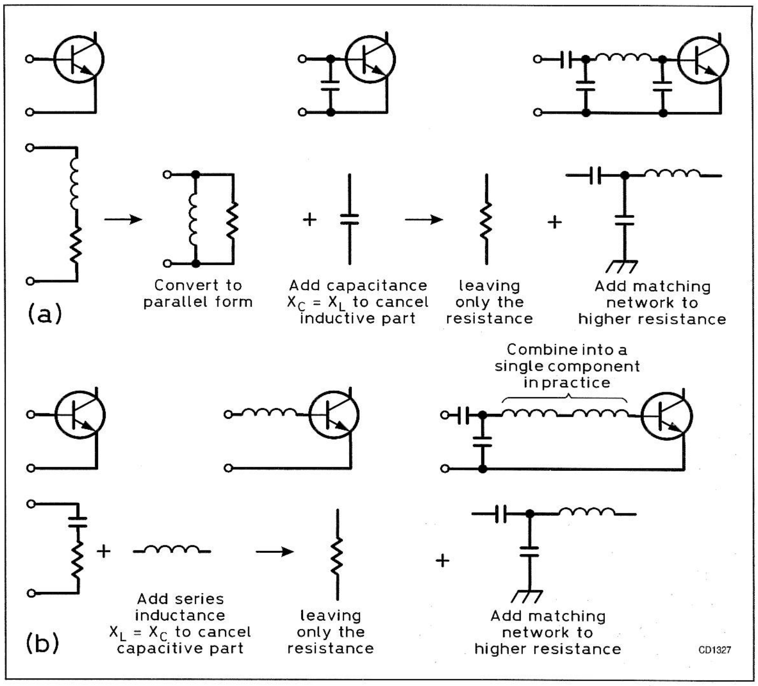

Fig 4.116. Transistor input matching. (a) Input impedance is inductive. (b) Input impedance is capacitive

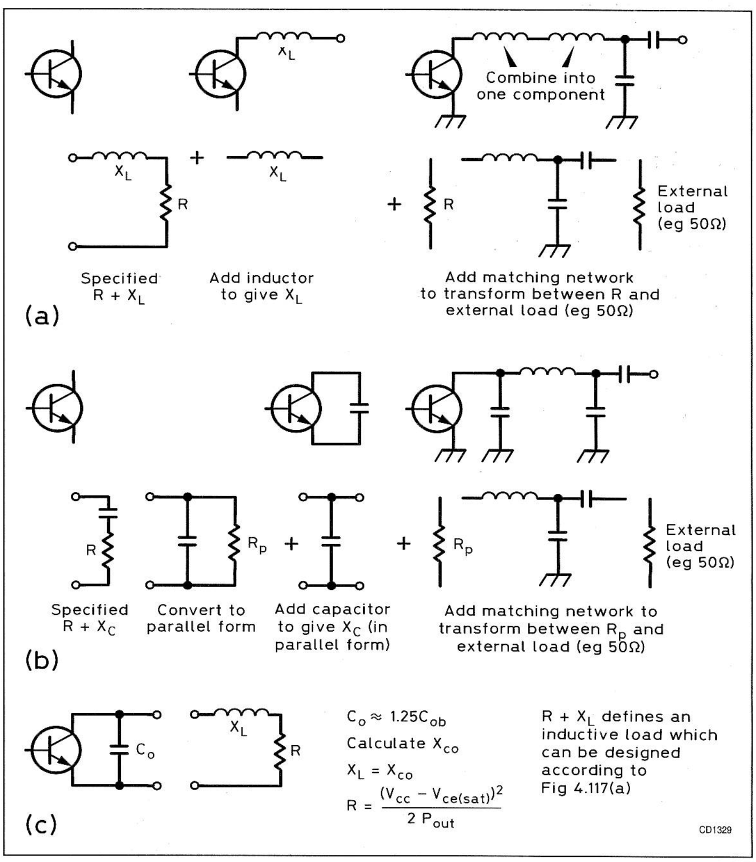

Fig 4.117. Transistor output matching. (a) Specified impedance is inductive. (b) Specified impedance is capacitive. (c) Estimated impedance

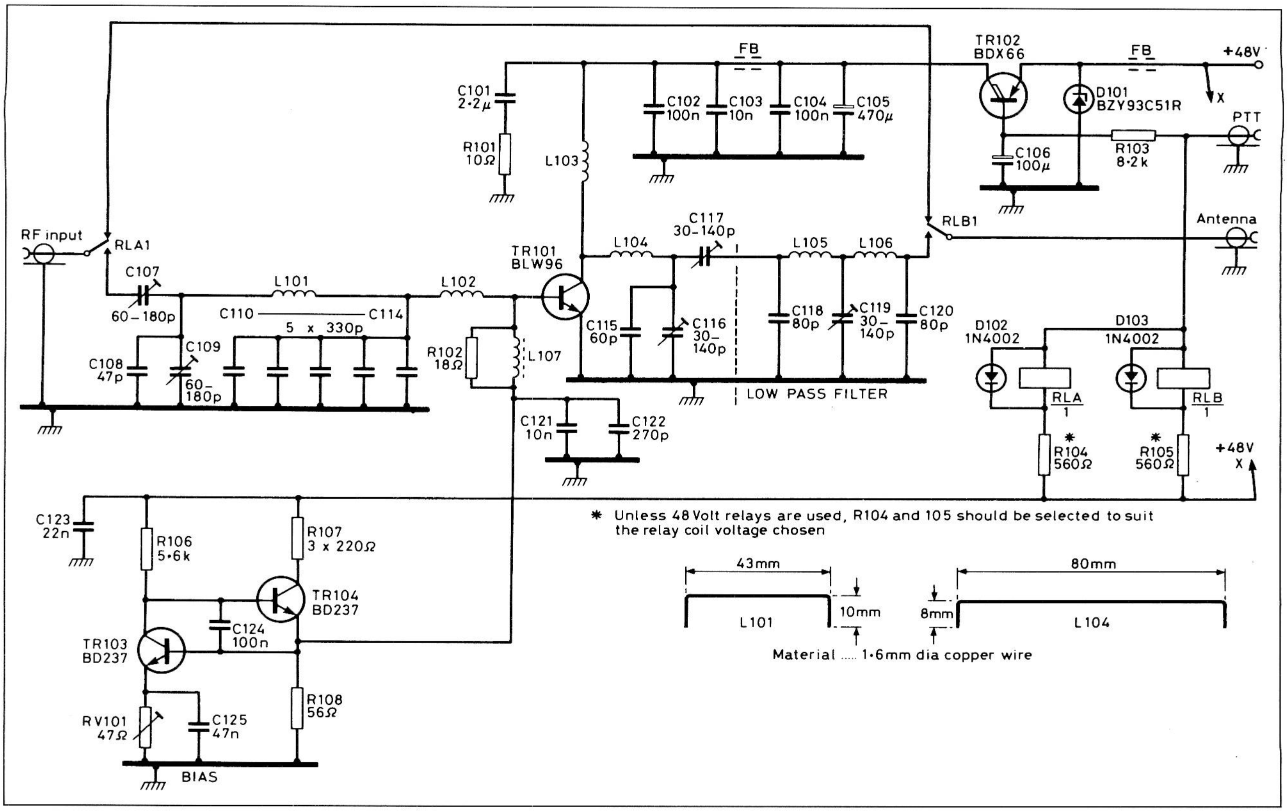

Fig 4.118. Circuit diagram of the amplifier

load resistance and add a parallel capacitor to give the required capacitive reactance.

Where the load impedance is being estimated, proceed as follows. The resistive part of the load is calculated from:

$$(V_{cc} - V_{cesat})^2/2P_0$$

V_{cesat} should be chosen according to the supply voltage and mode of operation; for 12V transistors this is 2V for FM only, and 4V for linear operation with SSB. For 28V transistors, the values are 4V and 6V respectively. The transistor output capacitance is difficult to define as it varies widely with collector voltage swing. If the transistor data shows a graph of C_{ob} versus voltage, use the value of C_{ob} at 60% of the supply voltage which will be used, otherwise use a value of about 1.25 times the C_{ob} figure given in the data sheet. In either case the desired load impedance is inductive, the inductive reactance being chosen to cancel the transistor's output capacitance.

Design example – 100W amplifier for 50MHz

G3WZT described a 100W linear amplifier for 50MHz [35] (Fig 4.118) which covered the process of designing input and output networks in some detail. Those portions of the article are repeated here as a comprehensive design example.

Input matching

The input matching consists of two T-networks which match the required drive impedance of 50Ω to the complex input impedance of the BLW96. T-networks are used in preference to the more simple L-arrangement, as there is no control of the working Q when using the latter. As the Q of an L-network increases with the ratio of transformed impedances, high circuit Qs can exist. This makes adjustment very critical and temperature sensitive, and also causes unwanted narrow bandwidth and high circulating RF currents. Inspection of the manufacturer's data for the BLW96 shows an input impedance of 0.37 + j0.15Ω (equivalent series components) at 50MHz. The reactive part of the input is very small (0.47nH) and in this instance may be ignored. However, other devices may have a much higher reactive component in the input impedance, and for this reason it will be taken into account as part of the input matching network in order to demonstrate the method.

To start the design, it is first necessary to decide the intermediate resistance between the two T-match sections (see Fig 4.119). Assuming both sections have the same Q, the intermediate resistance is the geometric mean value of source and load resistance R1 and R2′. Therefore, intermediate resistance:

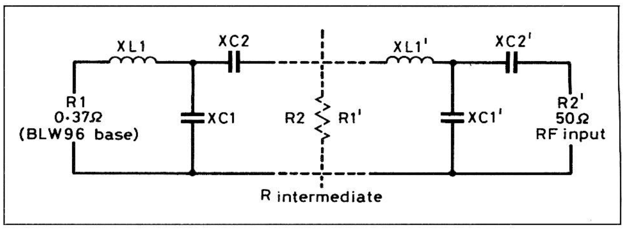

Fig 4.119. Input matching arrangement

$$R_I = \sqrt{0.37 \times 50} = 4.3\Omega$$

Working from the base, the first T-match section transforms from 0.37Ω to 4.3Ω, and the second section from 4.3Ω to the design source impedance of 50Ω.

The next requirement is to define the working Q of the sections. If too low a value is selected, component values become impractical; too high a value leads to the unwanted problems described earlier. The minimum working Q for given values of R1 and R2′ is when:

$$R_1(Q^2 + 1)/R_2 > 1$$

If variables are selected which give a value of less than 1, the value of A cannot be solved. In order to satisfy this requirement and avoid high Q for the reasons given earlier, a Q of 4 in both sections is used for this design.

Referring to Fig 4.119, the following formulae are applied to obtain the values for X_L, X_{C1} and X_{C2}:

$$X_L = QR_1 \quad (1)$$
$$X_{C2} = AR_2 \quad (2)$$
$$X_{C1} = B/Q - A \quad (3)$$

where:

$$A = (B/R_2 - 1)^{0.5} \quad (4)$$
$$B = R_1(Q^2 + 1) \quad (5)$$

Inserting values into the formulae gives the following values:

From (5) $B = 0.37(4^2 + 1) = 6.29$
From (4) $A = (6.29/4.3 - 1)^{0.5} = 0.68$
From (3) $X_{C1} = 6.29/4 - 0.68 = 1.89\Omega$
From (2) $X_{C2} = 0.68 \times 4.3 = 2.92\Omega$
From (1) $X_{L1} = 4 \times 0.37 = 1.48\Omega$

To calculate the component values:

$$C = 1/2\pi f X_C \quad (6)$$
$$L = X_L/2\pi f \quad (7)$$

Therefore:

$$C_1 = 1/2\pi \times 50 \times 10^6 \times 1.89 = 1684\text{pF}$$
$$C_2 = 1/2\pi \times 50 \text{ x}10^6 \times 2.92 = 1090\text{pF}$$
$$L_1 = 1.48/2\pi \times 50 \text{ x}10^6 = 4.71\text{nH}$$

The final configuration and values for the first matching section are shown in Fig 4.120(a). Although in this case the inductive reactance of the transistor input is insignificant from a practical point of view, it will be subtracted from the value of L1 to keep the example correct. The value of L1 then becomes:

$$4.7 - 0.47 = 4.23\text{nH } (X_{L1} = 1.48 - 0.15\Omega)$$

The second matching section, transforming the intermediate value of 4.3Ω up to the required driving impedance of 50Ω, is obtained in the same way. A working Q of 4 is used again. Refer to Fig 4.119.

From (5) $B = 4.3(4^2 + 1) = 73.1$
From (4) $A = (73.1/50 - 1)^{0.5} = 0.68$
From (3) $X_{C1'} = 73.1/4 - 0.68 = 22\Omega$
From (2) $X_{C2'} = 0.68 \times 50 = 34\Omega$
From (1) $X_{L1'} = 4 \times 4.3 = 17.2\Omega$
From (6) $C_{1'} = 1/2\pi \times 50 \times 10^6 \times 22 = 145\text{pF}$
$C_{2'} = 1/2\pi \times 50 \text{ x}10^6 \times 34 = 94\text{pF}$

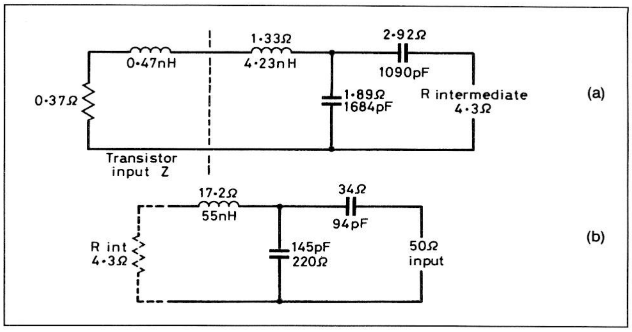

Fig 4.120. (a) First T-match section (input). (b) Second T-match section (input)

From (7) $L_{1'} = 17.2/2\pi \times 50 \times 10^6 = 55\text{nH}$

The circuit configuration, with values, is shown in Fig 4.120(b). In practice the series components C2 and L1′ are combined into one component. The capacitive reactance (2.92Ω) is subtracted from the value of inductive reactance (17.2Ω). This leaves an effective inductive reactance of 14.28Ω, which represents a value of 45.4nH. The values shown in the circuit diagram represent this effective inductance. The calculated values tend to be rather impractical, of course, and far from standard-value components. Where possible, mica compression trimmers are used, based on the calculated component values.

Output matching circuit

Before work on the output matching is started, it will be necessary to determine the load impedance for the BLW96. This may be done in two ways. The first, by taking the values directly from the manufacturer's data sheets (only valid for a specific power level); or second, if these are not available, by means of a simple calculation.

Most data sheets include a simple graph of resistance and reactance plotted against frequency for a given output power. For the BLW96 at 50MHz the equivalent series load impedance is 4 + j3Ω. Unfortunately this value is quoted at the wrong power level for this design. It should be made clear that these values are the complex conjugate of the transistor load impedance and represent the *load* required to match the device correctly. In this particular case, the transistor is represented by a 4Ω resistor in series with a 1060pF capacitor. If a full data sheet is not available or, as in this example, values are quoted at the wrong power level, a close approximation may be made by using the following formula in conjunction with the output capacitance:

$$R_L = (V_{cc} - V_{sat})^2/2 \times P_{out}$$

Based on a saturation voltage of 2V and a power output of 100W PEP, the value of R_L is:

$$(48 - 2)^2/2 \times 100 = 10.58\Omega$$

The collector capacitance against voltage will normally be shown in the form of a graph or table. As large changes in capacitance occur over the range of collector voltages, a general rule-of-thumb is to take the value shown at 50% of the supply. In this case C_c amounts to 350pF ($X_c = 9\Omega$) at a V_{cb} of 25V. This value is in parallel with the load resistance of 10Ω previously calculated.

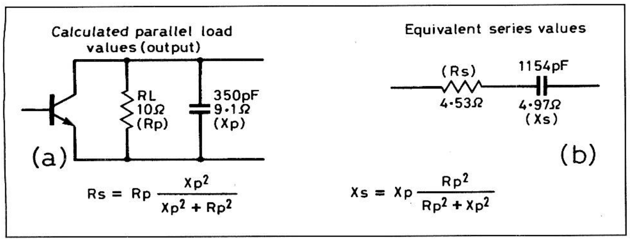

Fig 4.121. Output load conversions. (a) Calculated parallel load values. (b) Equivalent series values

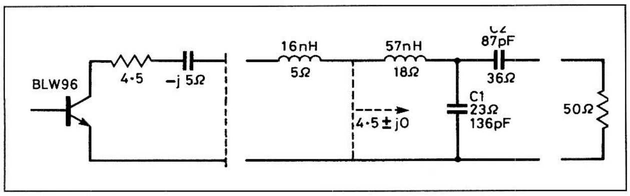

Fig 4.122. Output matching with values

For ease of matching, and to enable a comparison to be made with the published figures, the parallel circuit must be converted into an equivalent series circuit. The conversion formulae with the calculated values are shown in Fig 4.121. It will be seen that these figures differ slightly from the values in the data sheet as the calculation was carried out at a different power level. A good degree of accuracy is obtained if no other data is available at a specific power level.

Now that the required collector load impedance is defined, the output matching circuit can be designed to match from 4.5 – j5Ω to the required output of 50Ω. Unlike the input matching, only one T-match section will be required as the impedance step-up ratio is lower.

Referring back to equations (1) – (5) and using a Q of 4:

$B = 4.5(4^2 + 1) = 76.5$

From (5) $A = (76.5/50 - 1)^{0.5} = 0.728$

From (1) $X_L = 4 \times 4.5 = 18\Omega$

From (2) $X_{C2} = 0.728 \times 50 = 36.4\Omega$

From (3) $X_{C1} = 76.5/4 - 0.728 = 23.3\Omega$

From (6) $C_1 = 1/2\pi \times 50 \times 10^6 \times 23.38 = 136\text{pF}$

Table 4.9. Components for G3WZT 50MHz amplifier

R101	10R 0.5W carbon film
R102	18R 0.5W carbon film
R103	8k2 0.25W carbon film
R104, 105	Select to suit relays used
R106	5k6 0.5W carbon film
R107	3 × 220R 6W wire wound
R108	56R 0.5W carbon film
RV101	47R cermet trimpot
C101	2µ2 63V polycarbonate
C102, 104, 124	100n 100V monolithic ceramic
C103, 121	10n 100V monolithic ceramic
C105	470µ 63V tubular elec
C106	100µ 63V tubular elec
C107, 109	60–180p mica compression trimmer
C108	47p 50V ceramic chip
C110–114	330p 50V ceramic chip
C115	60p 250V Unelco mica or ATC
C116, 117, 119	30–140p mica compression trimmer
C118, 120	80p 250V Unelco mica or ATC
C122	270p 100V monolithic ceramic
C123	22n 100V monolithic ceramic
C125	47n 100V monolithic ceramic
D101	BZY93 C51R zener
D102, 103	1N4002
TR101	BLW96
TR102	BDX66 PNP Darlington
TR103, 104	BD237
FB	Suppression bead. Material, 3S2 (blue)
L101	See Fig 4.118
L102	15 × 7mm pad on PCB
L103	12t 1.2mm copper wire, 9mm ID, 28mm long
L104	See Fig 4.118
L105, 106	4½t 1.2mm copper wire, 10mm ID
L107	2½t 0.5mm enam copper wire wound through 6-hole ferrite bead
RLA, RLB	50R coaxial, type CX120P

$C_2 = 1/2\pi \times 50 \times 10^6 \times 36.4 = 87.4\text{pF}$

From (7) $L_1 = 18/2\pi \times 50 \times 10^6 = 57.3\text{nH}$

The final matching circuit values are shown in Fig 4.122. An additional 5Ω must be included in the value of X_L, making a total of 23Ω. This additional reactance, being of opposite sign, cancels the capacitive reactance part of the transistor output impedance (–j5Ω). Both the 16nH and 57nH inductors are combined into a single component.

Fig 4.123 shows the component layout for the amplifier. A component list is given in Table 4.9.

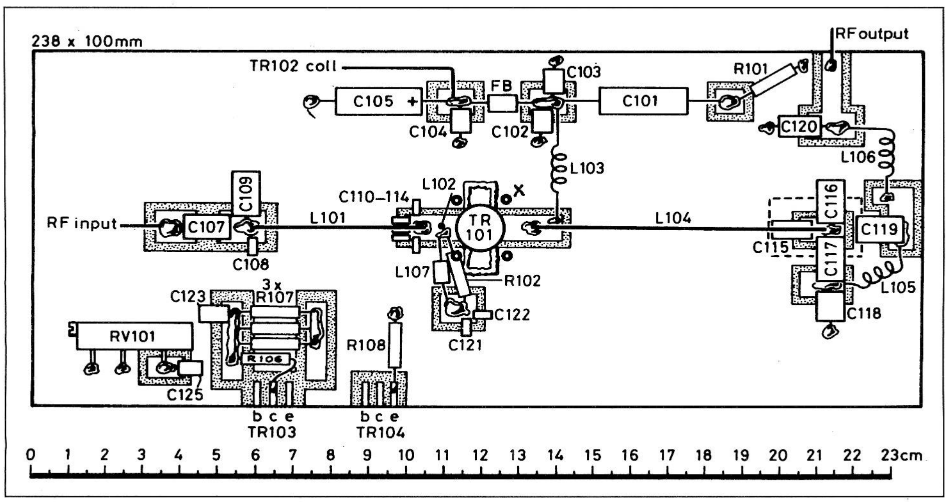

Fig 4.123. Main board layout. Connect top and bottom ground planes with pins at four points marked 'x'. Remove underside copper in dotted area

Combining for higher powers

A further method of reaching higher power levels is to combine a number of complete amplifiers. Normally these amplifiers will be identical, with very similar gains. In these

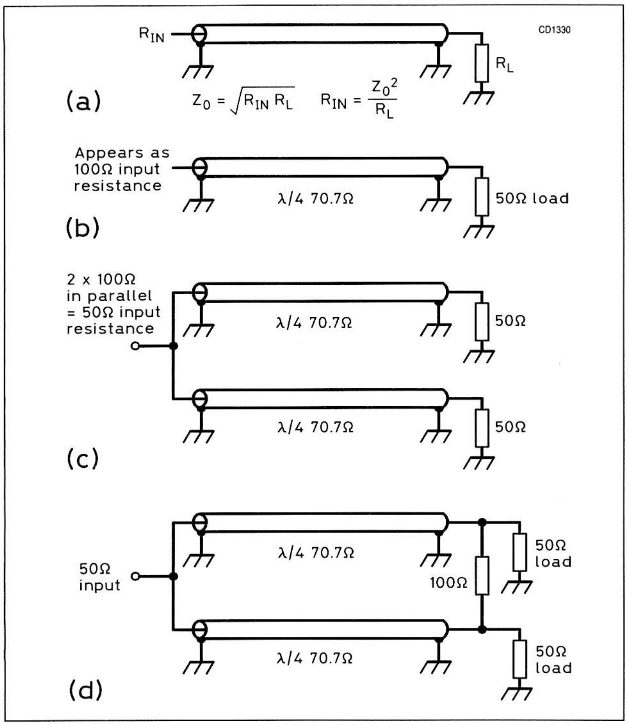

Fig 4.124. (a–c) Coaxial splitter/combiner and derivation. (d) Wilkinson splitter/combiner

examples, it is assumed that each amplifier is designed to work at 50Ω input and load impedances.

Splitters and combiners are basically the same; a splitter has one input and two or more outputs. Used in reverse the same circuit becomes a combiner with two or more inputs and one output. The combiner might use different components to cope with the higher power, but the circuit will be the same as the splitter. For convenience, the term 'combiner' is used here to cover both splitters and combiners.

In all of the examples discussed in this section, the lengths of cables and transmission lines relate to the electrical lengths, after taking account of the velocity factor.

The simple combiner in Fig 4.124(b) can be made using λ/4 coaxial cable transformers, as used to combine antennas. While simple, this suffers from the disadvantage of offering no isolation between the amplifiers. This can result in instability and, if the amplifier characteristics are not absolutely identical, loss of output power. The Wilkinson combiner is formed by adding a resistor between the divided points. This dissipates any imbalance and provides isolation between the two amplifiers; each one operates completely independently of the other. To get full isolation, it is important that the resistor is purely resistive and high-power flange mounted components are usually used. At UHF and microwaves, the capacitance of these resistors becomes too high for use in this configuration and it is better to use a configuration where the isolating resistor is a 50Ω load connected to ground. One such is the 'rat-race' (Fig 4.125), so called because of its form if made with printed transmission lines. This can be seen as an extension of the Wilkinson design. In these, the optimum impedance is 70.7Ω but in practice 75Ω coaxial cable works perfectly well.

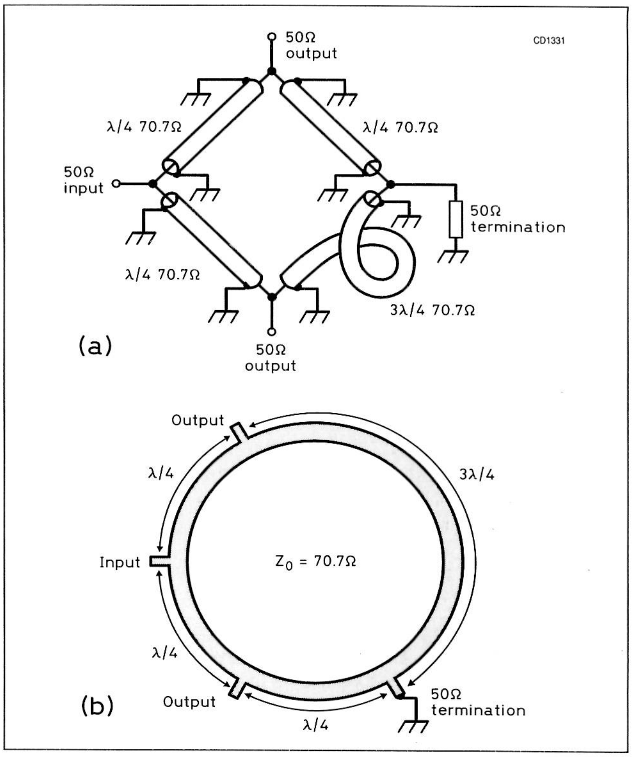

Fig 4.125. 'Rat-race' splitter/combiner. (a) Coaxial cable version. (b) Branch hybrid PCB form

Another family of combiners produce outputs which are have a 90° phase offset. For this reason they are often called quadrature couplers. These have the advantage that, when used as a splitter, the input impedance is always 50Ω as long as the impedances at the two outputs are the same (whatever the actual impedances are). This is of great value when making wide-band amplifiers but has limited importance in amateur applications.

The simplest, but most expensive, version is Sage Wireline™ (Fig 4.126). This can be viewed as a directional coupler

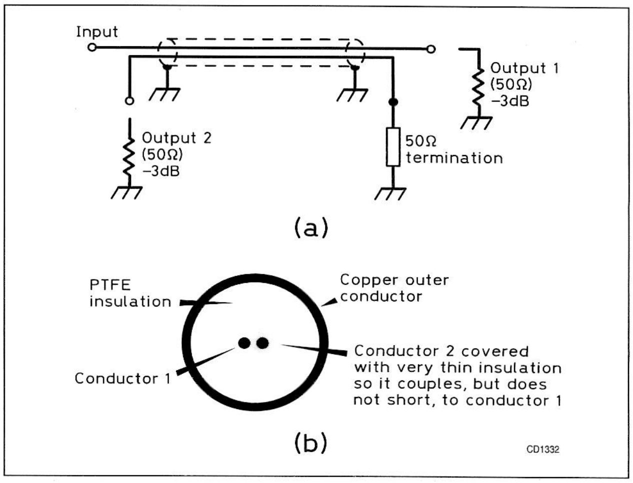

Fig 4.126. Sage Wireline™. (a) Shown schematically, as a directional coupler. (b) Shown mechanically as a cross-section through the cable

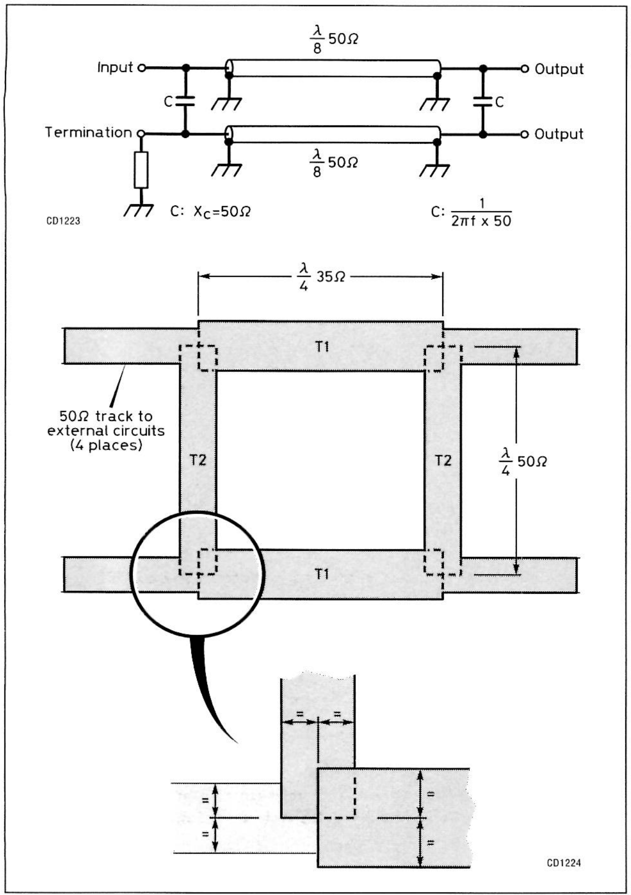

Fig 4.127. Eighth-wave splitter/combiner – coaxial cable and PCB versions

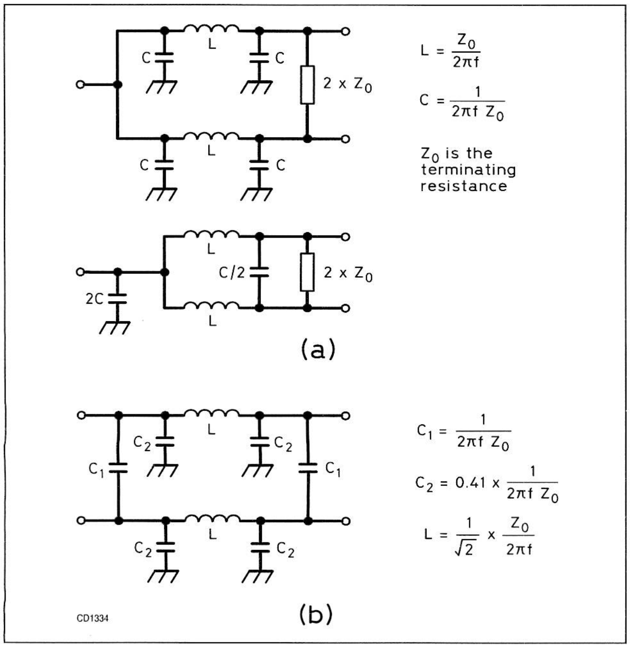

Fig 4.128. Lumped-component splitter/combiners. (a) Wilkinson. (b) Eighth-wave

where the coupling is 3dB when the electrical length is λ/4; at this frequency half of the input power transfers to the coupled line. Wireline also has the advantage of wide bandwidth; typically the coupling varies by ±0.5dB (±12%) over a 2:1 frequency range.

Two other forms of quadrature couplers shown in Fig 4.127 are the *eighth-wave* (or *capacitively coupled hybrid*) and *branch hybrid* designs. These have similar characteristics to Wireline, but operate over a narrower frequency range. The branch hybrid design is widely used at microwave frequencies as the whole combiner can be microstripline without discrete components to introduce unpredictable parasitic elements. Additionally, the split and combining connections are on the same side which is very convenient for circuit layout.

The Wilkinson and eighth-wave designs can be made using lumped components in place of the transmission lines (Fig 4.128). The bandwidth is reduced, but is still adequate for amateur band coverage. This is especially useful for the lower VHF bands where the coaxial cable lengths become unwieldy.

Fig 4.129 shows how amplifiers are connected using these combiners. The orientation of input and output on quadrature combiners is important; the wrong way round will result in all the output power going into the terminating resistor.

Hybrid modules

Modules are available for all VHF bands. They provide a simple and almost foolproof solution where low and medium output power amplifiers are required. A wide variety of output power and supply voltage ratings are available to suit a great many situations.

Philips, Motorola and Mitsubishi all produce ranges of modules for FM use in amateur or PMR transmitters and Mitsubishi, Icom and Toshiba also supply linear modules for all-mode use in the amateur bands. Major advantages of modules include small size, high gain within a single package, ease of use (no alignment) and guaranteed stability and ruggedness with normal loads. Prices are usually higher than for the equivalent amplifier in discrete components.

The use of the term 'hybrid' arises from the internal construction where a mixture of techniques is used. The RF power transistors and other semiconductors are often connected into the circuit with bond wires (about 0.001in/0.025mm diameter) directly from the silicon die to the circuit tracks, without the familiar package as an intermediate step. Capacitors are either conventional chip types, or silicon MOS capacitors which are manufactured on silicon wafers in a similar fashion to a transistor die. One connection is the surface which is bonded directly onto the track; the other connection is made by a wire bond between the other surface. The track resistors are usually thick-film types, formed directly on the surface of the substrate.

The substrate material is usually alumina (Al_2O_3) which has a dielectric constant of about 10, compared with 4.5 for normal epoxy glass PCB and 2.5 for PTFE-based materials. This is a key factor in miniaturising the circuitry as track widths and lengths in the matching circuitry both reduce significantly. Alumina has the disadvantage of being a brittle ceramic material and problems can arise if the module is subjected to shock or stress. For example, shock can arise if the module is dropped onto a hard surface, or marked with a centre punch.

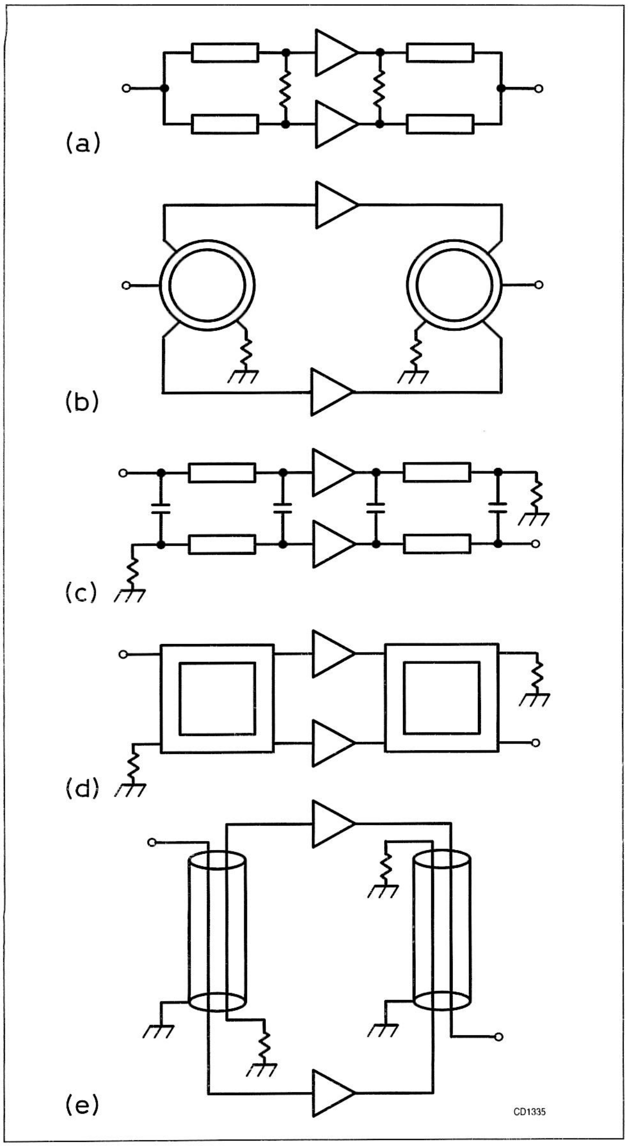

Fig 4.129. Combining amplifiers. The arrangement of input, output and termination connections is important in all cases. (a) Wilkinson. (b) Rat-race. (c) Eighth-wave. (d) Branch line. (e) Wireline™

Stress often arises when the module is fixed to a heatsink. If the heatsink surface is convex instead of flat, the situations shown in Fig 4.130 can arise; one fixing screw has been tightened, leaving the other end proud. If the other screw is then tightened, the flange of the hybrid will be bent, cracking the substrate. Motorola give detailed instructions explaining the need for the surface to be flat within 0.005in/0.12mm. Figs

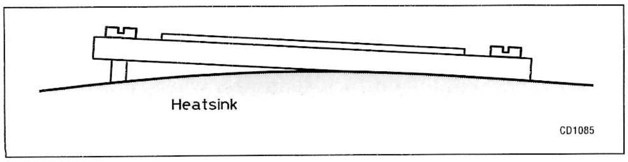

Fig 4.130. Hybrid module on convex heatsink with one fixing screw tightened

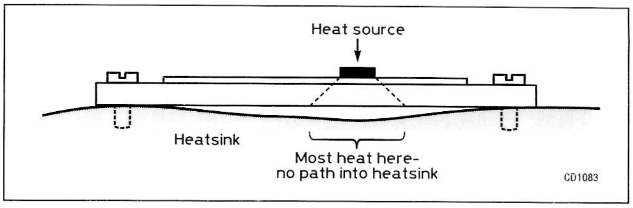

Fig 4.131. Here there is insufficient contact between a concave heatsink and the hot portion of the module

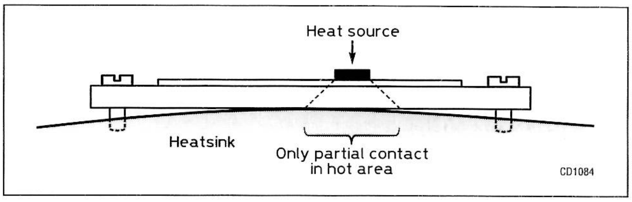

Fig 4.132. Here there is only partial contact with a convex heatsink

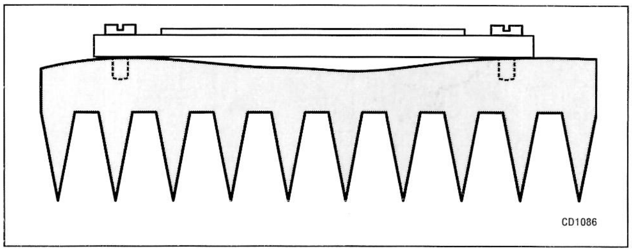

Fig 4.133. Exaggerated view of heatsink which is not flat across the fins

4.131 and 4.132 show problems of insufficient cooling which also occur when the mounting surface is not flat.

As discussed earlier, heatsinks are not normally flat across the fins, as shown in Fig 4.133. This can be seen, and corrected, by sanding very gently against abrasive paper placed face up on a flat surface. The aim is to end up with a flatter surface without making it heavily grooved; this is possible with gentle pressure and fine abrasives. Greater effort is needed than with discrete transistors because of the larger contact area. The module should be fitted to the heatsink using white thermal compound, just sufficient to give continuous contact between the module and the heatsink; the compound is intended to fill microscopic gaps between the surfaces and will not make up for surfaces which are not flat to start with.

Fig 4.134 shows typical circuits using modules. In all cases, LF decoupling should be added externally at each supply pin. In FM use, the output power can varied by adjusting the supply voltage to the driver stages (pin 3 of the BGY36 or V_{cc1} in Fig 4.134). For SSB, this is not usable because of the distortion which arises; the output power is controlled by adjusting the RF input power. Modules intended for SSB use have an additional supply input for the bias circuitry. This is usually a lower voltage than the main supplies so a stable bias source can be derived from a regulated voltage. In SSB use, the output power should not exceed 50% of the rated power of the module in order to ensure that IMD levels are acceptable. Fig 4.134 shows the bias supply for the linear module as 9V; for lower power modules designed for use with lower supply

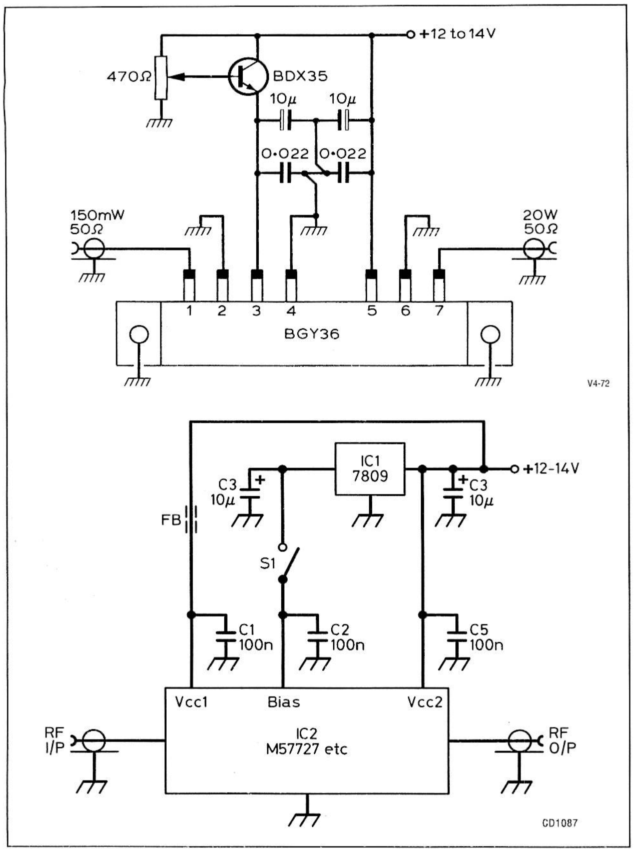

Fig 4.134. Typical hybrid module circuits

Table 4.10. A list of some hybrid modules suitable for use in amateur bands

Part no	Freq range (MHz)	Power (W)	Voltage (V)	Mode	Manufacturer
MHW710-1	400–440	13	12.5	FM	Motorola
MHW720-1	400–440	20	12.5	FM	
MHW720A1	400–440	20	12.5	FM	
BGY32	68–88	18	12.5	FM	Philips
BGY35	132–156	18	12.5	FM	
BGY135	132–156	18	12.5	FM	
BGY145A	68–88	28	12.5	FM	
BGY145B	146–174	28	12.5	FM	
BGY46A	400–440	1.4	9.6	FM	
BGY47A	400–440	3.2	9.6	FM	
BGY113A	400–440	7	7.5	FM	
M57735	50–54	10	12.5	SSB	Mitsubishi
M57796MA	144–148	5	7	FM	
M57713	144–148	10	12.5	SSB	
M57727	144–148	25	12.5	SSB	
M57726	144–148	35	12.5	FM	
M67727	144–148	45	12.5	SSB	
M57786M	430–470	5	7.2	FM	
M57716	430–450	10	12.5	SSB	
M57729	430–450	25	12.5	FM	
M57745	430–450	25	12.5	SSB	
M67715	1240–1300	1	8	SSB	
M57762	1240–1300	10	12.5	SSB	
SAV7	144–148	28	12.5	FM	Toshiba
SAU4	430–450	10	12.5	SSB	

FM modules have some stages biased in class C and are suitable for cw/FM only. SSB modules are linear and can be used with all voice/data/cw modes.

Table 4.11. Component notes for Fig 4.135

R_{E1}, R_{E2}	10R 0.1W chip (0805 size)
R_F	390R 0.5W carbon composition with 0.25in leads. It should be installed just above the transistor. 0.05in of lead length are in contact with the board, so the parasitic lead length is 0.20in on each side of the resistor
L_{E1}, L_{E2}	Parasitic inductances that affect the stability of the amplifier
RFC1, 2	RF chokes, roughly 300μH. 11t 26 AWG enam closewound, 0.166in ID. Use a No 19 drill bit as a mandrel
TRL1–6	50Ω microstriplines etched on 1/16in G-10 or FR-4 glass-epoxy circuit board
TR1	Motorola MRF581. This device is rated at a total device dissipation of 2.5W if the collector lead next to the package is kept at or below 50°C
J1, 2	SMA connectors were used in the prototype for convenience and ruggedness. Coaxial cable may be soldered directly to the board

voltages the bias supply is usually 5V, needing a 7805 regulator. Check the data before connecting power. A list of modules suitable for in amateur bands is shown in Table 4.10.

RF input requirements for modules with 1–25W output are typically in the region of 50–300mW. Figs 4.135–4.140 show some suitable linear driver circuits [36, 37] which can in turn be driven by MMICs typically used in modern transverter designs. Many crystal oscillator or synthesised sources will need only a single amplifier stage to reach the desired level for driving a module.

The M57762, giving 10W linear, 20W CW is particularly attractive for 1.3GHz use, where there are few alternative choices available. Some circuits have been published [38, 39] using specially produced transistors which NEC made some years ago, but these are long obsolete. Other transistors which have been used are usually intended for use at frequencies up to 800–900MHz and are unstable and/or of low gain at 1.3GHz. For higher power, modules can be combined. *VHF Communications* published articles [40, 41] which describe two and four modules combined; the two-module amplifier [40] uses combiners which provide no isolation between the modules, and the other [41] uses commercial four-way combiners which will typically cost at least as much as the modules themselves. Suitable combiner designs are discussed elsewhere in this chapter.

AMPLIFIER CIRCUITS

25W amplifier for 144MHz

Fig 4.141 shows this amplifier, based on reference [42], which is suitable for adding to low-power handheld transceivers; the amplifier includes RF driven T/R switching for automatic operation. The RF switching can be copied for use with other designs. The B25-12 transistor (CTC/Acrian) is no longer manufactured but similar types will work with little or no modification to L1 and L2. Construction can be in the form of Fig 4.105.

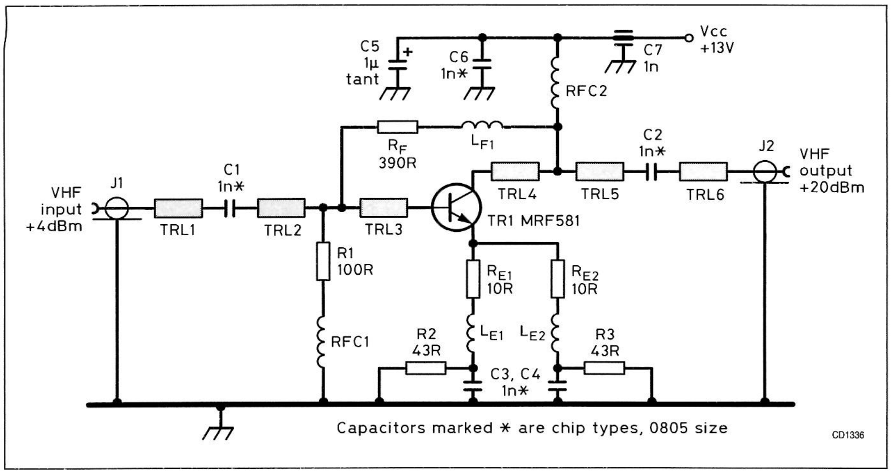

Fig 4.135. Circuit of the hybrid module driver for 50–144MHz. For component notes see Table 4.11. For the PCB layout and component layout see Figs 4.136 and 4.137 (Appendix 1) *(QEX)*

A small amount of the RF input signal is rectified by D1/2 and turns on TR1 and TR2, operating RL1–3. RL1/2 bypass the amplifier during receive and connect it into circuit during transmit. RL3 connects power to the bias circuit for TR3. D4 should be mounted in contact with TR3 to provide thermal compensation for the bias voltage. The current in D4 is adjusted with R4 to set the bias conditions for TR3.

Alignment

Disconnect R6 from L2 and apply 12V to the relay end of R5 but not to the rest of the circuit. Check the voltage across D4; varying R4 should allow the voltage to be adjusted smoothly over a range which includes 0.6–0.8V. This test is important as too high a voltage from a faulty bias circuit will destroy the RF transistor instantly. If the voltage is correct, reconnect R6 to L2. Set all variable capacitors to 50%, set R4 to minimum

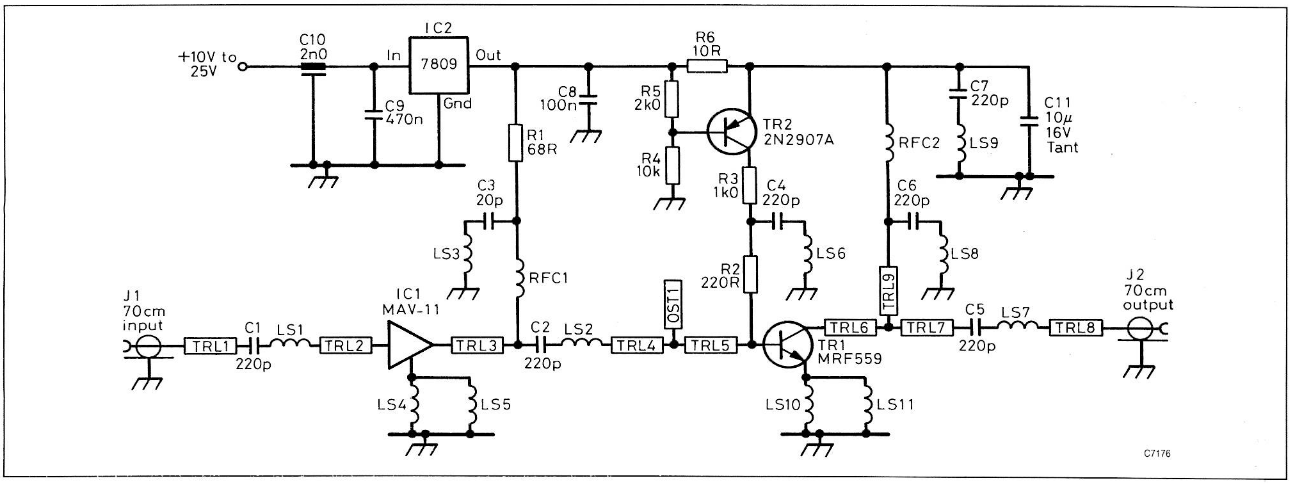

Fig 4.138. Circuit diagram shows the microstrip tuning elements used in the computer analysis. For component layout and PCB etching pattern see Figs 4.139 and 4.140 (Appendix 1) *(QEX)*

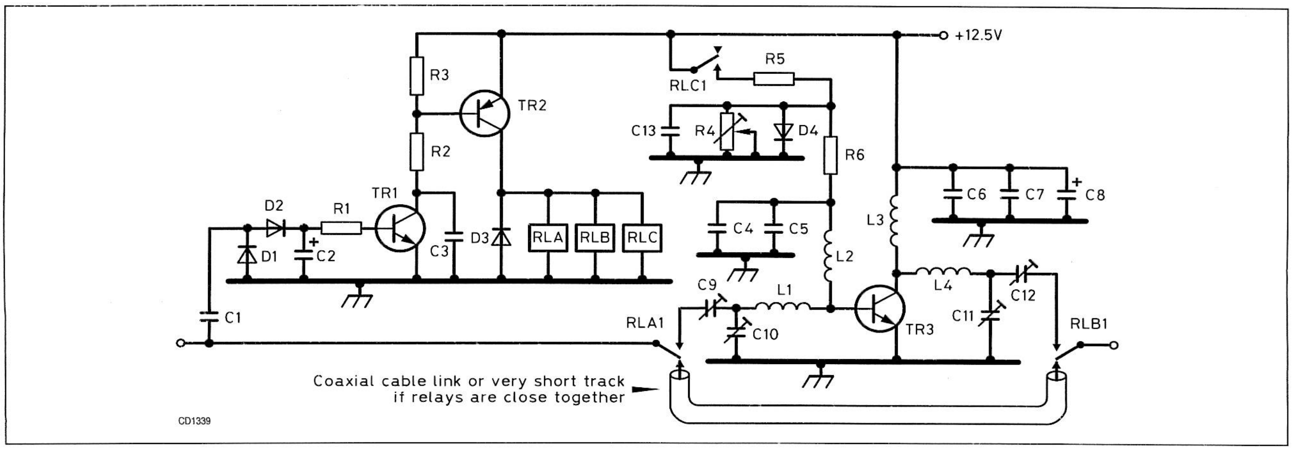

Fig 4.141. 144MHz 25W amplifier. For component values see Table 4.12

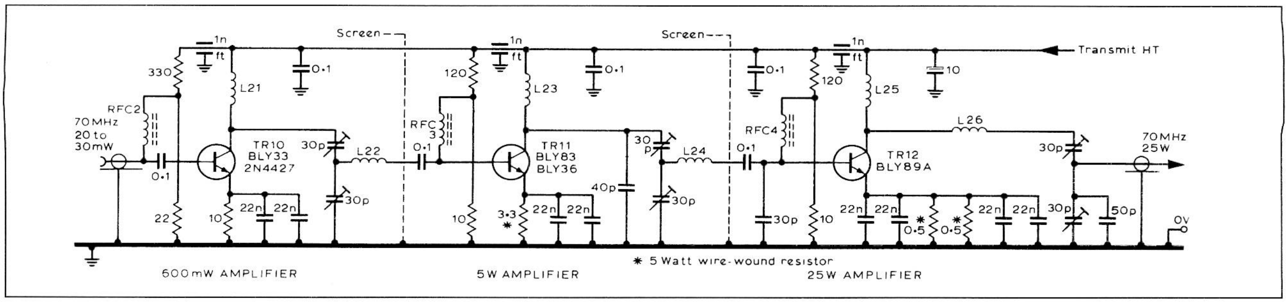

Fig 4.142. 20/30mW to 25W transmit amplifier

Table 4.12. Components list for Fig 4.141

R1	22k
R2, R3	1k
R4	200R pot
R5	39R 5W
R6	1R
C1	3p3
C2	4µ7
C3, C4, C6	1n
C5, C7, C13	100n
C8	47µ
C9–12	Philips C808 60p (yellow)
D1–3	1N4148
D4	1N4007
TR1	BC109/BC547
TR2	BC182/BC557
TR3	B25-12, TP2320, MRF222 etc
RLA, RLB, RLC	Miniature 12V relay
L1	2t 1.6mm enamelled copper wire, 7mm ID
L2	As L1 but 3t
L3	0.5mm enam copper wire threaded through 6-hole ferrite bead
L4	6t 0.5mm enam copper wire, 4mm ID

resistance and connect 12V. The current should be low; a few milliamps maximum. Connect TR1 collector to earth so that the relays operate; the current should increase to about 300mA. Note this current and adjust R4 for an increase of about 100mA. Remove power and the earth at TR1 collector. Restore 12V and apply about 0.5W RF drive; the relays should operate cleanly. Adjust the capacitors for maximum output power, about 4–8W, then increase the RF input to the full level of 2–3W. At this power level adjust the output capacitors for maximum output power and the input capacitors for minimum input VSWR. The output power and supply current should vary smoothly as the circuit is tuned; sudden jumps indicate instability.

This basic procedure can be applied to all amplifiers; check the bias, check DC conditions without RF, tune up at low power, and finally tune at high power.

C2 provides a delay before the amplifier reverts to receive so that the relays do not change over during every pause in speech when using SSB. For FM use this capacitor can be reduced to 0.01µF for rapid switching; for dual-mode use the larger capacitor can be switched in and out of circuit.

Amplifiers for 70MHz

These designs are taken from a transverter design by G3XBY and G3WOS [43]. The lower-power part (Fig 4.142) uses three cascaded stages to amplify from about 20mW to 25W. Bias in all the stages is set using emitter resistors which avoids the need for alignment and provides good thermal stability and improved linearity. The emitter decoupling capacitors should be mounted close to the transistor body with minimum lead length.

The 100W amplifier (Figs 4.143 and 4.144) uses two 2N6084 or BLY89 in parallel. The circuit uses an operational amplifier and emitter followers to give a very-low-impedance bias voltage supply. IC2 provides a negative supply voltage for the operational amplifier as the type used in the original (741) has input and output voltage ranges which cannot go close enough to ground without a negative supply voltage. Some later operational amplifiers can function in this application without the negative supply, but operation should be tested carefully before connecting to the RF transistors.

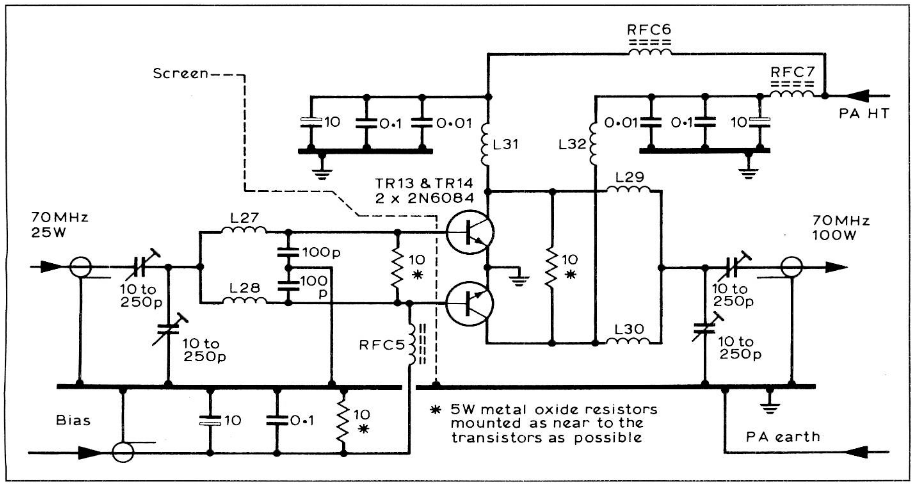

Fig 4.143. 100W transmit amplifier

50MHz 25W amplifier

The 25W 70MHz amplifier was modified by GW3XYW for use on 50MHz [44]. The 2N6080 and

Table 4.13. Components for 70MHz amplifiers

L21	6t 18 SWG 3/8in ID, length 5/8in
L22	5½t 18 SWG, 3/8in ID, length ½in
L23	7t 18 SWG, 3/8in ID, length ¾in
L24	3t 18SWG, 3/8in ID, length 3/8in
L25	7t 16 SWG, 3/8in ID, close wound
L26	6t 16 SWG, 3/8in ID, close wound
L27, L28	2t 16 SWG, ½in ID, ¼in leads
L29, L30	1t 16 SWG, ½in ID, ¼in leads
L31, L32	5t 16 SWG, ½in ID, close wound
RFC2–5, 10	6t on Mullard FX1898 6-hole ferrite bead
RFC 6–8	10t 18 SWG on 5/8in diam toroid
RFC9	1.5mH choke

2N6082 transistors chosen by GW3XYW can also be used in the 70MHz design and might prove easier to obtain than the Philips types in the original. The 100W amplifier could likewise be modified for 50MHz use.

Figs 4.145 and 4.146 show the circuit and layout. Construction uses single-sided PCB with Veropins in tight-fitting holes to give anchorage points for the components. Where the connection has to be insulated from the ground plane, a small circle of copper is removed with a drill bit before the pin is fitted. Some components are fitted underneath the PCB so the power transistors have to be mounted to the heatsink on aluminium or copper spacers to provide clearance. These spacers are important in maintaining proper cooling for the power transistors and so should be as wide as possible with very flat contact surfaces. C126 and C127 are shown as 30pF air-spaced trimmers; 60pF rotary foil types should work perfectly well as an alternative.

144MHz 10W amplifier

A similar amplifier for 144MHz [45] is shown in Fig 4.147. The output power is about 10W for 10mW input and an attenuator at the input can be used to adjust the input level if greater drive is available. For drive levels of 100mW or more, the first amplifier stage can be omitted and the input connected to the alternative input point as shown.

144MHz 100W amplifier

This amplifier was designed as the driver stage of a 400W 144MHz amplifier [46]. In that application the circuit

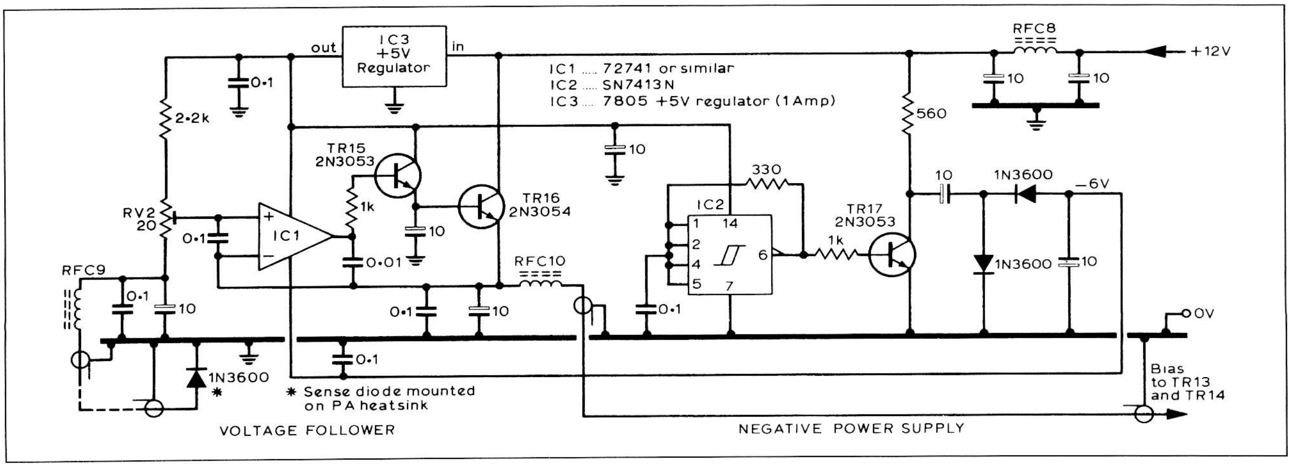

Fig 4.144. PA bias generator

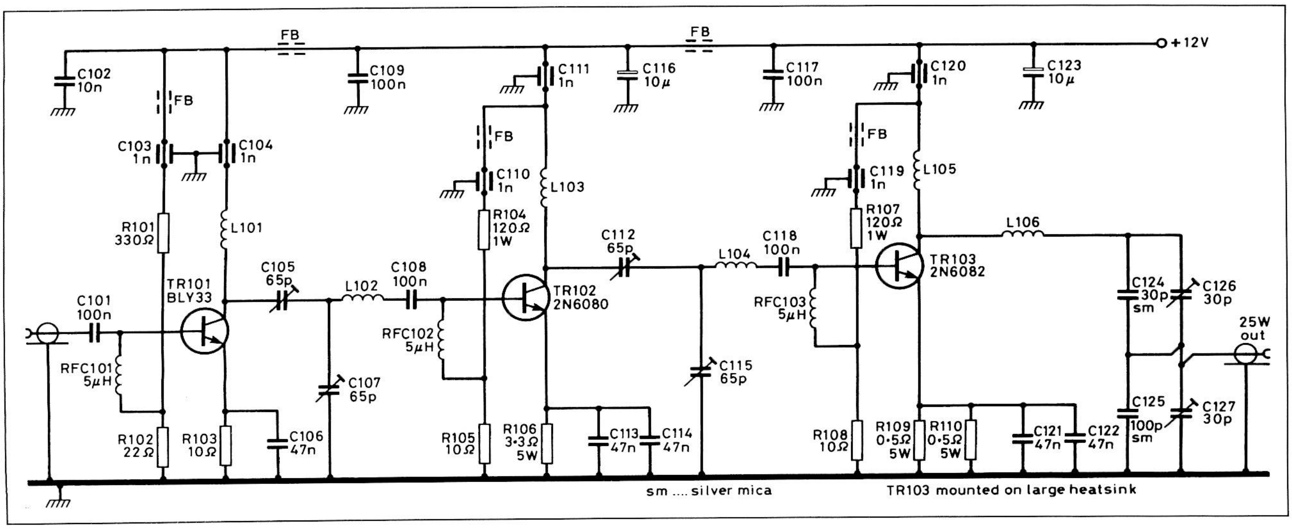

Fig 4.145. Linear amplifier circuit diagram