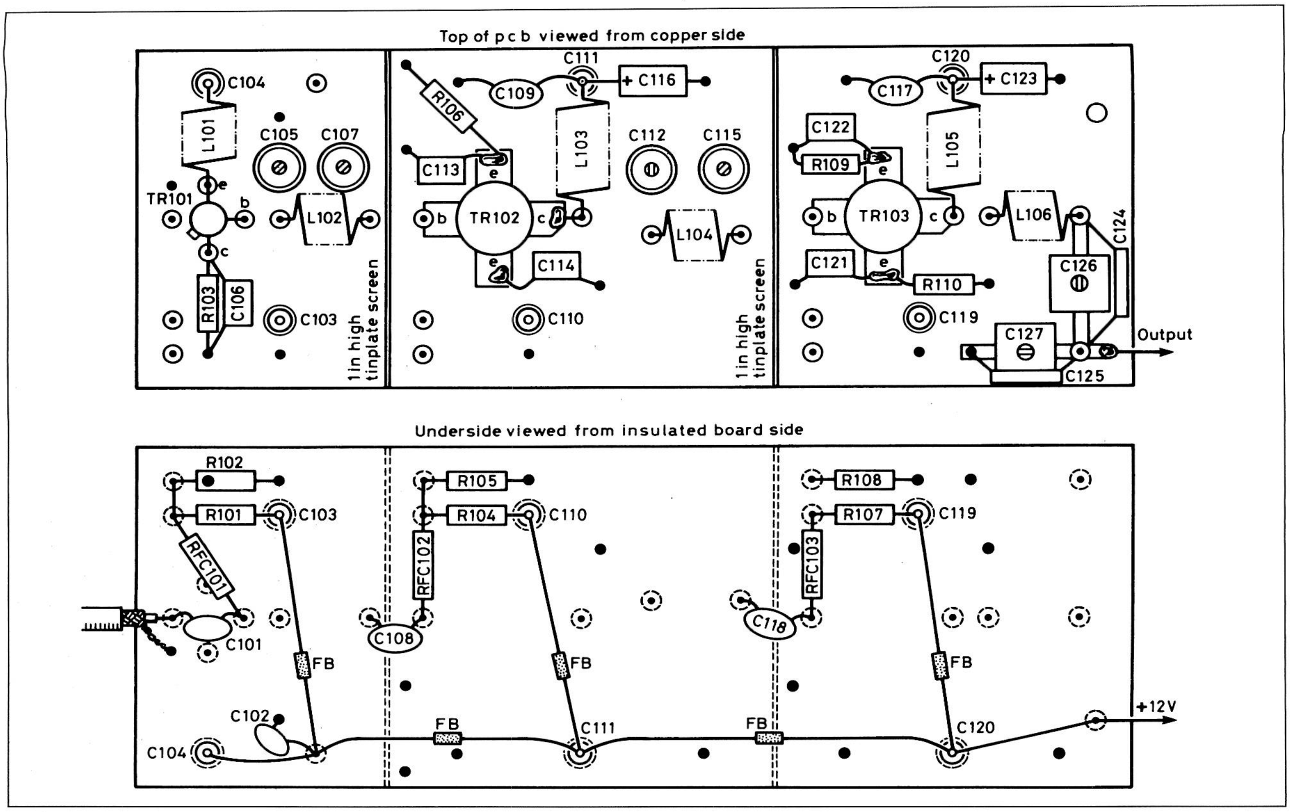

Fig 4.146. Linear amplifier component layout

operated at 75W output with very good linearity so that overall linearity was dominated by the output stages. Used alone, linearity will be acceptable at 100–120W output. The amplifier (Fig 4.148) uses a pair of BLW86 in push pull. The load resistance required is about 6.5Ω per device or 13Ω side to side. This is conveniently achieved by using a λ/4 transformer of 25Ω cable (2 × 50Ω in parallel) which converts from 50 to 12.5Ω. The cable also provides the balun function for push-pull operation. C8 optimises the load impedance for best linearity.

The 270pF capacitors cancel each transistor's input inductance to leave a resistance of 2Ω. A T-section network with a *Q* of 4 matches this to 6.25Ω and a λ/4 25Ω coaxial transformer matches this to 50Ω.

The bias circuit uses TR3 and TR4. TR3 base-emitter junction provides thermal compensation for the RF power transistors and TR4 supplies the output current. The two transistors also act as a voltage regulator, holding TR3 emitter voltage constant with varying bias current.

70MHz 200W FM amplifier

This amplifier (Fig 4.149) by G0MRF [47] uses a pair of transistors in parallel. These transistors are specified for use at 30MHz, 100W output but they still provide useful performance at higher frequencies and give excellent value for money. Unfortunately, not all transistors specified for HF use will work well at higher frequencies.

Clearly, this amplifier can be readily modified for 50MHz use, and bias added for linear operation (see Fig 4.106 for recommended decoupling arrangements if bias is added).

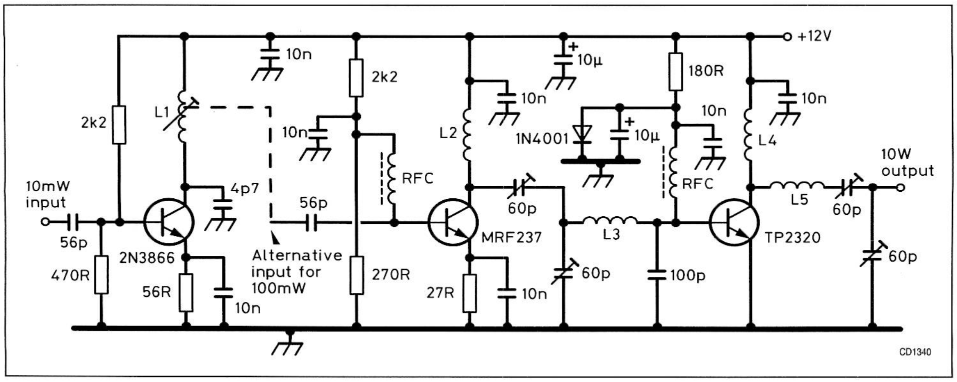

Fig 4.147. 144MHz 10W amplifier. All fixed capacitors are miniature ceramic. All trimmers are 60pF rotary, eg Philips C808 or similar. L1: Toko S18, 4½t, add tap at 2¼t. L2–L5 are 1.6mm enam Cu wire, 6.5mm ID. L2: 4t. L3: 2t. L4: 4t. L5: 3½t *(Radio and Electronics World)*

432MHz 4W FM amplifier

This simple amplifier (Figs 4.150 and also 4.151 in

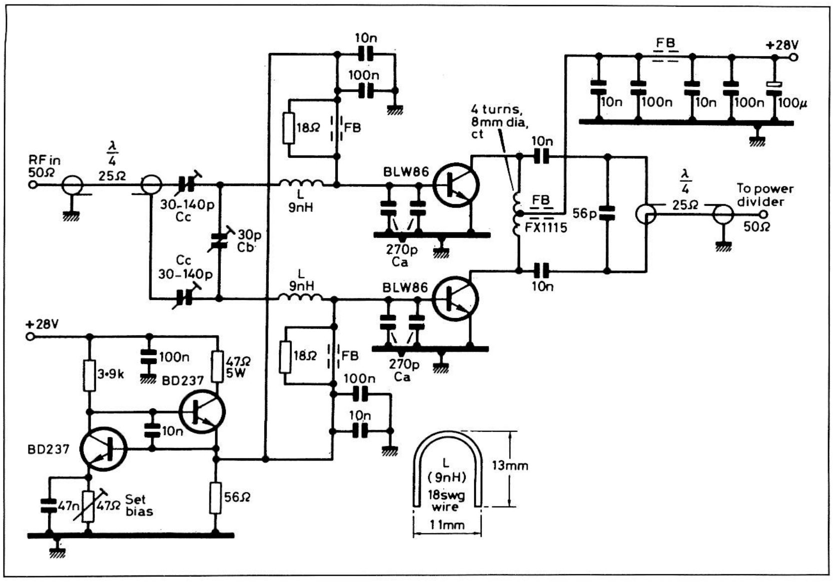

Fig 4.148. 75W drive amplifier circuit. Ca, Cb, Cc are mica compression types. Unmarked FB are 3S2 (blue). Coaxial transformers are two parallel 350mm lengths of UR110. 56p compensation capacitor is glass or ATC porcelain type. All other capacitors are chip ceramic. Quiescent current is set for 160mA total

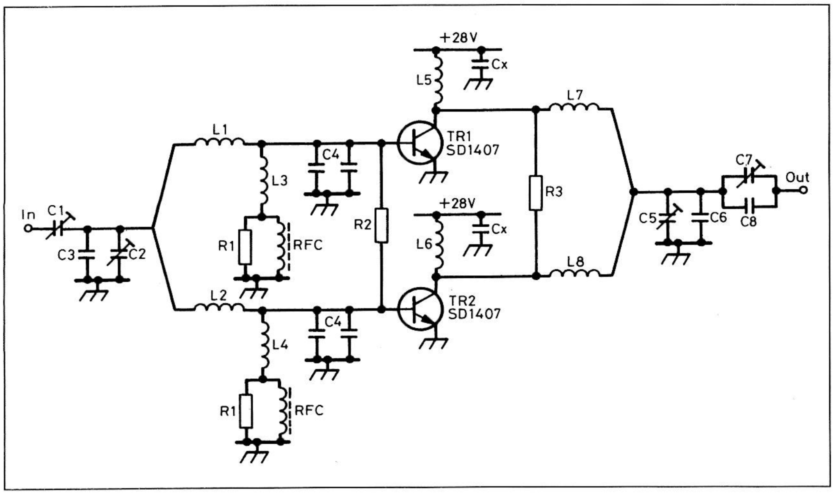

Fig 4.149. G0MRF's 200W FM/CW amplifier for 70MHz. Components are given in Table 4.14

Table 4.14. Components list for Fig 4.149

R1	10R 0.5W
R2	47R 1W
R3	100R 2W (R1–3 carbon film)
C1	6–65p film trimmer
C2, 5, 7	10–100p high-voltage rotary type with PTFE film dielectric or mica compression
C3, 6, 8	47p silver mica, 350V
C4	4 ×100p (each base has 400pF to ground)
Cx	100μ + 0.01μ + 1n in parallel
L1, 2	2t 18SWG, 10mm ID
L3, 4	8t 22SWG, 2.5mm ID
L5, 6	6t 18SWG, 6.5mm ID
L7, 8	2t 16SWG, 10mm ID
RFC	2t 22SWG enam Cu through FX1115 ferrite bead
TR1, 2	SD1407

Appendix 1) [48] can be a useful addition to a low-power exciter or, with the addition of bias, an output stage to follow the circuit of Fig 4.138 in a transverter. The circuit diagram is similar to the other amplifiers, but the matching inductors have been replaced by printed tracks; hand-wound components tend to be too imprecise at higher frequencies. This amplifier is built on double sided 1.6mm glassfibre/epoxy PCB; for higher powers the loss can be excessive and glassfibre/PTFE is often used. At 0.4W input, output power is 3W at 12V and 4W at 14V supply. Collector efficiency is about 50%.

Broad-band 300W FET amplifier

With the addition of appropriate harmonic filters this amplifier covers all bands from 20m to 2m with up to 300W (about 200W linear) output, 15dB (30×) gain and without any tuning. Figs 4.152, 4.153 and Table 4.15 show the circuit and constructional detail. The design techniques are radically different to those used in the other amplifiers described here and are described in considerable detail in the original article in *Microwaves and RF* [49]. Reference to the full article is essential for anyone considering building this amplifier. The amplifier is available in kit form [50] and offers an interesting and maybe cost-effective alternative to a series of individual amplifiers for each band. Basically, a 9:1 impedance transformer is used to match the transistor input impedance to 50Ω and a 4:1 transformer gives a load impedance of about 12.5Ω side-side, or 6.25Ω for each transistor. Two identical transistors are mounted side by side on a single flange so that the push-pull circuit is readily constructed and fully balanced. The transformers use a novel construction to give broad-band VHF performance. The windings are made from pieces of semi-rigid coaxial cable which have a solid copper outer conductor. The outers are soldered together to make a single low-impedance turn and the inners are connected in series to give a turns ratio of 2:1 (input) or 3:1 (output). 'E' and 'I' cross-section ferrite cores fit around the cables to extend the low-frequency response (which is about 70MHz without the ferrite). R8 and R9 introduce feedback which stabilises impedances and controls the gain at lower frequencies. For prolonged use at higher frequencies the ferrite can get very hot and fan cooling of the upper surface of the circuit might be needed.

AMPLIFIERS USING VALVES

The 4CX250 family of tetrode valves has been the most popular choice by far for high power amplifiers for some time, at

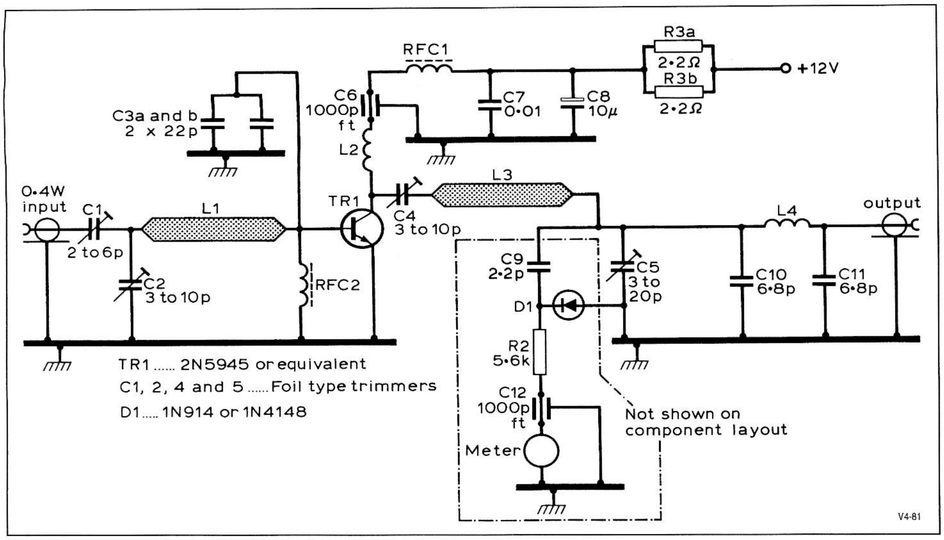

Fig 4.150. 432MHz power amplifier giving 3–4W output. L1: stripline. L2: 7t 6mm dia 22SWG wound. L3: stripline. L4 1t 7mm dia 18SWG. RFC1: 2t 22SWG ferrite bead. RFC2: 3t 20SWG large ferrite bead

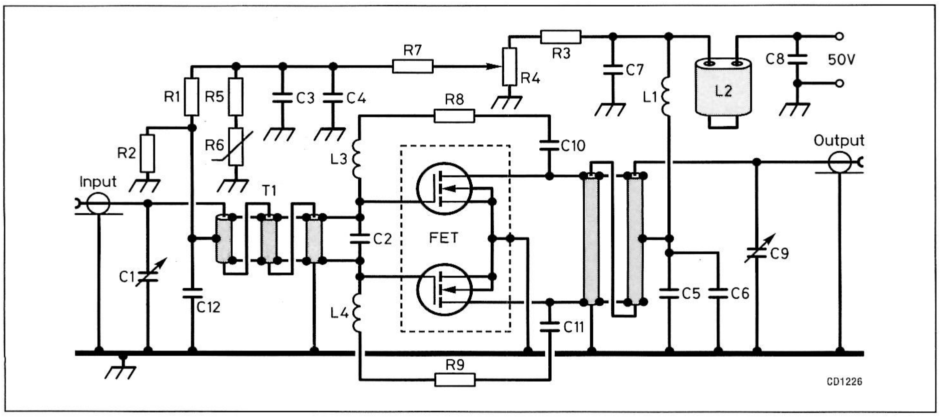

Fig 4.152. At the heart of the 300W amplifier is the Gemini push-pull transistor configuration. This broad-band amplifier operates in the 10 to 175MHz range *(Microwaves & RF)*

least partially because of ready availability of valves and bases on the surplus market. Recently alternative triodes have become more available at comparable prices.

Few amplifiers are built as exact replicas of published

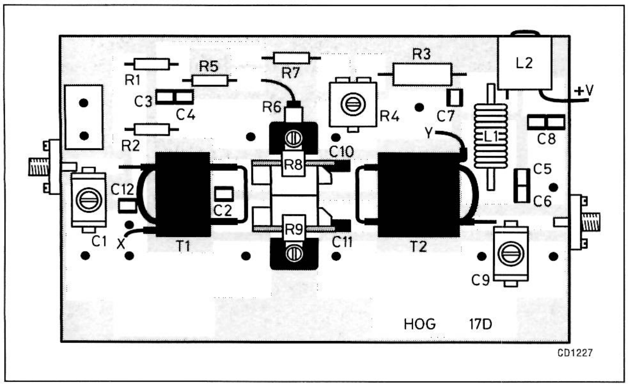

Fig 4.153. Component layout of the 300W amplifier *(Microwaves & RF)*

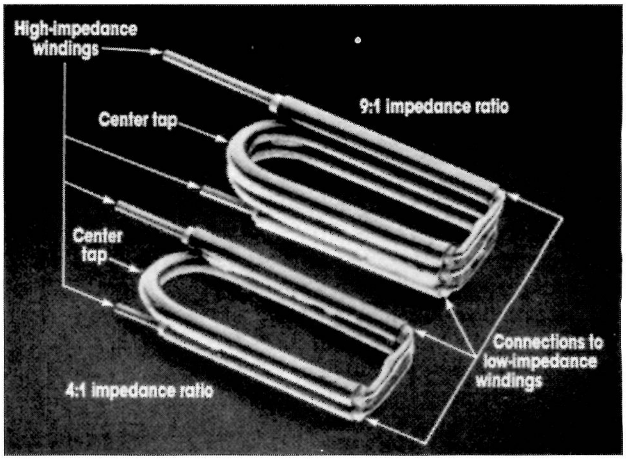

Fig 4.154. Construction of the wide-band transformer *(Microwaves & RF)*

designs; there is usually some variation from the use of materials to hand. The examples chosen here are intended to demonstrate a variety of different techniques, as well as providing designs for duplication.

Cooling

Valves are like almost every other electronic component; if they overheat they die young.

Cooling power valves effectively has long been a neglected part of amateur amplifier building. Most fans are designed to move air against little or no obstruction and cannot force air to flow through the restriction of a valve base and anode. See Fig 4.155 which shows the suggested airflow arrangements, with cold air cooling the valve base, heater pins and grid spigot before going through the anode. To cool a 4CX250 at full power (250W dissipation), a minimum airflow of 6.4cfm (3l/s) is required [51] and this requires pressure in the grid compartment. This pressure, usually called the *back pressure*, is normally defined in terms of *water gauge*, that is the displacement seen in a water manometer connected to the pressurised area. The pressures required for the 4CX250 equate to 0.82in (21mm) WG. Allowing for screening mesh over the apertures, and airflow within the cabinet, the blower needs to

Table 4.15. Components list for the broad-band amplifier

R1	1k, 0.5W
R2	1k5, 0.5W
R3	1k5, 2W
R4	1k trimpot
R5	6k8–8k2, 0.25W (depends on FET G_{fs}
R6	Thermistor, 10k @ 25°C, 2k5 @ 75°C
R7	2k, 0.5W
R8, R9	50R power resitor, EMC Technology Type 5310 or KDI Pyrofilm type PPR 515-20-3
C1, C9	8–60p, ARCO 404 or equiv
C2	130p ceramic chip
C3, C10, C11	100n cermic chip
C4, C5, C12	1000p ceramic chip
C6, C7	5000p ceramic chip
C8	470n ceramic chip, or lower values in paralle to match the value indicated
L1	10t 16 AWG enam, 5mm ID
L2	Ferrite beads, 1.5μH total
L3, L4	Lead lengths of R8 and R9, 20mm total
T1	9:1 RF transformer, 25Ω, 0.062in OD, semi-rigid coaxial cable
T2	1:4 RF transformer, 25Ω, 0.090in OD, semi-rigid coaxial cable
TR1	MRF151G

Notes: For T1, two type 75-26 E and I Micrometals powdered iron cores are required. For T2, three type 100-8 E and I Micrometals powdered iron cores are required.

All chip capacitors of 5000p or less are ATC type 100 or equiv.

be able to work at least 1in (25mm) WG. Some fans/blowers are shown in Fig 4.157. The 120mm axial fan cannot generate this pressure, regardless of airflow, and is completely

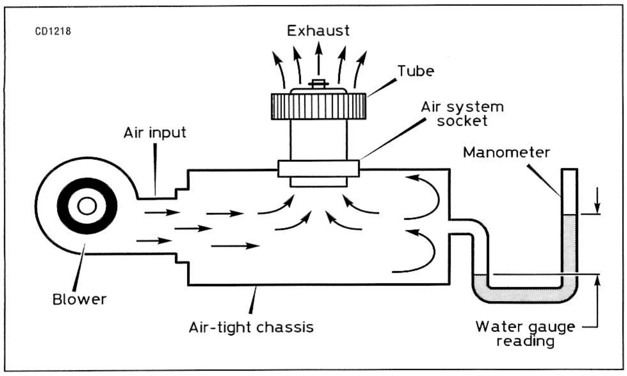

Fig 4.155. Grid blown cooling

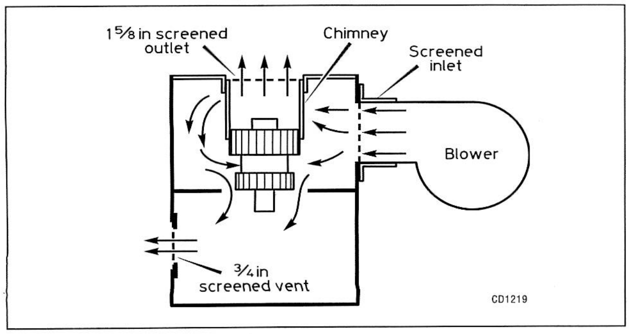

Fig 4.156. Anode blown cooling

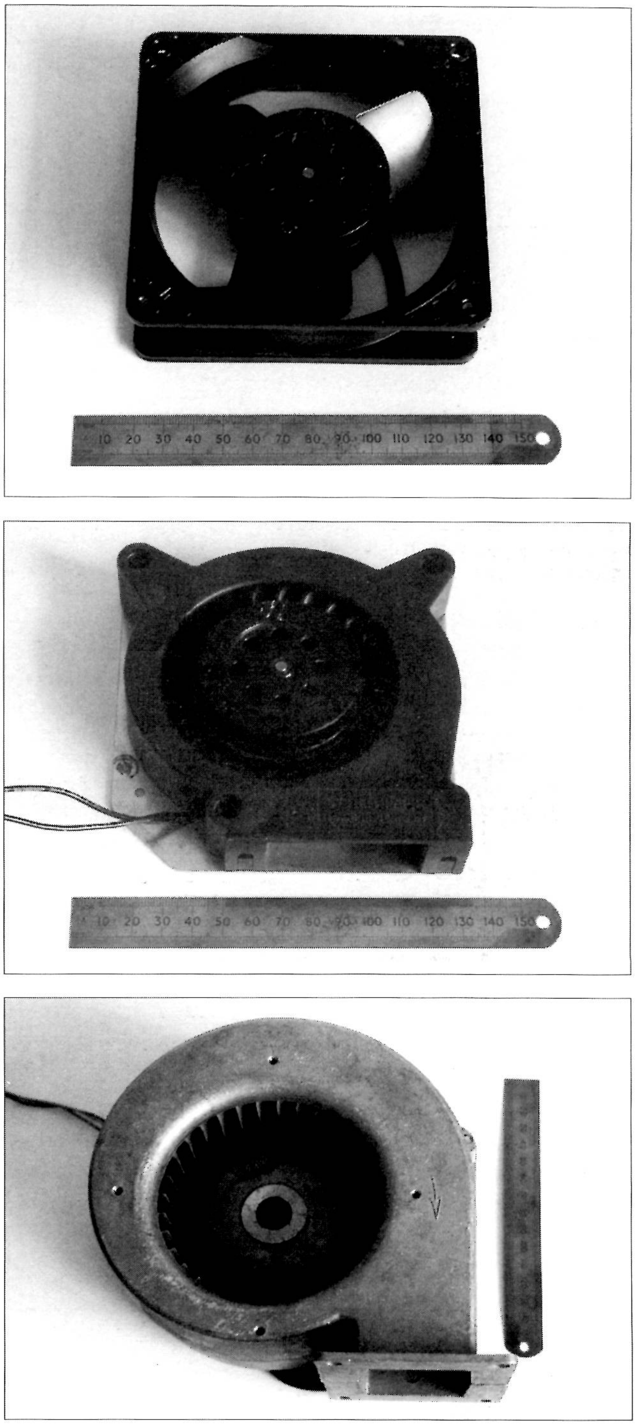

Fig 4.157. Top: An 120mm axial fan. Middle: A small blower, the Papst RG90/18-50. Bottom: EBM G2E 1120 blower

unsuitable. The RG90/18-50 will produce slight airflow, sufficient for about 100W anode dissipation. TheEBM G2E 1120 is well suited to the job, capable of 1.3in (33mm) WG and over 80cfm (170l/s) at 1in. Any blower capable of delivering adequate back pressure will almost invariably meet the volume requirement. See Table 4.16 for some suggested blowers.

Table 4.16. Blowers suitable for cooling external anode valves

Type	Pressure	Price	Supplier
Papst RG160-28/56S	30mm	£59	Farnell Electronic Components 0113 263 6311 order code 474400
Papst RG160-28/56S	30mm	£61	Electromail 01536 204555 order code 813569
Ziehl-EBM G2E 108-AA01-50 *	20mm	£92	Electromail 01536 204555 order code 223089
Ziehl-EBM G2E 120-AA12-50	33mm	£114	Electromail 01536 204555 order code 581262
ACI VBL5/3	28mm	£104	Air control Installations 01460 67171
ACI VBM5	30mm	£114	Air Control Installations 01460 67171
Etri 620CAZ016DC13	30mm	£180	Fan Technology 01403 275131

* Suitable for use as in Fig 4.156 only.

Cooling should be applied before the heaters are powered and ideally continued for some minutes after the heaters are switched off.

Circuit configurations

At VHF/UHF triodes are always used in grounded grid configuration with RF applied to the cathode (or heater if directly heated). For convenience, where 'cathode' is mentioned in the text, the alternative in brackets should be taken as read. If RF input is applied to the grid, the effect of the feedback capacitance from anode to grid has to be negated in order to prevent oscillation. This can be done by providing an antiphase signal (neutralisation) from anode to grid to cancel the feedback, or by reducing the resistance at the grid to a level where the circuit is stable. The latter results in low power gain and is frequency dependent. Neutralisation is used in small-signal triode amplifiers, but is impractical for the amateur in high-power circuits.

In the grounded-grid circuit, the grid forms a screen between the input at the cathode and the output at the anode, reducing the feedback to a small level which permits stable operation. The input impedance at the cathode is lower than when driving the grid in grounded cathode, so the circuit power gain is lower than offered by tetrodes such as 4CX250, but the power supply requirements are somewhat simpler. The circuits in grounded grid amplifiers can appear slightly confusing, and Fig 4.158 shows how the conventional grounded cathode circuit is developed to work with grounded grid connection, omitting the input and output matching circuits. Fig 4.158(a) shows a conventional circuit with RF input applied between the grid and cathode. M1 reads the grid current and M2 the anode current. M2′ also reads the anode current, but in a much safer position where the voltage is close to ground.

In Fig 4.158(b) D1 holds the cathode positive with respect to the grid, equivalent to holding the grid negative with respect to a grounded cathode. C1 bypasses D1 for RF signals, so the RF input is applied between grid and cathode. Valves intended for grounded-grid operation are designed to need low bias voltage, typically in the range 0–12V, so D1 is a readily available power zener diode.

In Fig 4.158(c), (b) is rearranged so that the grid connection is grounded. The RFC moves so as to isolate the cathode from D1 and C1 in (b) is not needed. RF input is still between the grid and cathode. S1 is added to show transmit/receive switching. On receive, S1 is open-circuit, stopping all current flow. In practice, R1 is left in circuit to prevent the cathode floating.

In a tetrode, the problem of anode-to-grid feedback capacitance is overcome by placing an additional grid between them. This is grounded to RF signals and screens the anode from the grid, hence the name *screen grid.*

The low level of feedback allows the tetrode to be used in grounded-cathode configuration, with the RF input signal applied to the control grid. The higher impedance at the control grid means that a lower power is needed to provide the necessary grid-to-cathode voltage swing, so the power gain of the circuit is higher than for grounded grid.

Fig 4.159 shows the normal circuit arrangement. g1 bias is typically −40 to −60V for quiescent bias when transmitting and about −100V to shut the valve off when receiving. The screen grid bias is in the regions of 350V (TX) and 0V (RX). The 4CX250 family has screen grid characteristics which can result in both positive and negative screen grid current (the screen grid can source as well as draw current) under normal operating conditions as the output power varies. The screen grid supply must be able to maintain a constant voltage under all conditions, and so must be able to sink (absorb) as well as source (supply) current.

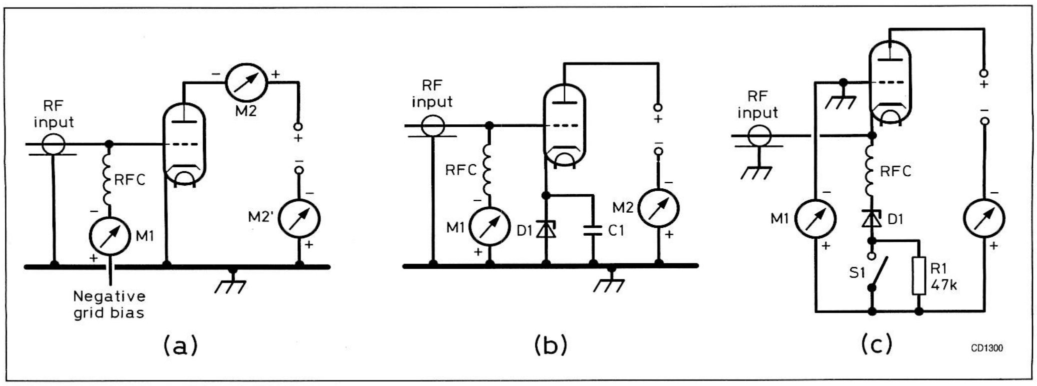

Fig 4.158. The evolution of a grounded-grid amplifier. (a) Conventional grounded-cathode circuit. (b) Cathode held positive with respect to the grid. (c) Rearrangement of (b) so the grid is grounded directly

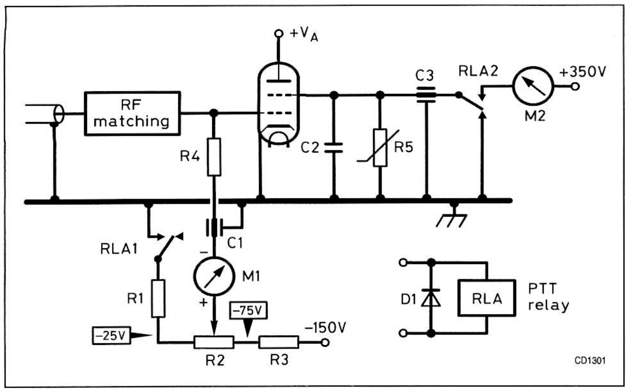

Fig 4.159. Typical tetrode amplifier circuit

Tetrode gain characteristics are dependent to a large degree on the screen grid voltage. Gain increases as the screen grid voltage is increased, but linearity is better with lower voltage. A value in the region of 325–350V gives an optimum balance for general use.

The control grid supply can be very simple, assuming that no grid current is allowed in normal operation. This will always be the case for linear use with SSB signals. If grid current is permitted, higher efficiency (at the expense of gain) can be achieved with increased RF input and different anode loading for CW and FM signals. This case will not be considered here as, under all normal circumstances, an amplifier set up for linear operation will give satisfactory operation with all types of signal.

On receive, S1 is open and the full negative voltage is applied to the grid via the various resistors. On transmit, R2 adjusts the voltage applied to the grid to allow the desired quiescent voltage to be set. In practice, if grid current is drawn, the change in grid voltage from the voltage drop in the resistors has little effect on operation. References [54] and [55] show more complex circuits which will hold the grid voltage constant when grid current is drawn.

R4 feeds the bias voltage to the control grid and also serves to load the grid circuit, effectively setting the gain. At high anode and screen grid voltages, a value of R4 higher than 2–3kΩ can give very high gain, leading to instability unless the amplifier is neutralised.

Output matching networks

An important factor in designing the matching network is the loaded *Q* (not to be confused with the *Q* of individual components). This is the ratio of reactance to resistance (see Fig 4.160(a)), and will vary from point to point in the matching network. The highest value at any point defines the value for the whole network.

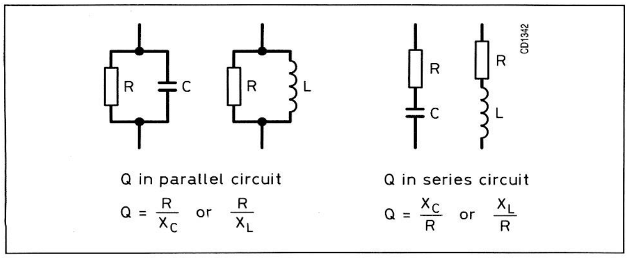

Fig 4.160(a). Q values in series and parallel circuits

Table 4.17. Components list for Fig 4.159

R1	2k2 1W
R2	5k 2W
R3	6k8 2W
R4	See text
R5	MOV V275 LA40B or equiv (275V RMS, 140J rating)
C1, 2	1000–5000p 500V feedthrough
C2	Screen grid capacitor built into valveholder
D1	1N4001
RLA	DPCO coil voltage to suit available supplies
M1	5mA
M2	10-0-10mA centre-zero

Typically, a loaded *Q* of around 10–15 is used when designing the output matching network. This is usually chosen as giving a balance between loss and harmonic rejection, with both factors increasing with increasing loaded *Q*. Given that every transmitter should have effective low-pass harmonic filtering before the antenna, values of *Q* below 10 can be used to give lower losses, provided that the component values remain reasonable.

Depending on the anode voltage and output power, the load resistance required at the anode is usually in the range 2000–5000Ω.

At lower frequencies a pi-network, as in Fig 4.160(b), is commonly used. At VHF the operating (or loaded) *Q* of the circuit tends to become high, leading to difficulties in adjustment and higher circuit losses. The minimum loaded *Q* is determined by the minimum value of C1 which is in turn defined by the valve output capacitance and circuit strays. It can be seen that the problem gets worse with increasing frequency.

Fig 4.161(a) shows the calculations for a 4CX250 at 50MHz and 144MHz.

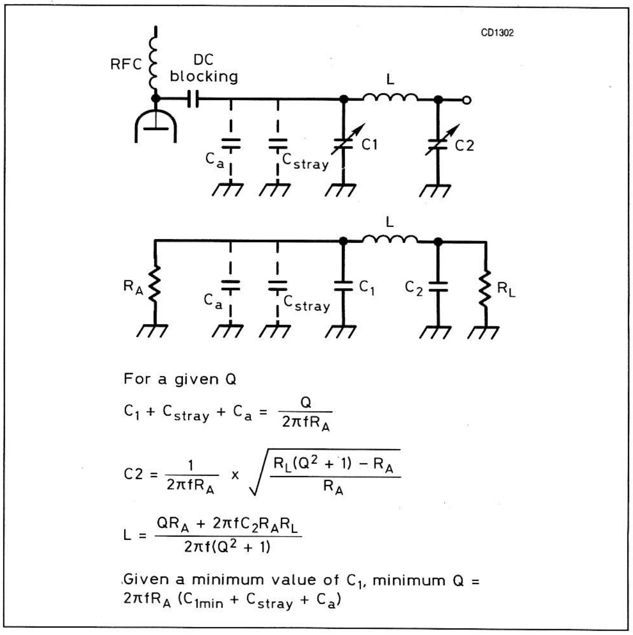

Fig 4.160(b). Pi network

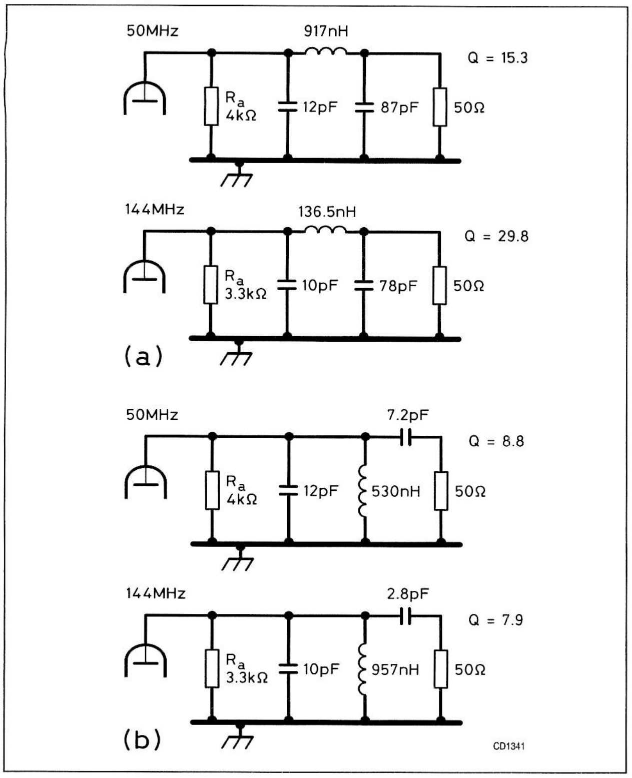

Fig 4.161. *Q* calculations

The usual solution is to incorporate the valve and associated capacitances into a tuned circuit with the valve anode at the high-impedance point. The resistance at the anode (R_a) is set by tapping the load (in this case the antenna) into the circuit at the appropriate point, as shown in Fig 4.162(a). In practice, the 'tap' can be implemented in several other ways, some of which are shown in Fig 4.162(b)–(d). Circuits (a)–(c) give monotonic performance; the anode resistance is lowered (heavier loading) as the tapping point moves towards the anode in (a), or C2 is increased in (b) or the output coil is moved closer to the anode coil in (c). K2RIW shows [56] how the circuit in (d) can give misleading indications whereby C2 appears to have a definite tuning point, but the desired loading is not achieved. In this case, it is necessary to adjust the coupling between the anode circuit and the output coupling link until C2 can be adjusted to give the correct loading. This can be seen when older designs, optimised for maximum CW output, are used in linear service with SSB signals; the loading is often too light (anode resistance too high) for good linearity.

Using the previous example, the tuned circuit configuration allows a lower loaded Q; Fig 4.161(b) shows that a value of less than 10 is easily achieved.

At higher frequencies the lumped components in the tuned circuits can be replaced by lengths of transmission line tuned to resonance with the valve capacitance and a small amount of tuning capacitance. At 144MHz the resonant circuit is usually $\lambda/4$, grounded at one end, while at higher frequencies $\lambda/2$ is used as $\lambda/4$ is impractically short when chosen to resonate with the valve output capacitance, as shown in Fig 4.163. Any of the coupling methods shown in Fig 4.162 can be used in conjunction with a transmission line anode circuit.

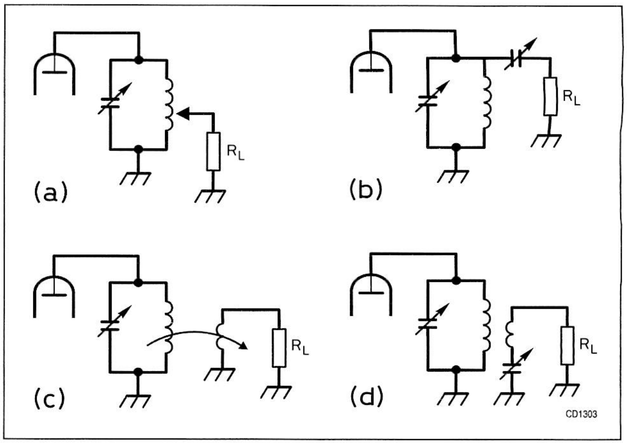

Fig 4.162. Various methods of tapping the load into the anode circuit

Valve bases

With tetrodes reliant on the screen grid RF grounding for stability, and with a bias voltage applied at the same time, special valve bases are needed. Fig 4.164 shows a 4CX250 tetrode; the screen grid is connected to one pin and also to the lower annular ring. Fig 4.165 shows a cross section of the contact spring finger and the capacitor which provides the RF grounding. The contact fingers introduce some series inductance, reducing the effectiveness of the grounding as the frequency increases. This can lead to instability, especially where high gain and maximum output power are sought.

Bases for HF use often have no built-in screen grid contact and capacitor, relying on decoupling through the pin connection. This will not work at VHF/UHF. All bases with built-in capacitors and contact fingers are likely to work without problems for 50, 70 and 144MHz, but the only reliable choice for UHF is the Eimac SK630 [57–59] which was specially designed with low-inductance screen grid contacts and additional screening to minimise feedback.

Grounded-grid circuits likewise need low-inductance

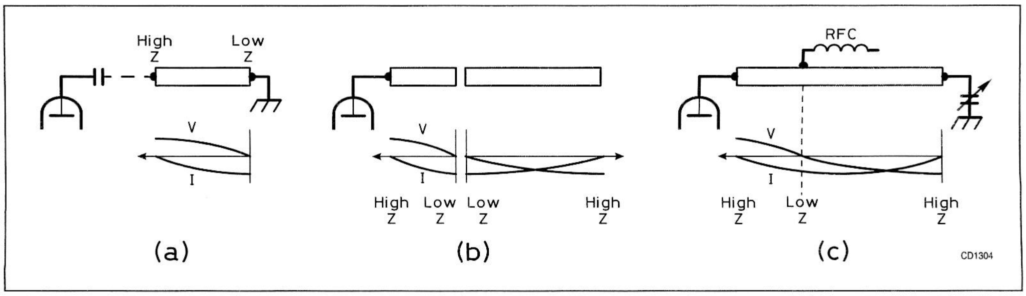

Fig 4.163. Using transmission line instead of lumped components

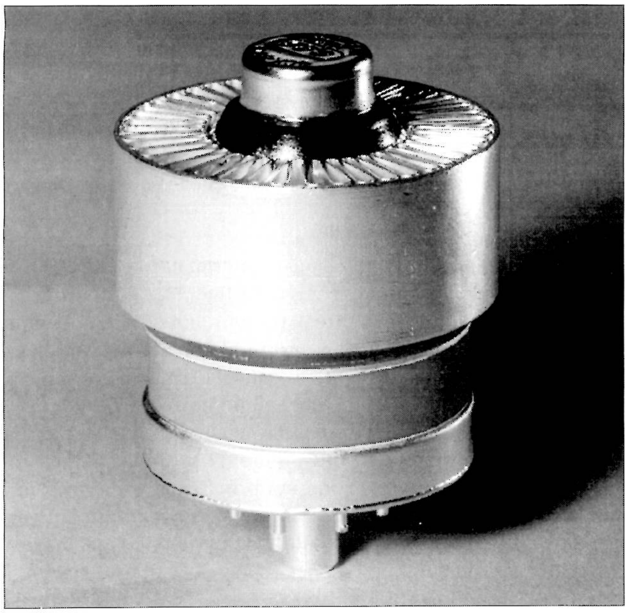

Fig 4.164. The 4CX250

grounding of the grid for correct operation. This is more easily achieved as there is no voltage to apply and so a direct connection to the chassis can be made.

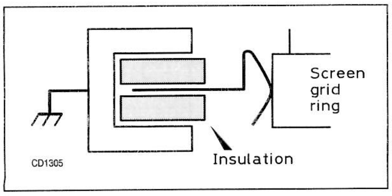

Fig 4.165. Cross-section through screen grid capacitor in valveholder

Power supplies

Never forget that high voltage WILL KILL YOU if you let it. Danger is not limited to anode supplies or mains; the grid bias supplies in a tetrode amplifier can be lethal too.

A good PSU will take at least as much effort as the amplifier it supplies. PSU problems are often behind disasters which are blamed on the RF section of the amplifier.

Anode supply

Depending on the valve(s) in use and the output power, the anode supply is likely to be in the region of 1–2.5kV at up to 0.5A. Anyone unfamiliar with working on such supplies should consider that this is equivalent to 12V at 100A, only MUCH more dangerous. A 12V 100A PSU is a major undertaking, and so is the high-voltage equivalent, albeit with different problems.

Large power supplies tend to be built around items which become available rather than from a list of catalogue items. This applies to transformers in particular, and other components in general.

The most important job before working on any power supply or amplifier is to make an insulated probe for earthing anything and everything to be worked on, even if there is apparently no chance of any voltage being present. The probe need be nothing more than a plastic rod with a bare wire hook at one end and a stout insulated wire which runs from the hook to the equipment chassis. The wire must be bolted or soldered in place at both ends; spring clips can drop off too easily. The hook allows the probe to hung onto circuits for extra security while working on them, but don't forget to unhook it before switching on again! Fig 4.166 shows an another design by G8DPS which both discharges and indicates the presence of voltage above about 200V.

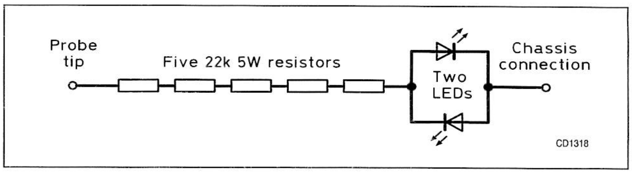

Fig 4.166. Discharge and indicator probe

Fig 4.167 and Table 4.18 show a typical anode voltage supply taken from reference [60] which includes a number of protection features which will help to avoid permanent damage in the event of problems occurring. Echoing the sentiments of GW4FRX in the article, the mains transformer will probably be the item which defines the mechanics of the PSU. Should it have to be replaced the chances of finding another

Table 4.18. Components list for Fig 4.171

R1, 4	MOV 275V AC 140J
R2	2k2 2W
R3	470R 50W
R5–12	100k 2W
R13	68R 50W
R14	Select for M1 to read desired FSD
R15	1R0 25W for 0.5A FSD, 2R2 25W for 0.25A FSD
C1	10,000μ 25V
D1	1N5408
D2	Bridge rectifier, 20 ×1N5408 and 20 ×680k 2W
D3	50V, 60A diode or 35A bridge rectifier
RLA	24V, 300-400Ω coil, 16A min contacts
FS1	5A mains fuse
FS2, 3	1A high rupture current (HRC) fuses, 1¼in ceramic body

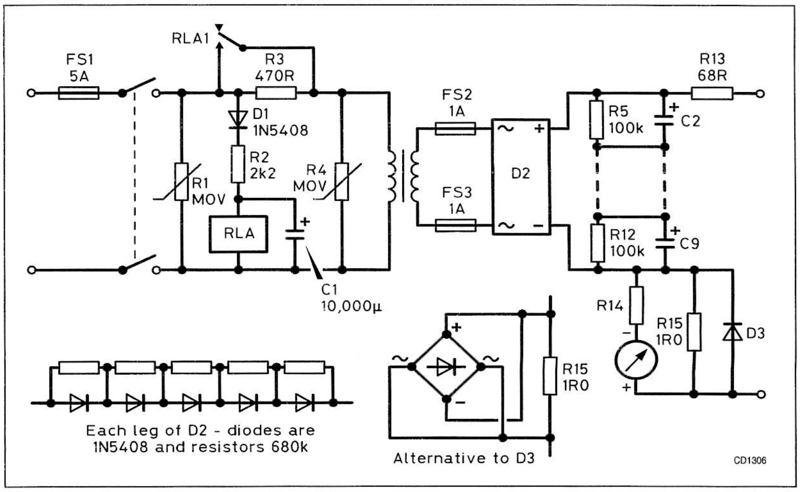

Fig 4.167. Typical anode voltage supply *(VHF/UHF DX Book)*

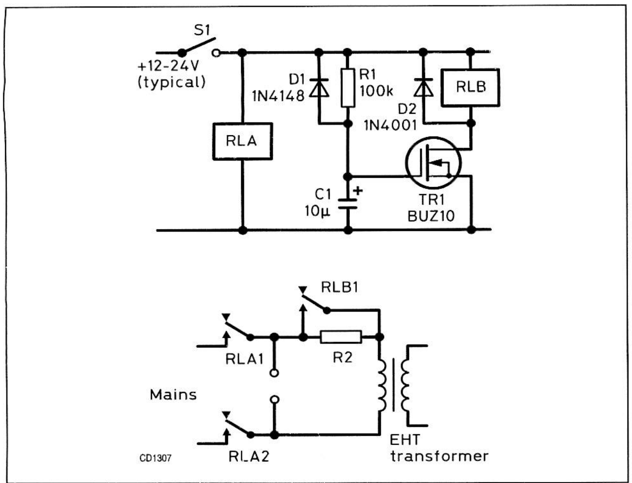

Fig 4.168. Alternative delay circuit. R1/C1 chosen to give delay of 1–5 seconds. S1: EHT on/off. RLA: DPNO 16A coil to suit DC supply. RLB: SPNO 16A coil to suit DC supply. TR1: N-channel power FET (eg BUZ10)

which is electrically and mechanically compatible is remote, unless it is a brand-new item, in which case it will have been very expensive. Either way, it is worth going to some lengths to ensure that the transformer is protected under all circumstances.

The input filter (not shown) and R1 remove noise and spikes on the mains supply, both of which can be significant when operating from generators. R1 is a metal oxide varistor (MOV) which behaves something like a zener diode; it has a high resistance up to threshold voltage, and then is able to conduct high current and absorb high transient energy, limiting the voltage peak. The surge limiting circuit around R3/RLA1 is much more gentle than most, limiting the primary current for about 10s rather than the few cycles allowed by most circuits. This allows the use of a lower value mains fuse than needed if surge currents have to catered for.

An alternative delay circuit is shown in Fig 4.168. This uses a low-voltage supply which is usually present in most PSUs.

An important point here is that the circuits are simply delay timers. Some designs use feedback from, for example, the anode supply voltage monitor to switch the resistor out of circuit. This is fine until a fault occurs which loads the secondary side excessively so that the switch does not operate; then the limiting resistor has full mains voltage across it permanently and can overheat or burn out. With a timed switch, the mains fuse will blow if there is a problem.

Both secondary connections to the rectifier should be fused. This protects the transformer against shorts in the rectifier or capacitors and the voltage/current will be zero at some point each cycle which allows any arcing to extinguish. Use only ceramic HRC (high rupture current) fuses and do not use normal panel-mount fuseholders.

Modern diodes have improved reverse breakdown characteristics and the voltage equalising resistors shown in older designs are not essential, although unlikely to cause harm. Capacitors across each diode should not be used; transient protection should be provided in the primary circuit.

Smoothing capacitors should be chosen to give a voltage rating of at least twice the output voltage; a total value of around 40µF will suit most applications. A string of resistors passing about 2mA provides a discharge path and equalises the voltage distribution between capacitors.

R13 limits the peak current in the event of a flashover, while dropping only 20–30V in normal operation. A fuse here will not provide adequate protection for either the PSU or the amplifier.

Current is measured in the return path so that metering is close to ground voltage. R15 must stay intact in the event of a flashover otherwise there is a risk of the chassis floating to +2kV. Choose a high-power resistor which will drop 0.5V at the desired meter FSD. A high-current diode across the resistor will carry the current in the event of a flashover, protecting the resistor and meter. A cheap and convenient solution is a high-current bridge wired as shown.

In addition to these measures, there must always be an independent heavy duty earthing strap between the amplifier and PSU chassis. Otherwise, if an invisible internal connection fails, you might only find out when you have an earthed connector in one hand and a chassis at +2kV in the other.

Heater supplies

For best life, heater voltages should be regulated to ±5% of rated value, measured at the valve pin as shown in Fig 4.169. The monitor point can be a small feedthrough capacitor or miniature socket on the grid compartment to allow the voltage to be measured without disturbing the airflow. The error introduced by the series resistor is minimal if a high resistance DVM is used for the measurement and no damage occurs if the test point is accidentally shorted to ground. Eimac have also recommended [61] tighter control of +0/–5%, with further reductions to 5.5V at frequencies above 300MHz to allow for cathode heating from 'back bombardment'. For domestic use with reliable mains supplies, 'select on test' resistors are a typical solution.

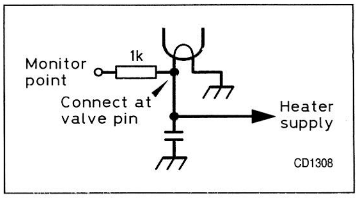

Fig 4.169. Monitoring the heater voltage

For conditions where the mains supply is not stable enough, a small Variac supplying the heater transformer or more complex regulated supplies, such as given in reference [62], will be needed. An alternative circuit is shown in Fig 4.170 [63]. TR1 acts as a variable resistance in series with the transformer primaries, diodes D1–4 steering the current correctly regardless of mains polarity. TR2–4 control TR1 so as to maintain a constant voltage at the transformer primaries, and thus a constant voltage to the heaters. WARNING – the circuitry is at mains potential and TR1 will need adequate cooling.

Cold heaters have a low resistance so initial current surge can be quite high. It is beneficial to arrange some form of current limiting or gradual increase in voltage. K1FO [64] recommends choosing a transformer with about 20% too much voltage and using primary and secondary resistors to drop the output by about 10% each, as shown in Fig 4.171.

The minimum warm-up times of 30 seconds for 4CX250 and three minutes for all other types should be strictly

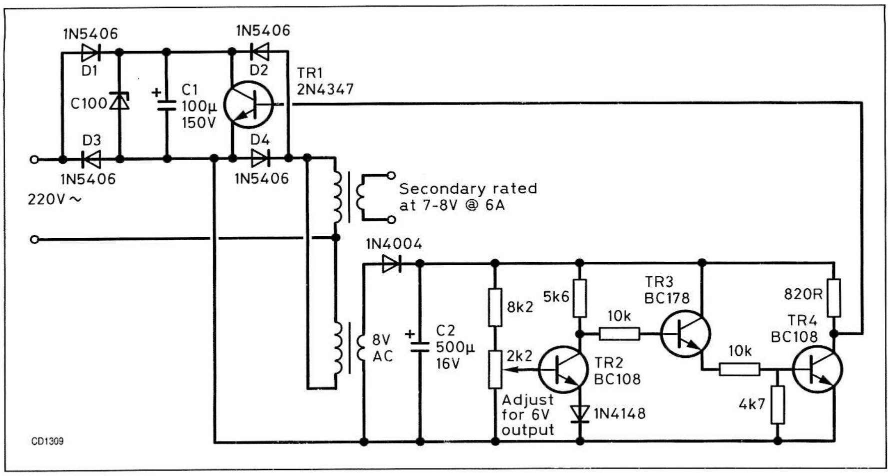

Fig 4.170. PA0LMD's AC regulator for stabilising the heater voltage applied to valves such as the 4CX250. Note that the electronics of the regulator are not isolated from the mains supply and suitable precautions should be taken

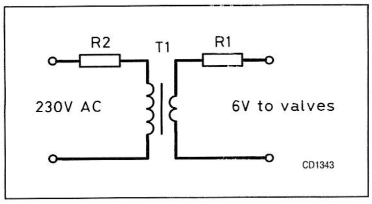

Fig 4.171. Use of primary and secondary resistors with the heater transformer as suggested by K1FO. R1: 0R47 10W for single 4CX250, 0R22 10W for pair of 4CX250s. R2: Select on test for final output voltage. T1: 9V secondary, 25VA min for single 4CX250, 50VA min for pair of 4CX250s

observed; if the cathode is not at a uniform temperature, emission is concentrated in the hot areas, and the current density can be high enough to strip the cathode with permanent loss of current capability.

Fig 4.172(a) shows a manual switching circuit which forces the correct turn-on sequence, but relies on the operator to allow the correct warm-up time for the heaters. Fig 4.172(b) [65] shows a simple circuit which can be used to automatically delay other supplies turning on until the heaters have warmed up.

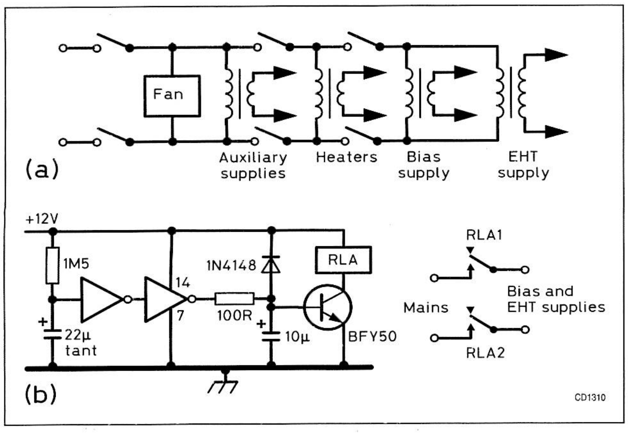

Fig 4.172. (a) Manual switching circuit which allows the correct sequence. (b) Simple circuit to delay switch-on automatically until the heaters have warmed up. The two invertors are CMOS types, eg CD4069) or inverting gates used as invertors. The 12V supply comes on when the heaters are switched on

Screen grid supplies

As discussed earlier, screen grid supplies for 4CX250s have to be able to source and sink current of about 10mA/valve while maintaining a constant voltage. This requirement lends itself to some form of shunt stabiliser. The simplest form in widespread use is a string of zener diodes in series, as in Fig 4.173. R2 provides a current of about 20mA through the diodes and this sets the maximum current which can be drawn by the valve before the voltage drops. The screen dissipation is thus limited to a maximum of 7W under normal operation and less than 12W (maximum rating) under worst-case fault conditions. This method of screen supply gives reasonable linearity but better results are achieved with active regulators which give much better regulation of the output voltage. Two such circuits are shown in Figs 4.174 and 4.175. The first is by G4XZL, used in the amplifier described in [66], the second is by G4IDE [67]. These and the zener diode circuit can suffer a problem if an anode-to-screen flashover occurs. A MOV is used to protect the capacitors in the valveholder. The type listed is specified not to conduct at voltages below 370V, but will start to conduct heavily at about 450V, with the capacity to carry thousands of amps for a few microseconds. For the duration of the flashover, the MOV acts to hold the screen grid voltage to about 650V (see Fig 4.176). The shunt regulator conducts heavily, trying to keep the voltage down to the set level, about 350V. This results in a brief high-current pulse. The regulator usually survives this, but meter shunt resistors and meters have been known to burn out. D1 can help to provide a measure of protection. In the case of the active regulators, the modification in Fig 4.177 introduces a current limit of about 70mA which protects the shunt regulator and meter from excess dissipation. References [68–70] contain alternative high-performance circuits.

Tuning up

Tuning procedures are different for tetrodes and triodes, although both start with 'heavy' loading, that is maximum coupling between the load and the tuned circuit so as to give a low anode resistance.

Correct loading for a tetrode is indicated by screen grid conditions. Apply a small amount of RF input, adjust the grid

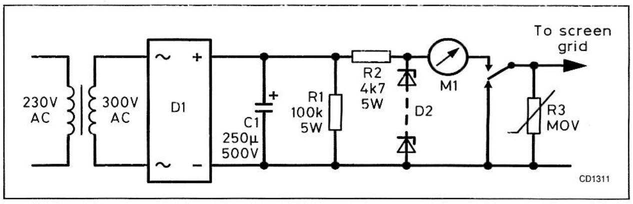

Fig 4.173. Simple screen grid supply for 4CX250. R3: 275V MOV. D1 1000V 1A bridge (4 × 1N4007). D2: 10 × 33V 5W zener diodes in series. M1: 10-0-10mA centre-zero. RLA: TX/RX switching relay

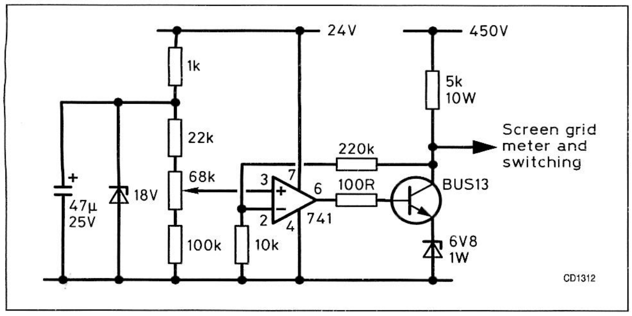

Fig 4.174. G4XZL active regulator

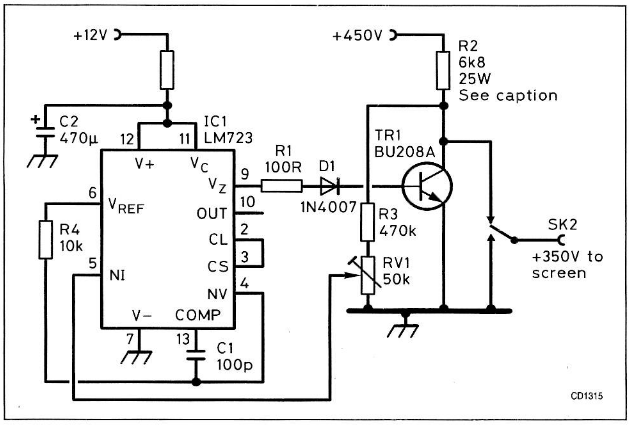

Fig 4.175. G4IDE active regulator. Set RV1 to give 250V for the screen at the collector of TR1. R2 should be chosen to give 25–30mA through the BU208A when the screen is not being supplied

and anode tuning for maximum output power and the grid input matching for lowest VSWR. Adjust the loading for most negative screen grid current possible, retuning the anode for maximum output power at each adjustment. Gradually increase RF input, continue retuning, adjusting the loading for most negative screen grid current until you reach approximately 50% of the intended peak anode current/input/output power. Do not alter the loading control again. Increase RF input until the screen grid current rises to about +5mA (per valve), and make final adjustments to anode and grid tuning for maximum output and minimum input VSWR. This is the maximum output at which the amplifier should be run. With low anode voltage, about 1500V or less, the screen grid current might not go negative and can be allowed to increase to 10mA peak. Below about 1000V, 20–30mA is acceptable. Table 4.19 shows typical values.

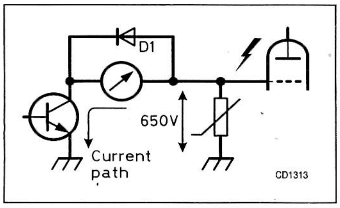

Fig 4.176. Use of MOV to give flashover protection

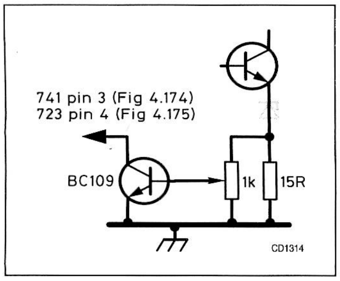

Fig 4.177. Modification to Figs 4.174 and 4.175 to give a current limit of about 70mA to protect the shunt regulator and meter from excess dissipation

Table 4.19. Typical operating conditions for a single 4CX250 at VHF

Anode voltage	900	1250	1900	V
Anode current	245	200	310	mA
Quiescent current	100	100	70	mA
Screen voltage	300	315	350	V
Screen current	30	10	5	mA
RF input power	3.5	4	3	W
Output power	110	160	330	W

In grounded-grid amplifiers the loading is adjusted to give the required output power and efficiency, so a reasonably accurate RF power meter is essential. All tuning should be carried out iteratively as the input power is gradually increased, with final adjustments at full power because the input impedance varies with input power. The actual value of grid current is not significant provided it is within the limits specified for the valve. The output power from a triode amplifier does not saturate in the same way as a tetrode; it is possible to keep increasing the input power and get more output power up to the point where damage occurs.

Aiming for about 50% anode efficiency (45% at UHF) will give reliable, linear operation.

In grounded-grid amplifiers, the grid can be damaged if RF input is applied without anode voltage. An interlock circuit which detects anode voltage should always be used to protect the valve by preventing the amplifier switching into transmit mode if the anode voltage is absent. Fig 4.178 shows a simple circuit based on a design by G4IDE [71].

VALVE AMPLIFIER CIRCUITS

Simple amplifiers capable of 50–100W output from 1–2W drive can be built easily around QQV06-40/07-50 *double tetrode* valves [72], which can still be readily found in surplus equipment.

In recent years there has been a tendency to design amplifiers (especially those using 4CX250) with the intent of achieving the full output capability of the valve, about 275W clean PEP in the case of the 4CX250. This need not be difficult to achieve, but it does require a good valve, 2kV or more anode voltage and a number of protection measures to avoid

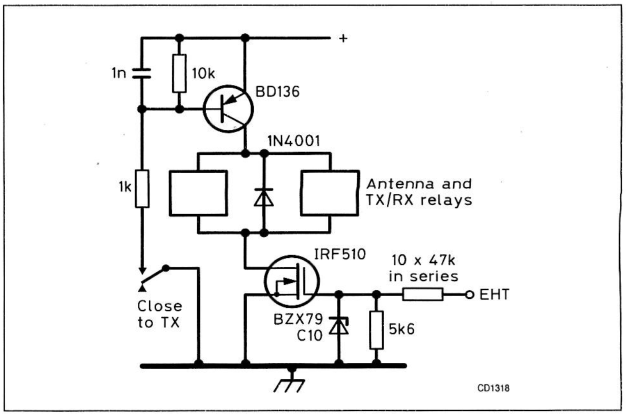

Fig 4.178. Simple grid protection circuit

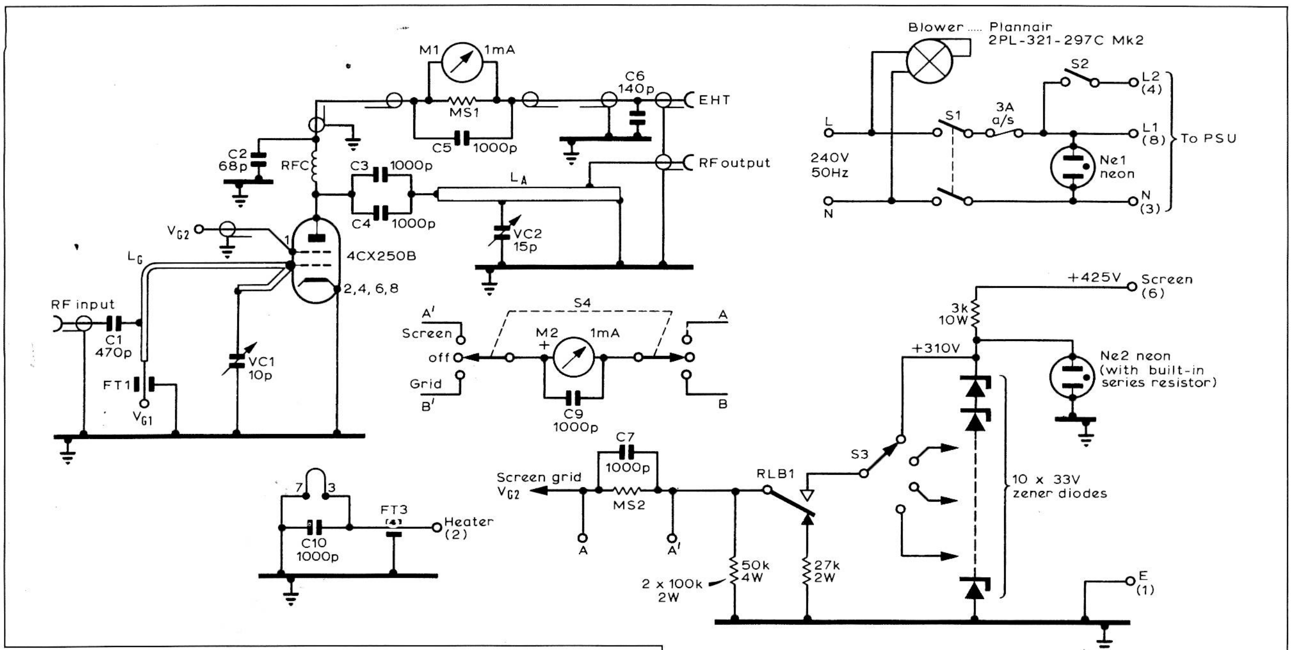
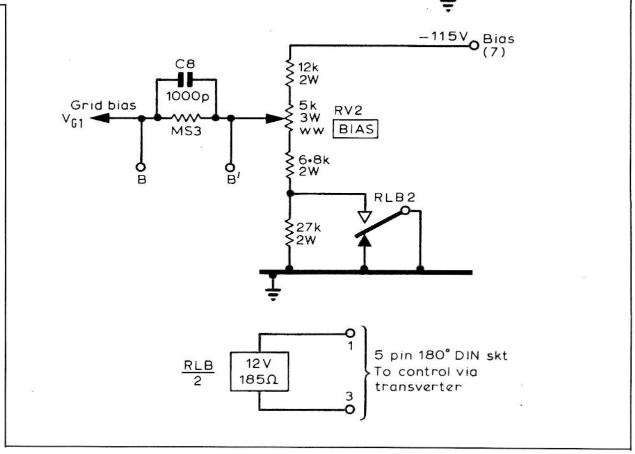

Fig 4.179. 144MHz linear amplifier. RFC: 2t 18SWG 3/8in ID, 1in long. Note that C2 should be 680pF not 68pF as shown. Alternatively a high-voltage 1000pF feedthrough can be used if available

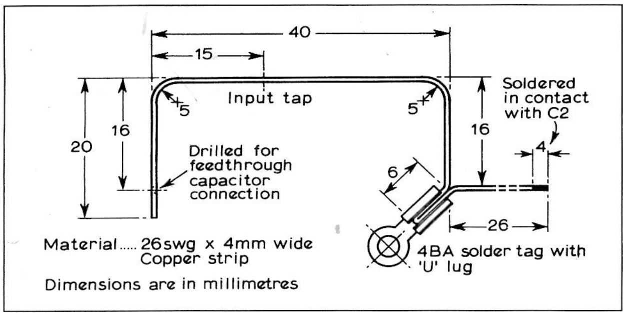

Fig 4.180. Grid inductor. 'C2' refers to Fig 4.187 or VC1 in Fig 4.179

total destruction in the event of an anode voltage flashover. As an alternative, aiming for an output power of 150–200W (only one-third to half an S point less) can simplify things. For example, anode voltage can be reduced to 1–1.5kV; the lower secondary voltage makes a transformer easier to find, fewer capacitors are needed and the stored energy (0.5 CV^2) is much lower. The STC data sheet for the 4CX250 specifies over 200W PEP output with acceptable linearity with an anode supply of 1500V. Very few old amplifier designs running at these voltages and powers had the range of protection circuits now considered essential. The reason: flashover and other problems were relatively rare. Amplifier gain was usually lower too, making the layout and choice of valve base less critical. For these reasons, the inexperienced builder is strongly recommended to aim at these lower targets initially. Nevertheless, the various protection measures outlined in the PSU section previously should be included for peace of mind at minimal cost.

Suitable designs have been featured in various editions of the *VHF/UHF Manual* and are repeated here. Despite their age, more recent designs differ only in detail. Figs 4.179–4.186 and Table 4.20 show a 144MHz amplifier [73] whose origins can be traced to GW3ZTH. The anode circuit uses a shortened λ/4 transmission line tuned to resonance with the valve anode capacitance and VC2. Output loading is set by the position of the output connection on the transmission line. While this does not provide a fully variable adjustment, in most situations the loading does not need to be adjusted once the optimum setting has been found. Great care is needed and

Table 4.20. Operating conditions for 144MHz amplifier

	No signal	Single tone
Anode voltage (V)	1350	1200
current (mA)	100	200
Screen voltage (V)	315	315
current (mA)	—	10
Grid voltage	−32	−32*
current (μA)	—	<100
Power output (W)	—	150–170

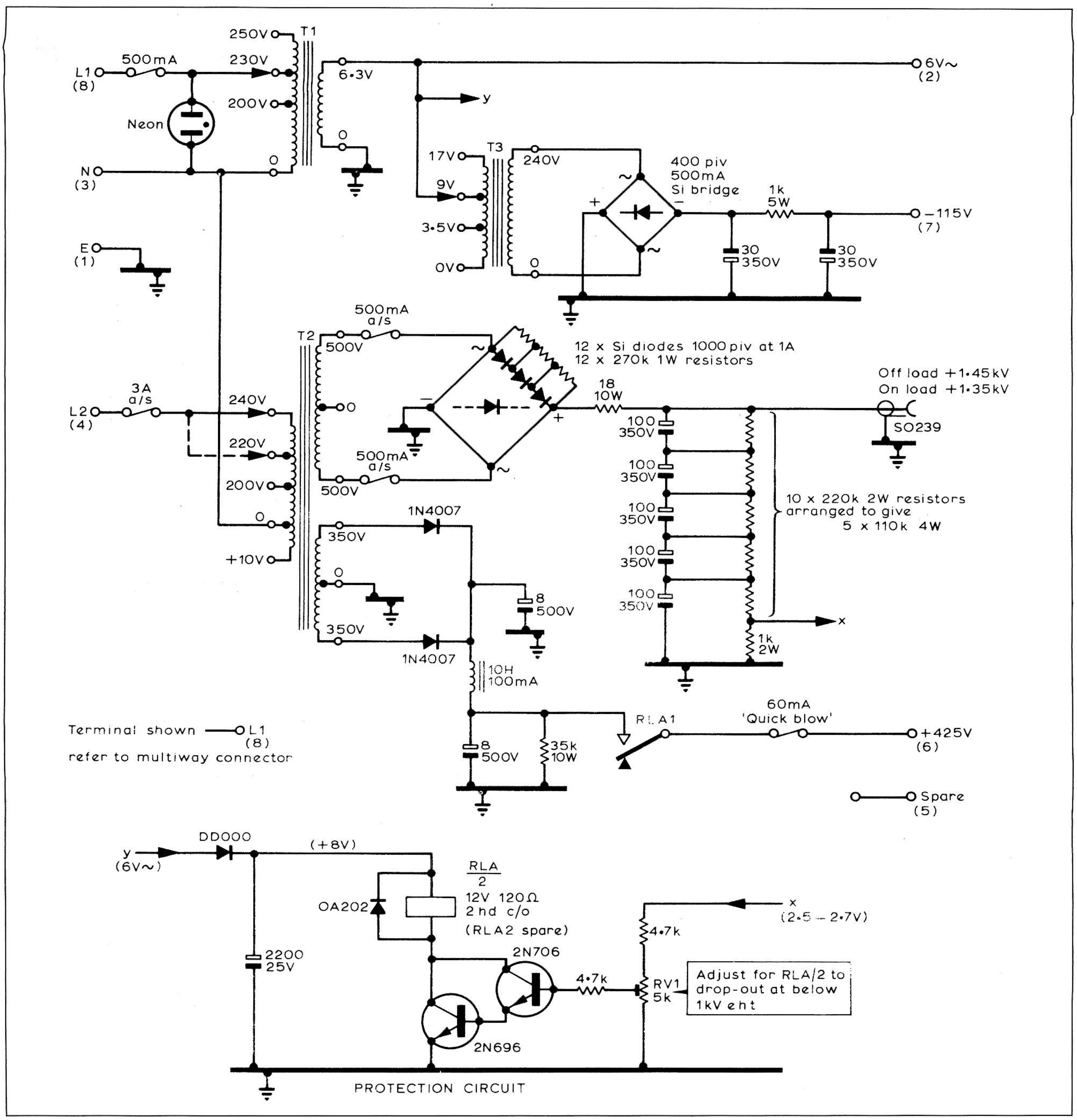

Fig 4.181. Remote power supply unit for 144MHz linear

the valve anode must be solidly earthed when working inside the anode compartment. Do not rely on disconnecting the anode supply; C3 and C4 can hold charge for a long time. The grid circuit uses a shaped strip as the inductor instead of a wound coil. The shape and dimensions fit nicely into a standard die-cast box, and should be followed closely to ensure correct tuning. A small variation in the length of the connection to VC1 should not affect operation. Input VSWR is optimised by adjusting the connection point of C1 onto L_G.

Figs 4.187–4.190 and Tables 4.20 and 4.21 show another design [74] which uses a lumped-component version of the λ/2 circuit in Fig 4.163. Output coupling is adjusted by varying the position of the coupling coil with respect to the tuning coil. The low output power figures for this amplifier reflect the licence conditions of the time, which limited DC input to 150W maximum for FM or CW. Given the same operating conditions as the previous design, the same performance should be achieved.

An amplifier for 432MHz [75] is shown in Figs 4.191–4.193; this was also designed at a time when DC input power was more restricted. Die-cast boxes are used as cheap, convenient ready-made housings for the grid and anode circuits.

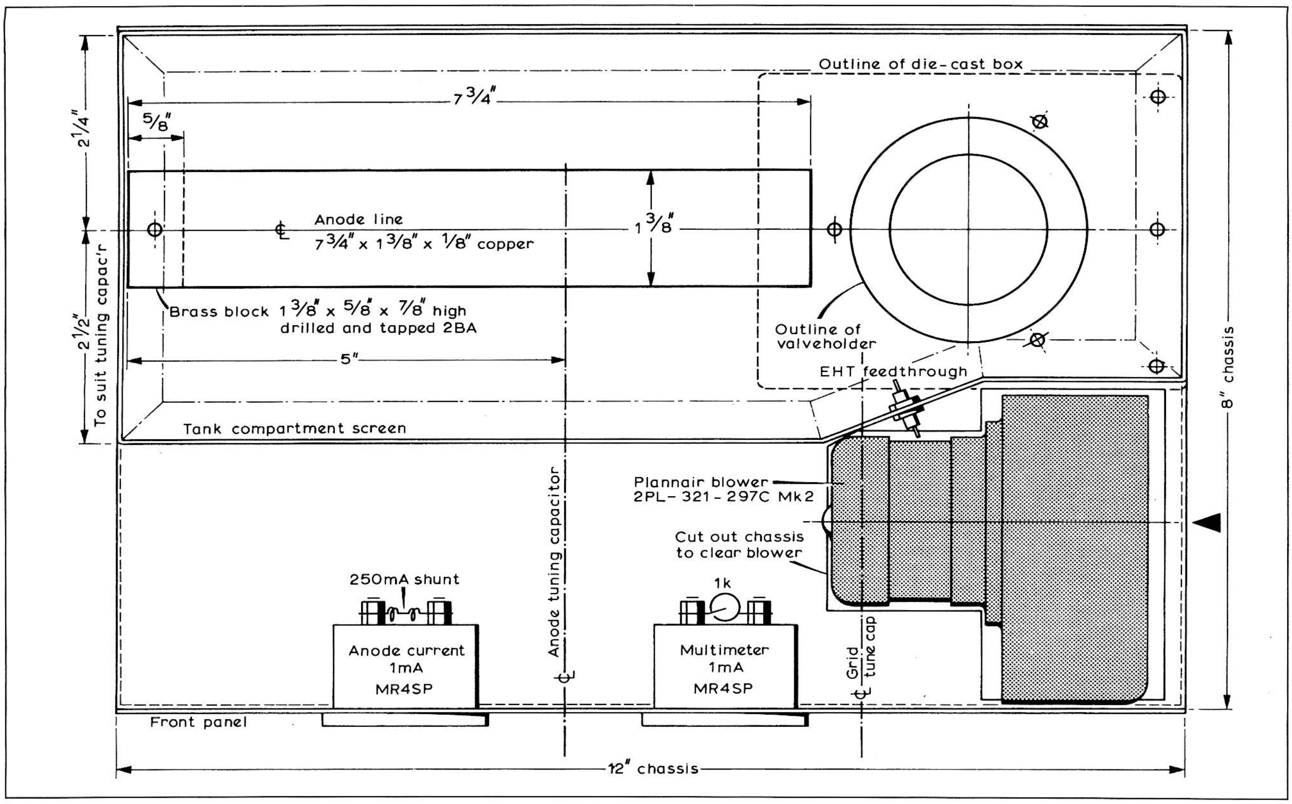

Fig 4.182. Above-chassis outline

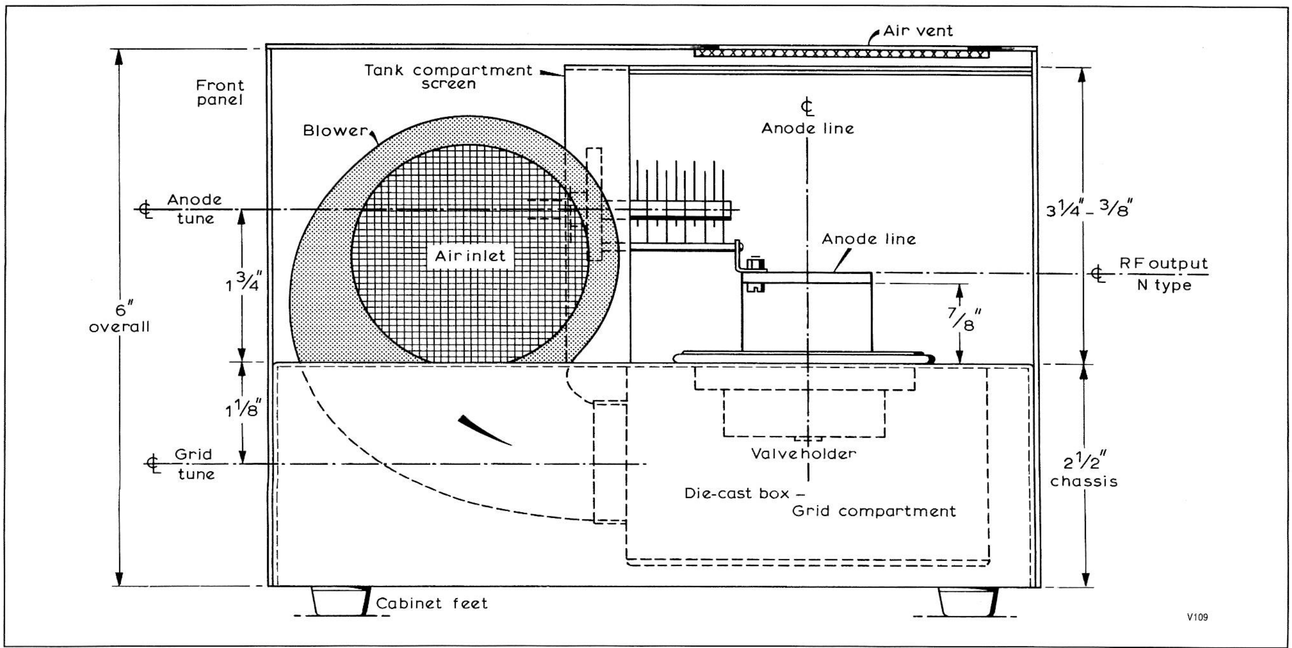

Fig 4.183. Side elevation

The design is capable of higher output power but the die-cast metal is not the best choice in high power RF fields and unpredictable effects might start to occur as the RF currents increase, although problems are unlikely to be seen at powers below about 175W. The output coupling might need some alteration to optimise the loading for linear operation, as described earlier. It would be beneficial to change the capacitor in the output coupling to an air-spaced piston type (eg

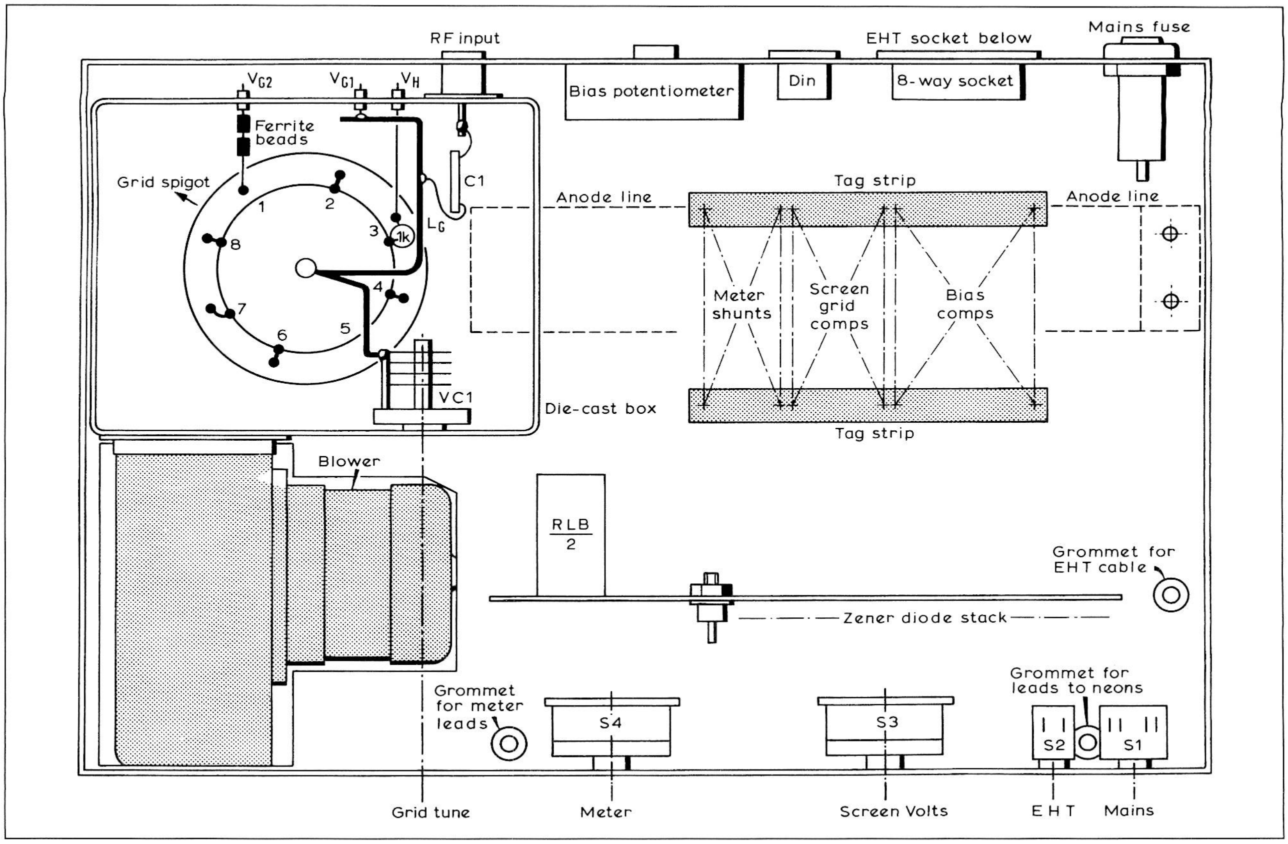

Fig 4.184. Below-chassis outline

Johansson) rather than the ceramic one used in the original. As an alternative, capacitive output coupling could be tried. Although widely used in other designs it must be stressed that this is untried in this amplifier. Estimated dimensions are a brass/bronze strip ½in wide over the anode line, 2.5in from the tuning capacitor end. The strip should extend to the mid-line of the anode line. See Fig 4.216 as an example.

A 'string' can be used as an alternative mechanism to adjust the anode tuning capacitor.

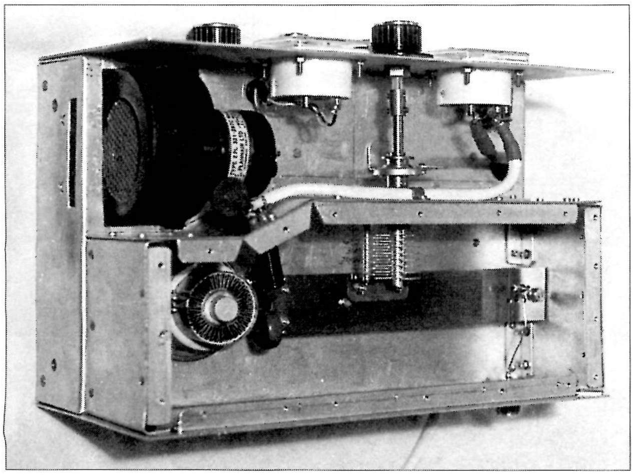

Fig 4.185. Interior showing fan, strip line and tuning capacitor

Higher-power amplifiers

In 1971 *QST* published an amplifier by W1QVF (later W1SL) and W1HDQ [76] using a pair of 4CX250s in push-pull, with λ/2 grid and λ/4 anode tuned circuits. This set a design standard which is regularly duplicated today. It achieved widespread popularity after being included in the *ARRL Handbook* for some years. An amplifier by G8LT using the same principles is shown in Figs 4.195–4.198 and Table 4.23.

A tuned circuit feeds RF voltage to the grids. The circuit is loaded only by the input impedance of the valves, which is

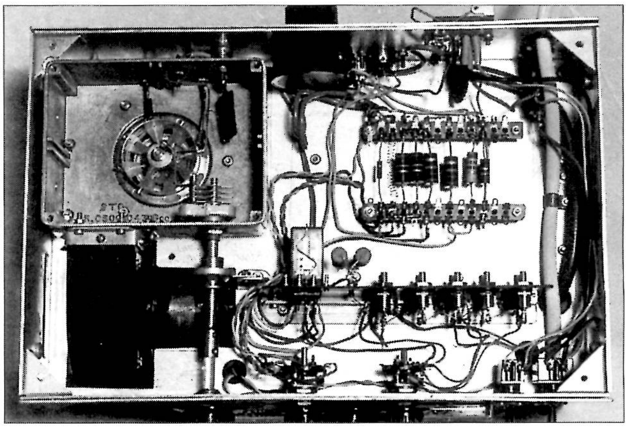

Fig 4.186. Underside view

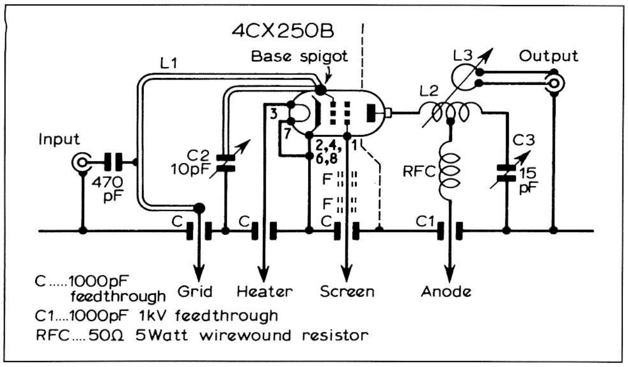

Fig 4.187. Circuit of compact 150W amplifier for 144MHz

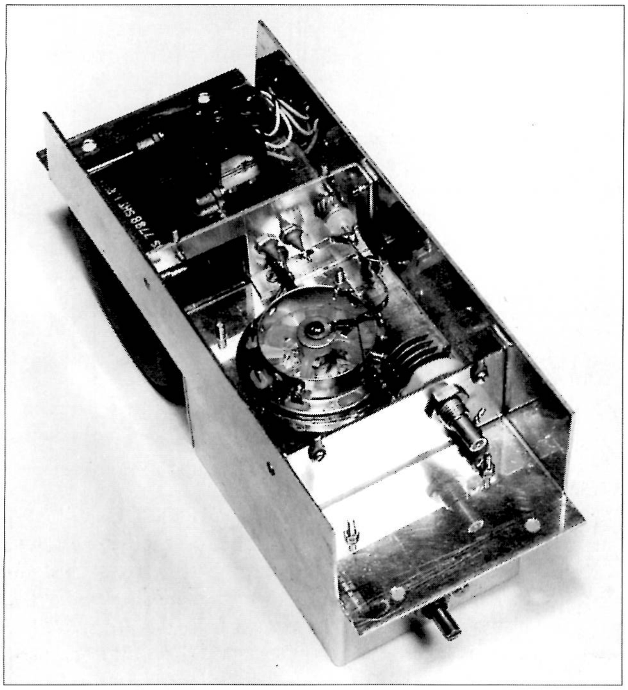

Fig 4.188. Underside view of compact amplifier for 144MHz

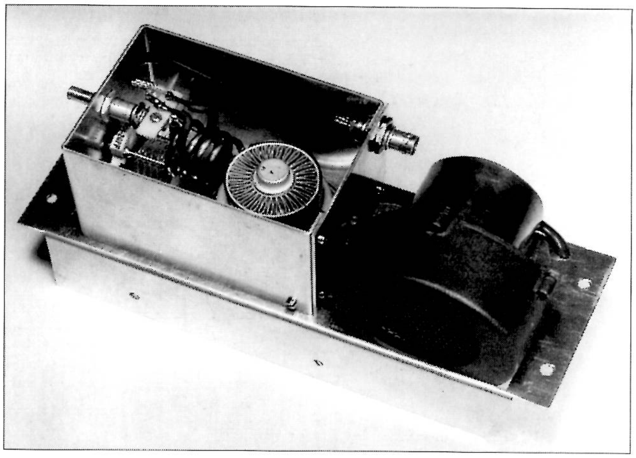

Fig 4.189. Compact amplifier for 144MHz

Table 4.21. 150W amplifier components list

C	1000p feedthrough
C1	1000p feedthrough 1kV type
C2	10p C804 Jackson
C3	15p C804 Jackson
F	Ferrite bead
RFC	50R 5W wirewound
L1	Copper strip loop (see Fig 4.180)
L2	3½t ¾in ID 1/8in diam copper
L3	1t ¾in ID insulated
Valve socket	Eimac or AEI
Blower	Planair type 2PL 321-284C Mk3

Table 4.22. Performance of 150W amplifier

Anode voltage (V)	750	800
Anode current (mA)	200	200
Screen voltage (V)	250	250
Screen current (mA)	5	8
Grid voltage (V)	–100	–100
Grid current (mA)	6.6	8
Drive power (W)	2.6	3.0
Output power load (W)	90	100

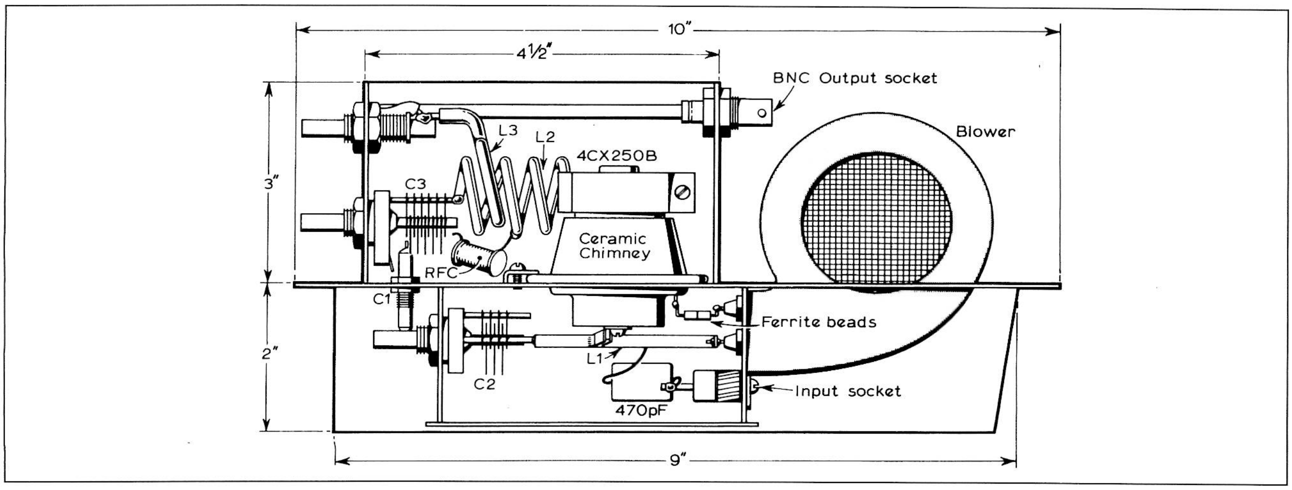

Fig 4.190. Side view of compact amplifier for 144MHz

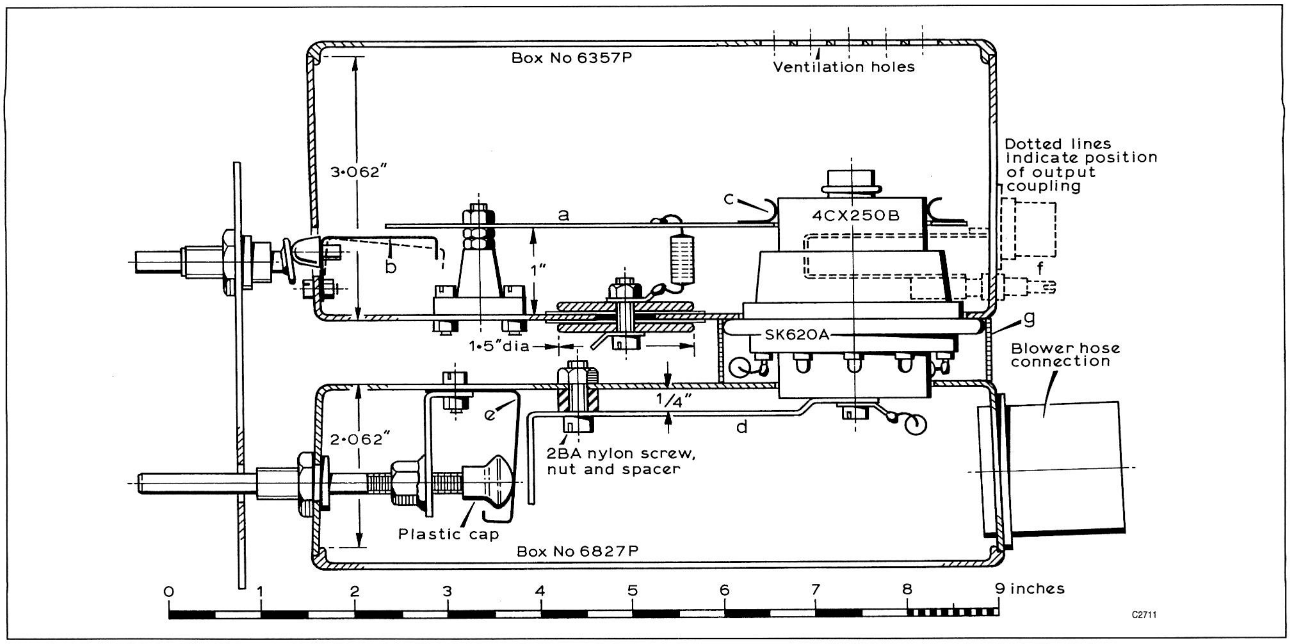

Fig 4.191. Section of complete assembly showing principal components for the 432MHz slab-line power amplifier

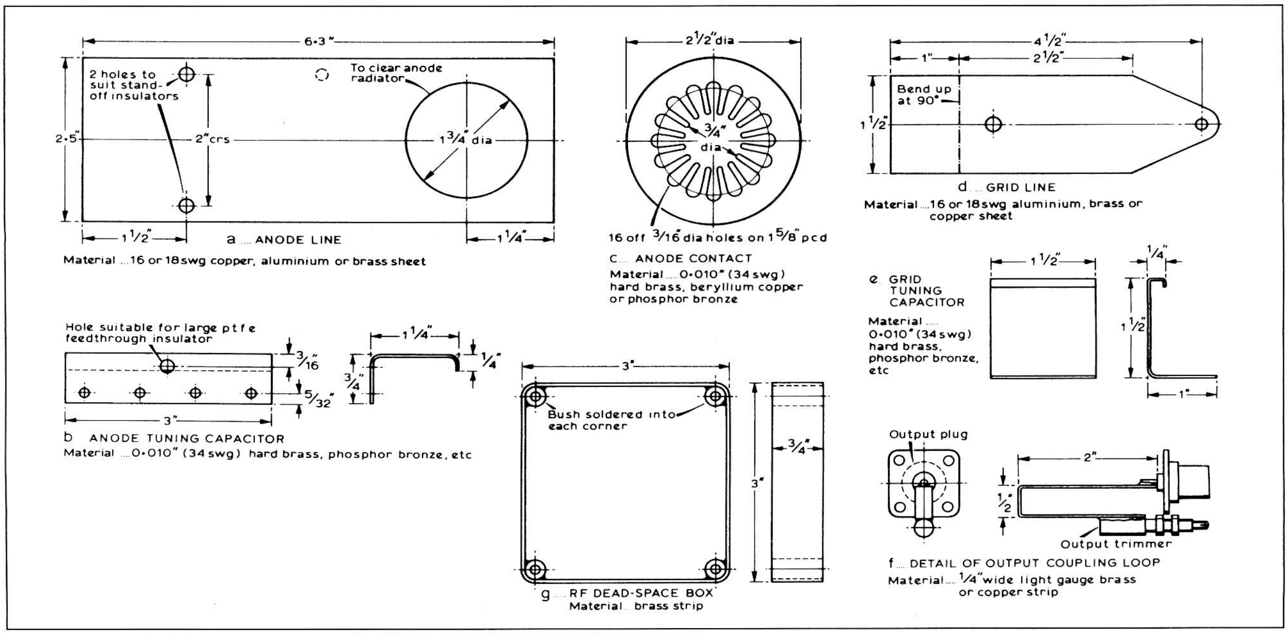

Fig 4.192. Details of components

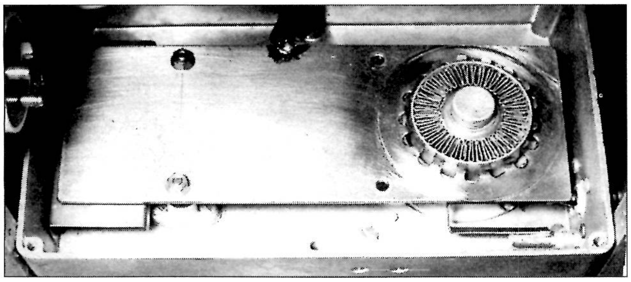

Fig 4.193. The anode in its holder and the output loop

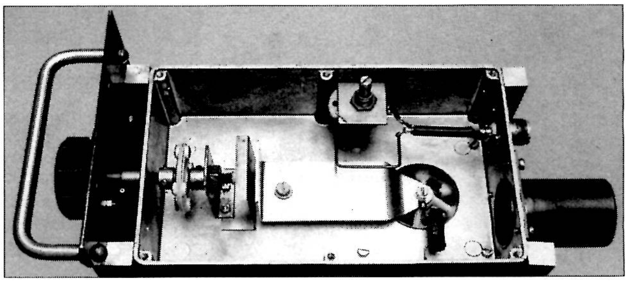

Fig 4.194. The grid line, input coupling loop and blower hose

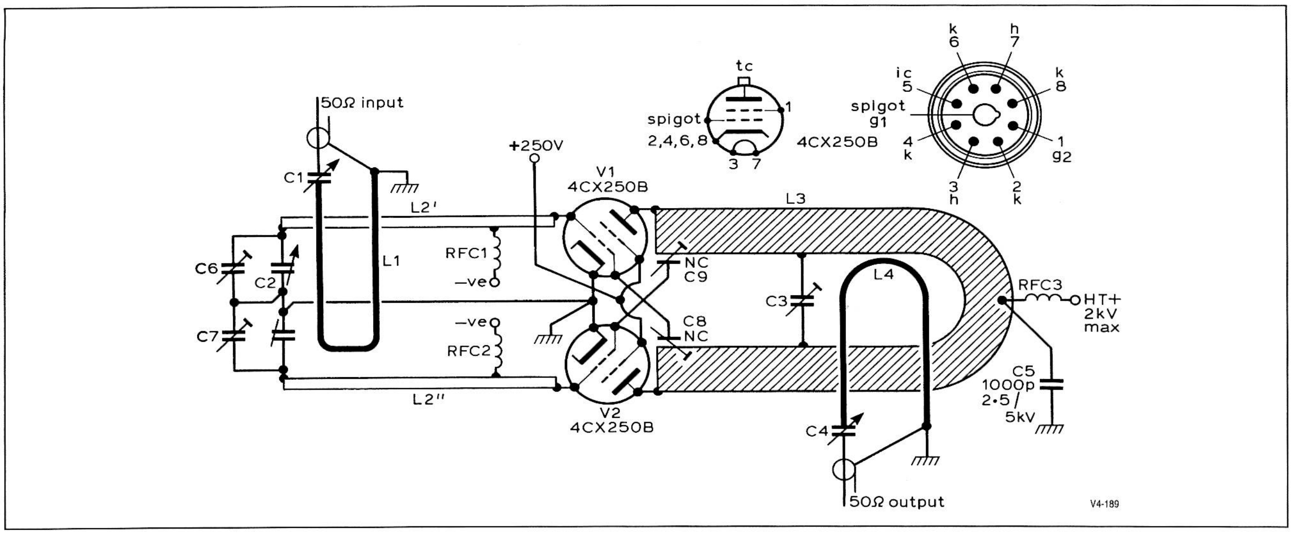

Fig 4.195. High-power 144MHz amplifier using a pair of 4CX250B valves

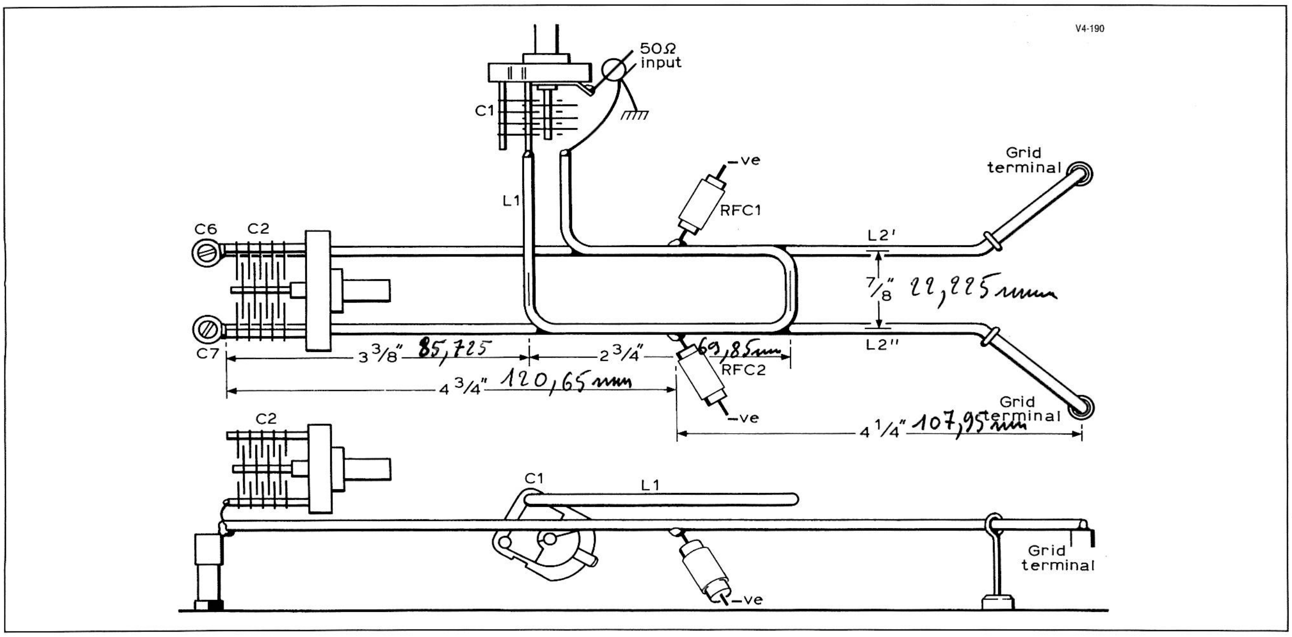

Fig 4.196. Grid circuit

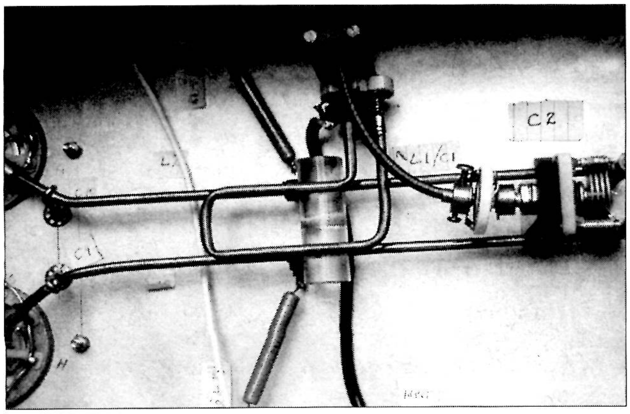

Table 4.23. Components for G8LT amplifier	
C1, C4	50p (Polar C4.04)
C2	15 + 15p (Polar C8.52/1)
C3	Disc, 1½in diam
C5	1000p, 2.5–5kV bypass
C6, C7	10p piston trimmer
C8, C9	NC, plate attached to 16 SWG wire
L1	Input coupling loop (see Fig 4.196)
L2', L2''	1/8in diam copper 9in long (see Fig 4.196)
L3	1in × 1/6in copper strip (see Fig 4.198)

Left: **Fig 4.197. Close-up view of grid circuit with components labelled**

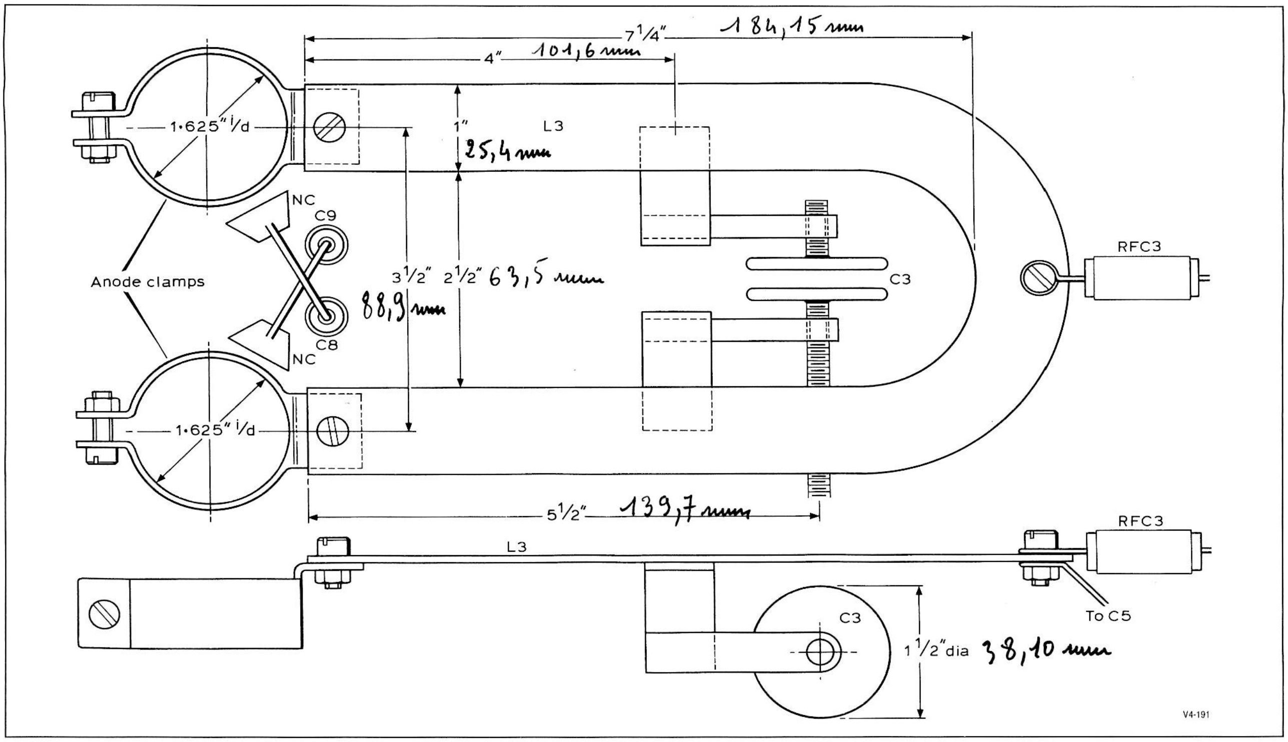

Fig 4.198. Anode circuit

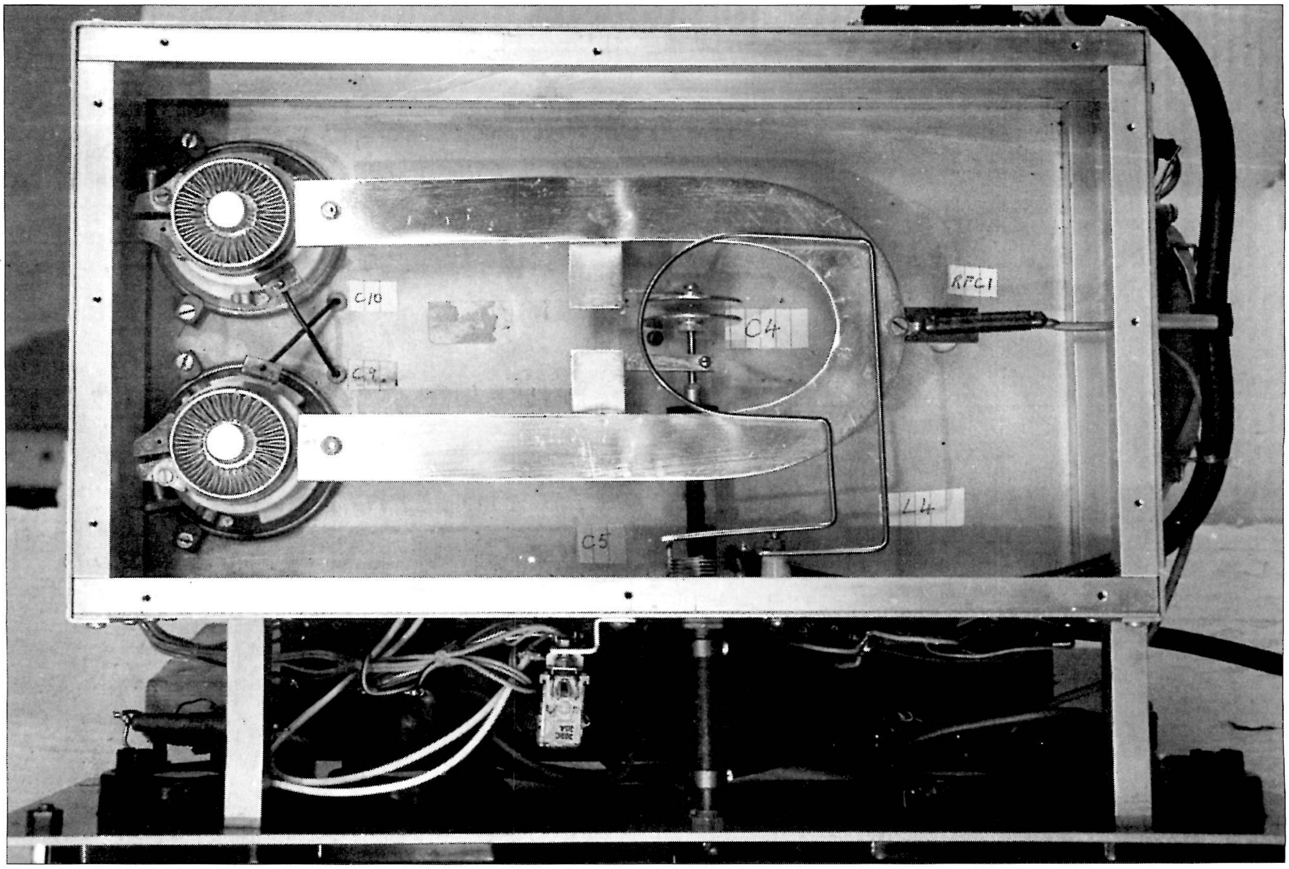

Fig 4.199. Top view of amplifier. Component numbers do not correspond to Fig 4.195

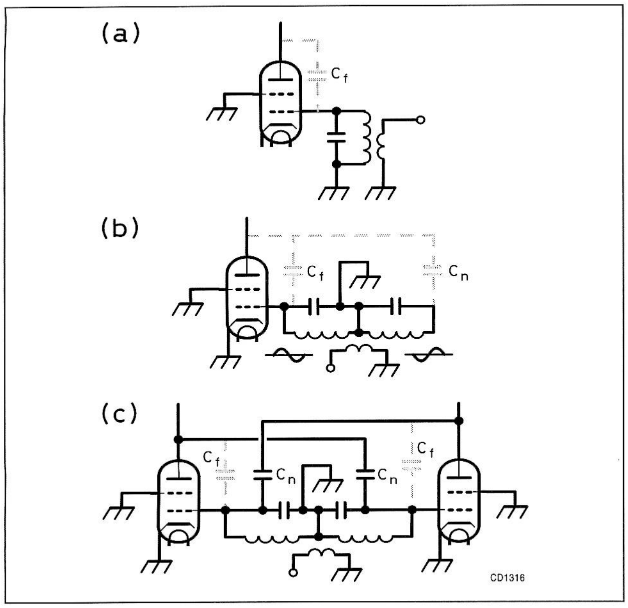

Fig 4.200. Evolution of neutralising circuit for a pair of valves. (a) Unwanted feedback. (b) Use of neutralising capacitor and additional circuitry for one valve. (c) Final circuit for two valves

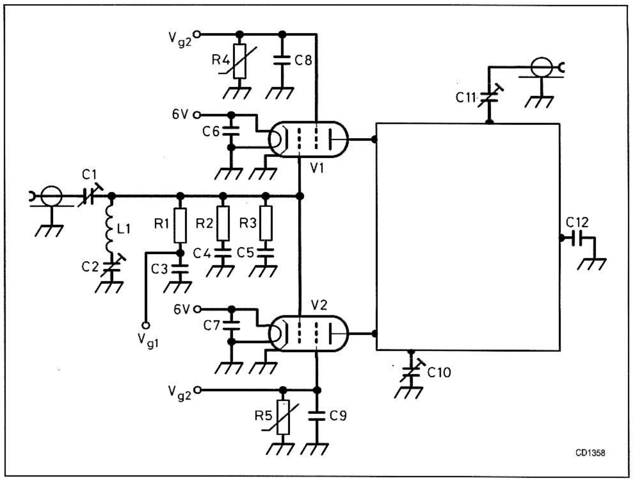

Fig 4.201. W2GN amplifier circuit. R1–3: 2×2k2, 3W in parallel. R4, 5: MOV, 275V AC, 140J. L1: 3t 1mm Cu wire, 16mm diam. C1, 2: 15p var. C3–7, 1000p 500V ceram. C8, 9: built into valve base. C10, 11: Mechanical flapper plates – see Fig 4.210. C12: formed from sandwich of metal plates and PTFE sheet – see Fig 4.212

fairly high, so there is a high RF voltage step-up ratio. This results in a high power gain for the amplifier; at least 20dB or 100× can be expected. The high gain means that the amplifier has to be neutralised. In Fig 4.200(a), the unwanted feedback signal path is represented by the anode-to-grid1 capacitor C_f. In Fig 4.200(b) an additional grid tuned circuit, coupled to the first one, is added and if the input signal is injected at the centre, the opposite ends will be in antiphase. By adding an external 'neutralising' capacitor C_n, an equal and opposite feedback signal is introduced which cancels that from C_f. The value of C_f and C_n is about 0.05pF, so C_n is not a physical component, but is made from a small brass tab held near the valve anode. The second valve can be drawn as a mirror image of Fig 4.200(b), and by sharing the same grid tuned circuit, the final circuit of Fig 4.200(c) is realised.

Neutralisation adjustments are straightforward, but can take a little time to get right [77]. Connect a load to the output, and leave the screen grid and anode supplies open-circuit. With the heaters on and control grid voltage set to about –35V (lower than the value normally needed to bias the valves), apply some RF input until slight grid current is seen. Peak the grid current with C2, reducing the RF input as necessary to get a few milliamps of grid current. Use C6 and C7 to balance the current between the valves. As C3 is adjusted to tune the anode circuit to the input frequency, some energy will be drawn into the anode circuit and the grid current will alter. The neutralising capacitors C8 and C9 are adjusted by changing their positions to minimise and eventually completely remove any effect of anode tuning on the grid current. The anode compartment cover must be in place when the neutralisation is checked. Depending on the valves and bases in use, the position and size of the tabs might be quite different from that shown in the diagrams. Initial changes should be made to keep C8 and C9 symmetrical but final adjustments might require small changes to one or other alone. It is important to persevere until there is absolutely no effect on the grid current at any combination of anode tuning and loading controls. Once this is achieved, the amplifier can be tuned up as described previously. With an anode voltage of 1.5–2kV, and 100mA per valve quiescent current, a clean, linear output of 400–500W should be achieved for about 4W drive.

The axial fan used on this amplifier is rare and non-standard, capable of high-pressure operation. Readily available axial fans are incapable of cooling such an amplifier, and a proper high-pressure blower should be used, as discussed earlier.

W2GN described an alternative design [78] which uses two valves in parallel. The anode circuit is a $\lambda/4$ transmission line as in Fig 4.163(a), tuned to resonance with fixed and variable capacitive plates. Output coupling is through a 'flapper' capacitor adjusted with a screw through the box lid. Photographs and drawings of a version built by G8GSQ are shown in Figs 4.201–4.207. The resistors which load the grid circuit are split

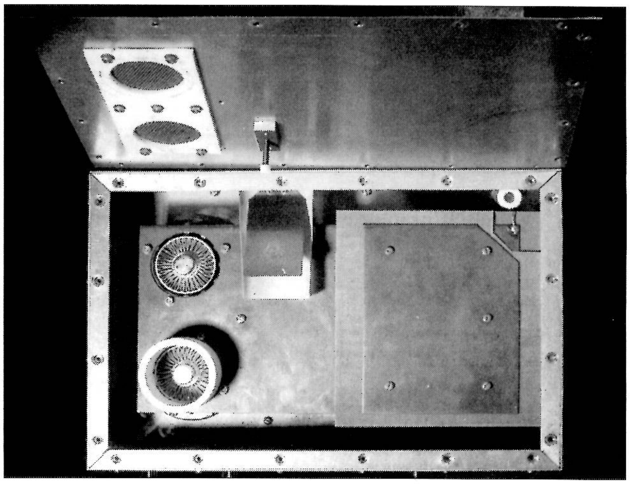

Fig 4.202. Anode compartment

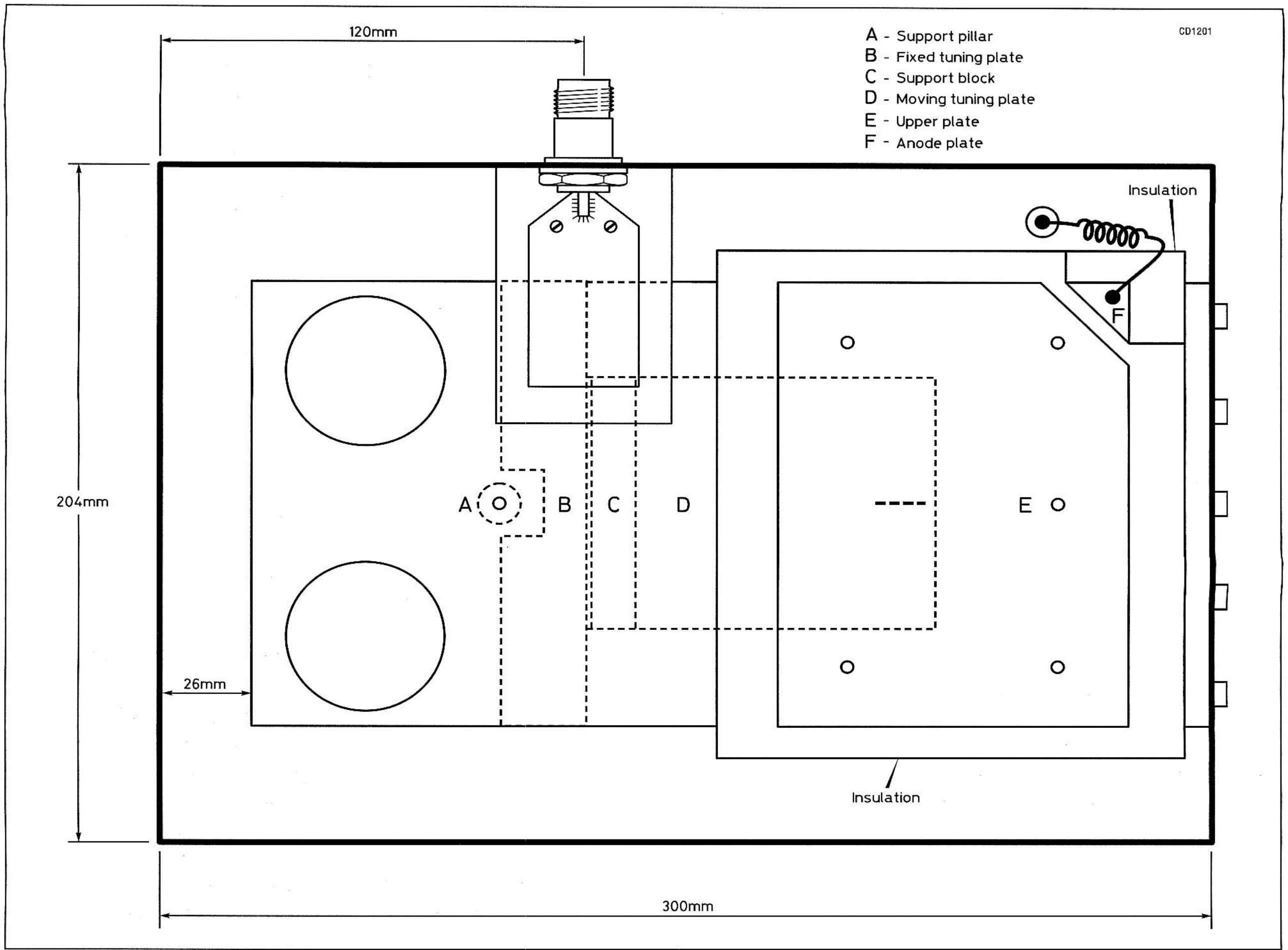

Fig 4.203. Overal view of anode compartment. A – support pillar, B – fixed tuning plate, C – support block, D – moving tuning probe

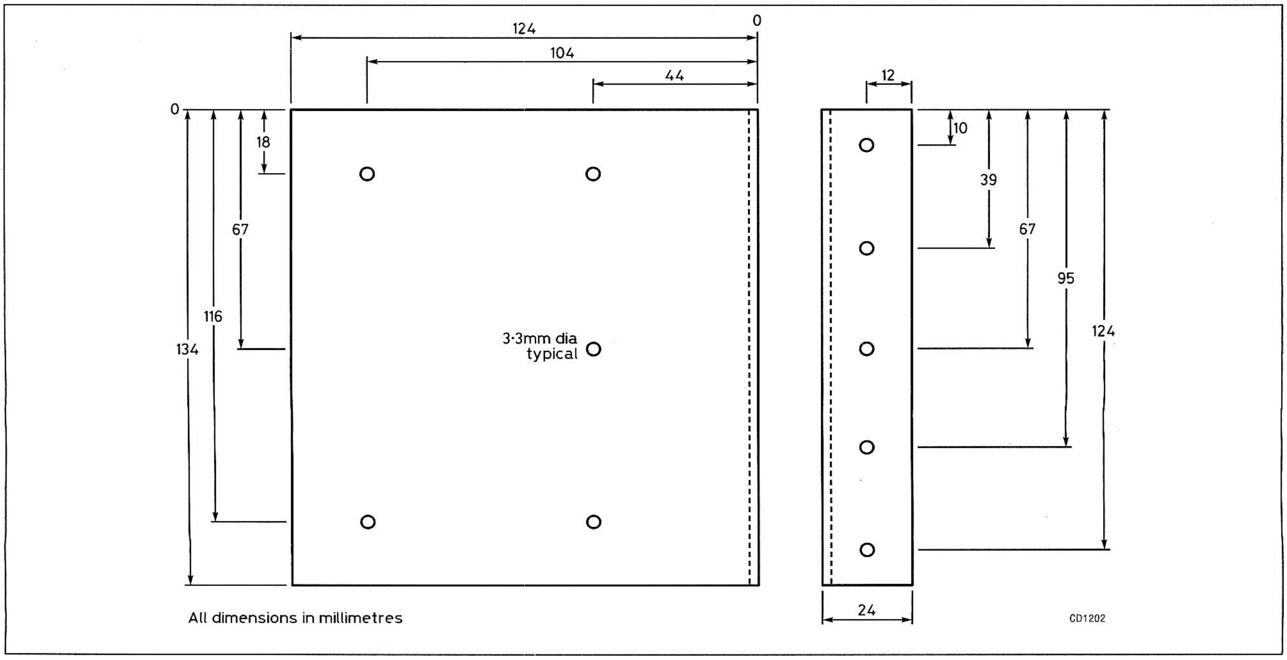

Fig 4.204. Lower plate, made of 1.2mm brass

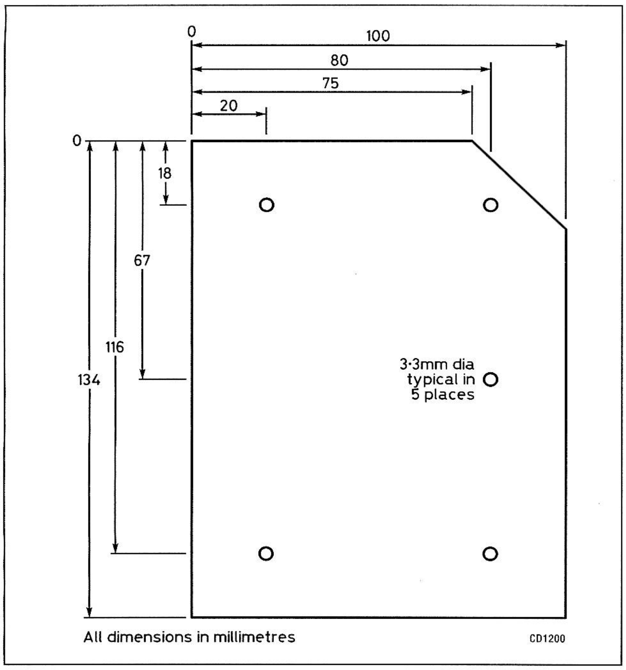

Fig 4.205. Upper plate, 1.2mm brass

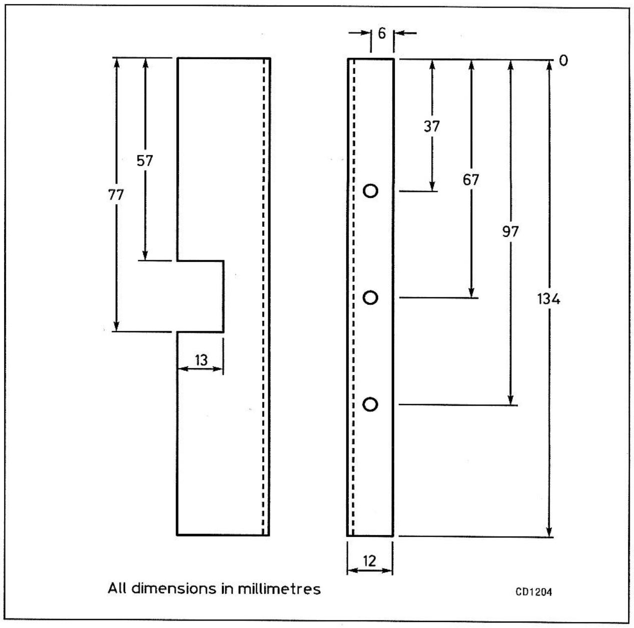

Fig 4.206. Fixed tuning plate, made of 0.5mm brass or copper

into three pairs to make use of available components. Because of the resistive grid loading the gain is reduced and neutralisation is not required. An alternative grid circuit shown in Fig 4.213 is a worthwhile modification so that the valves have individual bias control and so do not have to be selected as matched pairs.

The bases used in this version are ex-equipment types which have a PTFE chimney and a Y-shaped clip to hold the valve in place. The chimney and anode metal surround have been discarded. Following the example of the article, air is blown into the anode compartment; this was necessary as the original blower had marginal performance. A better blower has since been fitted and had this been available at the outset, the grid compartment would have been used as the air inlet. Originally the chimneys carrying the air from the anode to outside

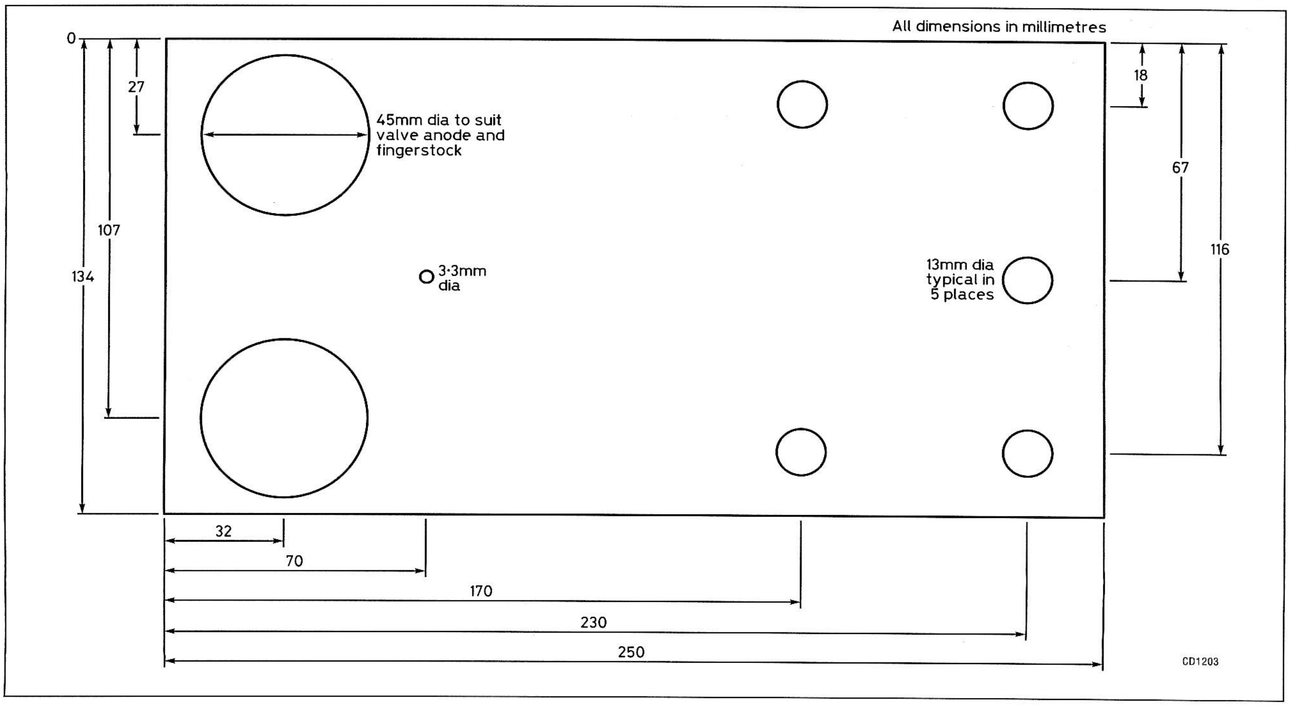

Fig 4.207. Anode line, 1.2mm brass

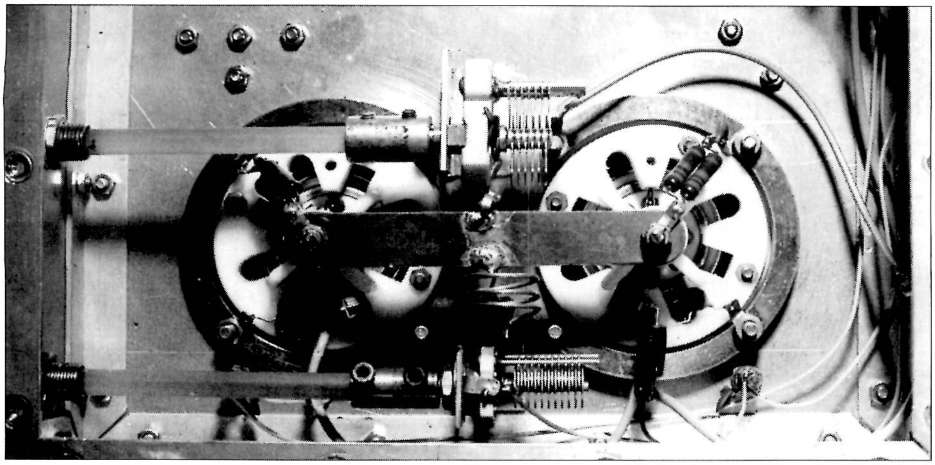

Fig 4.208. Close-up view of grid circuit

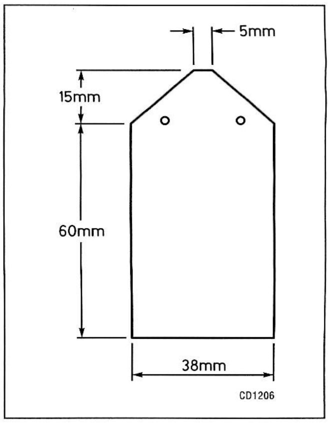

Fig 4.209. Output coupling, 0.4mm brass or beryllium copper

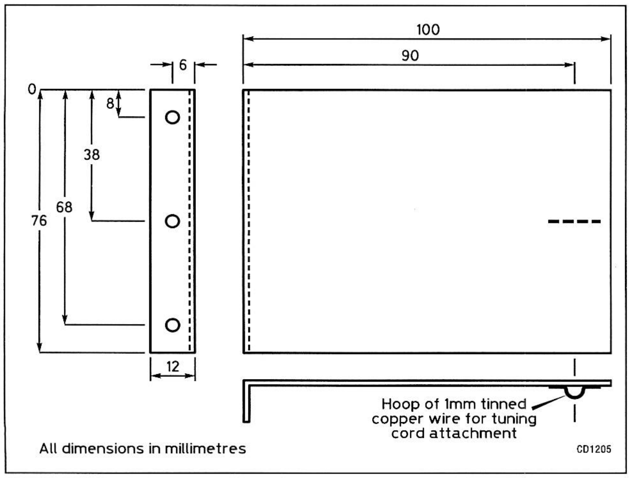

Fig 4.210. Moving tuning plate, 0.4mm brass or beryllium copper

were made from thin PTFE sheet rolled into tubes. Subsequently some suitably sized PTFE tube came to hand and this was cut to be a snug fit between the anode plate and the box

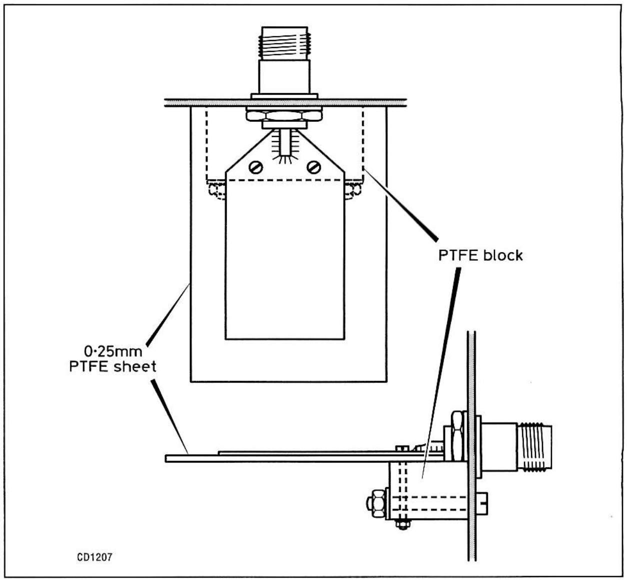

Fig 4.211. Output coupling support block

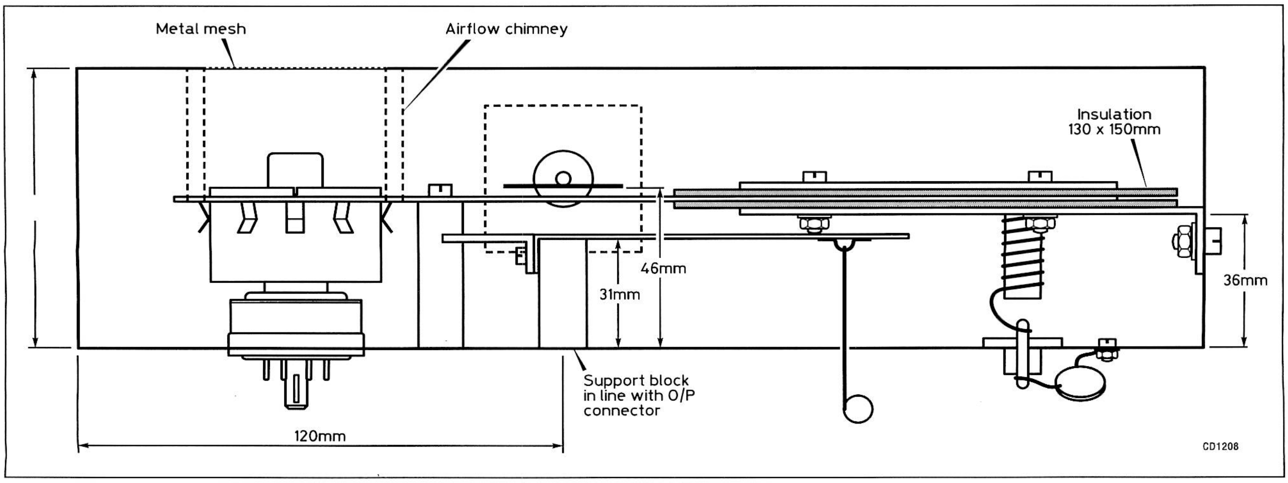

Fig 4.212. Side view of anode compartment

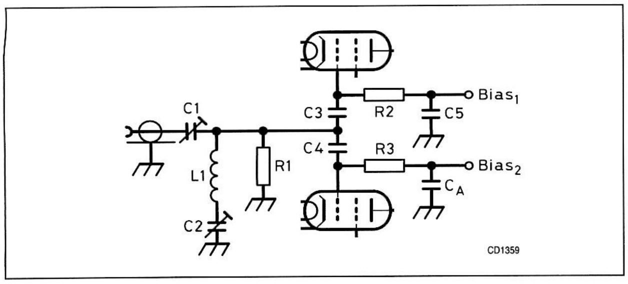

Fig 4.213. Alternative grid circuit

lid. A peculiar effect was seen in continuous small variations (about 1%) in output power. This was eventually traced to the PTFE insulation on the output coupling capacitor flapping in the airflow.

The DC blocking capacitor in the anode line is made from 0.8mm PTFE/glass PCB with all the copper removed. The original design used 0.25mm PTFE sheet but this was replaced after a flashover caused by a microscopic piece of swarf which had punctured the soft PTFE. PTFE will give excellent long-term service provided extreme care is taken to remove all swarf and rough edges.

432MHz high-power amplifier

In 1972 K2RIW published an article [79] describing an amplifier using a pair of 4CX250s in parallel. A large number of these amplifiers have been built since, with varying degrees of success. The means of achieving satisfactory and reliable operation are now largely understood: a good-quality pair of valves, genuine Eimac SK630 bases and attention to detail in the mechanical components. If all of this comes together, linear output power of around 500W can be expected; investing in high-quality cable will achieve legal limit power at the antenna. The K2RIW design is not an amplifier to build using random bits and pieces without access to test equipment (and plenty of time) for debugging. The cost of a pair of good new 4CX250s is significant, and the correct bases are likely to cost somewhat more than the valves.

In 1979, K1FO published a 1kW input, 530W output amplifier [80] using the 8874 triode. This became very popular in the USA where the valves were more readily available. The grounded-grid circuit gives lower gain than the K2RIW design but the PSU is greatly simplified and there are no stability problems. In 1986 G4ODA adapted this design to use the 3CX800; coincidentally, K1FO published a similarly modified design in 1987 [81]. While the cost of a 3CX800 appears initially high, bases are readily and cheaply available. The increased output power capability allows cheaper, lossier cable to be used while maintaining the power delivered to the antenna. The overall difference in cost is lower than might be thought initially.

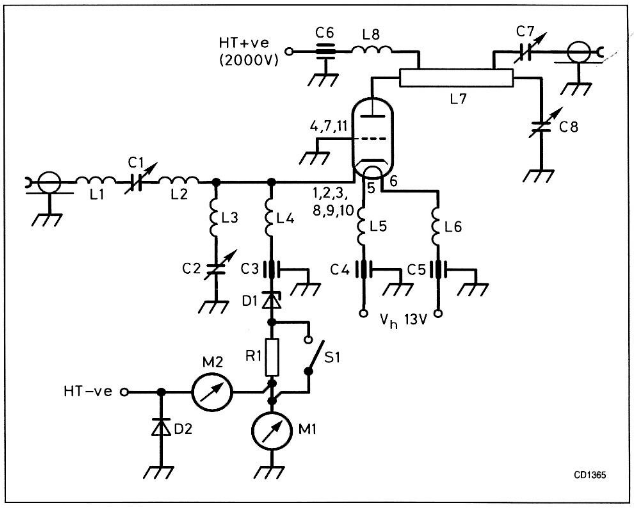

Fig 4.214. 3CX800 70cm amplifier circuit

G4ODA's design is shown in Figs 4.214–4.224. The grid contact ring on the valve is the same size as the screen grid ring on the 4CX250. The grid contact used here was taken from a scrap 4CX250 base, and its choice appears to have been fortuitous. The design was duplicated using an Eimac grid connector which has 36 closely spaced contacts. This amplifier, while not unstable, can show signs of feedback at some tuning settings. K1FO describes how to modify the contact ring to raise the inductance and 'tune' the point of minimum feedback close to 432MHz by breaking off every other contact to leave 18, and then remove or insulate a further six to leave six pairs which closely resemble the pattern of the contact ring used by G4ODA.

The option of insulating the final six contacts is to allow for individual tuning, although K1FO observes that the characteristics appear to vary little between valves, and the contacts can safely be removed.

1.3GHz

The 2C39/3CX100 triode is the 'standard' valve for this band. Two designs which represent the best aspects of amplifier design are by G3SEK/G4PMK [82] and N6CA [83]. N6CA's

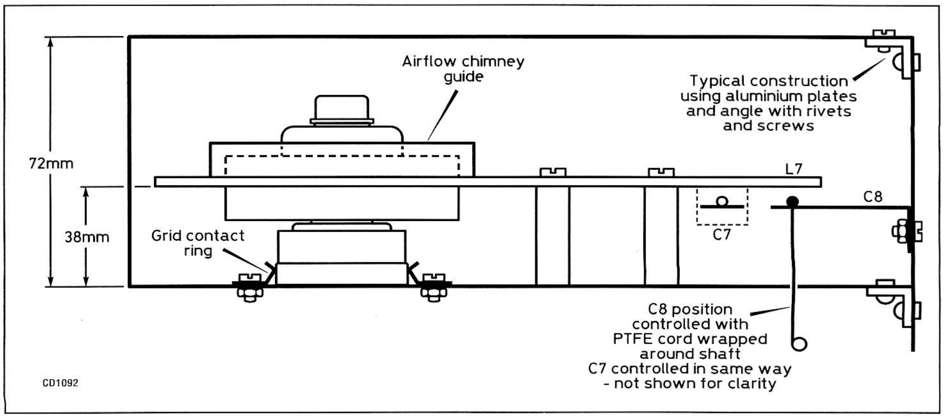

Fig 4.215. Cross-section of anode compartment. C7 controlled in same way as C8 but omitted for clarity. Also omitted for clarity are the EHT connections (L8/C6) and the airflow chimney

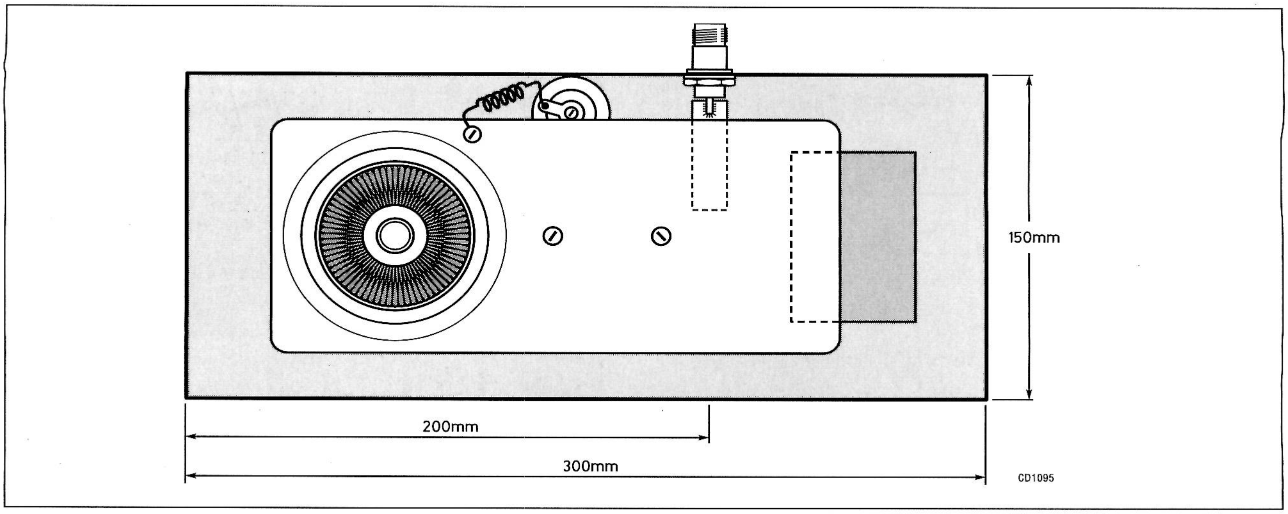

Fig 4.216. Overall view of anode compartment

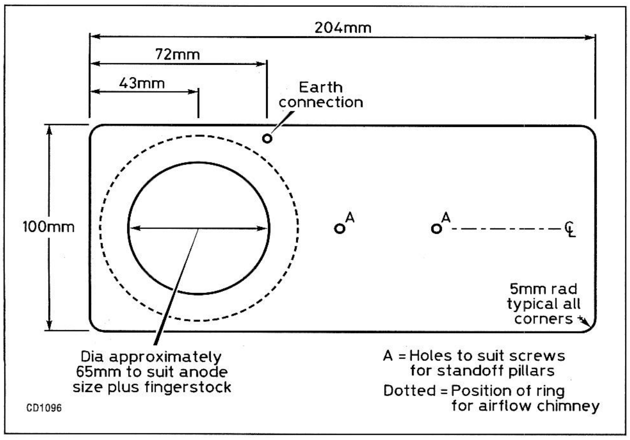

Fig 4.217. Anode transmission line L7

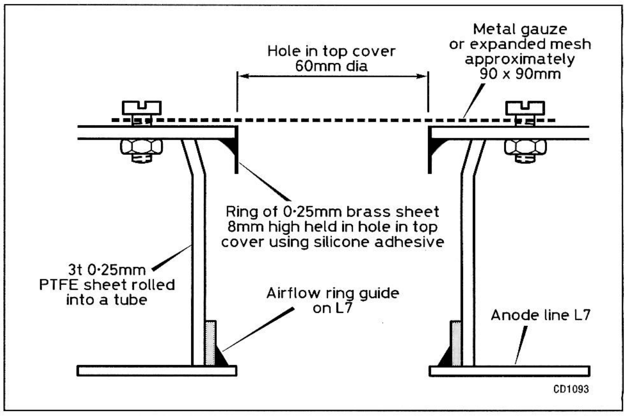

Fig 4.219. Airflow chimney

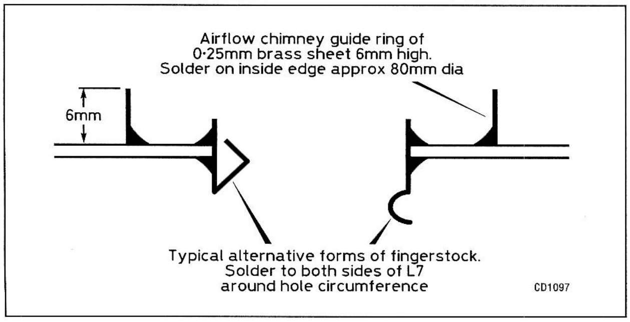

Fig 4.218. Cross-section through anode clearance hole

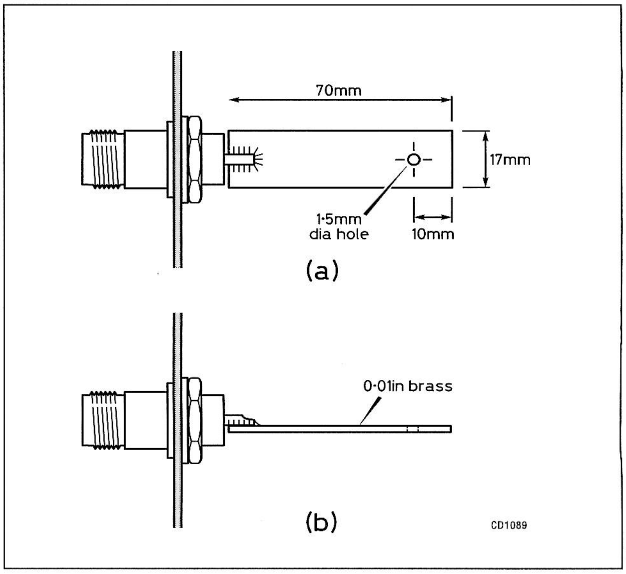

Fig 4.220. Output coupling flap C7

article is enormously detailed and emphasises achieving maximum output power.

The G3SEK/G4PMK design concentrates on maximising gain at modest output power, making it ideal to follow a transverter. In this design the depth of the cavity is fixed at 0.75in* and a coarse tuning screw is provided as described by

*All dimensions in this design are given in inches, because at the time brass strip and tubing were more readily available in inch than in metric sizes.

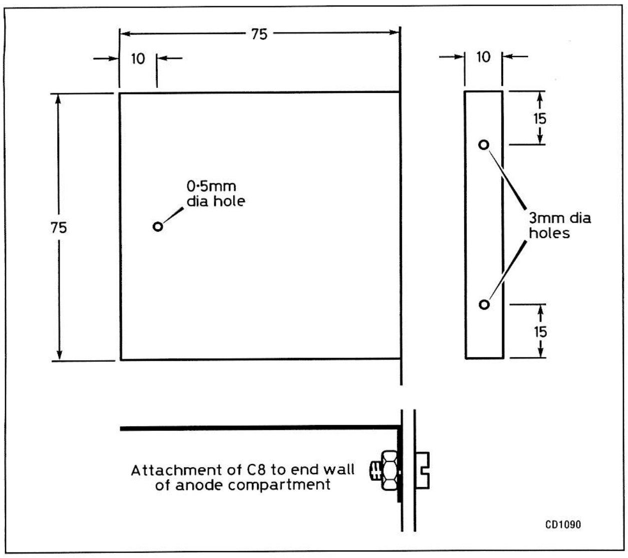

Fig 4.221. Tuning flap C8, 0.015 bronze or brass

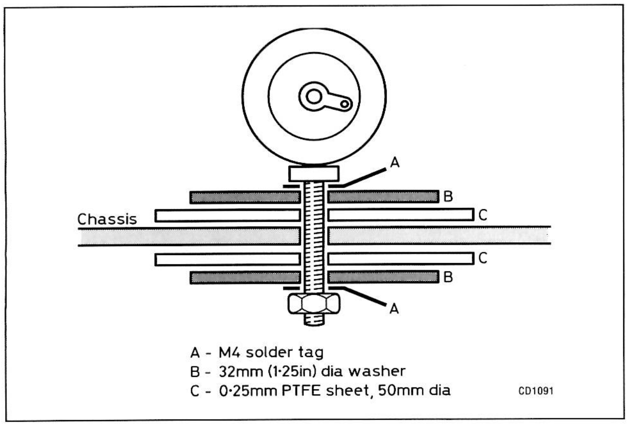

Fig 4.222. HT capacitor C6

Table 4.24. Components list for Fig 4.214

C1	5p variable
C2	15p variable
C3–5	1000p feedthrough
C6	EHT feedthrough capacitor – see Fig 4.222
C7	0.25mm brass sheet 17 x 70mm – see Fig 4.220
C8	0.4mm (0.015in) brass or bronze sheet – see Fig 4.221
L1	2mm enamelled copper wire, 15mm long
L2	2mm enamelled copper wire, 30mm long
L3	0.25mm brass plate 40 x 34mm
L4–6	10t 2mm enam copper wire, 6.5mm ID
L7	1.6mm double-sided PCB 204 x 100mm – see Fig 4.217
L8	4t 1mm enam copper wire, 6.5mm ID
R1	10k 1W
D1	5.6V 10W zener diode
D2	50V 30A diode (flashover protection)
M1	100mA
M2	1A
S1	close to transmit

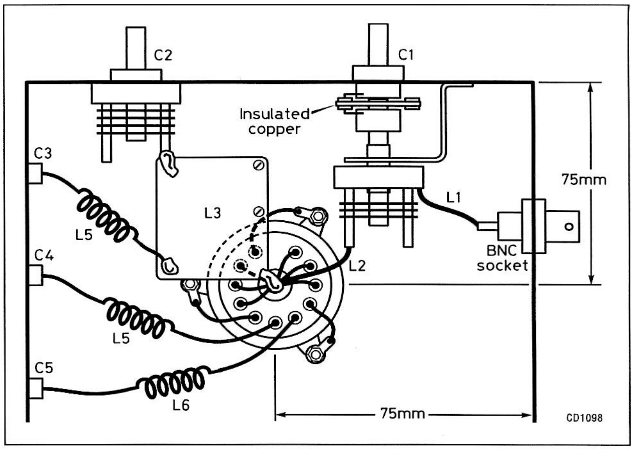

Fig 4.223. Grid circuit. C2 is two fixed, three moving vanes, air-spaced 0.8mm

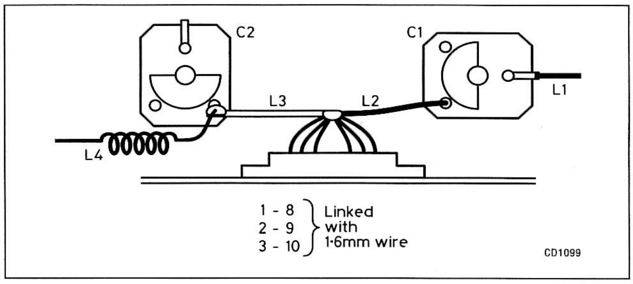

Fig 4.224. Partial section view showing cathode connections

Table 4.25. Typical operating conditions for 3CX800 amplifier

Anode voltage on load	2100V
Anode current	550mA
RF output power	595W
Grid current	25mA (varies between valves)
Drive power	25W
Quiescent current	50mA (varies between valves)

G3LTF and G3WDG [84]. The base plate is soldered to the walls, and the top plate, which regrettably has to be removable for access, is secured by no less than 20 screws. The result is a rigid, low-loss assembly.

There are several valves in the '2C39' family, which can differ mechanically as well as electrically. Not all variants have the grid sleeve: the other example in Fig 4.226(b) (a 7289/3CX100A5) has its grid contact ring only at the bottom (cathode) end. Inspection of the drawings for a professional amplifier shows that if the grid contact is made at the bottom of the grid sleeve and in the base plane of a 0.75in-deep cavity, then the bottom of the anode sleeve is flush with the top of the cavity.

This arrangement, shown schematically in Fig 4.226(b), was taken to be the 'correct' way to locate the valve in the cavity.

Grid and anode contacts

The greatest single problem in amateur designs using the 2C39 series of valves has always been the contact rings for the grid and anode. To insist on an extremely low-inductance grid contact, in the plane of the base plate, makes matters worse than ever!

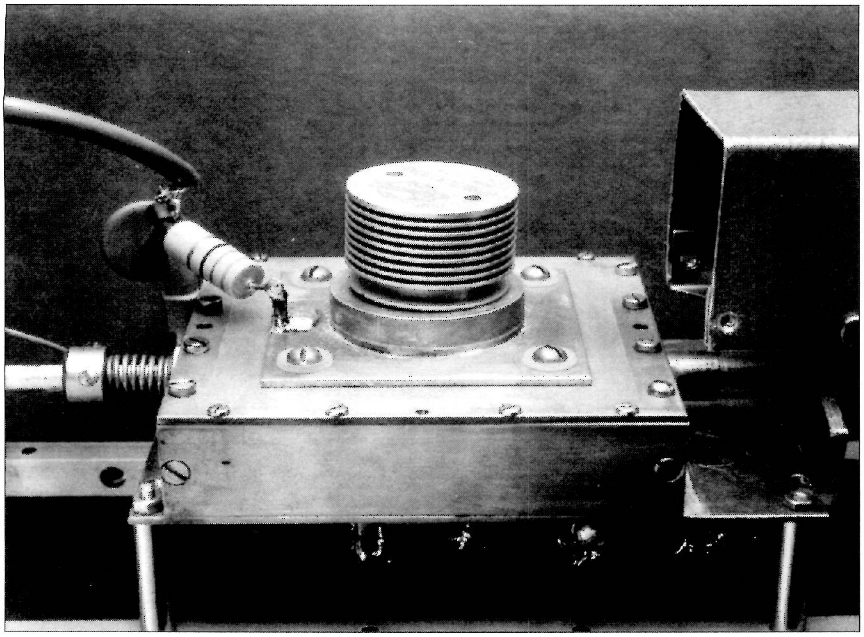

Fig 4.225. The G4PMK/G3SEK 1.3GHz amplifier

'Straight' finger-stock (Fig 4.226(a)) is ruled out because it projects either into or out of the anode cavity. Folded-over finger-stock is used in commercial preformed grid rings, but it is not readily available; and N6CA's experiments [85] suggest that the inductance of the resulting contact is barely low enough.

The solution is to use a ring of spiral spring to contact the valve, the spring-ring itself being held in a collet (Figs 4.226(b), 4.227 and 4.228). In effect, the valve is contacted by several quarter-turns of the spring, all of which are electrically in parallel and combine to make a contact of extremely low inductance. The collet can be let into the base of the cavity, so that the contact is made in the correct plane.

The ideal spring-ring material is a loosely-wound, silver-plated spiral spring of about 0.25in diameter. A perfectly acceptable home-made substitute is a spiral wound from narrow (eg 0.1in wide) phosphor-bronze strip such as draught excluder; this gives fewer contacts to the valve but each turn of the strip has lower individual inductance. Fig 4.227 shows the two alternative types of spring-ring in their collets.

Precise dimensions of the collet depend to a large extent on the available spring-ring material, and the prototypes were turned by 'cut-and-try' out of old brass vacuum fittings. The first step is to bore out the blank to just clear the grid sleeve of the valve. Then the internal groove is formed using a small boring tool (inset, Fig 4.228), repeatedly trying first the spring-ring alone for size, and in the later stages both the spring-ring and the valve. The fit of the valve can also be adjusted by pulling or squeezing the spring-ring. When all is well, the valve will be gripped gently but uniformly as it is twisted into place. Owing to the 'lay' of the turns of the spring-ring, the valve can only be twisted in one direction – the same for insertion and removal – so if spring-rings are used for both the grid and anode connectors they must be wound in the same sense.

The entire machining and fitting process is far easier than it looks, because no individual dimension is critical. The two

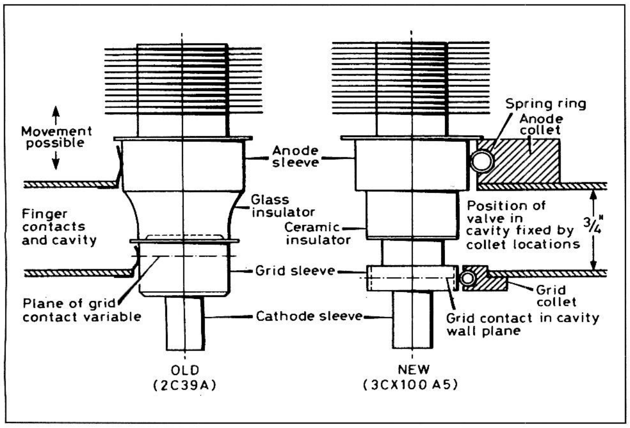

Fig 4.226. Old and new methods of mounting valves in a cavity (omitting details of anode DC supply)

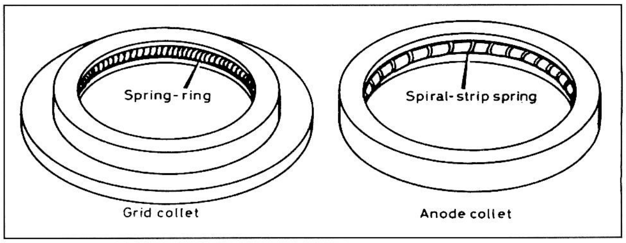

Fig 4.227. Grid and anode collets showing the two alternative types of spring material

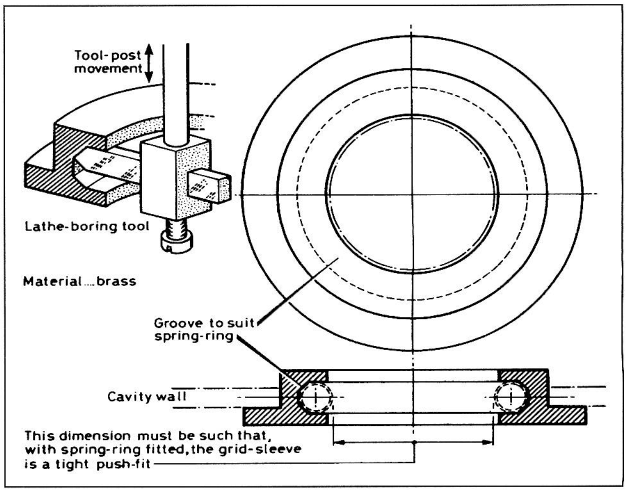

Fig 4.228. Details of grid collet

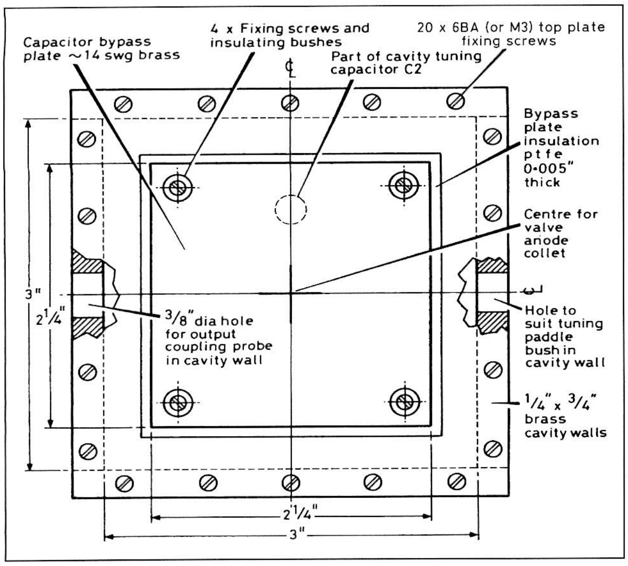

Fig 4.229. Top view of anode cavity assembly. The cavity top-plate has a 1.25in diameter hole in the centre to clear the valve anode sleeve

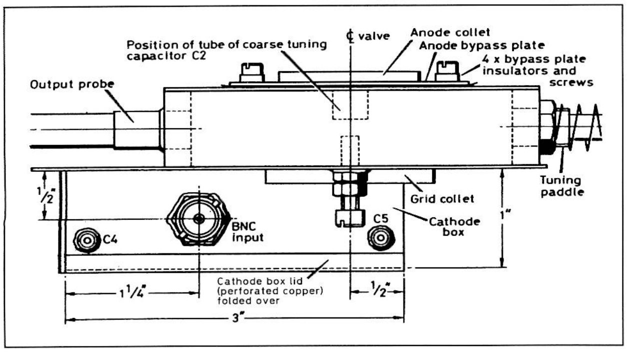

Fig 4.230. Side view

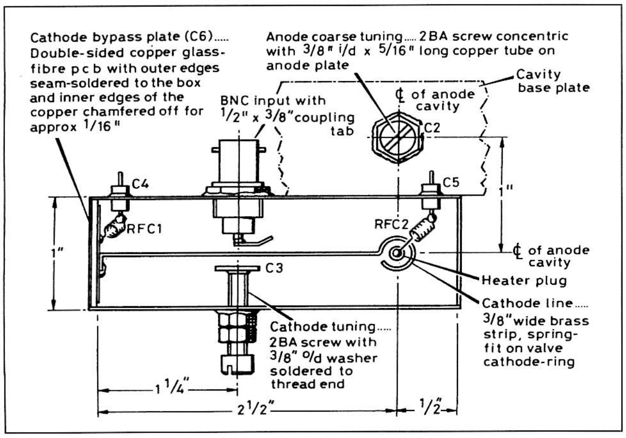

Fig 4.231. Details of cathode box and anode coarse tuning capacitor

prototype grid contact assemblies were produced in one lunch-time by a machinist with no delusions of competence!

The anode contact is much more forgiving of stray

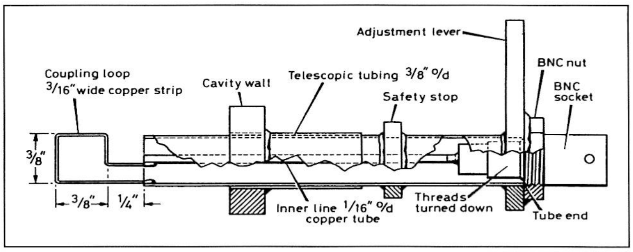

Fig 4.232. Output coupling probe

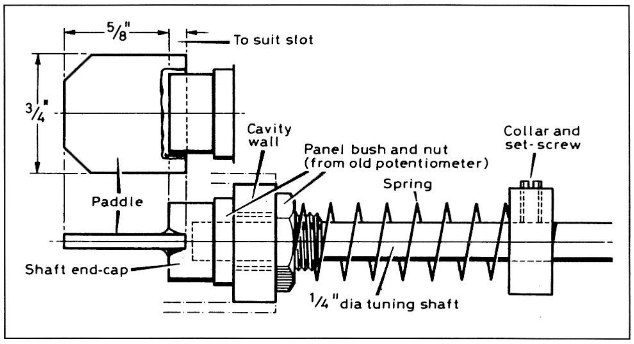

Fig 4.233. Tuning paddle

inductance than the grid contact, so a ring of ordinary finger-stock would probably suffice [84].

Correct axial location of the valve can be ensured if the flare at the top of the anode ring bears on the top of the collet, the latter being turned down to the correct thickness.

The cathode input circuitry closely follows an early design [84], the coarse-tuning screw has already been mentioned, and the fine-tuning paddle and coupling loop are also as before. Coupling with the magnetic field in the cavity is strongest when the loop is almost fully withdrawn to the cavity wall and at right-angles to the base plate. Coarse loading adjustment is by sliding the loop in and out, and fine adjustment by rotating it.

Dimensions and assembly

A general view of the amplifier is shown in the photograph, and leading dimensions are given in Figs 4.229–4.233. Non-critical dimensions are not given, being at the discretion of the constructor. Over-dimensioning the drawing would give a false impression that all dimensions must be slavishly followed: if that were true, we probably could not have built the prototypes! As noted earlier, the dimensions of the anode and grid connectors are only critical in that they must be adjusted to provide a good fit to the valve. However, the two collets must be coaxial in order to avoid shear forces on the valve, and detailed assembly instructions for the anode cavity are as follows:

1. Mark out the locations of the side walls and the centre of the cavity on the base plate. Drill a *pilot* hole in the centre of the base plate.
2. Solder the side walls into position. If the ends of the bars can be faced-off square (not impossible by hand or with a three-jaw lathe chuck) they may be pre-assembled into a square frame before soldering. After soldering, stone the top face of the side-walls flat.

3. Mark out and pilot drill the 20 fixing holes in the cavity top plate. Do *not* drill the centre hole yet. Tape the top plate accurately into position on the side walls, and on a drill press *drill two holes* in diagonally-opposite positions through the top plate and into the side walls. Tap these two holes and secure the top plate more firmly before drilling and tapping the rest of the fixing holes.
4. Again on a drill press, drill square through the pilot hole in the base plate, and through the top plate.
5. Use the pilot hole in the cavity top plate to locate the centre of the anode bypass plate, when marking and drilling through the latter for the four retaining screws.
6. Open out all pilot holes to full size. Be careful to retain concentricity.
7. Drill the four holes in the anode bypass plate slightly oversize for the shoulders of the available insulating bushes. Leave the retaining screws slack until the valve has been fitted squarely into place for the first time; then tighten them.

The cathode circuitry below the base plate (Figs 4.230 and 4.231) is assembled after the grid collet has been soldered into place.

Rather than fabricating the RF bypass capacitor for the 'cold' end of the cathode stripline [84] G4PMK and G3SEK chose to make the entire end wall act as a capacitor by making it from double-sided glass-fibre PC board (Fig 4.231), chamfering the copper from the inside edges to prevent a DC short-circuit.

The sliding loop coupling probe (Fig 4.232) is made using telescoping brass tubing available from good model shops. A safety stop must be provided to prevent the loop from touching the anode sleeve of the valve. The tuning paddle (Fig 4.233) needs to be well grounded to RF; this can be ensured by a strong compression spring over the shaft which maintains a firm contact between the paddle and shaft bushing. It is helpful if the external controls indicate the true orientations of the loop and paddle within the cavity.

As an optional extra, all the components can be silver-plated. The brass parts of the two prototypes were given an ultra-thin but tenacious coating of silver by the following method, which is a simple and effective way of applying a very thin but tenacious coating of silver to copper or brass, without resorting to electroplating and cyanide solutions. Although the coating is extremely thin, possibly less than the 'skin depth' for RF currents at 1.3GHz, tarnishing of silver affects its electrical properties far less than would tarnishing of untreated copper or brass, so the coating is worthwhile if only as a preservative.

Mix together two parts by weight of finely ground sodium chloride (common salt), two of potassium hydrogen tartrate (cream of tartar) and one of silver chloride. Store the mixture away from moisture or strong sunlight. To silver-plate an article, dampen a little of the powder with water and apply the resulting paste with a cloth using a vigorous rubbing action (wear rubber gloves). The abrasive nature of the paste will help remove any slight tarnish. When finished, wash the article thoroughly and dry it.

Cooling

The multiple fins of the anode cooler present a large surface area for efficient heat transfer, though only if the cooling air is forced between them. If air is merely blown in their general direction, it will take the easy way round the outside and will not cool the anode! A transverse-finned cooler is not as effective as a ducted axial-flow cooler (eg that of the 4CX250B) but a suitable air duct can be made from a variety of easily worked materials such as Perspex or Formica.

At the higher power levels, overheating of the grid can cause electron emission, leading to DC instability and shortened valve life. The problem can be avoided by efficient cooling of the anode (which otherwise tends to heat up the whole valve) and of the grid/cathode region. The spring-ring grid connector is a good conductor of heat as well as RF, and helps keep the grid cool. If the whole amplifier is mounted upside down, the cathode cavity is adequately ventilated by natural convection through the perforated cover. The anode cavities of the prototype amplifiers were not ventilated at all, a point which should be considered if this design were to be used at power levels of more than a few tens of watts.

All amplifiers of this general type can suffer from the problem that the different thermal loadings on transmit and receive lead to changes in the internal capacitances of the valve, and hence to drift in the output level as the amplifier warms up. The problem is obvious enough at high power levels, but it also occurs at very low drive levels because the valve is only lightly loading the high-*Q* cavity, making it more susceptible to drift. The difficulties can be largely overcome by cooling the valve adequately on transmit and reducing or removing the airflow on receive [86] so that its temperature remains more nearly constant. For example, it is possible to set the tuning paddle either so that the amplifier achieves maximum output within a few seconds and then drifts off tune after about one minute, or alternatively so that it takes about 20 seconds to reach full output and stays in tune for several minutes – for contests and ragchews respectively! A further improvement could be expected from the use of one of the modern temperature-compensated derivatives of the 2C39, eg the 7855.

Operating conditions

The circuit diagram of the amplifier is very simple (Fig 4.234). For maximum gain, a fairly high standing current of the order of 50mA is required, ie DC efficiency has to be sacrificed. At low drive levels the amplifier will operate at virtually constant anode current, so simple cathode-resistor biasing will suffice. During development of the amplifiers a 250Ω wirewound potentiometer proved perfectly satisfactory, and a 22kΩ resistor connected from the cathode bypass to ground allows the valve to cut off safely during receive periods or if the bias resistor fails. At higher drive levels, constant-voltage biasing must be used in order to maintain linearity on SSB, and an arrangement in which a single transistor acts as both bias regulator and T/R switch is shown in Fig 4.234. The zener diode sets the cut-off bias on receive and limits the transistor's collector voltage to below V_{ceo}.

The usual precautions regarding heater voltage should be observed when using the amplifier; at no time must the heater voltage exceed 6V. It is important that the cathode of the valve be allowed to reach full operating temperature before the anode voltage is applied. A delay of 60–90 seconds is adequate.

The power gain achievable will depend on the type of valve, and on its operating history if it is second-hand. One of the

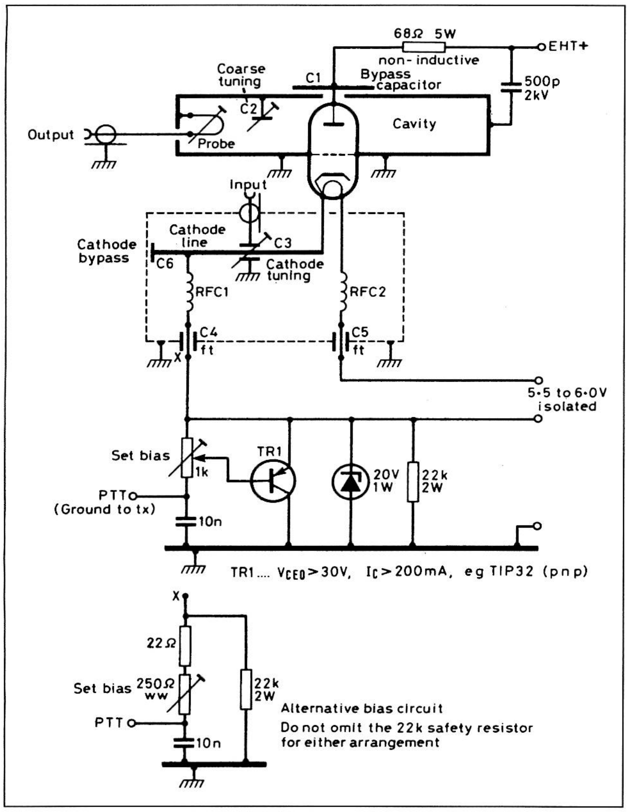

Fig 4.234. Circuit diagram of the amplifier

prototype amplifiers, using a 'good' but not remarkable 7289, gave the following measured performance with an EHT supply of 1kV, when the input and output matching were re-optimised to suit the available level of drive power.

Available drive power (W)	Output power (W)	Power gain (dB)
0.35	27	19
0.5	32	18
1.0	40	16

The lower power gains at higher drive levels do not imply that the amplifier is non-linear. If the adjustments had been optimised at the 1W drive level, then reduction of drive to 0.5W would linearly halve the output power to 20W. However, if no more than 0.5W drive were available, the amplifier could be reoptimised to give 32W of linear RF output. The prototypes were developed using EHT supplies of 1–1.1kV. Some reduction in gain was found at 800V, and in a brief test using 1.5kV one of the prototypes gave 60W RF output for 1W of drive.

An alternative anode circuit [87] using a $\lambda/2$ stripline is shown in Figs 4.235–4.238. This offers reduced gain, output power and efficiency compared with the cavity design, but it also demands less effort and metalwork in construction.

The anode circuitry fits inside a standard die-cast box (Eddystone 26908PSL) and the amplifier is assembled onto a 1.6mm copper plate which replaces the box lid (the plate fits over, not inside, the box). The plate and lid need to be in good contact for best performance and this is aided by making a gasket from thin copper foil folded into four layers. Grid

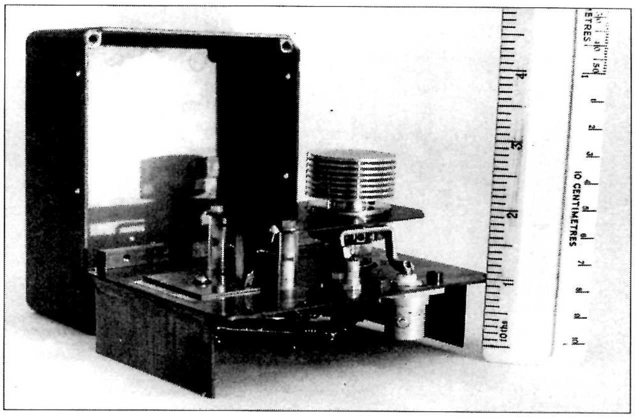

Fig 4.235. The tripler

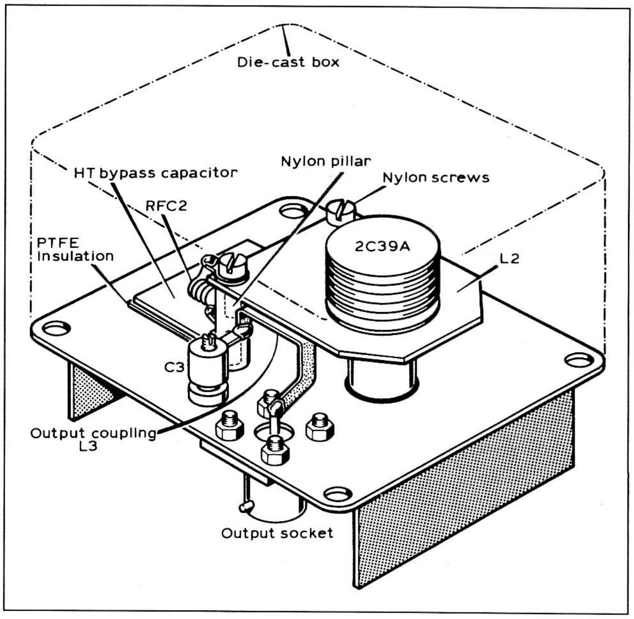

Fig 4.236. General arrangement of anode circuit

and cathode circuitry can be as the previous example. The support pillars for the anode line are 21mm high, and are preferably PTFE or ceramic rather than the nylon shown in Fig 4.236. With the alternative output probe G4DGU [88] reported 75W output at 8dB gain from 900V anode supply. At these power levels, forced cooling is essential and screened holes will have to be provided in the box sides to allow a blower to supply cooling air.

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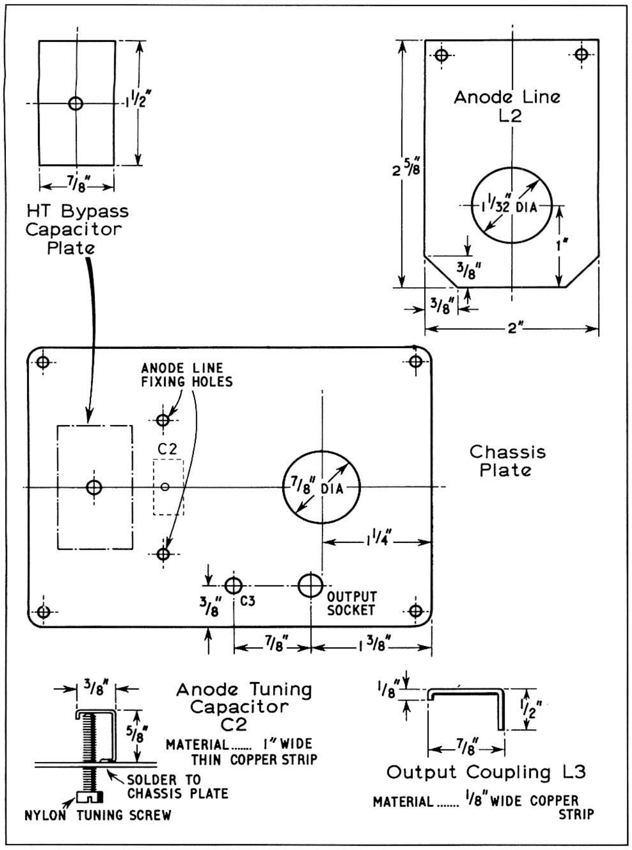

Fig 4.237. Construction details of anode circuit

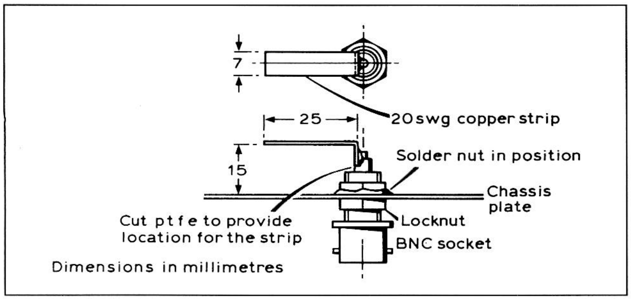

Fig 4.238. An alternative output probe and position

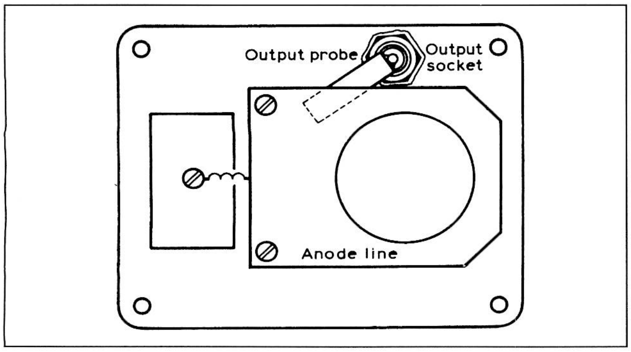

Fig 4.239. Approximate position of the output probe for 80Ω output impedance

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5 Antennas and transmission lines

THE antenna and its associated feeder or transmission line are arguably the most important elements of any VHF or UHF station, but are frequently considered least in its assembly. Without a good antenna system, and equally good feeder arrangements, much of the RF power generated by the transmitter will be dissipated as heat before it reaches the antenna. If the antenna is badly sited, or unsuitable for the location, the power reaching it may be radiated in the wrong direction, or scattered from other nearby antennas and structures, again wasting the RF generated by the transmitter. The problems of line loss increase with increasing frequency, and greater care is required to make the most of the power available – and it is generally more difficult (or expensive) to generate more power as the frequency increases.

Similar arguments apply to reception – why throw away a large percentage of the signal captured by the antenna before it reaches the receiver? Solutions using mast-head preamplifiers are discussed in other parts of this book, but they entail some complexity and expense if the antenna is to be used for both transmission and reception.

Of course, it is not always possible to find and erect an ideal antenna in an ideal location, particularly in an urban environment. This chapter describes the principles of antenna and transmission lines for the VHF and UHF amateur bands, such that the reader can make informed choices about installations that are both practical and effective for their environment. The designs for home construction are practical, and can be built with confidence that they will work *if the details are followed closely*. However, at VHF/UHF small deviations from detail can affect the performance of the antenna quite markedly, and if suitable measuring apparatus is not available, much frustration can ensue in trying to make the antenna work. This chapter includes some more complex antennas that require some measuring facilities to optimise performance after construction; these are for the enthusiastic antenna experimenter.

However, before tackling the selection and installation of antennas, it is useful to understand some of the underlying theory of their operation and use.

ANTENNA FUNDAMENTALS

All antennas have certain basic properties which can be well defined. These are:

- Radiation pattern
- Polarisation
- Gain
- Input impedance
- Impedance and radiation pattern bandwidths

Of these, the radiation pattern is generally the most basic parameter used for selecting an antenna for a particular purpose. Many amateurs use highly directional antennas which require rotation to point in the desired direction of communication. However, antennas providing all-round (omnidirectional) coverage are usually required for repeater stations, and are of course essential for vehicle installations.

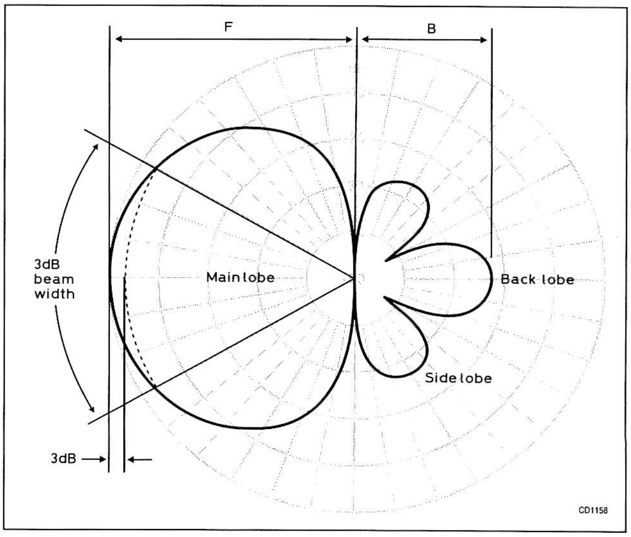


Fig 5.1. Typical polar diagram of a VHF Yagi antenna

Radiation pattern

The *radiation pattern* describes the spatial distribution of the power radiated by the antenna, that is, the directions in which the signal is transmitted or from which it is received. The key characteristics of directional antennas are usually expressed as the beamwidth in two principal planes at right-angles to each other, known as the *E plane* and the *H plane*, and described in the next section. The beamwidth in these principal planes is usually defined as the angle including the main beam at which the radiated energy falls to one-half the maximum level. This is called the *half-power beamwidth*, and the points on the radiation pattern are often called the *3dB* or *half-power points* of the radiation pattern, being 3dB below the main beam. See Fig 5.1.

Fig 5.1 shows the spatial distribution of power in one principal plane for a typical directional VHF antenna. This representation is called a *polar diagram*, where the power in a given direction is indicated by the distance of the curve from the centre of the diagram (in this case, where the *lobes* touch), and can be thought of as a plan view of the variation in radiated power. The radiation pattern can be presented on a variety of polar diagram charts, the principal difference being the

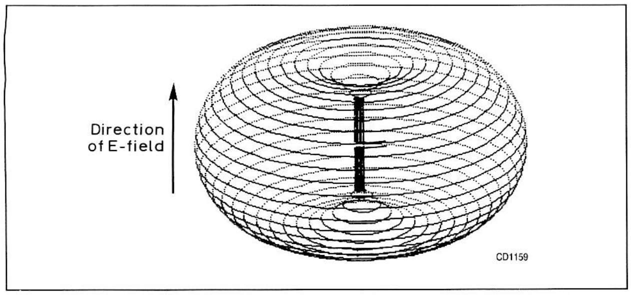


Fig 5.2. Radiation from, and polarisation of, a dipole antenna

arrangement of the radial scale. The most usual forms are *linear*, where the radius of the pattern is directly proportional to the radiated power in a given direction, and *logarithmic*, where the radius represents the relative power in decibels.

Key features of the radiation patterns of the antenna shown in Fig 5.1 are the main lobe or main beam, and the presence of several minor lobes including one pointing in the opposite direction to the main lobe. The *front-to-back (F/B) ratio* is the ratio of the energy radiated by the peak of the main lobe to that in the opposite direction, and is often used as an estimate of the 'goodness' of a beam antenna. This ratio is usually expressed in decibels. As more power is radiated in minor lobes, less power is available in the main lobe, and the *gain* of the antenna is reduced (see below). Omnidirectional antennas ideally have a circular radiation pattern in one plane, but will still have a shaped radiation pattern in the other principal plane, at right-angles to the first.

The linearly scaled graph is useful for measuring the beamwidth of the main lobe accurately, whereas the logarithmically scaled chart more clearly shows the levels of the sidelobes, which may be barely visible on the linear chart. The ARRL has promoted the use of a hybrid chart which combines features of both types of graph, by using a quasi-logarithmic radial scale marked in decibels [2].

Polarisation

Radio waves comprise both electric and magnetic fields mutually coupled at right-angles to each other and at right-angles to the direction of propagation. From this derive the two *principal planes* used in describing radiation patterns; the *E-plane* lies parallel to the electric vector or E-field in the main lobe, and the *H-plane* lies parallel to the magnetic vector or H-field in the main lobe. Accordingly, these two principal planes will be at right-angles to each other.

The polarisation of an antenna is defined in terms of the orientation of the electric field vector in the direction of maximum radiation. The maximum radiation from a dipole occurs in a plane bisecting its centre and at right-angles to the dipole arms. The electric field vector in this plane lies parallel to the arms of the dipole, see Fig 5.2.

Thus a dipole mounted horizontally above the ground is said to radiate *horizontally polarised* signals, and the same dipole mounted vertically would radiate *vertically polarised* signals.

Whilst many amateurs use vertical or horizontal *linear* polarisation for terrestrial communications, satellite users often use *circular* polarisation to reduce the effects of propagation, ground reflections or the spinning of the satellites on the signals. The effect of circular polarisation can be visualised as a signal emanating from a dipole rotating about its centre at the frequency of radiation. The tip of the electric vector traces out a corkscrew as it propagates away from the antenna, and like a corkscrew, the polarisation is described as *right-* or *left-handed circular*, dependent on the direction of rotation of the electric vector as seen from the transmitter.

A fixed linear dipole will receive an equal signal from a circularly polarised wave whether it is mounted vertically, horizontally or in an intermediate position. The signal strength will be 3dB less than if a circularly polarised antenna of the same sense is used; however, a circularly polarised antenna of the opposite sense will receive no signals. Both these effects are due to *polarisation mismatch* between the wave and the receive antenna.

In practice, an antenna may radiate unwanted polarisations in a variety of directions, including the main lobe. This is called *cross-polarised radiation*, and for linearly polarised antennas it will be perpendicular to the wanted radiation. In circularly polarised antennas, the cross-polarised element is that part of the signal that is radiated as circular polarisation of the opposite sense to that intended. The relationship between wanted and unwanted signals is often expressed as an *axial ratio* or *ellipticity*, the definitions of which can be found in reference [1]. The smaller this figure, the better.

It is worth noting that with linearly polarised antennas, particularly beams with complex polar diagrams, radiation from the sidelobes can be of the opposite polarisation to the main beam, and will often be complex or elliptical, especially outside the principal planes. Hence the reception of signals from a cross-polarised station may often be stronger with the beam pointing away from, or at an angle to, the transmitting station.

Gain and directivity

The gain of an antenna is a basic property which is frequently used as a figure of merit. It is defined as the maximum signal radiated in a given direction relative to that of an *isotropic radiator* fed with the same power. An isotropic radiator is a hypothetical, lossless antenna which radiates equally in all directions. In practice, a half-wave dipole is often used as the reference radiator; if the dipole is lossless, it has a maximum gain of 1.64 (or $10 \times \log_{10} 1.64 = 2.15$ decibels) relative to the isotropic antenna.

The *directivity* of an antenna is defined purely in terms of its radiation pattern, as the radiation intensity in a given direction to the radiation intensity averaged over all directions. A practical antenna may have good *directivity*, but poor *gain* if the antenna is lossy through poor design, use of lossy components or poor mechanical construction. If the antenna is lossless, the gain and directivity will be the same.

High directivity is achieved by compressing or focusing the radiated power into a small *solid angle*, the product of the half-power beamwidths in the two principal planes. An isotropic radiator radiates equally in all directions, which can be imagined as equal illumination over the surface of a sphere. A good, directive VHF antenna will confine most of its radiation to a few tens of degrees around the main beam, corresponding to the beam of a pencil torch illuminating the inside of the sphere.

If antenna losses are small, and the side and backlobes are also much smaller than the main lobe (which we would expect for a well-designed beam antenna), there is an approximate formula which relates the 3dB beamwidth of the antenna in the two principal planes (E and H) to the gain of the antenna:

$$\text{Gain relative to a } \lambda/2 \text{ dipole} = \frac{27{,}000}{\theta_E \theta_H}$$

where θ_E is the angular width, in degrees, between the half-power points in the E-plane, and θ_H is the angular width, in degrees, between the half-power points in the H-plane. The gain can be expressed in decibels relative to a half-wave dipole (dBD) by taking the logarithm of the expression:

$$G_{dBD} = 10\log_{10}\left[\frac{27{,}000}{\theta_E \theta_H}\right]$$

This formula is reasonably accurate (within 2dB) for well-designed, efficient antennas with gains greater than 10dBi (8dBD), and can be useful for estimating the beamwidth where a radiation pattern is only available for one plane and the gain of the antenna is also known. The gain in decibels relative to an *isotropic* radiator is found by adding 2.15 to G_{dBD}.

Input impedance

The impedance presented at the feedpoint by an antenna is a complex function of the size and shape of the antenna, the frequency of operation, and its environment. The impedance is affected by the proximity of other conducting objects, where the induction of RF currents alters the impedance through *mutual coupling* between the antenna and object. The elements of a Yagi antenna are mutually coupled together, and the driven element would present a very different impedance if measured in isolation from the rest of the structure.

Input impedance is usually complex. The resistive part is composed of the radiation resistance, which can be thought of as dissipating power by radiating it as electromagnetic energy (desirable), and loss resistance, which dissipates power as heat (not desirable). The reactive part arises from the behaviour of antenna elements as resonators, or tuned circuits, and it can change rapidly with variations of frequency.

Impedance bandwidth

The impedance bandwidth of an antenna is defined as the frequency range over which the antenna impedance results in a voltage standing wave ratio (VSWR) less than some arbitrary limit. This may be typically 1.5:1 for amateur operation with solid-state transmitters or higher values for other applications. Ideally, an antenna should be impedance matched to the feedline and thence to the transmitter or receiver. Although *tuned feed* arrangements are sometimes used at HF, where a high standing wave ratio may be acceptable on the feedline, the losses in VHF feeders and tuning components usually preclude this approach at VHF and UHF. Feeders and matching arrangements are discussed later in this chapter.

Radiation pattern bandwidth

Antenna radiation patterns are also dependent upon the operating frequency. Using the analogy of the Yagi antenna's elements as tuned circuits, the loss of resonance away from the design frequency results in small currents in the elements and a consequently severe loss of performance. In the case of the Yagi antenna this results in a sharp decrease of gain, and destruction of the desired radiation pattern. For beam antennas, such as the Yagi, the radiation pattern bandwidth is often defined as the frequency range over which the main lobe gain decreases to 1dB less than its maximum.

It should be noted that the impedance bandwidth and radiation pattern bandwidth are independent of each other. It is quite possible for the impedance bandwidth to be greater than the radiation pattern bandwidth, especially with high-gain antennas, and to be able to feed power into an antenna that then wastes it by radiating it in other than the desired direction!

ANTENNA ARRAYS

Purpose

The gain achievable with any antenna structure is ultimately limited by the fundamentals of its operation. However, higher gains can be achieved by using several antenna elements in an *array*. The array can comprise antennas *stacked* vertically above each other, or arranged side by side in *bays*, or a combination of both. These are *broadside* arrays, where most of the radiated power is projected at right-angles to the plane in which the elements lie. An array can also be formed where the main beam is projected along the array of elements; these are *endfire* arrays, of which the HB9CV and Yagi antennas are examples.

An array of elements has a narrower beamwidth, and hence a higher gain, than the individual antennas. The maximum achievable gain could be N times greater than one element fed with the same power ($10 \log_{10} N$ decibels) if there are N elements in the array. However, more complex feed arrangements can reduce the VSWR bandwidth and introduce losses, reducing the array gain. Arrays need care in construction and attention to detail, especially at UHF and above, but the results reward the effort expended.

Broadside arrays

If a pair of identical isotropic radiators or *point sources* are fed in phase with equal power, an interference pattern will be set up. The field will be a maximum at right-angles to the array. However, cancellation will occur at other angles where the wave from one antenna has travelled an odd number of half-wavelengths further than from the other antenna – see Fig 5.3.

The same principle can be applied to more than two point sources, which if they are equally spaced, produce a radiation pattern of the form:

$$E_\vartheta = \left[\frac{\sin\left(\frac{N}{2}\psi\right)}{\sin\left(\frac{1}{2}\psi\right)}\right]$$

$$\psi = 2\pi d \sin\theta + \beta$$

where θ is the angle measured normal to the line of the array, d is the separation of the point sources in wavelengths, β is the phase difference between the elements (usually zero) and N is the number of elements.

The electric field at any angle to the array is calculated by adding together the fields from each point source, taking into

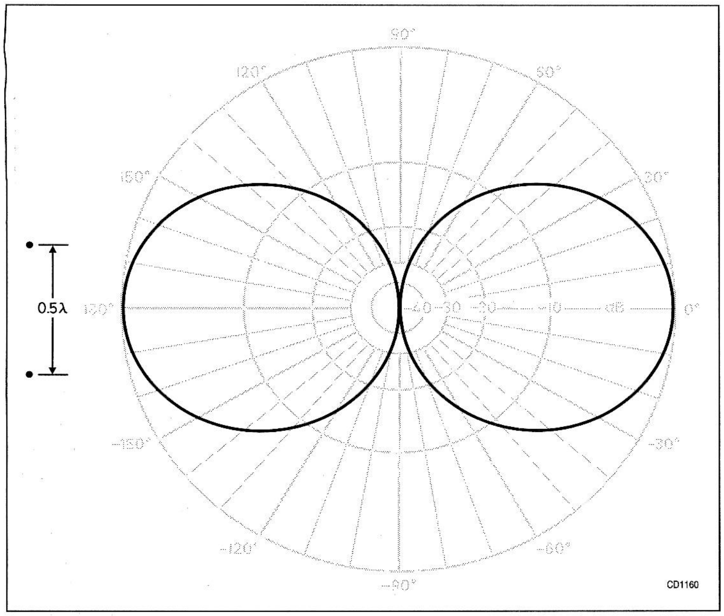


Fig 5.3. Array of two point sources and their radiation pattern

Table 5.1. Calculations for the radiation pattern of a four-element collinear dipole array

Angle relative to normal (°)	Array factor factor	Dipole pattern (voltage)	Radiation pattern
0	4.00	1.00	4.00
5	3.74	0.99	3.71
10	3.00	0.98	2.94
15	1.98	0.95	1.88
20	0.89	0.91	0.81
25	−0.06	0.87	−0.05
30	−0.73	0.82	−0.59
35	−1.05	0.76	0.80
40	−1.06	0.69	−0.74

account the phase delay of the field from each source, ie *vector addition* of the fields. The resulting pattern for arrays of point sources is often known as an *array factor.*

In a practical array, each point source will be replaced by a real antenna, a half-wave dipole or perhaps a Yagi antenna. Provided that all the antennas are pointing in the same direction, the pattern produced by such an array can be found by multiplying together, at each angle of interest, the array factor just calculated and the *pattern factor* or radiation pattern of the antennas used in the array.

This means that any nulls in the element pattern will be reproduced in the array pattern, which is why arrays of complex elements often result in many sidelobes.

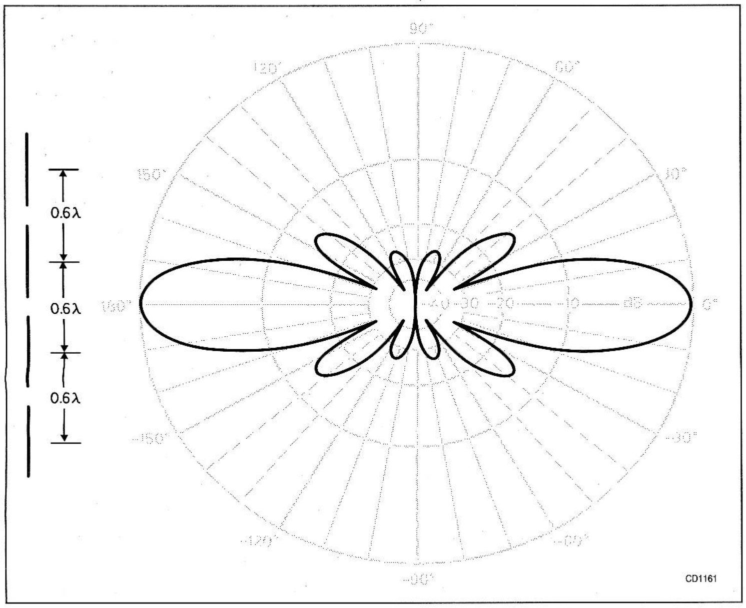


Fig 5.4. Plotted radiation pattern for a four-element collinear dipole array

The radiation pattern of such an array can be calculated by programming a computer, in a spreadsheet, or by hand by setting out a table as shown in Table 5.1.

Note that if working from antenna patterns or data expressed in decibels, the directivity at each angle *must* be converted to a fraction before multiplying by the array factor. For example, if the directivity at a given angle is −2.8dB relative to the peak of the main beam, the directivity in linear terms is:

$$10^{-2.8/20} = 0.7244$$

or $\text{antilog}_{10}(-2.8/20)$ for those with log tables. It is easiest to scale the directivity to the peak of the main lobe prior to carrying out the calculation.

Endfire arrays

If a pair of point source antennas are separated by one quarter-wavelength, and are fed with a phase difference of 90° between them, the radiated field from one antenna will reinforce that of the second in one direction, and will completely cancel the field from the second in the opposite direction – see Fig 5.5.

The equation for the radiation pattern is the same as shown for the broadside array above, except that the phase angle β is now 90°. Other spacings may be used, provided that the phase difference is adjusted to ensure that the radiation is cancelled in the desired direction. Antennas such as the HB9CV (shown later in this chapter) use this technique to provide directivity and a good front-to-back ratio from mechanically compact structures. Both elements are fed by the transmitter. The Yagi antenna generates its radiation pattern using similar phasing principles, but only one element is fed as described later in this chapter.

Antenna array theory can be found in almost any book devoted to antennas. However, a good treatment with many radiation pattern examples can be found in references [3] and [4].

Practical considerations and limitations of arrays

Stacking separation

High gain cannot be achieved by simply stacking many elements close together. If we consider a

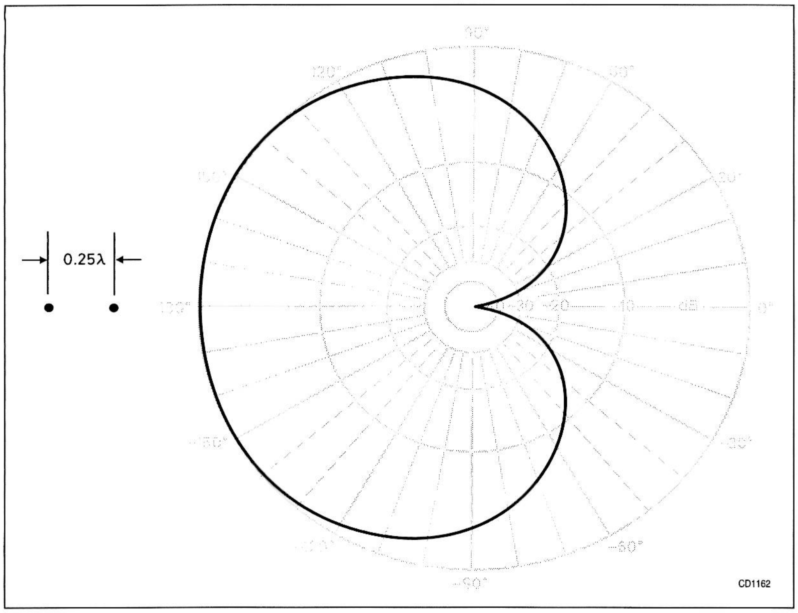


Fig 5.5. Radiation pattern of two point sources separated by one quarter-wavelength and fed in quadrature

dipole collecting power from an incident field for delivery to a load (receiver), it can be thought of as having a collecting area or *effective aperture* that is somewhat larger than the dipole itself. The higher the directivity of the antenna, the larger the effective aperture, as given by the relationship:

$$A_{eff} = \frac{\lambda^2}{4\pi} D$$

where D is the directivity of the antenna and λ is the working wavelength.

If the effective apertures of adjacent antennas overlap, the incoming RF energy is shared between them, and the maximum possible directivity (or gain) of the elements cannot be attained.

The optimum stacking distance is a function of the half-power beamwidth of the elements in the array, and is given by:

$$S_{opt} = \frac{\lambda}{\left[2\sin\left(\frac{\phi}{2}\right)\right]}$$

where ϕ is the half-power beamwidth. Note that this is usually different for the E and H planes, so that the spacing of the elements is also usually different in each plane.

Also, when antennas are placed close together, *mutual coupling* between elements occurs. This leads to changes in the current distribution on the elements, changing both the radiation pattern and the feed-point impedance of each element. The changes to the feed impedance often result in unequal powers being fed to the elements of the array, with consequential loss of gain. Optimum stacking rules are based on the assumption of minimum mutual influence which can be difficult to predict for composite antennas such as Yagis. However, antennas with low sidelobe levels are less susceptible than those with high sidelobes, as might be expected intuitively.

The coupling and effective aperture overlap problems cannot simply be solved by arbitrarily increasing the separation of the elements. As the element spacing increases, *grating sidelobes* appear, which reduce the forward gain. The grating lobes are due solely to the array dimensions, and can be seen by plotting the array factor for the chosen configuration.

Power divider and transmission line losses

The usual arrangements for feeding an array of antennas require each antenna to be fed an equal amount of power in the same phase as all the other elements in the array. There are several ways that this can be achieved, as shown in the section on antennas for construction later in this chapter. The power can be divided N ways at one point, from which equal length transmission lines feed each element, or groups of elements may be fed by several power dividers which are in turn fed by another power divider. Each system has its merits but losses incurred in the power dividers and cables erode the gain provided by the array.

Each time a cable connection is required, whether to a power divider or an antenna, the connection usually creates a small impedance mismatch. Power will be reflected from this mismatch and others in the system, reducing the gain of the array. The cumulative degradation of the power distribution to each antenna element can be startling, and of course the same degradation occurs when the antenna is receiving. In general, the simplest feed arrangements incur lowest losses and best performance.

Phasing errors

The small mismatches described above can result in errors in the phase of the current injected into the array elements, again leading to loss of gain and filling of nulls in the radiation pattern. Incorrect line lengths can result in the same effect. In constructing feeds for UHF and above, care must be taken in cutting and connecting cables.

A knowledge of the *velocity factor* of the actual cable used is essential if good results are to be achieved. The velocity factor is the rate at which the RF propagates along the cable relative to the speed of light (or RF) in a vacuum, and is modified by the dielectric constant of the cable insulator. A cutting error of 2mm in a solid polythene dielectric cable (such as URM67) will result in a phase error of 5° at 1296MHz. Measurement of the cable characteristics as described in references [5] and [6] can help eliminate many of the uncertainties of cable harness fabrication.

Alternatively, the velocity factor can be found with a dip meter coupled to a very small loop at the end of an open-circuit length of cable. At VHF, this loop should be no greater than 3mm radius. At higher frequencies, it is sufficient to trim the dielectric of the cable flush with the braid, fold the inner over the dielectric and solder to the braid, especially for larger-diameter cables. The dip will appear when the cable is an odd number of quarter-wavelengths long; if the frequency is checked with a counter or calibrated receiver, the velocity factor can be accurately calculated from:

$$v = \frac{4Lf}{300n}$$

where L is the length of the stub in metres, f is the resonant frequency in megahertz, and n is 1, 3, 5, the length of the stub in quarter-wavelengths. The lowest resonant frequency corresponds to $n = 1$.

Velocity factors are typically 0.66 for solid polyethylene, 0.72 for solid PTFE, and around 0.85 to 0.95 for foamed dielectrics or semi-airspaced cables. Very-low-loss Heliax-style cables which support the inner on small dielectric stand-offs can have velocity factors of 0.98.

Size, weight and wind loading

The size and weight of an array grows rapidly as the number of elements increases. The thoretical increase in gain over a single element follows the power law:

$$G_{max} = 10 \log_{10} N \quad \text{decibels}$$

where N is the number of elements.

Two elements provide 3dB gain, four elements 6dB gain, eight elements 9dB, 16 elements 20dB gain and so on, under ideal conditions. However, given the spacing constraints, the weight and wind loading of the array can quickly become unmanageable, especially if low-gain elements such as dipoles are used. It is for this reason that most high-gain antenna arrays constructed today use Yagi antennas for the array elements, as relatively few driven elements are required. This also simplifies the feed arrangements, which in turn reduces the losses and the cumulative phase and mismatch errors that tend to occur as the feed arrangements become more complex.

TRANSMISSION LINES

Antenna feeders or, more correctly, transmission lines, can make or break the performance of a station. At UHF and higher, the losses in the transmission lines feeding the antenna can be significant, dissipating RF power as heat, and requiring much larger antennas to achieve the desired radiated power. At these frequencies, there are essentially two useful types of transmission line for antennas: open wire and coaxial cable.

Open-wire line

The open-wire transmission line, comprising two parallel conductors held apart at intervals by spacers or spreaders, is still often used for feeding HF antennas. It provides a low-loss, easily constructed feeder capable of handling high powers, and the characteristic impedance can be adjusted by changing the wire diameter or spacing. The characteristic impedance is given by:

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

where D is the spacing between the wire centres, both wires having diameter d.

At VHF and higher frequencies, dielectric losses can become significant, but are minimised in the parallel-wire line. Apart from the spacers, the dielectric between the lines is air, and only a vacuum provides a better dielectric. The velocity factor is very close to unity. However, the spacing between wires must be much less than the wavelength if power is not to be lost through radiation. At VHF and above, this forces the selection of thinner wires or lower characteristic impedances if the feeder is to remain reasonably robust, and to some extent limits the uses of this type of transmission line to providing low-loss feeds to antennas in arrays, where the ability to adjust the impedance by varying the spacing is useful. Open-wire line is not really practicable for frequencies above 432MHz. This, and the difficulties of rigging long runs of closely spaced lines and of bringing them through the wall of a building (or round the antenna rotator) have probably discouraged their more general use.

Where open-wire line is required, use enamelled soft-drawn copper wire and solid PTFE rod for spacers if possible. The wire should be stretched to straighten and work-harden it immediately prior to assembly. Great care should be taken to ensure both wires are of equal length and made up/mounted symmetrically with respect both to dielectrics and conducting objects adjacent to the line.

Coaxial lines

Coaxial transmission lines, as their name implies, comprise an inner conductor mounted centrally within an outer conductor. The characteristic impedance for concentric circular conductors is given by:

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

where D is the inside diameter of the outer conductor, d is the diameter of the inner conductor and ε is the dielectric constant of the insulator (1 for air).

A square-section outer can be used to simplify connector mountings for home-constructed power dividers and transformer sections. The characteristic impedance is approximated by:

$$Z_0 = \left(\frac{138}{\sqrt{\varepsilon}}\right) \log_{10}\left(1.08\frac{D}{d}\right)$$

where D is now the inside dimension of the square outer conductor.

The principal advantage of coaxial transmission lines is that the surfaces carrying the RF current and the dielectric are inside, allowing robust, weather-resistant design and simple mounting on metal surfaces or masts. The disadvantages are dielectric losses (which increase rapidly with frequency), cost and weight.

Flexible cable designs use a braided outer conductor which, if it does not thoroughly cover the dielectric, will allow the RF to leak out through gaps in the braid. Cheap, so-called 'RG58' cable sold for Citizens' Band use should be avoided at all costs, as the braid coverage can be less than 50% – at VHF and above little power will reach the antenna. There are also cheap cables using a single wire and metallised plastic wrapping as 'braid' which are useless for VHF purposes.

Good-quality flexible coaxial cables have thick, close woven single or double outer braids. Genuine RG58 or URM67 cables provide flexibility with acceptable losses, especially for short lengths. Where longer cable runs are necessary, cables with semi-air spacing and copper-foil outer conductors provide better performance, although some care is needed in

Table 5.2. Attenuation of coaxial cables

Cable type	Diameter (mm)	Velocity factor	Attenuation (dB/100m) at 50MHz	70MHz	144MHz	432MHz	1296MHz
URM76, RG58CU	5.0	0.66	12	14	19	32	N/A
URM43	5.0	0.66	8.1	10.2	16.1	28.5	N/A
URM67, RG213U	10.3	0.66	4.6	5.6	8.3	15.5	27.0
Westflex 103	10.3	0.85	2.0	2.5	4.5	7.5	13.0
3/8in Flexwell	12.3	0.89	2.0	2.4	3.0	6.4	10.8
5/8in Flexwell	23.0	0.92	1.25	1.5	2.5	3.8	6.8
7/8in Flexwell	29.0	0.92	0.83	1.0	1.45	2.5	4.4

Table 5.3. Miniature and special-impedance cables for stubs and transformers

Cable type	Impedance (Ω)	Diameter (mm)	Dielectric	Velocity factor	Attenuation (dB/100m) at 100MHz	300MHz	1000MHz
URM95	50	2.3	Polythene	0.66	27	46	85
RG174U	50	2.3	Polythene	0.66	—	—	—
URM70	75	6	Polythene	0.67	15	27	52
URM111	75	2.3	PTFE	0.72	25	44	81
RG62AU	95	6	Air-spaced polythene	0.83	—	—	—

sealing the ends of these types to prevent ingress of moisture which will rapidly degrade the cable irretrievably. The ultimate in coaxial feeders are the Flexwell or Heliax types of cable, with a continuous corrugated copper outer, and air-spaced or PTFE foam dielectrics, together with special connectors to ensure good sealing and minimal mismatch. The performance of some typical 50Ω coaxial cables under impedance-matched conditions is shown in Table 5.2. Note that losses will be higher if appreciable standing waves exist on the cable. There is further information in Chapter 12 – 'General Data'.

An optimised installation would use rigid, low-loss cable for the fixed runs, with a short section of flexible cable to bridge the antenna rotator.

Matching stubs and transformer sections (discussed below) often require cables with characteristic impedances other than 50Ω. Also, miniature cables are sometimes desirable for constructing matching networks and filters. The higher loss is usually acceptable, as the length of cable used is small. Characteristics of a few readily available special cable types are shown in Table 5.3.

Impedance matching circuits

At VHF and above it is usual to use transmission lines, rather than lumped components, to obtain an impedance match between systems of different characteristic impedances. They can also be used to match arbitrary impedances (such as an antenna) to its feeder, of which a few techniques are shown below. Other methods addressing the design and calculation of matching circuits and components in detail are shown in references [7] and [8]. An excellent article on the behaviour of transmission lines and their use as circuit elements, together with some computer programs, can be found in reference [9].

Single stub matching

A short-circuited section of lossless transmission line behaves as a pure reactance at its input terminals. This impedance is given by the formula:

$$X_{in} = Z_0 \tan \beta l$$

where βl is the electrical length of the line (taking into account the velocity factor) and Z_0 is its characteristic impedance

The input reactance is inductive until the line length is a quarter-wavelength. The input impedance is then infinite, ie an open-circuit. As the line length is increased further, the reactance becomes negative, and the line behaves as a capacitor at the input terminals, although the remote end is a short-circuit. When the line is exactly one half-wavelength long, the input terminals appear to be short-circuited with no residual reactance. This cycle is repeated as the line is extended further.

An open-circuited stub behaves in a complementary manner to the short-circuited line. The input impedance is given by:

$$X_{in} = \frac{Z_0}{\tan \beta l}$$

Now the short length of line appears as a capacitor at its terminals, becoming a short-circuit at one quarter-wavelength, appearing inductive as it lengthens further, and becoming an open-circuit at one half-wavelength.

The term *matching* is used to describe the process of suitably modifying the effective load impedance to make it behave as a resistance and to ensure that this resistance has a value equal to the characteristic impedance of the feeder used. To make a complex load (ie a load possessing both resistance and reactance) behave as a resistance, it is necessary to introduce across it a reactance of equal value and opposite sign so that its reactance is effectively cancelled. The stubs described above provide the means to supply this reactance. Although there is no need to make the characteristic impedance of a stub equal to that of the transmission line, it may be desirable to do so for practical reasons.

In addition to tuning out the reactance, a match still has to be made to the transmission line characteristic impedance. The impedance at any point along the length of a $\lambda/4$ resonant stub varies from zero at the short-circuit to a very high impedance at the open end. If a load is connected to the open end and the power is fed into the stub at some point along its length, the stub may be used as an impedance transformer to give various values of impedance according the position of the feed point.

This is shown in Fig 5.6. The distance L is adjusted to tune the antenna to resonance and will be $\lambda/4$ long if the antenna is already resonant. The distance l is adjusted to obtain a match to the line. However, it can be convenient to have a stub with

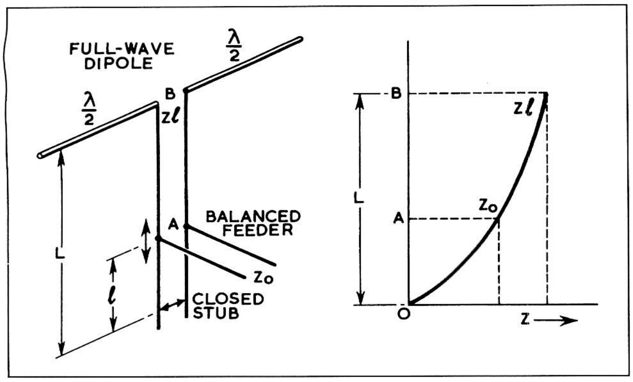


Fig 5.6. Stub matching applied to λ/2 dipole

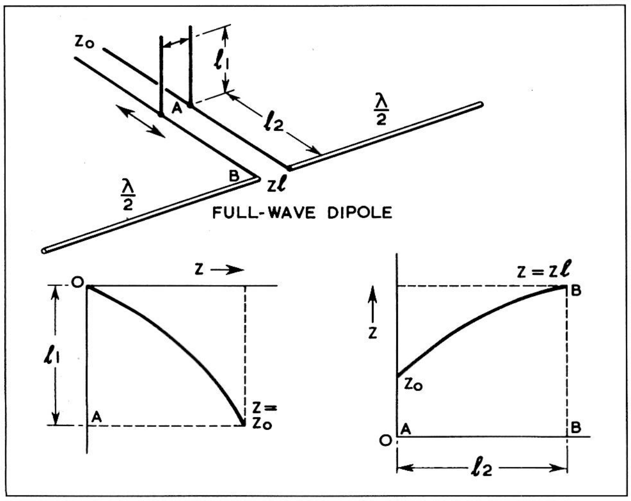


Fig 5.7. Stub matching with a moveable short-circuited stub

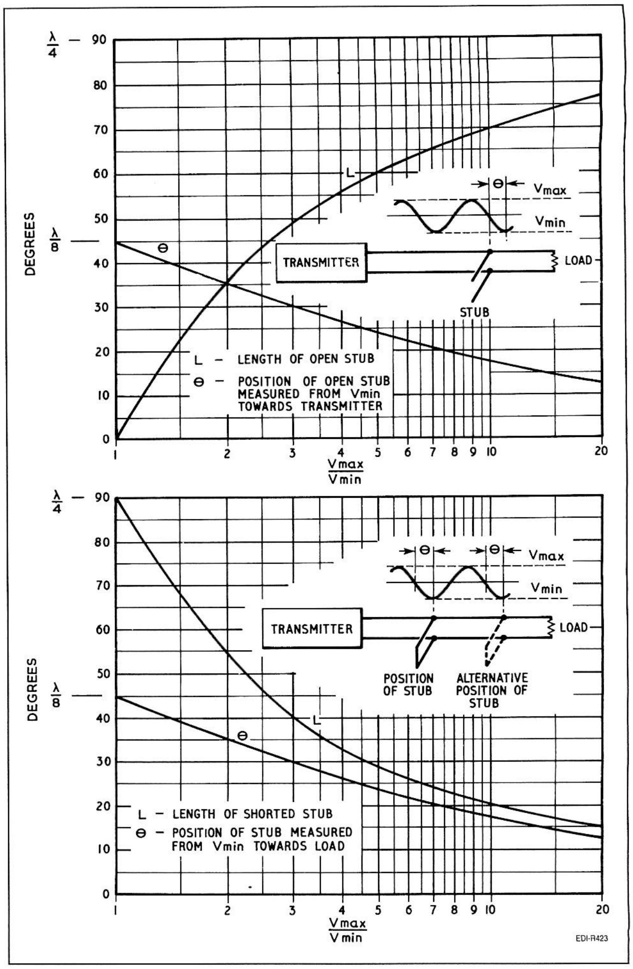


Fig 5.8. Impedance matching charts for single stub

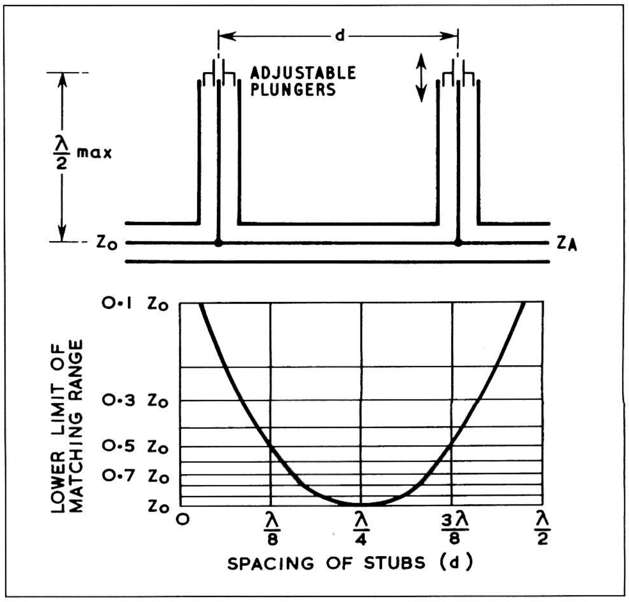


Fig 5.9. Two-stub coaxial tuner graph. Z_0 is the characteristic impedance of the feeder

an adjustable short-circuit which can slide along the transmission line (see Fig 5.7).

In practice, matching can be achieved entirely by the 'cut-and-try' method of adjusting the stub length and position until no standing waves can be detected. The feeder line is then said to be *flat*. However, the frequency range over which any single-stub matching device is effective is quite small, and where wide-band matching is required some other matching system may be needed. Fortunately, for most amateur purposes the bandwidth required is relatively narrow and the single-stub technique is usually sufficient. Fig 5.8 shows the positioning of open- and short-circuited stubs when the VSWR and the position of the VSWR minimum are known.

Two-stub matching

It is in making stub adjustments by 'cut-and-try' that the open-wire transmission line comes into its own because of the relative ease of repositioning the stub and the short-circuit. With coaxial line it is impracticable to construct a stub with an adjustable position. However, two fixed stubs spaced by a fraction of a wavelength can be used for matching purposes (see Fig 5.9).

The spacing usually employed is λ/8 or a multiple thereof. With this spacing, independent adjustment of the short-circuit stub lengths gives a matching range over 0.5 times the characteristic impedance (Z_0) of the transmission line upwards. As the spacing between the stubs is increased towards $\lambda/2$ or decreased towards zero, the matching range increases, but the adjustments become extremely critical and the bandwidth narrow. The theoretical matching range limits cannot be realised in practice because of finite losses in the stubs, so attention should be paid to providing reliable short-circuiting plungers in any home-built adjustable two-stub tuning units.

Adjustable stub tuners suitable for use at 144MHz have largely disappeared from the professional inventory, and are therefore extremely scarce in amateur circles. However, if the impedance of the device to be matched can be measured, the position of the stub can be calculated from Fig 5.9, or by the methods described in detail in references [7] and [8]. Coaxial feeders and stubs carefully cut to length and checked (see 'Practical considerations and limitations of arrays' above) will usually achieve a reasonable match, which can then be adjusted by trimming the stub for best results. If trimming worsens the match, replace the stub with a slightly longer one, and start trimming again.

Transmission line transformers

Quarter-wave transformer

A length of transmission line of a different characteristic impedance than the feeder can be used to transform impedance, providing an alternative technique which may be used to match a load to a transmission line. A special condition occurs when the length of the section of line is an odd number of quarter wavelengths long when the following formula applies:

$$Z_t = \sqrt{Z_0 Z_l}$$

where Z_t is the characteristic impedance of the section of the line and Z_0 and Z_l are the feeder and load impedance respectively. For example, if Z_0 is 80Ω and Z_l is 600Ω then:

$$Z_t = \sqrt{80 \times 600} = 219\Omega$$

This matching section is useful for transforming impedance and is called a *quarter-wave transformer* – see Fig 5.10. Note that the dimensions are in wavelengths and that allowance must be made for the velocity factor of the wave if dielectrics other than air are used to separate the conductors (see 'Practical considerations and limitations of arrays' above for typical velocity factors).

'Cot' transformer

The preceding methods require the use of special-impedance cable sections which, although they may be readily constructed in open-wire form, are difficult to realise in coaxial cable. There is another technique which can be used to transform between two cable systems of different impedance using short, equal length sections of the two types of cable in the sequence System 1-Z_2-Z_1-System 2 as shown in Fig 5.12.

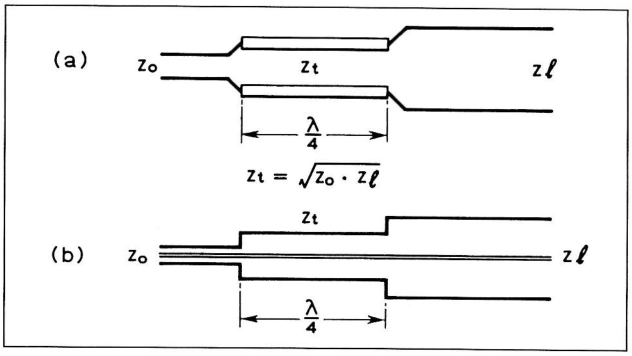


Fig 5.10. Quarter-wave transformer construction in (a) open wire and (b) coaxial forms

The formula for the electrical length of the matching sections has been simplified by G3KYH to:

$$\cot^2\theta = \frac{Z_1}{Z_2} + \frac{Z_2}{Z_1} + 1$$

where $\cot^2\theta = 1/\tan^2\theta$ and θ is the electrical length of each section in degrees.

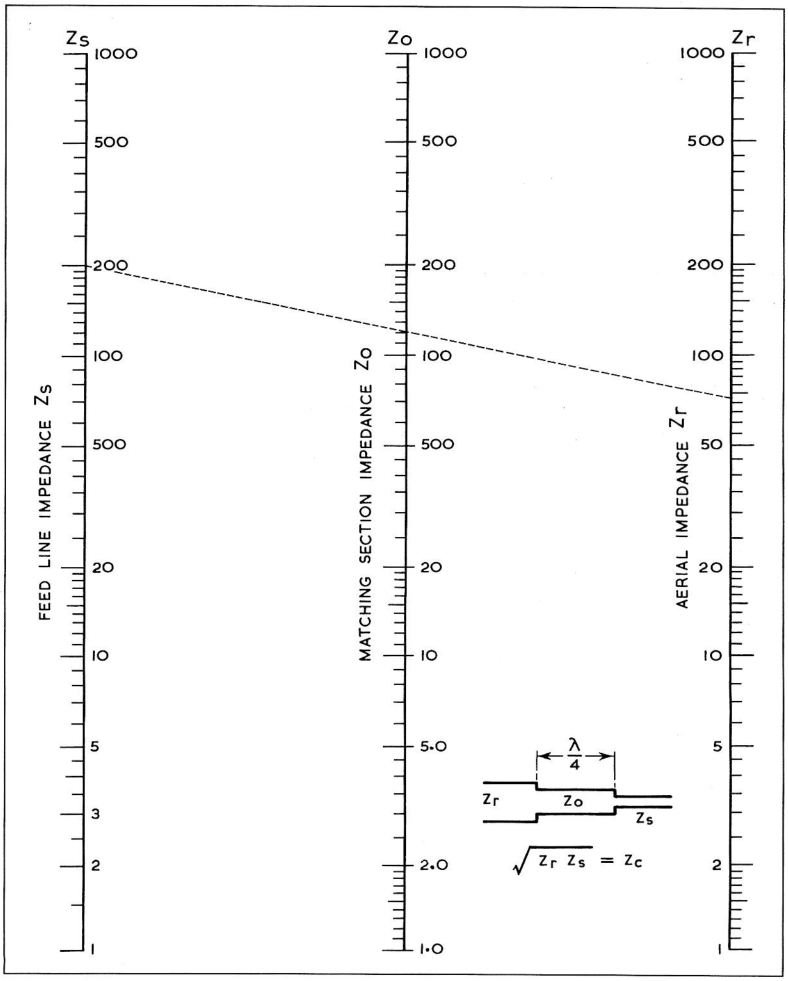


Fig 5.11. Chart showing impedance of quarter-wave transformer required to match between Z_s and Z_r

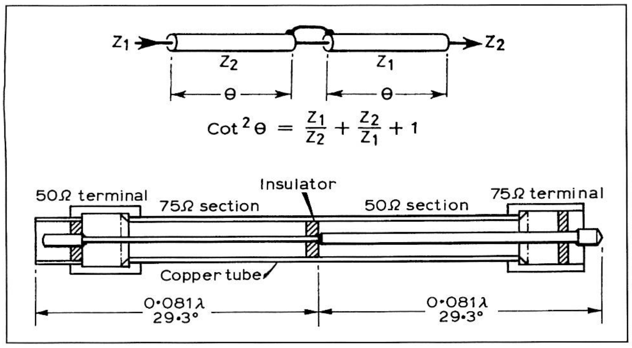


Fig 5.12. Transmission line transformer for matching 50Ω to 75Ω systems

To transform between 50Ω and 75Ω or vice versa, θ = 29.3°. The physical length must take into account the velocity factor of the sections imposed by their construction. One way of realising the transformer is shown in Fig 5.12.

Tapered line transformer

A section of tapered line can also be used to effect an impedance transformation. Again, a λ/4 section is only a special case, and to achieve a match in a particular installation the line length and the angle of taper should be varied until a perfect match is achieved. This form of matching device is called a *delta match* and is only really practical with open-wire feeders.

Power dividers

The quarter-wave transformers described above can be used to build power dividers to feed antenna arrays where interconnection cable lengths or available cable types will not permit transformation through the cables. This is especially valid for UHF systems, where it may be convenient to split the power at one place and feed the elements of the array with cables that have been cut to be of identical electrical length.

Two methods for achieving a two-way power divider are shown in Fig 5.13 below. Two outputs, connected together, can be fed by a suitable single λ/4 transformer section. This works well if the two loads are well matched both in magnitude and phase. The second method uses separate quarter-wave transformer sections to feed each output, which can provide better overall performance if the two loads are well matched but not absolutely identical, as may be the case with the outer elements in an array. The separate transformer sections ensure that identical in-phase currents will be provided at the outputs despite minor differences in load impedance.

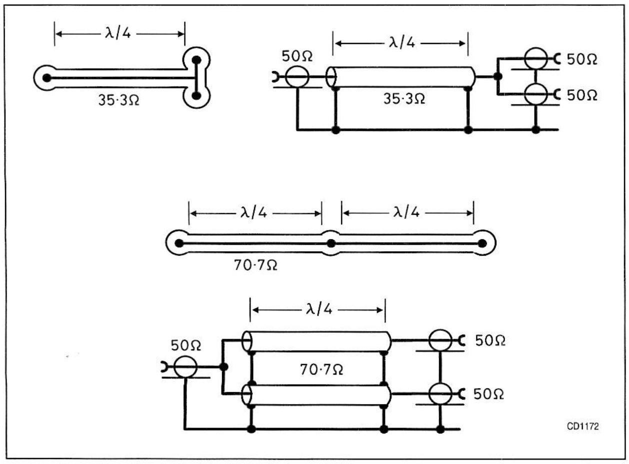


Fig 5.13. Two types of quarter-wave power dividers

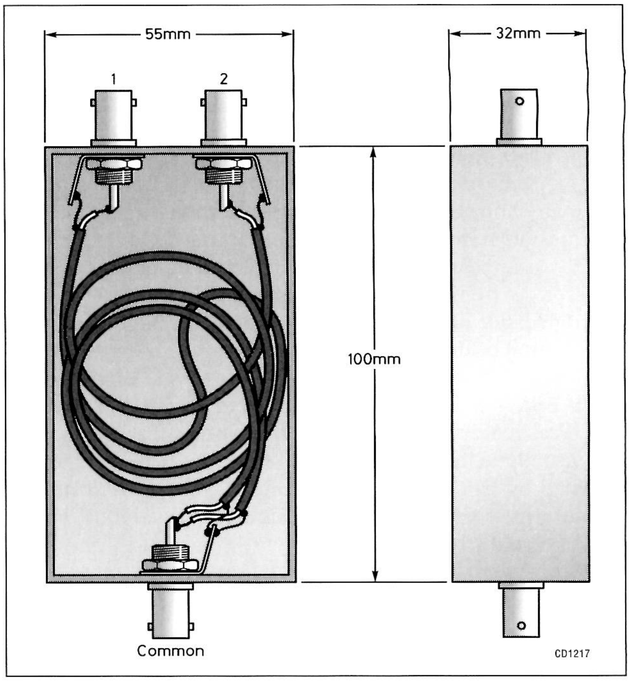


Fig 5.14. Compact two-way power divider for 145MHz

Construction of power dividers depends largely on the division ratio, the frequency of operation and the materials available. Square section tubing can provide faces for up to four output connectors if suitable inner conductors can be found to provide the correct transformation ratio. However, such dividers for 50 and 144MHz are large, and more compact but equally efficient dividers can be constructed from suitable coaxial cable, coiled up to fit within a box – see Fig 5.14. The example shown uses miniature 75Ω PTFE cable (URM111) for two-way power division at 145MHz for powers no greater than 50W. For higher powers, larger diameter, lower-loss cable should be used in a larger box.

Baluns

In many cases, antennas require a balanced feed with respect to ground, with equal and opposite currents in each leg of the feed. Coaxial cables are not symmetrical and, if they are connected directly to a balanced antenna, current will usually result on the *outside* of the braid – see Fig 5.15. The effects of this current is usually unwanted radiation, manifested as distorted radiation patterns, or interference with other electronics (EMC problems) where the feeder outer carries the unwanted RF to the susceptible equipment. In extreme cases, it can result in 'hot shack' effects – RF burns or changes to the impedance seen by the transmitter as other pieces of equipment are connected to the transmitter. These effects can be eliminated by suppressing the unwanted currents on the outer

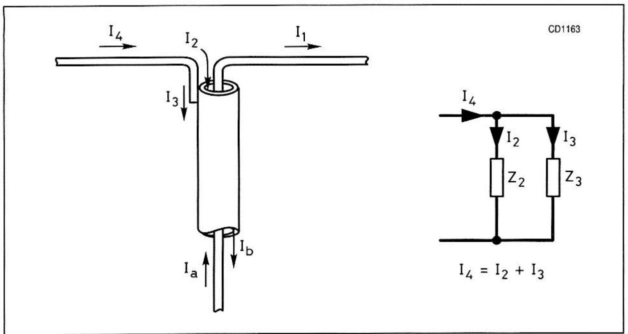


Fig 5.15. Currents inside and at the end of a coaxial cable

of the feed cable through the use of a balance-to-unbalance transformer (*balun*).

Under normal operation, the RF current flows on the *inside* of the outer conductor and the *outside* of the inner conductor of the coaxial cable. Under these circumstances, I_a and I_b are equal and opposite. However, at the end of the cable braid, the current I_4 may divide into two parts: I_2 flowing on the inner of the braid (and hence not capable of radiation or interference) and I_3 which flows on the *outside* of the braid and can be a potential source of trouble. The fraction of current on the outside of the braid is directly related to the impedance presented by the path on the outside of the braid (Z_3) to the characteristic impedance of the cable (Z_2), as shown in the figure. Balun designs increase the value of Z_3 and may also provide impedance transformations which are not part of the true balun action.

A coaxial sleeve balun is shown in Fig 5.16 below. The short-circuited quarter-wave stub surrounding the end of the coaxial cable presents an impedance of several thousand ohms to any currents that would flow on the cable outer. Most of the current then flows on the *inside* of the cable outer as required. Similar results are achieved with the Pawsey stub, Fig 5.17, which operates in exactly the same manner as the sleeve balun.

Good results can also be obtained with thin cables by coiling the cable close to the feed point to form an inductive choke with the outer of the cable, but care is necessary to ensure that the capacitance between turns (increased by the cable jacket) does not tune the choke below resonance which will prevent it from being effective. Ferrite beads or sleeves may be used on very small cables but this is more appropriate for low-power circuits than for antennas, where care is required in selecting the right materials; many ferrites are lossy at VHF and may melt the cable or shatter under even modest RF power.

An example of a much-used transformer balun is shown in Fig 5.18. This uses a half-wavelength of cable to invert the signal for the second leg of the balanced feed, and in the process also provides a 4:1 impedance increase. There is no connection between the balanced circuits and the outer of the feeder, hence no current flows on the outer of the feed. The length of the phasing cable should take into account the velocity factor of the cable, and all outer braids may be connected together close to the balanced output as shown.

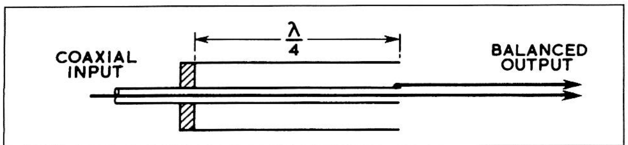


Fig 5.16. Coaxial sleeve balun

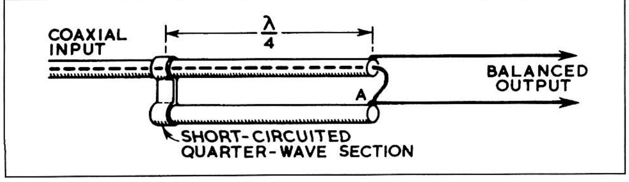


Fig 5.17. Quarter-wave open balun or Pawsey stub

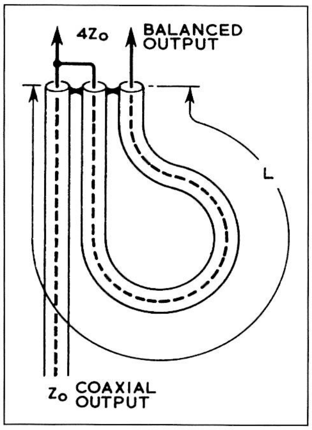


Fig 5.18. A coaxial balun giving a 4:1 impedance step-up

Transmission line filters

The characteristics of transmission line stubs and transformer sections can be used to create effective filters and diplexers. An excellent series of articles by G4SWX [15, 16] covers the design and adjustment of a range of harmonic and TVI filters, together with diplexers for several VHF/UHF bands. The diplexers can be useful where several transmitters and receivers are connected to a single broad-band antenna, or where concurrent transmission and reception on different bands are needed, as in certain types of satellite communications.

SELECTING AND INSTALLING ANTENNAS

This section deals with the essentials of choosing an antenna and its optimum location. Choice may be limited by the location of the station, planning considerations and the funds available, but is ultimately determined by the purpose of the station and the operator's interests.

Many of the decisions and trade-offs are between antenna beamwidth and gain. If interests lie with mobile stations or packet radio, steerable antennas can be a nuisance. Simple omnidirectional antennas, such as a monopole on a ground plane, will provide good local coverage, but higher-gain omnidirectional antennas, such as collinear arrays, may be desirable for hearing stations further afield. Most mobile and packet stations use vertical polarisation, placing another constraint on the choice of fixed antennas.

An antenna with very high gain will have a narrow beamwidth and few sidelobes. This is fine if you know where to point the array, but many stations will not be heard because they are outside the beam. Gain can be achieved with a relatively wide beamwidth in the horizontal plane by stacking elements vertically; the horizontal beamwidth is then determined by that of a single antenna element in the array, and the vertical beamwidth is narrowed by the array of elements in that plane. Slot-fed Yagi antennas are an efficient example of this type of antenna.

The converse may apply if using a steerable array for satellite communications; the azimuth angle is usually known, or can be found by steering the antenna, but the elevation varies as the satellite crosses the sky. A relatively narrow horizontal beamwidth and a broad vertical beamwidth with an antenna that is tilted upwards a few degrees will permit good communications without requiring the complexity and expense of elevation rotators and control equipment. Such an array can also be used for terrestrial communications with little loss of performance.

As the user's interests develop, the type of antenna or antennas required for optimum operation will become clearer. However, for beginners, relatively low-gain antennas offer a low-cost start with the greatest chances for success – the more complex the installation, the greater the chance it will not work first time, and the more difficult it is to find a fault which is often manifested as poor performance rather than total failure!

Polarisation

As stated above, most mobile and packet stations use vertical polarisation. However, at VHF and above, horizontal polarisation offers some advantages for long-distance propagation. This is due, to some extent, to the way that waves are scattered and diffracted by the ground, whether a plain, buildings or a hill edge, and also by atmospheric refraction effects. However, the use of horizontal polarisation for directional arrays is largely driven by the difficulties encountered in mounting vertically polarised Yagi arrays on metallic support masts without the mast interfering with the radiation pattern or compromising the mechanical integrity of the array. Whilst a horizontally polarised Yagi may be mounted mid-boom to a conductive vertical mast without ill effect, the radiation pattern of a vertical Yagi thus mounted would be completely destroyed. Dielectric masts of adequate strength also affect the radiation patterns at 144MHz and above, and so the antennas should ideally be supported from behind the reflector element. This presents considerable mechanical difficulties, especially with long Yagis, unless counterbalancing weights are fitted. The whole structure becomes much larger and heavier than necessary for horizontal polarisation, and is generally not used by amateurs.

Man-made interference, especially impulsive noise from motor vehicles and electric appliances, tends towards vertical polarisation, so the use of horizontal polarisation may also be beneficial in these circumstances.

Satellite communications do not require, but can be enhanced by, the use of circularly polarised antennas. Many satellites generate circularly polarised signals, and others may be spinning or tumbling whilst producing linear or mixed polarisations. A circularly polarised antenna will often reduce the short-term fading caused by satellite rotation, and by interfering rays scattered from the ground. There is a downside – if the available wave is largely of the opposite polarisation to the receive antenna, very little signal will be received (see 'Polarisation mismatch'), perhaps less than would be received by a linearly polarised antenna. Whilst the direction of polarisation can be reversed by suitable switching, this adds to the cost and complexity of the antenna. Circular polarisation can also be beneficial under conditions where the path is marginal and changing due to refraction or reflections, as in the case of communicating with mobile stations. However, the gain of a circularly polarised system against a linearly polarised source is 3dB less than would be obtained by correct linear polarisation at both ends under the same conditions.

What is of over-riding importance in choosing antenna systems is that the polarisations of the source and receiving antennas are the same, ie matched. Cross-polarisation between systems can result in losses of 15 to 20dB, which could completely negate the gain of the antenna system.

Height gain

When an antenna or array is mounted over ground, some radiation will strike the ground and be scattered or reflected by it. A remote receiving station may, in the simplest case, receive some power directly from the transmitting antenna, and some from the point of reflection – see Fig 5.19.

Similar effects occur when large reflecting objects such as electricity pylons or buildings are partially illuminated by the main lobe or side lobes of the transmitting antenna. The effect of the interfering ray will depend on its strength relative to the direct ray, the differential distance travelled, the nature of the reflecting object and other factors. However, if the reflection is relatively strong, the received signal will generally be improved as the height of the antenna is increased.

Several considerations apply when deciding the height of the antenna above ground. For optimum performance the antenna should be above any local screening from buildings and other obstacles. In addition, the rule-of-thumb figure of approximately 12m (40ft) is worth considering as it often raises the antenna above the layer of electrical interference and also the signal variations caused (at higher frequencies) by the heat layer above buildings. Such a height may also reduce the problems that arise when RF is coupled into house wiring, or directly into consumer electronic equipment, causing TVI and EMC problems.

If there is no screening by buildings, and assuming the antenna is over ground that is reasonably flat for several miles, the main lobe radiation from it will tend to be raised in the vertical plane. As the antenna height increases, the direction of the main lobe will level off in the required horizontal plane. Performance may be degraded by secondary lobes from the antenna. In general, however, as height increases, the pattern

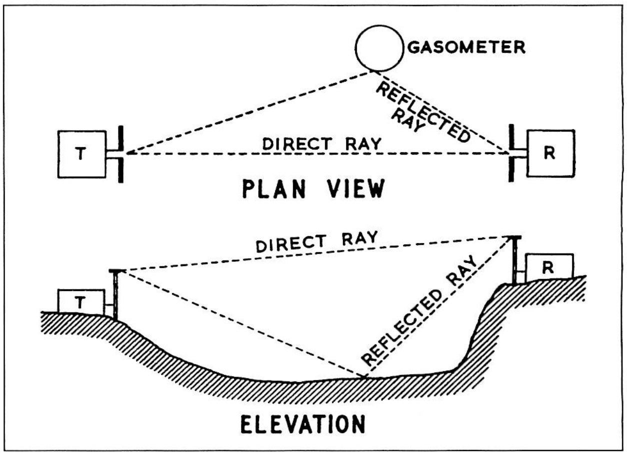


Fig 5.19. Interference between direct and indirect rays

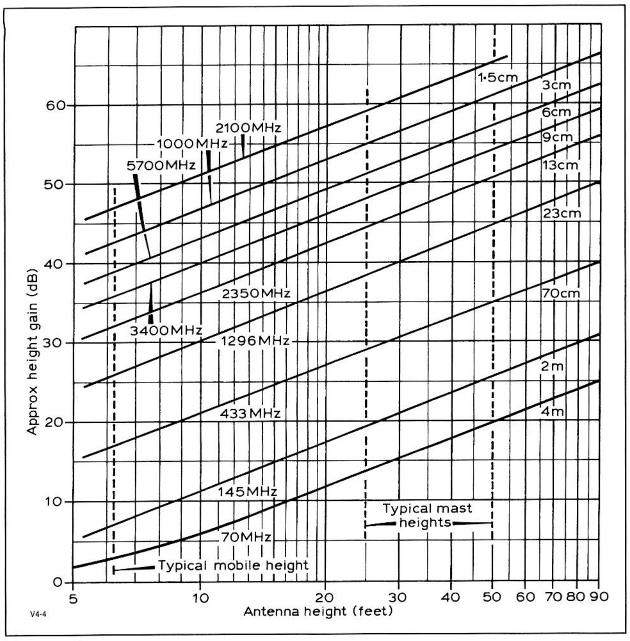


Fig 5.20. Antenna height gain correction factor

improves, and an additional gain of 6dB is obtained each time the antenna height doubles.

Fig 5.20 gives the approximate height gain obtained at various frequencies for various heights above ground. For heights greater than 12m, and assuming all obstacles have been cleared, a 24m (80ft) mast would be required to increase gain by a further 6dB. The additional expense for the mast may not be justified by the 6dB gain improvement, and care should be taken that losses in the additional feeder length required do not cancel out any gain that may be expected from the increased height. (See 'Coaxial lines' earlier for typical cable losses).

Should the station be well sited, on a hill for instance, increasing the mast height may make little or no improvement. The effective height above ground will relate to a point at the bottom of the hill, not the base of the mast under these circumstances. Conversely, a station in a valley or behind a hill may obtain a considerable increase in gain with height, much in excess of 6dB, as a more favourable angle to the hilltop or looking over it is achieved. A change of the vertical mounting angle (tilt) or a change of polarisation will also often provide a gain improvement.

The three basic configurations for receiving and transmitting antennas and the intervening ground are illustrated in Fig 5.21. The classic plane-earth case is shown in Fig 5.21(a), and under ideal conditions the signal received at the distant antenna follows the relationship:

$$e = \text{constant} \times \frac{h_T h_R}{\lambda d^2}$$

where h_T is the height of the transmitting antenna, h_R is the height of the receiving antenna, λ is the wavelength, and d is the distance between antennas. In this expression h_T, h_R and d must all be in the same units and d must be much larger than either h_T or h_R (by a factor of least 10) which is usually the case in practice.

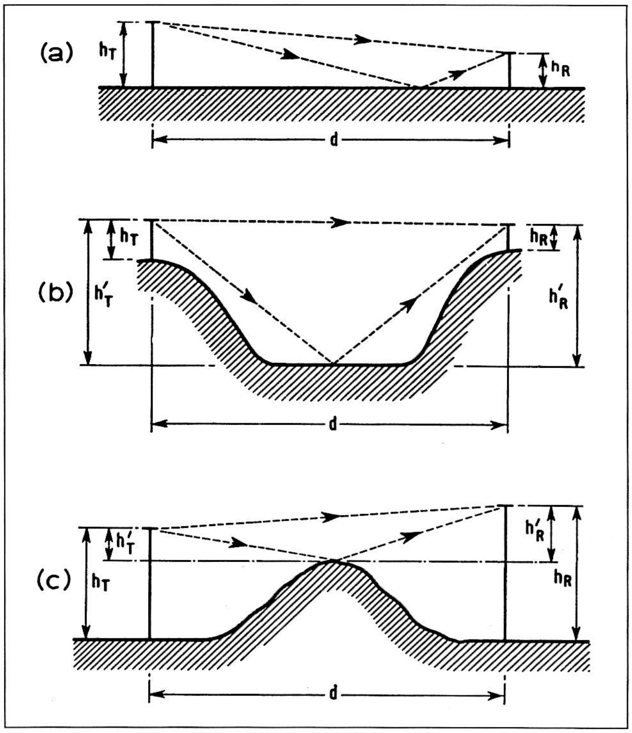


Fig 5.21. The effect of ground profile on direct and indirect rays

From this expression it is clear that an increase in either h_T or h_R will result in a corresponding increase in e, and doubling the height will give an increase of 6dB. This is the 6dB 'height gain' rule.

Fig 5.21(b) may be that of the operator who has selected a good hilltop site for 'portable' operation. Here, the antenna height above immediate ground is relatively small compared with the effective height above ground level at the point from where the indirect ray is reflected. Now,

$$e = \text{constant} \times \frac{h'_T h'_R}{\lambda d^2}$$

where h'_T and h'_R are the *effective* heights of the two antennas. There is still height gain to be achieved by increasing antenna height locally, but not at the same rate as in the first example. To obtain a gain of 6dB it is necessary to double h'_R, and this will require a manyfold increase in h_R. In the limit it clearly is not worthwhile seeking any great antenna height improvement; this is often the case for portable stations on hilltops, when the increased loss in the feeders is less than offset by the small additional signal to be obtained by raising the antenna.

The third case, Fig 5.21(c), is that of a station whose antenna is just able to 'see' over the surrounding higher ground, and is the reverse of case (b). The effective height h'_R is much less than h_R, and only a small increase in the height of the antenna is required to bring massive improvements in signal level.

To summarise, the antenna should ideally be positioned as high as possible, whilst taking into account the additional losses of the extra feed cable required. The signal improvements can be quite large if the station is in a location masked by hills or buildings.

EMC and location

The problems in achieving electromagnetic compatibility (EMC) are becoming ever more severe as ownership of electronic devices increase, together with housing densities. The problem is no longer solely one of television interference (TVI); amateur transmissions can interfere with the operation of telephones, audio equipment, and car security systems. There is also a growing reverse problem of domestic electronics interfering with amateur reception; line timebase noise from older large-screen television sets, and more recently from personal computers, is adding to the general pollution of the RF spectrum. Whilst legislation is now in place to reduce emissions from, and the susceptibility of, domestic equipment, it will be many years before some of these equipments are replaced. In the meantime, it is prudent to design installations to minimise the mutual interference that may occur.

In positioning antennas, the following should be considered.

Avoid:

- Placing the antenna close to your neighbour's (or your own) TV antenna!
- Allowing arrays to 'stare' directly in the same plane at adjacent antennas; raising the array a few feet may reduce the potential for interference dramatically.
- Locating antennas where they can easily couple into mains wiring or plumbing.
- Placing antennas where they can couple into overhead telephone wires.
- Running feeders next to mains wiring or plumbing where coupling may occur.

Do:

- Use coaxial feeders with good screening indoors.
- Use chokes and baluns to minimise any RF on the (coaxial) feeders.
- Place the antennas as far away as is practicable from other antennas, TV feeders, mains wiring or telephone cables.
- Make friends with your neighbours and explain what you are doing – it is then easier to resolve any difficulties that occur.
- Choose materials that are electrochemically compatible to minimise corrosion and harmonic generation through the 'rusty bolt' effect.
- Overhaul antenna systems regularly to prevent 'rusty bolt' effects and deteriorating cable connections that can cause interference (also a good idea to preclude the antenna from deteriorating mechanically).

See also Chapter 6 – 'EMC'.

Installations

Internal installations

Whilst external installations are preferable, adequate VHF and UHF antenna systems can be installed successfully within a loft space. Space constraints will limit the size of antenna, especially if a rotatable array is considered, but omnidirectional antennas usually present few problems.

The roofing material can have a marked effect on signal losses, especially when wet. Slates shed water and dry out fairly quickly, but old and porous tiles, although adequately waterproof, can scatter and absorb much of the signal. Measurements have shown a difference of more than 7dB for propagation through dry and wet weathered tiles at 435MHz.

Coupling into wiring and plumbing should be very carefully addressed and investigated with loft installations, and it is prudent to choose antennas that are not highly tuned or over-sensitive to the presence of adjacent objects.

Chimney installations

The nature of these installations is usually determined by the strength of the chimney, the courage of the erector, and his willingness to revisit the chimney regularly.

An end chimney can provide an excellent antenna site provided it is not too close to adjacent property and antennas. A short mast is usually lashed to the stack at two or three places, and a fixed or rotating array fitted immediately above. Space and safety considerations dictate the size of the antennas, and whether they are assembled *in situ*. If the chimney is relatively accessible, complex antennas and systems that require maintenance or tuning are practicable. However, simpler, fixed systems are better if the site is difficult (or expensive) to access. In choosing the antennas, the wind load and overturning moment should be carefully considered, and if in doubt, a survey of the chimney should be carried out by someone with the necessary knowledge and qualifications – this could prevent expensive roof repairs! Lightweight fixtures and fittings can be bought from television antenna supply companies. Heavier-duty fixtures are available from amateur antenna suppliers and professional antenna installers.

Masts and towers

Where suitable chimneys are not available, but wall and garden space can be used, a mast or tower may provide the best option, especially if the station is to be located in a garage or shed. Many varieties of mast are available commercially, both as free-standing and wall-mounted designs, and hardware for fastening masts to walls and footings is available from a number of suppliers.

A mast can offer the freedom to experiment with antennas if suitable equipment for raising and lowering is provided. However, planning permission is nearly always required, and considerable thought is needed before installing free-standing towers – the foundation requirements can be considerable, and are not easily relocated. If garden space permits, a guyed mast may provide a good solution, although assistance and a great deal of care is required when erecting such structures, especially when loaded with antennas. The reader is referred to the articles by G3ZPF [17] for information outlining the size of foundations and guy anchors required together with some methods for construction.

In addition, lightning protection should be installed to good earths with short, wide copper straps to minimise the current flowing into the radio equipment and house wiring if a lightning attachment does occur. Again, professional advice should be sought in this respect.

Wind loading

In all considerations for antenna design and erection, the wind loading of the array must be taken into account, so that the rotator (if used) and the supporting structures can be selected or designed to withstand the stresses to which the system may be subjected. It is prudent to allow a good margin for safety, to allow for wear and tear, corrosion and general deterioration in any calculation related to safety.

The articles by G3ZPF [17] provide the basis for estimating the loads experienced by antennas and masts. BS8100 and the earlier, but still useful, BSI CP3 Chapter 5 Part 2 provide information on basic wind speeds throughout the UK. This can be used with the given topography, ground roughness and height above ground factors to determine the dynamic pressure that is likely to be experienced by the structure under 'worst-case' conditions. Wind loading of masts and towers is now assessed under BS8100, and advice should be sought from the manufacturer for the appropriate information.

Safety

Masts and antennas are potentially dangerous structures especially during erection and dismantling. There are a number of safety rules that *must* be observed during these activities:

1. Thorough checks must be carried out to ensure that there is *no* possibility of the mast or antenna coming into contact with overhead power wires, however it may topple or collapse.
2. The job of erection or lowering must be planned carefully. Considerations must include the positioning of each part of the antenna and/or mast at every stage of the process. Enough persons should be available, wearing boots, gloves and safety helmets during the raising and lowering processes, and sufficient guiding ropes should be used to ensure control of the structure at all stages of the operation. There should be no possibility of tripping on ropes or equipment during the operation, which should not be attempted in strong winds or when it is getting dark.
3. Before raising or lowering, all components and fastenings should be double-checked for being fixed firmly and safely. The base of the mast must be firmly fixed to prevent slipping.
4. Everybody involved shall have their role clearly defined. Those not needed to assist should be kept well clear. If any children are nearby, a person should be tasked to keep them clear of the area of operations, including those areas where the mast might topple or fall, and animals should be kept under control.
5. One person should be in charge of the operation, and not take any part in the lifting activities themselves. He/she shall give clear concise instructions, which have been rehearsed before the actual lift or lowering.
6. Safety precautions must be continued until all mast fixings and guys are secured, and any temporary ropes and equipment have been removed and stowed away.
7. After erection, the mast should be inspected for tightness of bolts and for the integrity of any protective coverings. The mast and antennas should be regularly inspected for tightness of bolts, wear and/or damage to guys and fastenings, and integrity of protective coverings. The mast and antennas should be lowered for full inspection and overhaul at least every three years, or more often in exposed locations. Electrical continuity, sealing and painting/greasing should be checked, together with the replacement of items that weather or denature in sunlight (plastic covers and fittings).

Antennas – build or buy?

General considerations

A great deal of satisfaction can be obtained from building one's own antennas, either by following detailed instructions, or by experimenting with the materials available. However, as either the frequency or complexity of the antenna increases, so does the need for some test equipment or measuring facilities. There is little more frustrating than to have spent many hours in construction, and then be unable to make the antenna work.

Simple antennas such as whips and dipoles can be tuned with a low-power transmitter and VSWR meter, by 'pruning' element lengths for best VSWR.

As the frequency increases, variations in construction and the dielectric constant of materials used for insulators or spacers can have a considerable influence on the performance of antennas. Tools to at least estimate the input impedance (resistance *and* reactance) can become essential to find out exactly why an antenna is not working. Noise bridges can provide the necessary information up to the 70cm band, beyond which second-hand professional impedance bridges provide the best means for measuring antennas.

Single antennas with gains greater than, say, 10dB really require facilities for measuring changes of gain if experimentation is to be meaningful. This requires a reasonable amount of space with few unwanted reflections, and a calibrated signal generator and receiver. This is not as difficult as it may seem; the techniques, together with much other useful information on building test equipment and carrying out measurements are shown in *The Antenna Experimenter's Guide* by G3LDO [19].

Careful adherence to construction details will usually result in a working antenna with minimum tuning, so it can be well worth considering building your own if basic metal working facilities are available.

Arrays can also be assembled with a minimum of measuring equipment if ready-built antennas are obtained (all antennas in the array should be identical), and care is taken in measuring and making up the feed cables. Some manufacturers can supply ready-built (and phase matched) feeder harnesses for standard array configurations.

Decreasing cost and increasing power of personal computers has added a relatively new tool to the serious experimenter's armoury. A number of programs are available for modest cost which allow the electrical design of antennas on-screen, followed by analysis of the current distribution, radiation pattern and gain. They can be useful to evaluate the effect of changes in dimensions or configuration, or to try out completely new structures. Some knowledge of antenna theory is necessary – the 'garbage in-garbage out' syndrome certainly applies! A description of the techniques and the results that can be obtained have been described by G3SEK and G3HCT [20, 21]. Special programs are also available for optimisation of Yagi antennas.

Choice of materials

The use of dissimilar metals in an antenna system is likely to cause considerable trouble due to electrolytic corrosion. Each metal has its own electro-potential and, unless metals of similar potential are used, the difference will cause corrosion even when they are dry. When moisture is present, the effect

Magnesium	Anodic
Aluminium	
Duralumin	
Zinc	
Cadmium	
Iron	
Chromium iron alloys	
Chromium nickel iron alloys	
Soft solder tin-lead alloys	
Tin	
Lead	
Nickel	
Brasses	
Bronzes	
Nickel copper alloys	
Copper	
Silver solders	
Silver	
Gold	
Platinum	Cathodic

Fig 5.22. Electrochemical series for metals

will be much more severe and can be enhanced by atmospheric pollution.

If, for any reason, dissimilar metals must be used then considerable care should be taken to exclude moisture. The metals can be arranged in order of their electrochemical potential as shown in Fig 5.22.

Metals in each of the groups may be used together with little corrosive action, but metals from different groups will quickly corrode at the point of contact. The list is arranged in order, so that the greater the spacing between materials in the list, the greater the effect.

The materials in the lower part of the list will corrode those in the upper. For example, brass or copper screws in aluminium will corrode the aluminium, whereas cadmium-plated brass screws would cause less corrosion.

Corrosion can cause weakening of mechanical structures and also increases in contact resistance between elements and feeders, resulting in dissipation of transmitter power as heat. Under some circumstances, the joint between corroded materials behaves as a semiconductor, generating harmonics and intermodulation products that cause interference to other radio users, both in and out of band. For this reason, selection of materials to minimise corrosion is important, and all antenna joints and weather protection measures should be inspected and refurbished at least every two years.

DIRECTIONAL ANTENNAS FOR FIXED STATIONS

The Yagi antenna

This is one of the most useful antennas for VHF/UHF, as it can be compact, robust and provide good directivity and gain with a relatively simple structure. Unfortunately, it is also one of the most complex antenna structures to analyse, and it can be difficult to construct Yagis that really provide good performance, especially where high gain is required.

The original research was carried out by S Uda in Japan in 1926, but it was the review and translation into English by his professor, H Yagi, in 1928 that introduced the design to the West. The basic array comprises a driven dipole element with a passive dipole adjacent to it. If both elements are tuned to resonance, the currents in each element are approximately equal, and are in phase. By lengthening the passive (*parasitic)* element, the phase of the current is delayed, whilst the amplitude remains almost unchanged. When the phase delay complements the spacing between the elements, the radiated power will be directed away from the parasite, which is then known as a *reflector*. By similarly placing another, shorter *director* parasite in line with the driven element, but on the opposite side from the reflector, the directivity can be further enhanced. The principles for this phenomenon were addressed in the section on arrays.

Yagi antennas can be provided with large numbers of reflectors, the elements being excited by mutual coupling with the driven element and other elements according to their relative position. As the magnitude and phase of the current on each element is influenced by their relative positions and lengths, the permutations of dimensions that can produce satisfactory performance become very large.

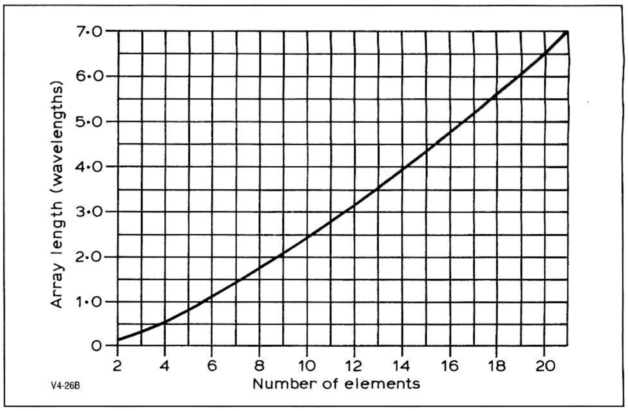


Fig 5.23. Optimum length of Yagi antenna as a function of number of elements *(ARRL Antenna Book)*

Until relatively recently, performance of the antenna was optimised by 'cut-and-try' methods, as the mathematical analysis of the problem was too complex to afford numerical solution other than for small numbers of elements. Theoretical methods indicated the limits of performance that could be expected of the array for given constraints, eg Fig 5.23 and Fig 5.24, showing boom length and number of elements for 'optimum' arrays. However, experimental work showed that these gains were rarely realised, usually falling short by 0.5 to 1dB, and very poor performance was sometimes obtained, especially from long Yagi antennas. Many independent investigations of multi-element Yagi antennas have shown that in general the gain of a Yagi is directly proportional to the array length provided the number, lengths and spacing of the elements are properly chosen. However, to constrain the number of variables, the concept of equal-length reflectors, or equally spaced reflectors was often used for elements well removed from the driven element.

A suite of results for several antenna geometries was published by P Viezbicke for the US Department of Commerce and National Bureau of Standards in 1976 [10], and it has become a reference document for many Yagi designers. It addresses: (a) the effect of reflector spacing on the gain of a

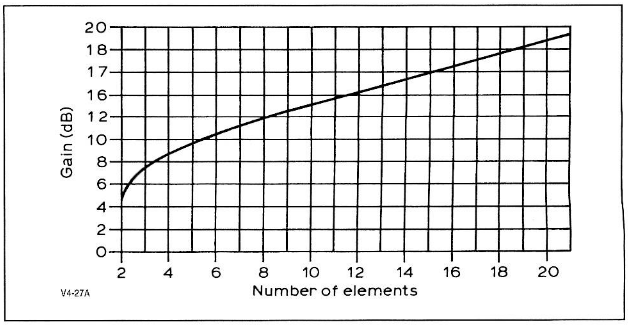
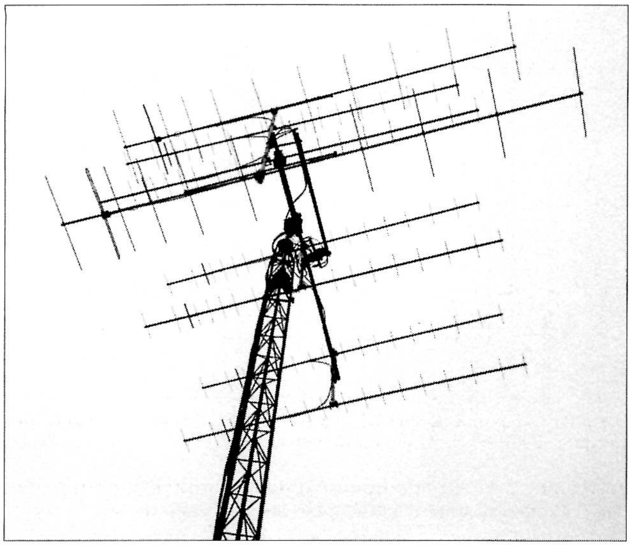


Fig 5.24. Gain (dB) over a $\lambda/2$ dipole versus the number of elements of a Yagi antenna *(ARRL Antenna Book)*



DL6WU has stacked two 11-element Yagis of his own design with six of the highly respected K2RIW 19-element Yagis for 432MHz

dipole; (b) effect of different length directors, their spacing and number on realisable gain; (c) effect of different diameters and lengths of direct, realisable gain; (d) effect of the size of a supporting boom on the optimum length of parasitic elements; (e) effect of stacking of antennas on gain; and (f) measured radiation patterns of different Yagi configurations.

However, as greater computing power has become available, it has been possible to investigate the theoretical optimisation of Yagi gain more closely, and to take into account the effects of mounting the elements on metallic and dielectric booms. Dr J Lawson, W2PV, carried out an extensive series of computations, collated in reference [11], which explain many of the disappointing results achieved by constructors. G Hoch, DL6WU, has especially studied the design and construction of long Yagis [12–14] and identified the pitfalls.

Generally, short Yagi antennas with less than six elements will perform reasonably well with a selection of materials and minor deviations from the optimum dimensions. However, higher-gain Yagis need to be carefully constructed with minimum deviation from the design if the gain is to be realised. If it is necessary to use different diameter tubing from that specified, the length of the element must be adjusted to compensate for the change in self-reactance that results. Hoch gives a formula for the reactance of an element of arbitrary length (L) and diameter (D) for a given wavelength:

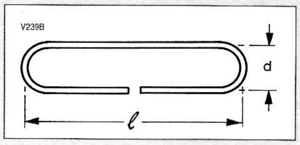
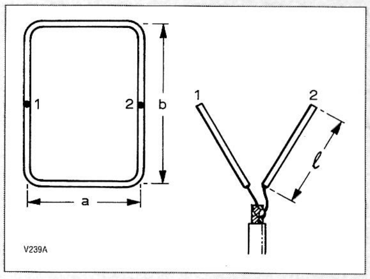
$$X = \left\{430.3\log_{10}\left(\frac{2\lambda}{D}\right) - 320\right\}\left(\frac{2L}{\lambda} - 1\right) + 40$$

The modified length L' for a new element diameter D' can be calculated by rearranging the formula:

$$L' = \left\{\frac{(X-40)}{\left\{430.3\log_{10}\left(\frac{2\lambda}{D'}\right) - 320\right\}} + 1\right\}\frac{\lambda}{2}$$

Table 5.4 shows typical component dimensions for a range of Yagi antennas for 4m, 2m and 70cm, using either open or

Table 5.4. Typical dimensions of Yagi array components

	Length		
	70.3MHz	145MHz	433MHz
Driven elements			
Dipole (for use with gamma match)	79 (2000)	38 (960)	12¾ (320)
Diameter range for length given	½–¾ (12.7–19.0)	¼–3/8 (6.35–9.5)	1/8–¼ (3.17–6.35)
			
Folded dipole 70Ω feed			
l length centre/centre	77½ (1970)	38½ (980)	12½ (318)
d spacing centre/centre	2½ (64)	7/8 (22)	½ (13)
Diameter of element	½ (12.7)	¼ (6.35)	1/8 (3.17)
			
a centre/centre	32 (810)	15 (390)	5 1/8 (132)
b centre/centre	96 (2440)	46 (1180)	152 (395)
Delta feed sections (length for 70Ω feed)	22½ (570)	12 (300)	42 (110)
Diameter of slot and delta feed material	¼ (6.35)	3/8 (9.5)	3/8 (9.5)
Parasitic elements			
Element			
Reflector	85½ (2170)	40 (1010)	13¼ (337)
Director D1	74 (1880)	35½ (902)	11¼ (286)
Director D2	73 (1854)	35¼ (895)	11 1/8 (282)
Director D3	72 (1830)	35 (890)	11 (279)
Succeeding directors	1in less (25)	½in less (13)	1/8in less (3)
Final director	2in less (50)	1in less (25)	¾in less
One wavelength (for reference)	168¾ (4286)	81½ (2069)	27¼ (693)
Diameter range for length given	½–¾ (12.7–19.0)	¼–3/8 (6.35–9.5)	1/8–¾ (3.17–6.35)
Spacing between elements			
Reflector to radiator	22½ (572)	17½ (445)	5½ (140)
Radiator to director 1	29 (737)	17½ (445)	5½ (140)
Director 1 to director 2	29 (737)	17½ (445)	7 (178)
Director 2 to director 3, etc	29 (737)	17½ (445)	7 (178)

Dimensions are in inches with millimetre equivalents in brackets.

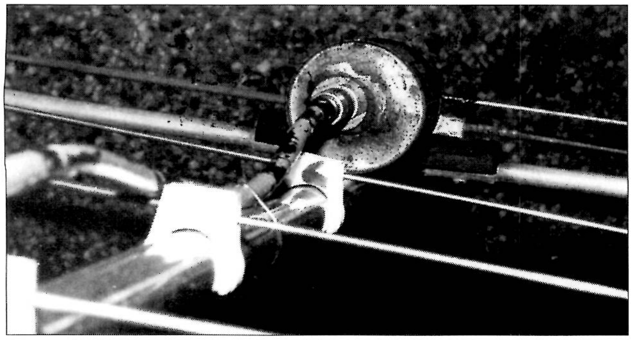


Photo showing how a driven element and first directors are typically mounted on a boom

folded dipole elements to drive the array. As stated above, antennas of this form can be expected to work reasonably well with up to six elements (12 elements for a skeleton slot array) if the dimensions are adhered to. Longer Yagis need to be constructed exactly as described, including the boom and fastenings used to secure the elements, if claimed performances are to be realised without recourse to antenna measurement ranges. The articles by DL6WU [12–14] address the construction of such antennas.

Skeleton-slot Yagi stack

The skeleton slot provides an ingenious means to feed two stacked Yagi antennas efficiently and achieve a good impedance match. The skeleton slot can be thought of as a pair of λ/2 dipoles spaced vertically by 5λ/8. Since most of the radiation is provided by the centre of the dipoles, their ends can be bent out of plane with little effect, and joined together with high-impedance feeder so that end feeding may take place. To feed both dipoles in phase, the feed point must be midway between them; the high impedance is transformed to more manageable levels by a tapered section or delta match as shown in Fig 5.25.

The overall slot Yagi structure is shown in Fig 5.26. Element dimensions are taken from Table 5.4 and the radiation pattern, together with those of other typical Yagi antennas, is shown in Fig 5.27.

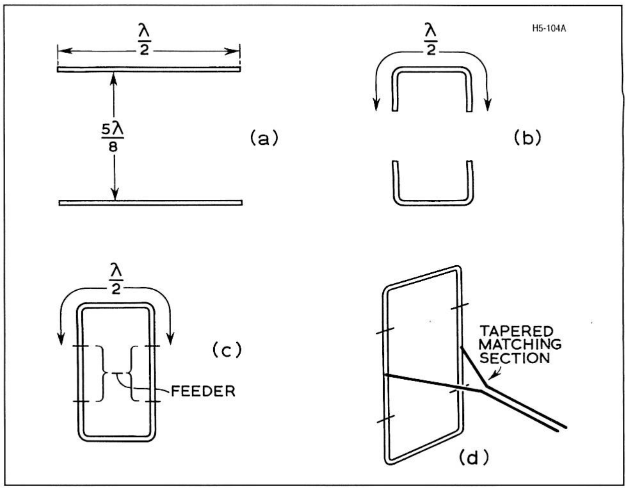


Fig 5.25. Development of the skeleton-slot radiator

Quad antennas and arrays

The quad antenna offers a useful horizontally polarised alternative to short Yagi antennas at VHF and above, being both compact and lightweight with just a driven element and reflector – see Fig 5.28. Gains of 5.5 to 6dB are readily obtained, together with good front-to-back ratio. Arrangements for two- and four-antenna arrays of quads are described below.

Typical dimensions for quad elements are shown in Table 5.5. The input impedance is strongly affected by the spacing between the driven element and the reflector, and will be between 180 and 230mm for an input impedance of 72Ω. The elements may be made from 3mm or 6mm aluminium rod or bar, and if the vertical dimensions of both elements are made the same, two short cross-pieces can be used to separate the elements and mount them to a mast. The cross-pieces may be metal so that the whole structure with the exception of the feedpoint and reflector stub (if used) can be solidly built and bonded together. A balun should be used at the feedpoint, although "this is not essential if the feeder is short and of low loss".

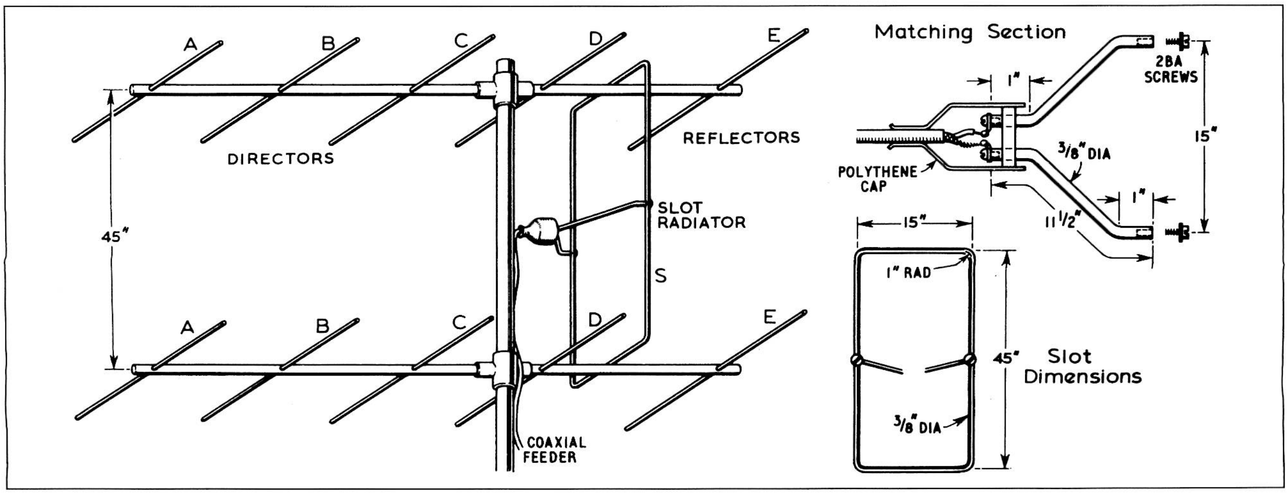


Fig 5.26. Six-over-six skeleton-slot Yagi for 2m

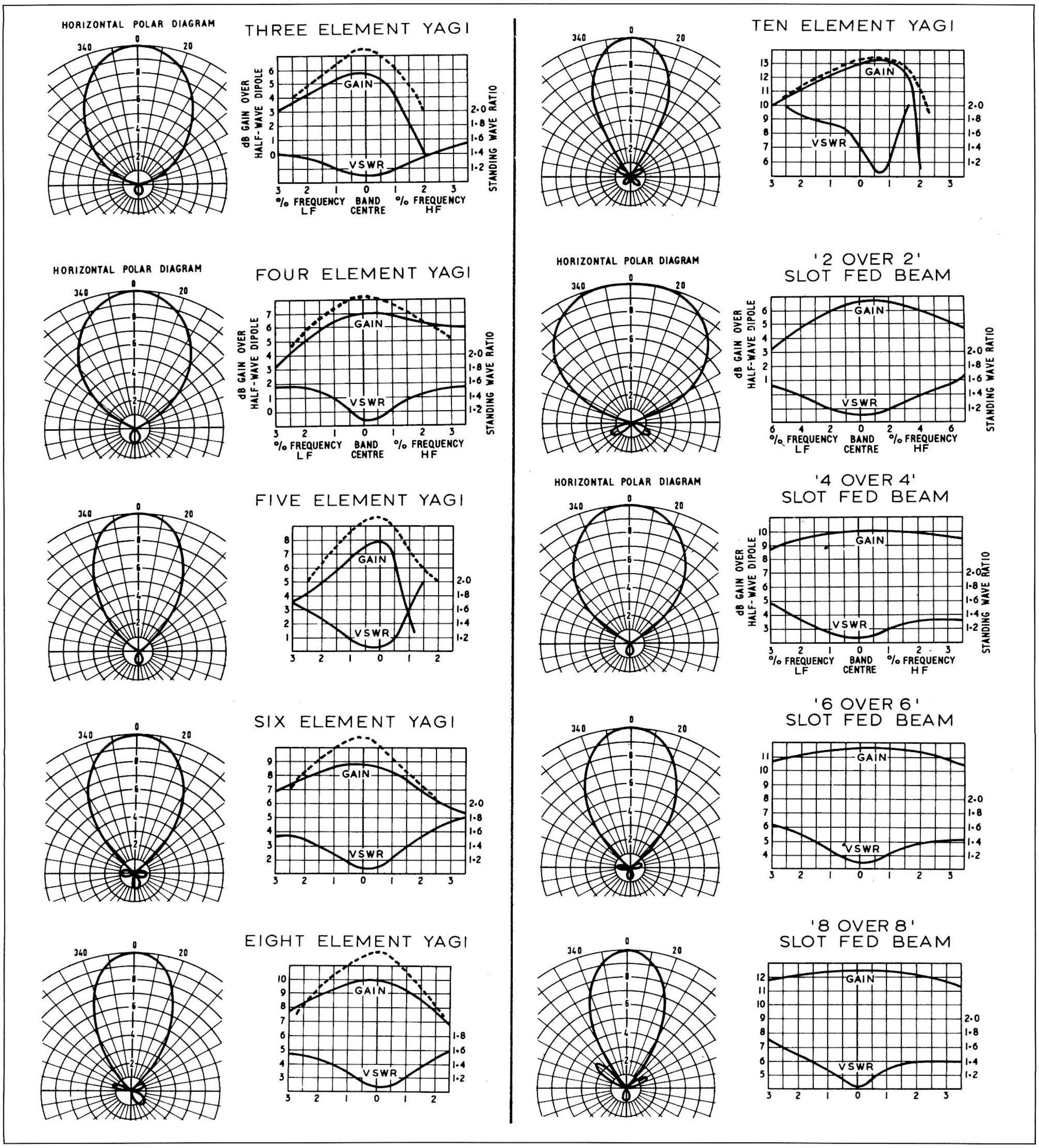


Fig 5.27. Charts showing voltage polar diagrams and gain against VSWR for various Yagi antennas with uniform directors. Dotted lines are for antennas optimised for maximum gain

The antenna can readily be configured as a two-element or four-element array. Each antenna for 144MHz has the dimensions shown in Fig 5.29, spaced 178mm between elements. Two quads stacked vertically should be spaced $5\lambda/8$ between centres and paralleled through a single $\lambda/4$, 51Ω transformer. To obtain an input impedance of 72Ω (for further baying as a four antenna array), it may be necessary to increase the separation of the reflector and driven element to 230mm to overcome the effects of mutual coupling. The pair of antennas should provide a gain of 8.2dBD (10.3dBi) with a front-to-back ratio of 20dB.

A four-antenna arrangement for the 2m band is shown in Fig 5.30. The layout is determined by the ease with which feeder cables can be run, and the minimisation of unsupported

Table 5.5. Design dimensions for 70 and 144MHz quad antennas

Band (MHz)	Reflector 1 total length	Reflector 2 total length	Director (if used)	Approx length of stubs if used: Reflector s/c	Director s/c
70 (a)	173 (4390)	165 (4190)	157 (3990)	—	—
70 (b)	165 (4190)	165 (4190)	165 (4190)	8 (203)	8 (203)
144 (a)	84 (2130)	80 (2030)	76 (1930)	—	—
144 (b)	80 (2030)	80 (2030)	80 (2030)	4 (101)	4 (101)

Dimensions are in inches with millimetre equivalents in brackets.

(unguyed) sections of mast. Reflector-director spacing is 230mm, the vertical spacing remains 1650mm between centres, and the horizontal spacing is 2070mm, one wavelength in free space. The feed arrangements are shown in Fig 5.31 below. The design is based on a 72Ω main feeder; for 50Ω array impedance, the transformer in the feedline should be replaced with a 42Ω section or a 'cot transformer' as described in the section on transmission line transformers. The more readily available URM57 may be substituted for UR1 cable with minor adjustments to the antenna element spacing.

The gain of the overall array should be 13.5dBD (15.6dBi)

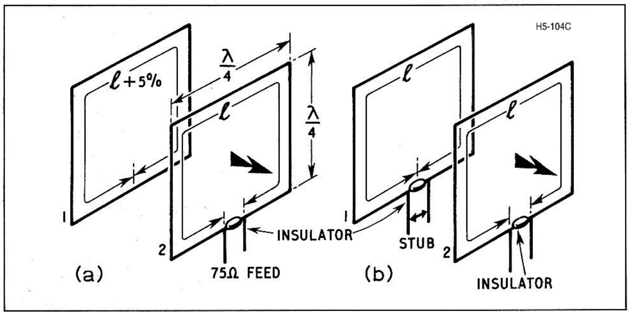


Fig 5.28. Quad antenna dimensions

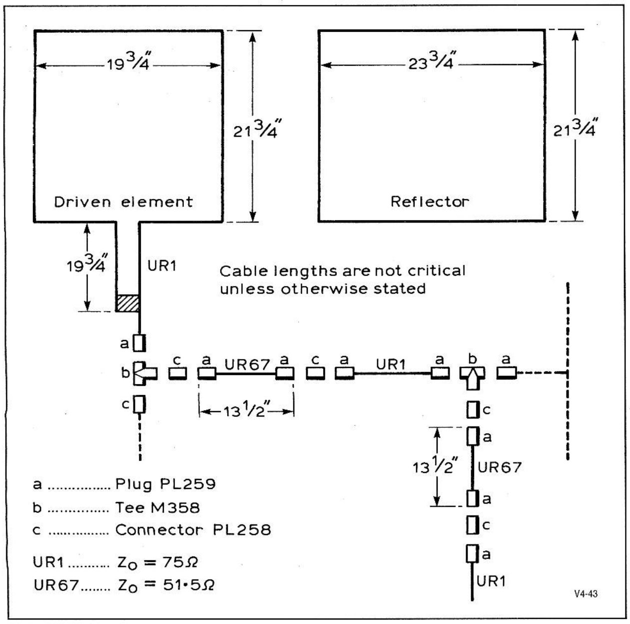


Fig 5.29. Arrangement of two- or four-antenna quad array with power divider and matching details

with a front-to-back ratio of 18dB. The radiation pattern is shown in Fig 5.32.

Quad-element Yagi (quagi)

The quad-element Yagi (Fig 5.36) offers better performance than a simple Yagi of comparable size, together with reduced sidelobes. Up to five elements will perform satisfactorily, although larger structures can be made with care. The relative performance of Yagis with circular (loop) and conventional straight elements is shown in Fig 5.33. Loop Yagis with square and circular elements have comparable characteristics. Comparative measured radiation

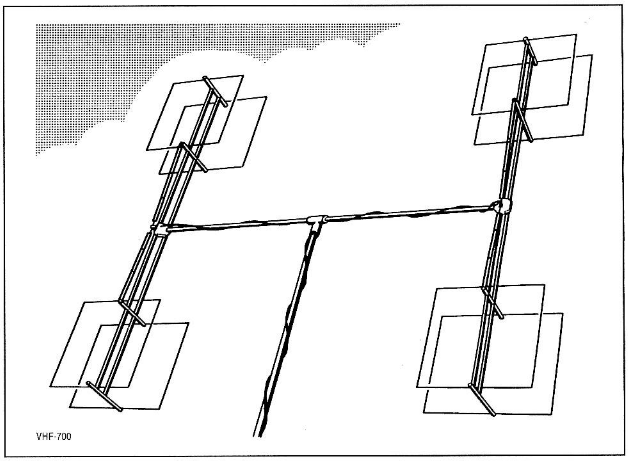


Fig 5.30. A 144MHz cubical quad array

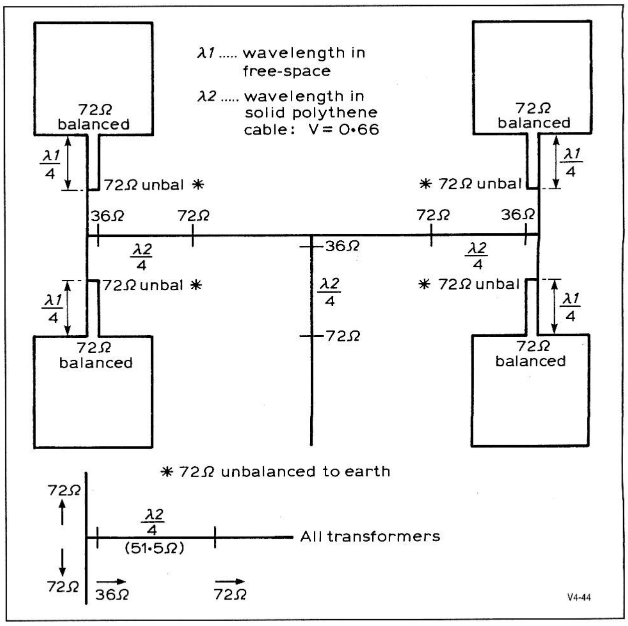


Fig 5.31. Matching and transformer system for four-element quad array

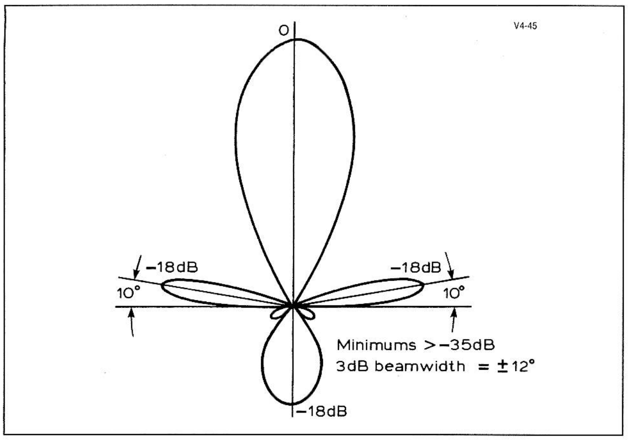


Fig 5.32. Horizontal radiation pattern of four-element quad antenna array

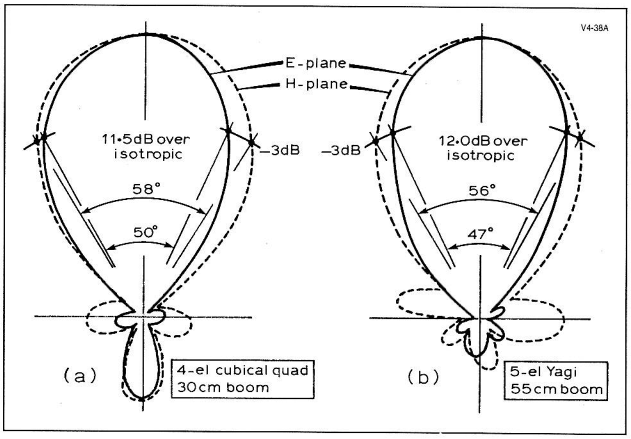


Fig 5.34. Measured voltage patterns of four-element quad and five-element Yagi showing approximately equivalent *beamwidths*

patterns for conventional and quagi antennas are shown in Fig 5.34 and Fig 5.35.

The only insulator required is that of the feed point, resulting in a simple and mechanically robust structure. 9mm aluminium rod or tube is satisfactory for elements for 144MHz and above.

Table 5.6 shows dimensions for several multi-element quagi antennas for 144MHz; dimensions can be scaled for 432MHz. This antenna is relatively easy to construct, and will work well.

A quadruple quad antenna

This collapsible antenna, designed for portable use [18] but equally useable as a fixed antenna for use indoors or in a loft, can achieve gains of between 10 and 11dBi on the 2m band. It is effectively a stacked quad using mutual coupling instead of a phasing harness to excite the outer elements. Constructional details are shown in Fig 5.37.

Each section has a circumference of around 1.04λ, which is not as would be expected for conventional quads. The dimensions are the result of experiments to obtain the best front-to-back ratio and least sensitivity to adjacent objects, which can be important for portable or loft operation, ensuring that the antenna will work without extensive adjustment.

Note that the antenna was designed for low-power (1W)

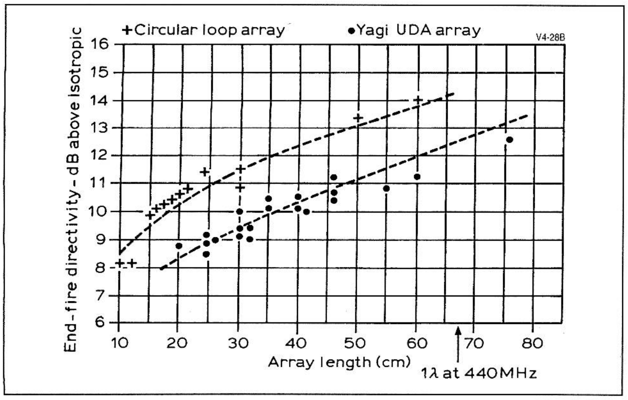


Fig 5.33. Comparative directivity of quad and conventional Yagi antennas as a function of array length (*ARRL Antenna Book*)

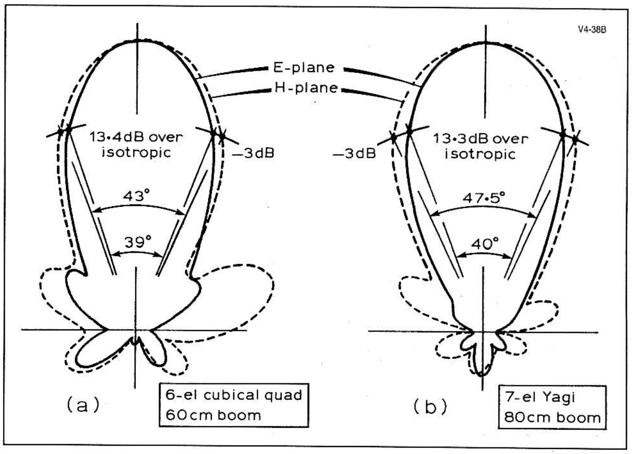


Fig 5.35. Measured voltage patterns of six-element quad and seven-element Yagi (*ARRL Antenna Book*)

operation; the ferrite bead must not be allowed to magnetically saturate, or non-linearities and harmonic generation may occur. The bead may also become hot and shatter.

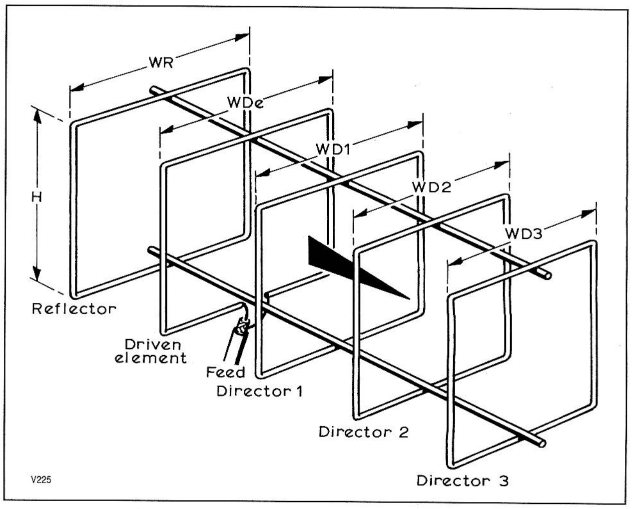


Fig 5.36. General arrangement of a multi-element quad

Table 5.6. Centre-to-centre dimensions for multi-element quad Yagi

Height H	21 (533)	21	21	21
Width reflector WR	24½ (622)	24½	24½	24½
Driven WD_e	20½ (520)	20½	20½	20½
Director 1 WD_1	—	18 (457)	18	18
Director 2 WD_2	—	—	16 (406)	16
Director 3 WD_3	—	—	—	14 (356)
Spacing				
Reflector to Driven	7 (178)	19 (483)	20 (508)	20
Driven to Director 1	—	12 (305)	14½ (368)	14½
Director 1 to Director 2	—	—	14½	14½
Director 2 to Director 3	—	—	—	14½
Approx gain (dBD)	5	7	10.5	12.5

Element diameters all ³/₈in (9.35mm). Feed impedance in all cases is 75Ω. Dimensions are in inches with millimetre equivalents in brackets.

For higher-power operation, ferrite rings could be considered for the balun transformer, or a sleeve balun constructed as appropriate.

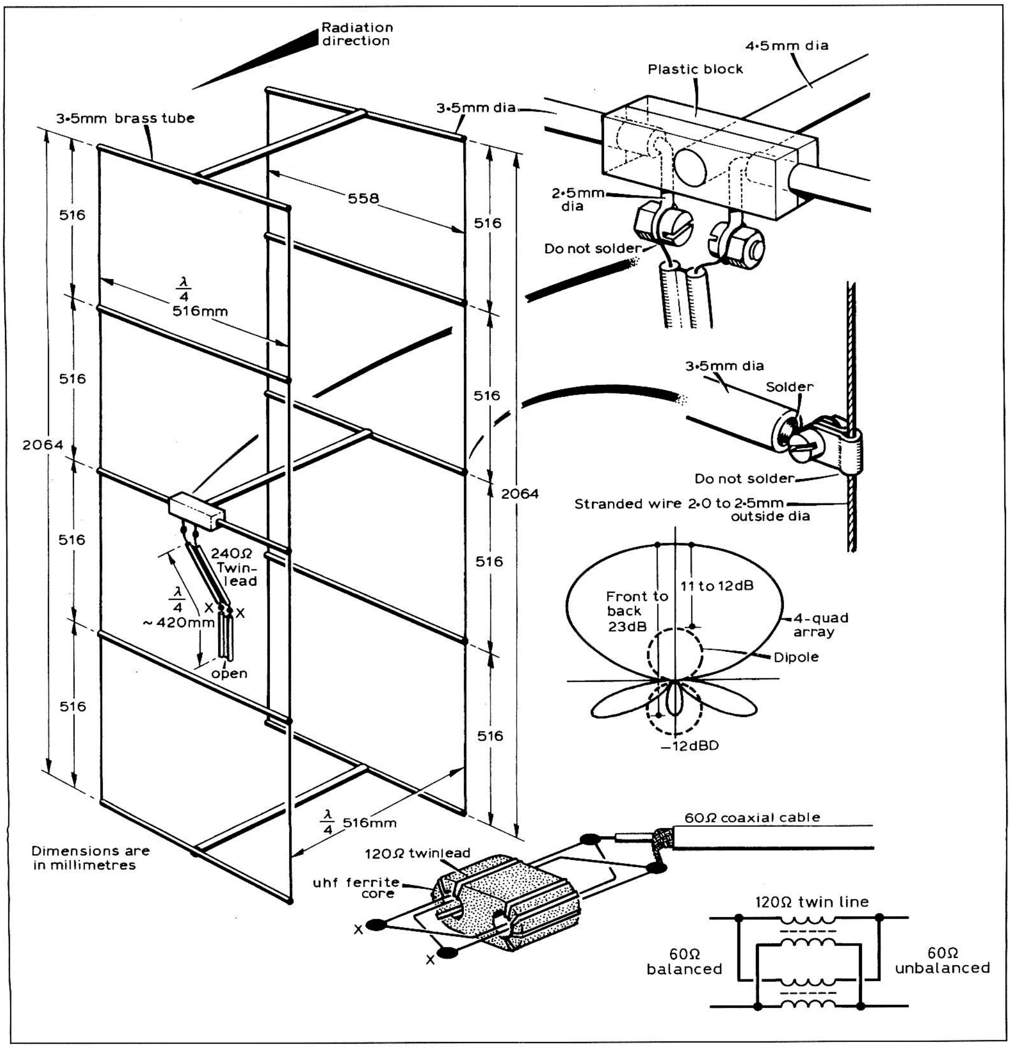


Fig 5.37. Quadruple quad. The match point xx should be found experimentally and will be approximately 200mm from the open end (*VHF Communications*)

Log-periodic antenna

This antenna was originally designed and proved at the University of Illinois in the USA in 1955. Since then the military, in particular, have made considerable use of it. Its particular properties are a very wide bandwidth, governed only by the number of elements used, and the directive qualities of a Yagi antenna.

Tables 5.7 and 5.8 show typical dimensions for element spacings and length for log-periodic arrays which are derived from a computer-aided design produced by W3DUQ in *Ham Radio,* August 1970. Other frequency bands can be produced by simple scaling of *all* dimensions.

The tabulated parameters have a 5% overshoot in the working frequency range at the low end and a 45% overshoot at the high-frequency end to maintain logarithmic response over the complete frequency range specified.

In log-periodic operation approximately four elements are active at any one specific frequency, hence the need for the high-frequency and low-frequency extension. The alpha or logarithmic element taper is 28° for all three antennas, which exhibit a forward gain of 6.55dBD with a front-to-back ratio of typically 15dB and a VSWR better than 1.8:1 over the specified frequency range.

The construction can be straightforward but it should be noted that the element lengths for the highest-frequency antenna were calculated for the elements to be inserted completely through the boom, flush with the far wall. The two lower-frequency antennas have element lengths calculated to butt flush against the element side of the boom. If the elements are to be inserted through the boom on the 21–55MHz and 50–150MHz antennas, the boom diameter must be added to the length of each element.

Table 5.7. Spacing and dimensions for log-periodic VHF antennas

Element	21–55MHz array Length (mm)	Diameter (mm)	Spacing (mm)	50–150MHz array Length (mm)	Diameter (mm)	Spacing (mm)	140–450MHz array Length (mm)	Diameter (mm)	Spacing (mm)
1	3731	38.1	1050	1602	2.54	630	535	6.7	225
2	3411	31.8	945	1444	2.54	567	479	6.7	202
3	3073	31.8	850	1303	2.54	510	397	6.7	182
4	2770	31.8	765	1175	19.1	459	383	6.7	164
5	2496	31.8	689	1060	19.1	413	341	6.7	148
6	2250	25.4	620	957	19.1	372	304	6.7	133
7	2029	25.4	558	864	19.1	335	271	6.7	119
8	1830	19.1	500	781	12.7	301	241	6.7	108
9	1650	19.1	452	705	12.7	271	215	6.7	97
10	1489	19.1	407	637	12.7	244	190	6.7	87
11	1344	19.1	366	576	12.7	219	169	6.7	78
12	1213	12.7	329	522	9.5	198	149	6.7	70
13	1095	12.7	0	472	9.5	178	131	6.7	63
14				428	9.5	160	115	6.7	57
15				388	9.5	0	101	6.7	52
16							88	6.7	0
Boom	7620	50.8	12.7	5090	38.1	152	1823	38.1	152

Table 5.8. Spacing and dimensions for log-periodic UHF antenna (420–1350MHz array)

Element	Length (mm)	Diameter (mm)	Spacing (mm)
1	178	2.1	75
2	159	2.1	67
3	133	2.1	61
4	127	2.1	55
5	114	2.1	49
6	101	2.1	44
7	91	2.1	40
8	80	2.1	36
9	72	2.1	32
10	63	2.1	29
11	56	2.1	26
12	50	2.1	23
13	44	2.1	21
14	38	2.1	19
15	34	2.1	17
16	30	2.1	0
Boom	607	12.7	

As the supporting booms are also the transmission line between the elements for a log-periodic antenna they must be supported with a dielectric spacing from the mast of at least twice the boom-to-boom spacing; otherwise discontinuities will be introduced into the feed system. Feed line connection and the arrangement to produce an 'infinite balun' is shown in Fig 5.39. Any change in the boom diameters will necessitate a change in the boom-to-boom spacing to maintain the feed impedance. The formula to achieve this is:

$$Z_0 = 273 \log_{10} D/d$$

where D is the distance between boom centres and d the diameter of the booms.

The antenna can be orientated either horizontally or vertically (if a non-metal mast section is used) to suit the

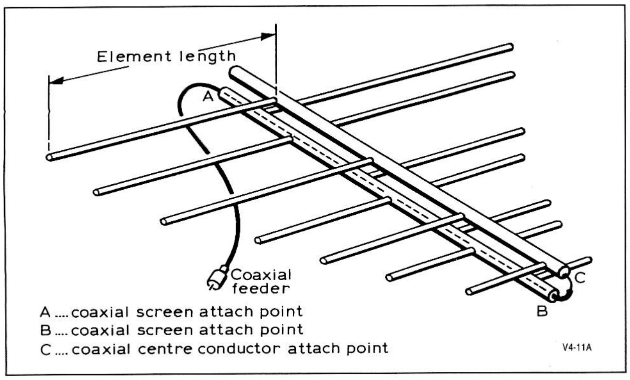


Fig 5.38. Typical log-periodic antenna. Note that the bottom is fed from the coaxial outer while the top boom is fed from the centre conductor (*Ham Radio*)

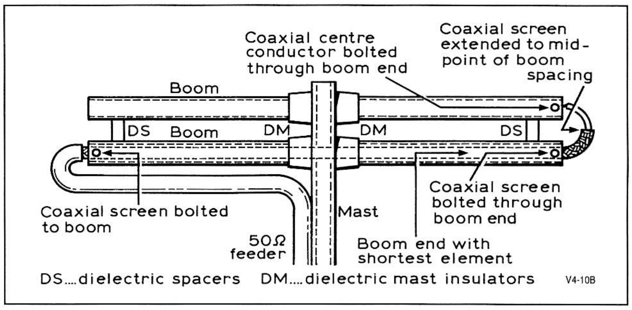


Fig 5.39. Feeding the log periodic is relatively simple. Remove the outer plastic jacket from the feedline for the entire length of the boom, so that the coaxial outer is permitted to short itself inside the boom as well as the solid electrical connections at each end of the boom (*Ham Radio*)

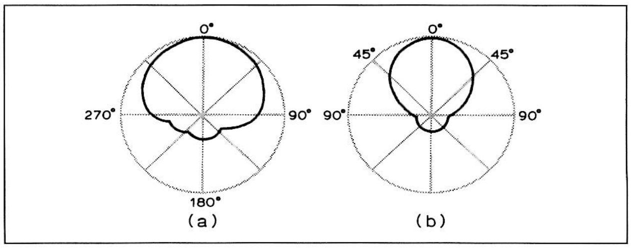
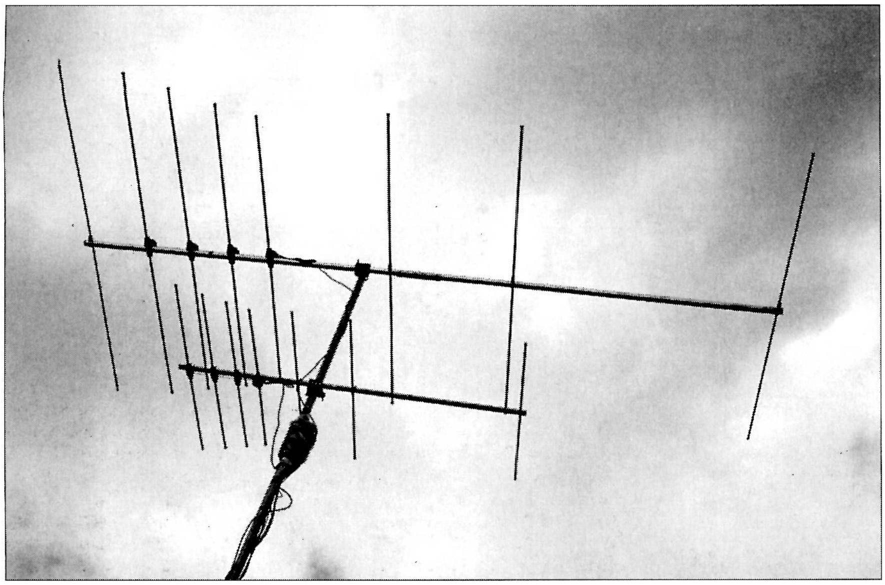


Fig 5.40. Typical log-periodic voltage radiation patterns: (a) horizontal, (b) vertical (*Ham Radio*)



50MHz and 144MHz log-periodic antennas

polarisation required. The horizontal half-power beamwidths will be typically 60° with a vertical half-power beamwidth of typically 100°.

Log-periodic Yagi bandpass antenna

This is an antenna with an interesting and useful band-pass characteristic, giving a flat response over a wide band, and significant attenuation outside. It is basically a combination of a log-periodic driven section with a parasitic Yagi section.

The prototype 50MHz design gave a gain of 12dBD and a bandwidth of 2MHz. The details given in Fig 5.41 are for 144MHz, where the bandwidth would be around 5MHz.

This type of characteristic offers obvious advantages in terms of reducing adjacent channel interference; and also giving a more constant performance over the whole 144 to 148MHz band. The simple Yagi, by comparison, is essentially a narrow-band antenna.

Dimensions and construction details for 50, 70 and 144MHz have been developed by G3FDW [22].

The corner reflector

The use of an aperiodic plane reflector spaced behind a radiating dipole has already been discussed. If this reflector is bent to form a V, as shown in Fig 5.43, a considerably higher gain is achieved. The critical factors in the design of such an antenna array are the corner angle α and the dipole/vertex spacing *S*. The curves in Fig 5.44 show that as α is reduced, the gain theoretically obtainable becomes progressively greater. However, at the same time

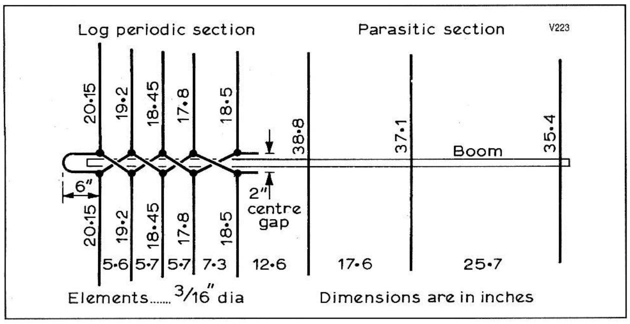


Fig 5.41. A log-periodic Yagi band-pass antenna for 145MHz

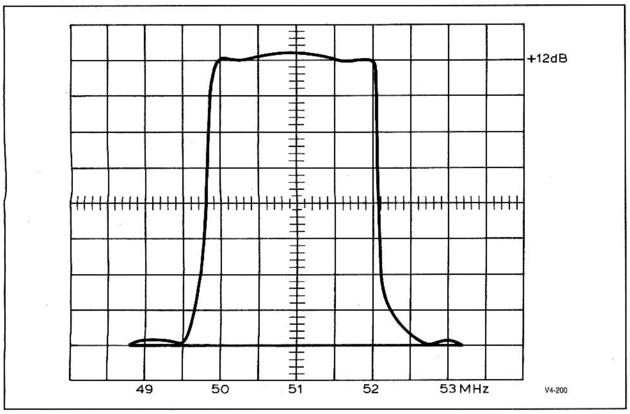


Fig 5.42. Gain versus frequency characteristic of the 50MHz log-periodic Yagi

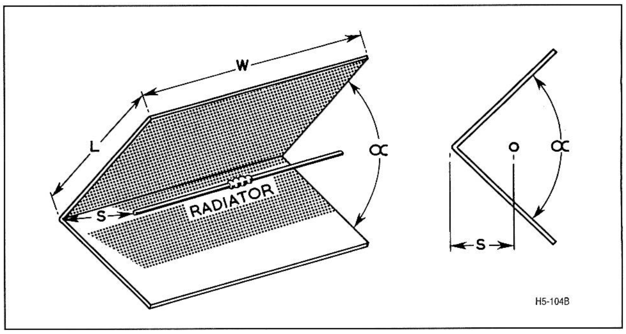


Fig 5.43. Corner reflector. The λ/2 dipole radiator is spaced parallel with the vertex of the reflector at distance S; its characteristics are shown in Figs 5.44 and 5.45

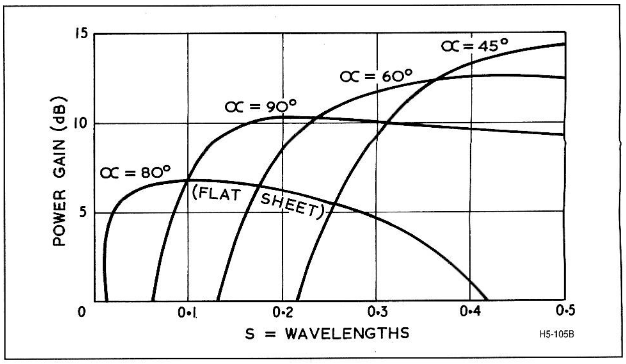


Fig 5.44. Theoretical power gain obtained by using a corner reflector with a λ/2 dipole radiator

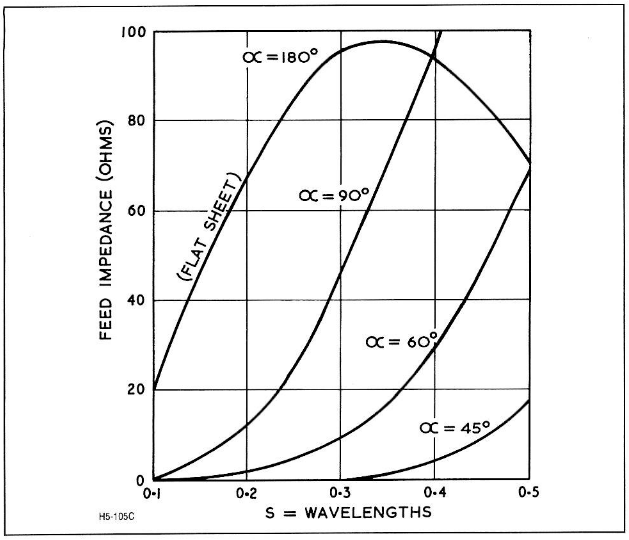


Fig 5.45. Feed impedance of a λ/2 dipole provided with a corner reflector: see Fig 5.43

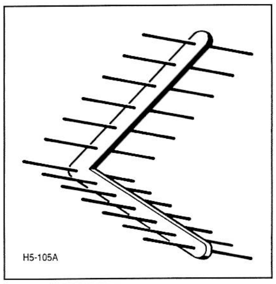


Fig 5.46. The corner reflector can be modified by using a set of metal spines arranged in V-formation to replace the sheet metal or wire-netting reflector

the feed impedance of the dipole radiator falls to a very low value, as can be seen from Fig 5.45. This makes matching difficult and hence a compromise has to be reached. In practice the angle α is usually made 90° or 60°; adjustments in a 60° corner are a little more critical although the maximum obtainable gain is higher. The final matching of the radiator to the line may be carried out by adjusting the distance *S*.

It does not greatly affect the gain over a useful range of variation but causes a considerable change in radiation resistance. A two-stub tuner may also prove helpful in making final adjustments.

The length *L* of the sides of the reflector should exceed 2λ to secure the characteristics indicated by Fig 5.44 and 5.46, and the reflector width *W* should be greater than 1λ for a λ/2 dipole radiator. The reflecting sheet may be constructed of wire netting as described previously or alternatively may be fabricated from metal spines arranged in a V-formation, all of them being parallel to the radiator: see Fig 5.46. The spacing between adjacent rods should not exceed 0.1λ.

A useful approximation for the power gain *G* referred to a λ/2 dipole is $G = 300/\alpha$ where α is the angle between the sides measured in degrees.

Table 5.9. Corner/trough reflector

Angle α (degrees)	Value of *S* for maximum gain (λ)	Gain (dBi)	*T* (λ)
90	1.5	13	1–1.25
60	1.25	15	1.0
45	2.0	17	1.9

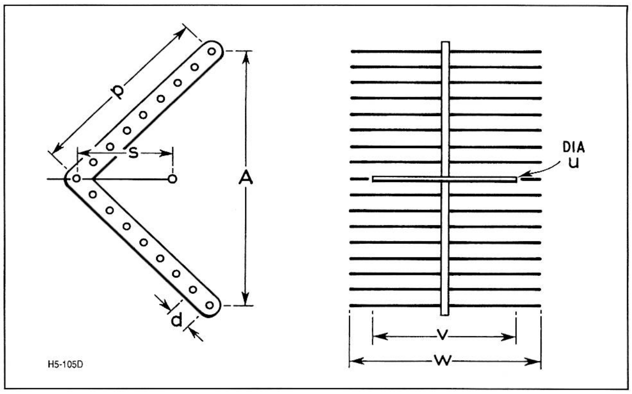


Fig 5.47. Dimensions for a 60° corner reflector antenna system giving a gain of about 13dBD. The feed impedance of the dipole radiator is 75Ω. The apex may be hinged for portable work

Dimensions in millimetres

Band	*p*	*s*	*d*	*v*	*w*	*A*	*u*	λ
144	2540	1016	152	965	1270	2540	9.5	2083
433	889	337	38	324	508	889	6.4	692
1296	305	114	12.7	102	203	305	3.2	232

The maximum dipole/vertex spacing *S* included in the curves shown is λ/2. Spacings greater than this would require rather cumbersome constructions at lower frequencies, but at the higher frequencies larger spacings become practicable, and higher gains than would be suggested by Fig 5.44 can then be obtained; see Table 5.9. This indicates that the corner reflector can become a specially attractive proposition for the 1.3GHz band, but the width across the opening should be in excess of 4λ to achieve the results shown.

HB9CV mini-beam

An antenna that falls into the category of horizontally or vertically polarised, portable rather than mobile, or for base station use, is the HB9CV mini-beam. Similar units are the *lazy H* and *ZL special* often used on the HF bands. The HB9CV version, however, has one or two mechanical advantages which makes it particularly suitable for VHF portable use.

Figs 5.48 (taken from 'The HB9CV Antenna for VHF and UHF', H J Franke, DK1PN, *VHF Communications* February 1969) and 5.49 show two methods of construction for the HB9CV antenna. A point that should be stressed is that a *series* capacitor of 3–15pF is required to adjust finally the gamma match/phasing combination to a VSWR of about 1.3:1 against 50Ω. The dimension of the element spacing and the transmission lines, particularly the spacing (5mm), is critical for optimum impedance matching and phasing, and hence gain and front-to-back ratio.

The principle of operation is as follows. If two dipoles at close spacing, typically 0.1–0.2λ, are fed out of phase, 'end fire' radiation will occur in a direction at right-angles to the line of the dipole elements. If the dipoles are resonant at the same frequency a bidirectional pattern with a gain of typically 3dB referred to a single dipole will be realised. However, if correct phasing between the elements is used, a unidirectional or beam pattern is produced. The different lengths found on most HB9CV antennas assist with bandwidth. The end at which the beam is fed designates the direction of radiation. A theoretical gain in excess of 6dBD should be possible. However, depending on the construction techniques,

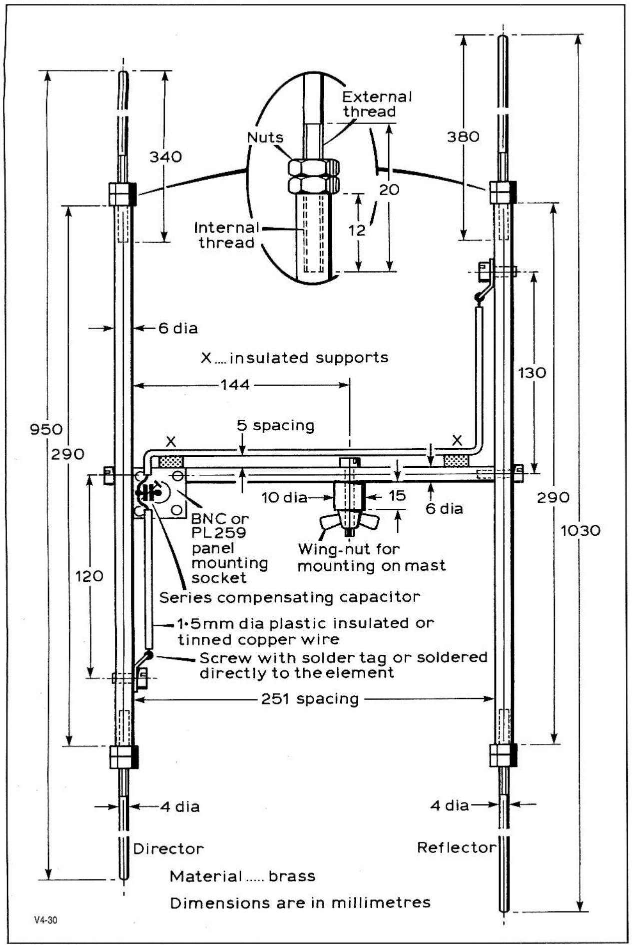


Fig 5.48. A collapsible HB9CV antenna for the 144MHz band (*VHF Communications*)

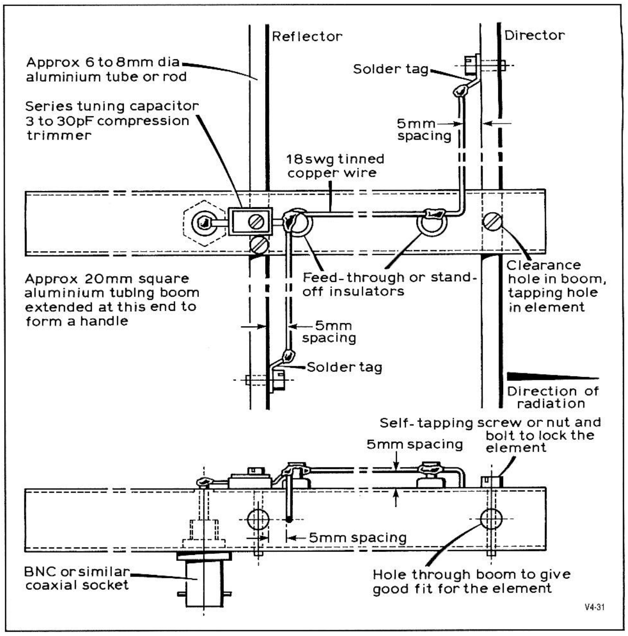


Fig 5.49. Alternative construction of the HB9CV. Dimensions as per Fig 5.48

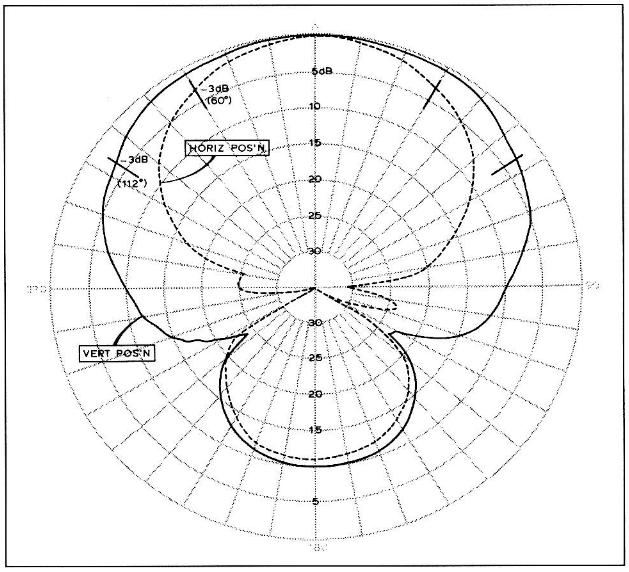


Fig 5.50. HB9CV antenna at 10m above ground

gains between 4 and 5.5dBD with front-to-back ratios between 10 and 20dB tend to be realised in practice. The radiation pattern shown in Fig 5.50 is for the antenna of Fig 5.48 which has a gain of typically 5dBD.

The HB9CV was mounted on a professional glassfibre radiation pattern measuring mast for the 10m test. This ensured a minimum disruption of the antenna radiation pattern when set up for vertical polarisation.

Crossed Yagi with adjustable polarisation

Vertical polarisation is popular for mobile operation in the UK, due to the basic fact that it is far easier to obtain omnidirectional radiation with a vertical antenna than it is with horizontal one. This is particularly important on a vehicle, where the mechanical simplicity of a short vertical rod considerably outweighs the complexity of a halo or crossed dipole, particularly when it is realised that the horizontal antenna must be at least λ/2 above the vehicle surface to ensure low-angle radiation.

Repeaters using vertical polarisation for much the same reason of simplicity of antenna design means that operation of a fixed station, either direct to mobiles or via repeaters, can only be satisfactorily accomplished if a means of changing polarisation is available. It is of course quite possible to use two antennas, and ideally two rotating systems, but the cost becomes rather formidable.

Space communication, where control of polarisation is difficult or impossible, has forced the use of circular polarisation and it is surprising that it is not used more between fixed stations for long-distance terrestrial work. The fundamental advantage of circular polarisation is that all reflections change the direction of polarisation, precluding the usual addition or subtraction of main and reflected signal; therefore there is far less fading and aircraft flutter when circular polarisation is used at each end of the link. The use of circular polarisation at one end only, with normal horizontal or vertical at the other

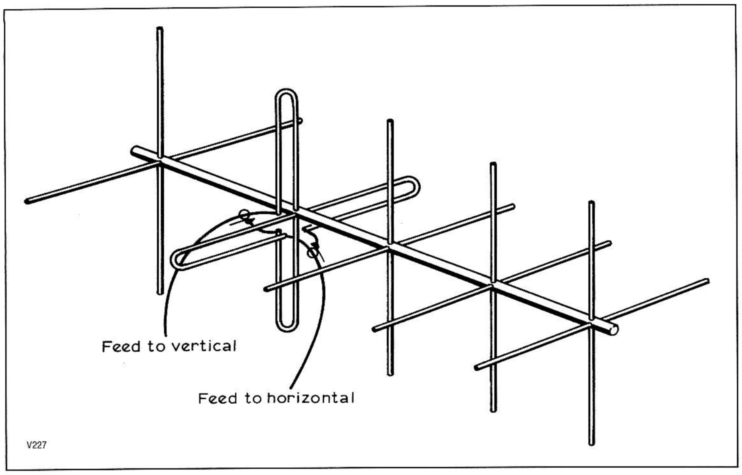


Fig 5.51. General arrangement of a crossed Yagi antenna

end of the link, naturally results in a 3dB loss, and therefore to achieve the full advantages of circular polarisation it is necessary for all stations to use it.

The usual practice when using circular polarisation for terrestrial communications is to standardise on clockwise or 'righthand' in the northern hemisphere, and this may well become standard for the amateur by its regular adoption. The direction of polarisation is referred to as viewed from the *rear* of the antenna.

Changing all VHF operations to circular polarisation is obviously not practical, but if a system of switching polarisations were in use at all stations it would soon become evident that circular offers advantages, and there would of course be the added bonus that vertical would be available for operation with mobiles. Having used a system of polarisation switching, big variations are found in polarisation from stations, in particular mobiles. Quite often a mobile using a vertical antenna has been found to be of equal strength on all polarisations and in some cases a definite advantage for circular has been shown.

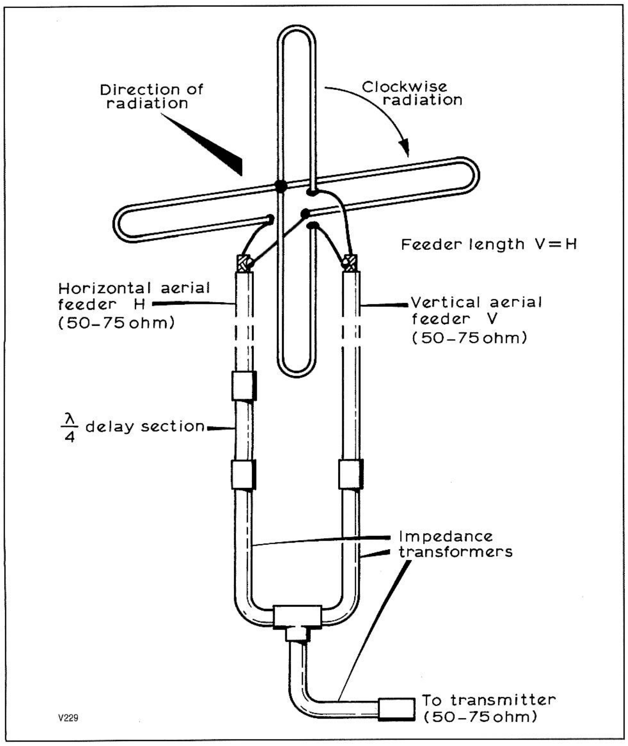


Fig 5.52. General arrangement of feeders with delay line (phasing) for clockwise radiation

Circular polarisation normally brings to mind the helix antenna, which can only produce modes of circularity, depending upon whether the thread of the antenna element is wound clockwise or anti-clockwise. Horizontal or vertical polarisation is possible from helix antennas, but only by the use of two helices and suitable phasing, with no real means of control. The simple means of changing polarisation is to mount a horizontal Yagi and a vertical Yagi on the same boom, giving the well-known *crossed Yagi*. Separate feed to each section of the Yagi brought down to the operating position will enable the user to switch to either horizontal or vertical, but it is perhaps not generally realised that it is a relatively simple matter to alter the phasing of the two Yagis in the shack and obtain four more polarisation options, namely two slant positions (45° and 135°), together with two circular positions (clockwise and anti-clockwise) which with horizontal and vertical gives six positions altogether. This capability is also of great assistance for transmission and reception through satellites.

Although vertical polarisation is mechanically and electrically advantageous when using a simple dipole type of antenna, the presence of the mast in the same plane as the vertical elements on a Yagi considerably detracts from performance. This can be very simply overcome with a crossed Yagi with polarisation switching, mounting the antenna with elements at 45°. The mast then has little effect on the input impedance of the two Yagis, and vertical and horizontal polarisations can still be produced by feeding both antennas in the correct phase relationship.

Assuming therefore that a crossed Yagi is mounted at 45° with individual feeders to the operating position, the polarisation available and the phasing required is as follows:

Slant position 45° and 135°	Antennas fed individually
Circular positions clockwise and anti-clockwise	Both antennas fed with 90°+ or 90°– phase relationship
Horizontal and vertical	Both antennas fed with 0° or 180° phase relationship

This all sounds very complicated, but in actual fact the desired result may be accomplished relatively simply with a three-gang six-position Yaxley-type wafer switch. A coaxial switch is the 'pure' way to do the job but, considering the cost of a three-gang six-way coaxial switch together with the necessary plugs and sockets, the difference in performance is just not worthwhile on 144MHz.

The first problem to overcome is simply that of providing the correct matching for feeding two antennas in parallel. Briefly, with 75Ω antennas the two feeders are simply paralleled, giving 37.5Ω, and λ/4 of 50Ω feeder used to transform

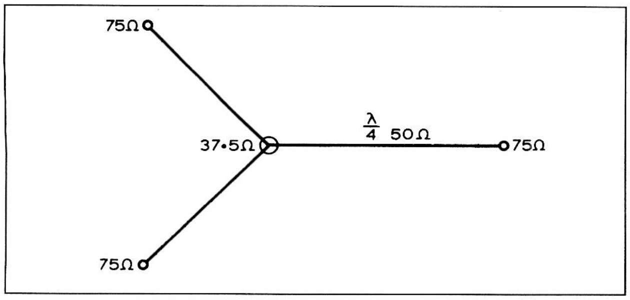


Fig 5.53. Matching two 75Ω antennas by paralleling to 37.5Ω and increasing impedance to 75Ω again

back to 75Ω, as illustrated in Fig 5.53. 50Ω antennas are treated in a slightly different way in that λ/4 of 75Ω feeder is used in each feeder to transform up to 100Ω and the two are placed in parallel to produce 50Ω again, as shown in Fig 5.54.

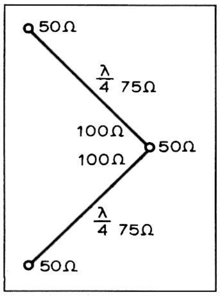


Fig 5.54. Matching two 50Ω antennas by increasing impedance to 100Ω and paralleling to 50Ω again

Phasing is simply a question of altering the length of the feeders to each half of the crossed Yagi as the polarisation is changed. Where a 90° phase shift is required, λ/4 of feeder is inserted and where a 180° phase shift is required, λ/2 feeder is inserted. The polarisation switch must therefore arrange for correct matching by switching in the appropriate λ/4 impedance transformer and correct phasing by switching in the appropriate length of feeder.

There is an added complication in that by no means all antenna systems are 50Ω, and a considerable number of 75Ω users still remain on VHF. 50Ω has become an international standard and is of course completely standard on low frequency; it can therefore only be a matter of time before all VHF installations are 50Ω.

Figs 5.55 and 5.56 show the necessary switching arrangements for 75Ω and 50Ω antennas respectively. The normal drawing of a switch makes the illustration of the 50Ω system extremely complicated, and Fig 5.56 is drawn as a side view of the Yaxley switch with the six contacts visible in a vertical line, the moving contact not being shown. It will be noticed that the 50Ω version is much simpler as there is no need to manufacture T-junctions in the cables.

It is very necessary for the phasing lengths of feeder to be accurately cut and this may be simply accomplished with a GDO. First, use the smallest possible diameter cable to minimise the mechanical problems of connection to the contacts of the switch. Types UR43 for 50Ω and UR70 for 75Ω are to be preferred and certainly a solid dielectric type should be used in the interests of uniformity. To obtain λ/4 of cable, cut off slightly more than the calculated length, which in the case of 144MHz will be 15in of solid dielectric cable, leave one end open-circuit, and short the other end with the shortest possible loop that will produce a dip on the GDO. It is surprising just *how small that* loop can be and, given a reasonably sensitive GDO, a virtual short-circuit will still couple. Check the dip frequency, which will probably be around 120MHz, and carefully clip pieces off the open end of the cable until the dip

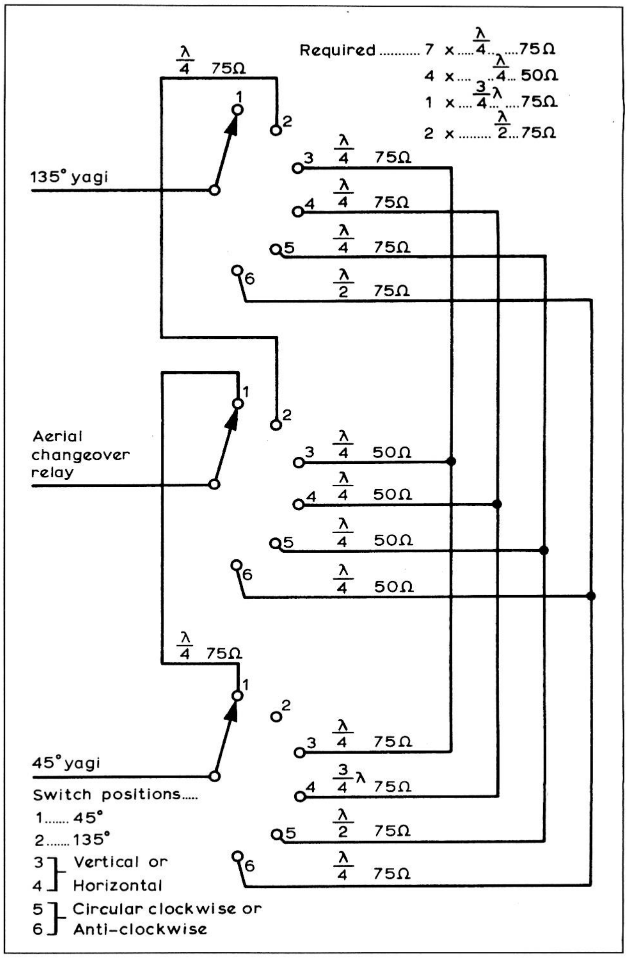


Fig 5.55. 75Ω phasing and matching switch

occurs at 145MHz. Assuming that a solid dielectric cable of similar size is used throughout the switch, there is no necessity to dip each length. The uniformity of the cable is sufficient simply to copy mechanically this λ/4 and to double or treble

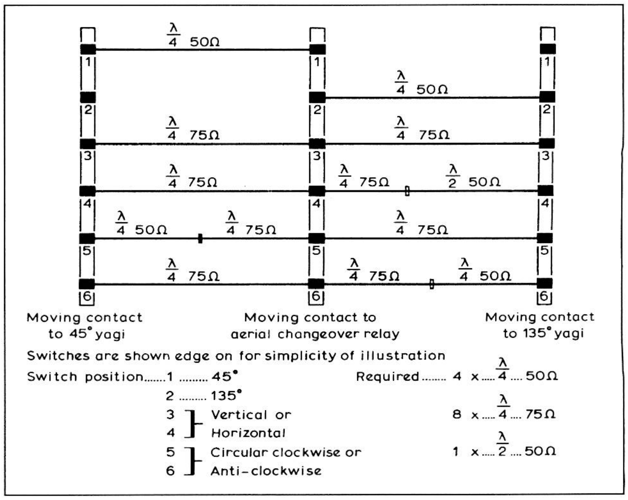


Fig 5.56. 50Ω phasing and matching switch

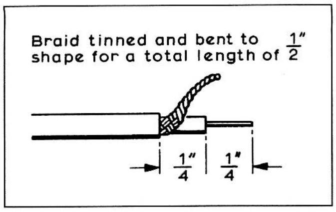


Fig 5.57. Method of 'tailing' coaxial cable

it where λ/2 or 3λ/4 is required. The slight shortening of the cables when they are prepared for connection is compensated by the length in the switch contacts.

Remember when wiring the switch that every effort should be made to maintain impedance and all cable ends should be made up as short as possible to the configuration shown in Fig 5.57. All outer braids on each wafer of the switch must be joined together by the shortest possible route and not connected to the frame of the switch. The use of miniature switches with small-diameter cable makes for a beautifully neat assembly, but very great care indeed is needed to deal with the many coaxial connections in a switch of this small size. The joining of a length of 50Ω and 75Ω is important, and here every effort should be made to maintain the coaxiality of the cable by pushing the braid back away from the inner, making the inner connections carefully, taping up with polythene tape to avoid any possible short-circuit, and then bringing the braids back again over the tape and binding securely with fine wire. Any attempt at soldering will probably be disastrous, as the polythene will undoubtedly melt with the risk of short-circuit. Further protection may be given by a layer of tape over the entire joint. Similarly, the T-junctions on the 75Ω switch may be made up by cutting small triangular sections of tinplate and quickly soldering the outers of each cable to the tin; in this case short-circuits may be seen and avoided. Fig 5.58 illustrates the method.

Assuming that the switch has been satisfactorily built, there is now the problem of whether the feeders to the halves of the crossed Yagi are of the correct individual length. Ideally, these feeders should be cut mechanically and electrically to equal length before installation, and the two halves of the crossed Yagi should be in exactly the same place on the boom. While the feeders may be cut accurately, it is mechanically difficult and almost impossible to mount the two halves of the Yagi in the same place. They inevitably have to be spaced by a few inches. It is therefore necessary to correct this mechanical displacement of phase by an equal displacement of length of the feeders, and in practice it is far easier to simply connect everything up with unknown lengths of feeders and adjust the length of one or both feeders until the switch operates correctly.

A convenient method of adjustment is to receive a horizontally polarised signal of constant amplitude from a local station, ensuring that the transmitting and receiving antennas

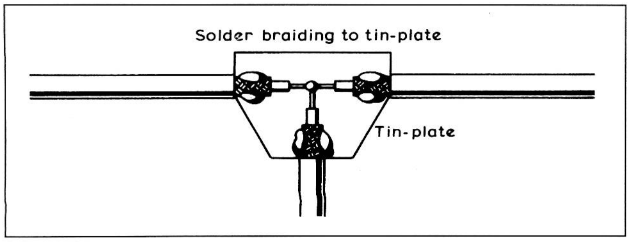


Fig 5.58. Method of joining three cables

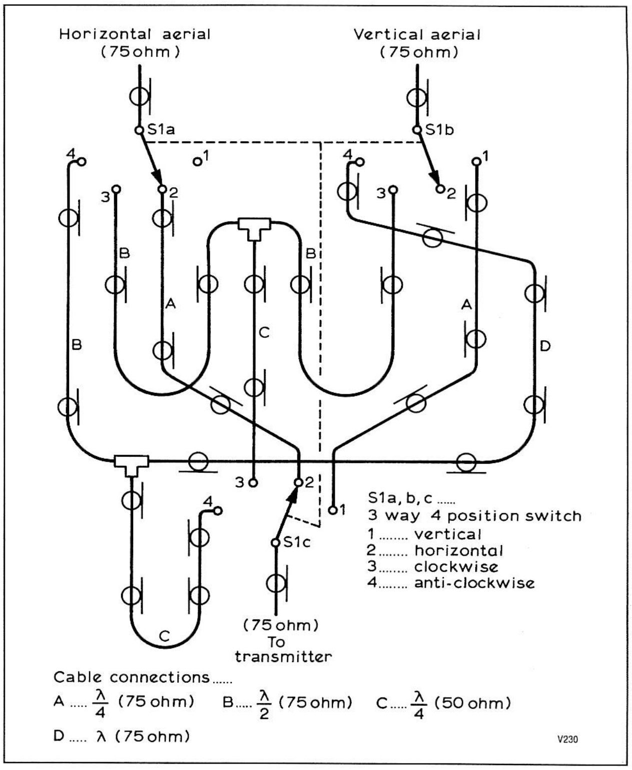


Fig 5.59. An alternative arrangement for feeding crossed Yagi antennas which provides various polarisations at the click of a switch

are beamed directly at each other. This point is vitally important – a beam antenna only radiates its intended polarisation from the main lobe – a fact which will become very evident in subsequent use of the switch. The feeder lengths should now be adjusted so that all slant and circular positions are equal, together with maximum rejection in the vertical position of the switch. The choice of which shall be the horizontal and vertical positions can now be taken. Accurate S-meter readings logged for each position of the switch after every feeder adjustment are essential. Typically, the slant or circular positions will be about one S-point down on the horizontal, while the vertical position will be some six S-points or 20 to 30dB down. To avoid the problem of the man trying to level the legs of a four-legged table and finishing up with a 3in-high table, when cutting feeder lengths cut only 1in at a time from one feeder. When the recorded readings indicate that the last cut as one too many, cut that last piece from the other feeder and the optimum situation will be restored.

With the Yagis mounted at 45°, it may appear surprising that a horizontal signal can produce differing signal strength on each antenna, but this will happen until the respective feeders are of equal length. The reason is the inevitable mismatch (sometimes deliberate to improve noise factor) which occurs at the input to the converter or receiver. Remember the object is *equal* signals, not maximum signals – converter mismatch can be compensated for and *maximum* signal strength achieved by altering the length of the main feeder after the switch, which will not affect the phase relationships between the antennas.

The question now arises as to which of the circular polarisation positions are clockwise or anti-clockwise. This subject

Table 5.10. Received signals expected with various switch connections

Switch position	Polarisation of signal (dB down)					
	Horiz-ontal	Vert-ical	45°	135°	Clock-wise	Anti-clock-wise
Horizontal	Max	20/30	3	3	3	3
Vertical	20/30	Max	3	3	3	3
45°	3	3	Max	20/30	3	3
135°	3	3	20/30	Max	3	3
Clockwise	3	3	3	3	Max	20/30
Anti-clockwise	3	3	3	3	20/30	Max

merits an entire article; it will be remembered that even the world's top telecommunication engineers got this one wrong on the first transatlantic TV broadcast via Telstar. Should the operator wish to define the circular positions, then with accurately cut equal feeders and an accurately made switch, position 5 will be clockwise and 6 anti-clockwise, providing the antenna connections are as shown in Fig 5.56. If the antenna connections are not known, then the only way to calibrate the switch is to receive a known circularly polarised signal, when the respective positions will be immediately evident.

A correctly wired and phased switch should perform as in Table 5.10.

Axial-mode helix

The helix antenna is a simple means of obtaining high gain and wide-band frequency characteristics. When the circumference of the helix is of the order of 1λ axial radiation occurs, ie the maximum field strength is found to lie along the axis of the helix. This radiation is circularly polarised, the sense of the polarisation depending on whether the helix has a right-hand or left-hand thread.

If a pick-up dipole is used to explore the field in the direction of maximum radiation, the signal received by this dipole will show no change of amplitude as it is rotated through 360°, thus indicating true circular polarisation. At any point to the side of the helix the wave will be elliptically polarised, ie the horizontal and vertical components will be of unequal strength.

A helix may be used to receive the circularly polarised waves radiated from a transmitting helix but care must be taken to ensure that the receiving helix has a thread of the same sense as the radiator. If a thread of the wrong sense is used, the received signal will be very considerably weaker.

The properties of the helical antenna are determined by the diameter of the spiral *D* and the pitch *P* (see Fig 5.60) and depends upon the resultant effect of the radiation taking place all along the helical conductor. The gain of the antenna depends on the number of turns in the helix. The diameter of the reflector *R* should be at least λ/2, the diameter of the helix *D* should be about λ/3 and the pitch *P* about λ/4.

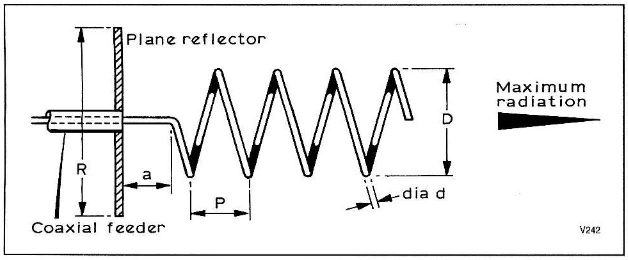


Fig 5.60. The helix antenna. The plane reflector may take the form of a dartboard type of wire grid. The dimensions given in Table 5.11 are based on a pitch angle of 12°. The helix, which may be wound of copper tube or wire, the actual diameter of which is not critical, must be supported by low-loss insulators

A helix of this design will have a feed impedance of about 140Ω; this may be transformed to the feeder impedance by means of a λ/4 transformer. A typical helical antenna having a seven-turn helix has a gain of approximately 12dBi over a 2:1 frequency range. However, to achieve this gain fully it is necessary to use a circularly polarised antenna (eg a helix of the same sense) for reception. If a plane-polarised antenna, such as a dipole, is used there will be a loss of 3dB.

A practical helix antenna for 144MHz

The greatest problem to be overcome in this type of antenna for 144MHz, with its relatively large helix diameter of 24½in, is the provision of a suitable support structure.

Fig 5.61 shows a general arrangement, in which three supports per turn (120° spacing) are shown, and details of suitable drilling of the central boom are given in Fig 5.62.

The helix may be made of copper, brass, or aluminium tube or rod, or coaxial cable. This latter alternative is an attractive material to use, being covered and substantially weatherproofed. If coaxial cable is used the inner conductor should connect to the outer at each end, or be removed completely.

The reflector is located at a distance *a* behind the start of the first turn, and is supported by crossed supports from the central boom. The material for the reflector can be any kind of metal mesh such as chicken netting or plastic-coated garden mesh.

The central boom should be sufficiently rigid to adequately support the whole structure, and should ideally at the same time be of a non-metallic material such as wood, thick-wall plastic tube or thick-wall glassfibre. Although glassfibre is

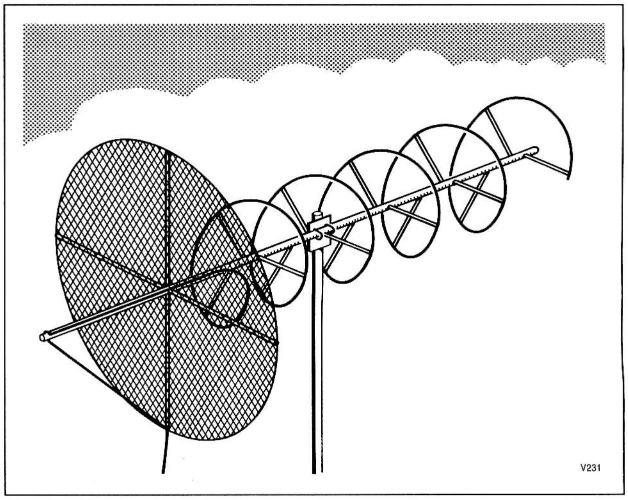


Fig 5.61. General arrangement ot support structure for a five-turn helical antenna for 144MHz

more expensive it would undoubtedly be worthwhile for a permanent installation.

The length of the final turn of the helix can be adjusted to obtain optimum circularity. This would entail rotating a dipole set up in line with the helix at a distance of, say, 10m, to be outside the near field and clear of all objects. The signal obtained from the dipole will be constant for all points of rotation when the helix is optimised for circular polarisation. Any variation of the signal is known as the *polarisation axial ratio* or *boresight ellipticity,* and is usually expressed as a ratio or decibel figure. Helix antennas for higher frequencies are easier to construct and require little adjustment. Detailed instructions for building a 435MHz helix have been published by G3RUH [23].

Table 5.11. General dimensions for 144, 433 and 1296MHz helix antennas

Band	Dimensions D	R	P	a	d
General	0.32λ	0.8λ	0.22λ	0.12λ	
144MHz	25½ (648)	64 (1626)	17¾ (450)	8¾ (222)	½ (12.7)
433MHz	8¾ (222)	22 (559)	6 (152)	3 (76)	3/16–½ (4.8–12.7)
1296MHz	3 (76)	7 (178)	2 (50)	1 1/8 (28)	¼–1/8 (3.2–6.4)
Turns	6	8	10	12	20
Gain	12dB	14dB	15dB	16dB	17dB
Beamwidth	47°	41°	36°	31°	24°

Dimensions in inches, millimetres are given in brackets. The gain and beamwidth of the helical antenna are dependent upon the total number of turns as shown above.

Bandwidth $= 0.75$ to 1.3λ

Feed impedance $= 140 \times \dfrac{\text{circumference}}{\lambda}$ ohms

(Note: λ and circumference must be in the same units.)

Beamwidth (degrees) $= \sqrt{\dfrac{12{,}300}{\text{No of turns}}}$

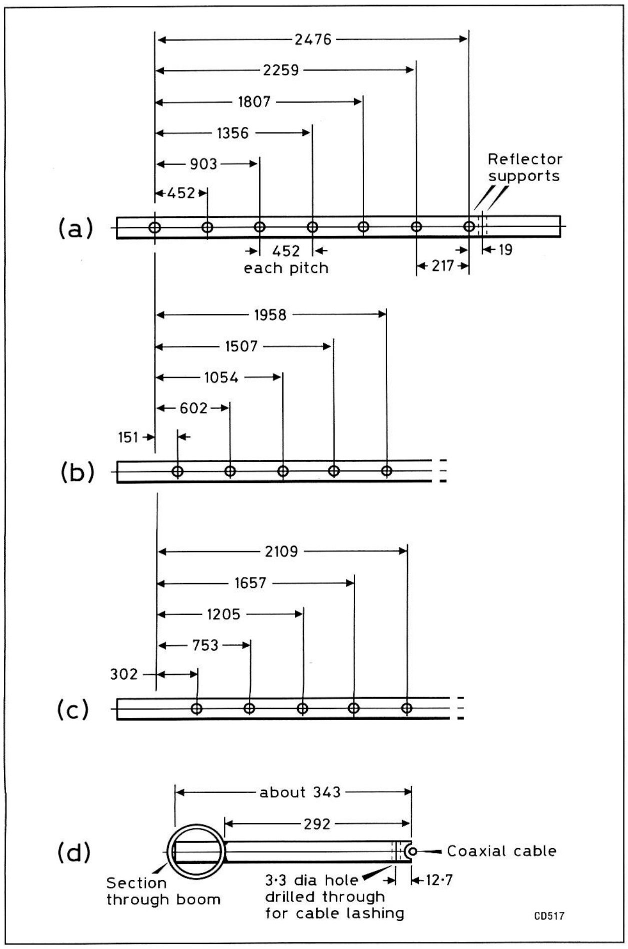


Fig 5.62. (a) First side drilling dimensions, reflector support holes are drilled at right-angles; (b) and (c) are drilled at intervals of 120° and 240° respectively from (a). (d) Cutting and filing dimensions for the element stand-offs

OMNIDIRECTIONAL ANTENNAS FOR FIXED STATIONS

The horizontally polarised omni-V

This antenna consists of a pair of λ/2 dipoles. The ends of the dipoles are physically displaced to produce quadrature radiation and are supported on a λ/4 shorted stub. A pair of Q bars are tapped down the stubs to a point where the impedance is 600Ω so that when the two units are fed in parallel they produce an impedance of 300Ω at the centre. A 4:1 balance-to-unbalance coaxial transformer is fitted to the centre point of the Q bars so that a standard 75Ω coaxial cable feeder may be used.

A 50Ω feed can be arranged by repositioning the Q bars on the antenna stubs. This can best be achieved by monitoring the VSWR on the coaxial feeder whilst adjusting the Q bar position by small but equal amounts on both stubs.

The general arrangement is shown in Fig 5.63(a). Fig 5.63(b) shows how the antenna may be arranged to give a bidirectional radiation pattern.

Simple crossed dipoles

The ordinary turnstile, also known as *crossed dipoles,* provides a simple yet very effective horizontally polarised, omnidirectional antenna. It consists of two horizontal dipoles

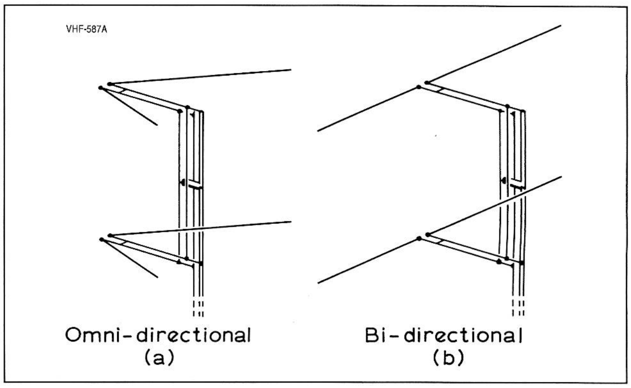


Fig 5.63. Formation of the omni-V antenna

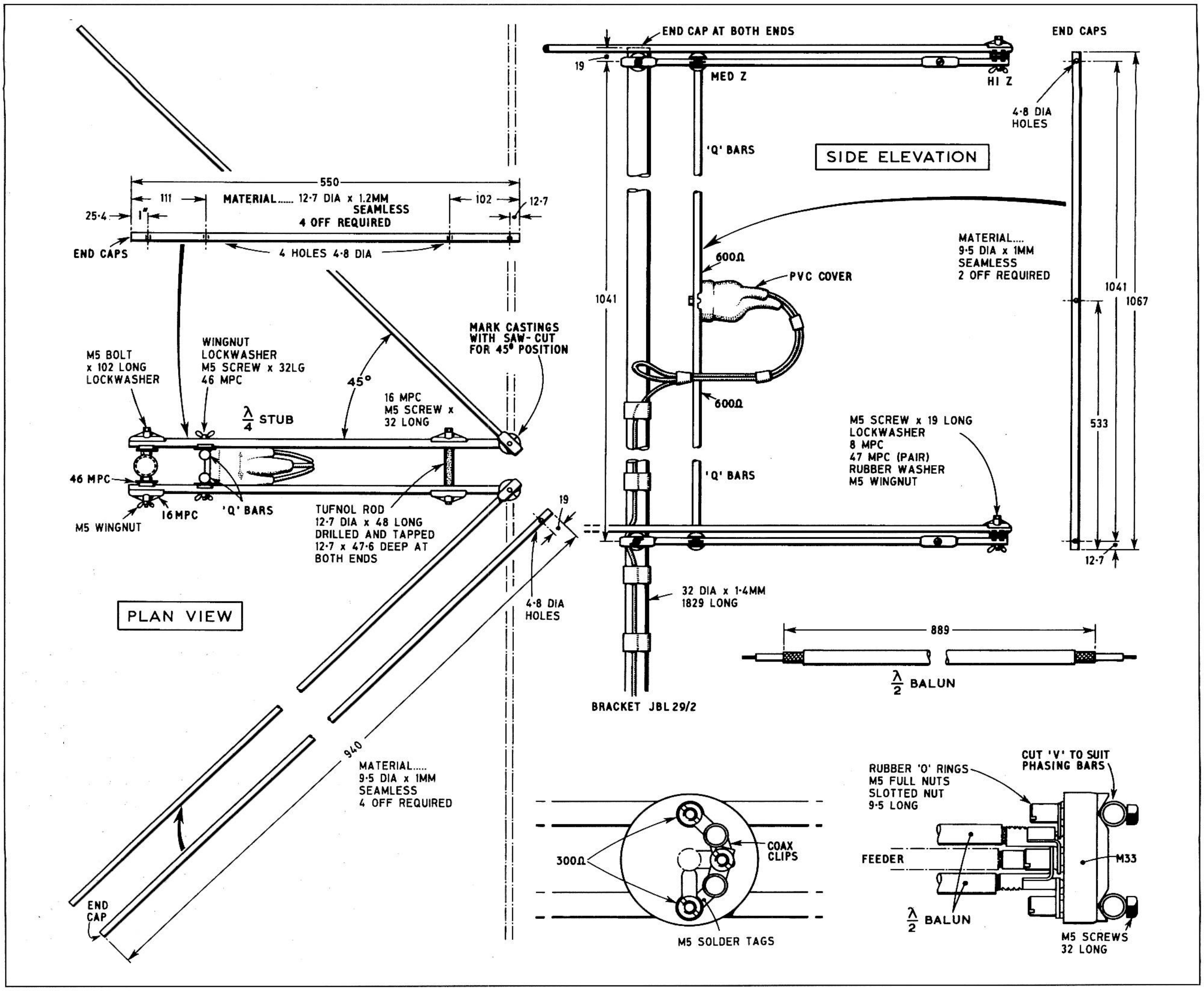


Fig 5.64. The mechanical details of the construction of the omni-V (dimensions are in millimetres)

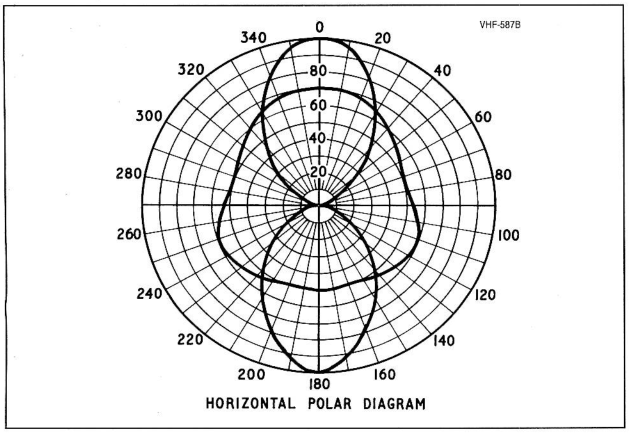


Fig 5.65. The horizontal polar diagram for an average antenna, *showing* both the bidirectional and omnidirectional charts

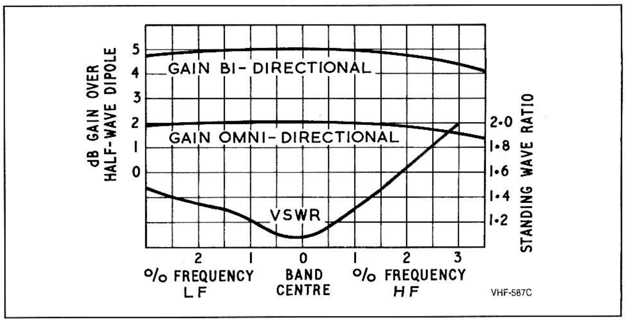


Fig 5.66. Chart showing gain versus VSWR on the omni-V just described

mounted at right-angles and fed with equal power but at 90° phase difference (Fig 5.67). Matching to 75Ω is quite simple and with a little adjustment a very low SWR can be obtained. The radiation pattern produced tends to be square with the 'corners' at element ends being about 1dB up on a dipole's maximum radiation.

The 90° phase difference is readily obtained with a resonant λ/4 feeder between the dipoles and a further resonant

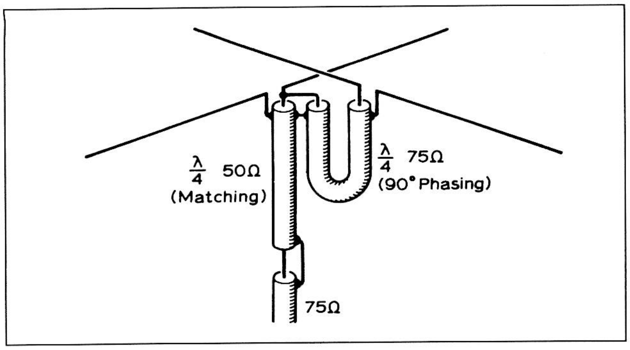


Fig 5.67. Phasing and matching arrangement of crossed dipoles

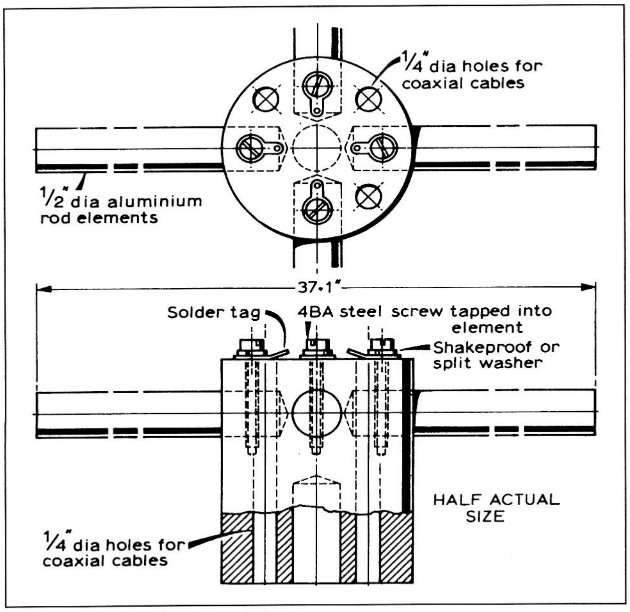


Fig 5.68. Details of central insulator

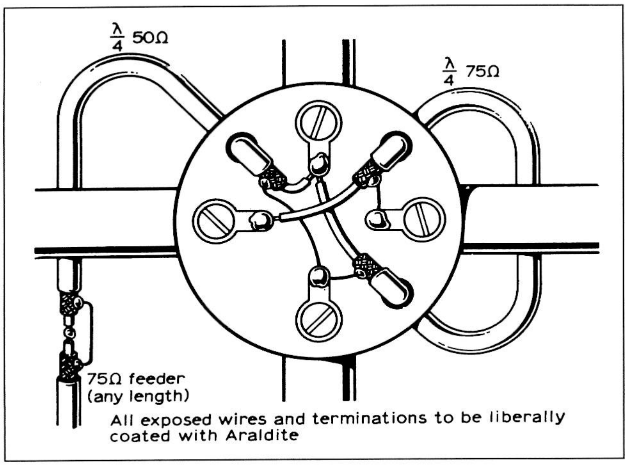


Fig 5.69. Connections of coaxial sections

series λ/4 section as a matching transformer for the characteristic line impedance. The antenna should be mounted at least 0.5λ above any conducting surface, otherwise much of the signal will be radiated upwards.

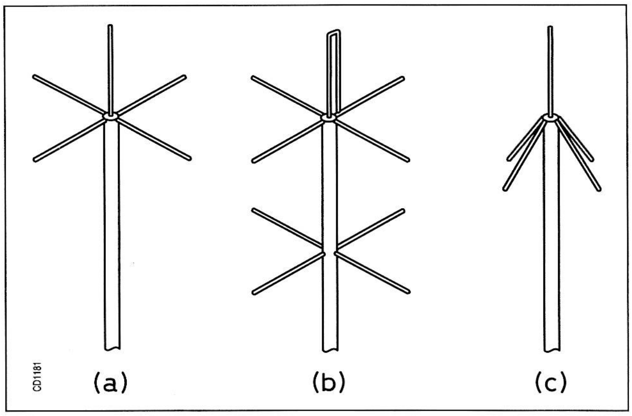


Fig 5.70. Quarter-wave ground-plane antennas. The folded monopole radiator may be used with any of the ground-plane configurations

The λ/4 ground-plane antenna

This is one of the simplest omnidirectional antennas to construct and usually yields good results. However, some unexpected effects may occur when the antenna is mounted on a conductive mast, or if RF current is allowed to flow on the outside of the coaxial feeder.

In its simplest form, the ground-plane antenna comprises a λ/4 extension to the inner of a coaxial cable, with several wires extending radially away from the end of the outer of the coaxial cable: Fig 5.70(a). The input resistance will be quite low, of the order of 20Ω, although this may be transformed to a higher impedance by using a folded monopole radiator as shown in Fig 5.70(b). Equal-diameter elements provide a 4:1 step-up ratio to around 80Ω, and a smaller diameter grounded leg can reduce the input impedance to 50Ω.

The feedpoint impedance can be modified by bending the ground plane rods downwards from the horizontal: Fig 5.70(c). If the radiating element and the ground plane rods are all λ/4 long, the input resistance is approximately:

$$R = 18(1 + \sin\theta)^2 \quad \text{ohms}$$

where θ is the ground-plane rod angle below the horizontal, in degrees. A 50Ω resistance is achieved when θ is 42°.

The ends of the ground-plane rods are sometimes joined together with a conductive ring to provide additional mechanical stability. The ring increases the electrical size of the ground plane, and the length of the radials can be reduced by about 5%.

The few rods forming the ground plane usually do not prevent current flowing on any conductive supporting mast or on the outside of a coaxial feeder. The mast or feeder can become a long radiating element which may enhance or destroy the radiation pattern of the antenna, dependent upon the magnitude and phase of the mast currents relative to that on the antenna. An example of this is shown in Fig 5.71(a), where the monopole and ground plane is mounted on a 5λ mast (about 10m). The corresponding radiation patterns without mast or cable influences are shown in Fig 5.71(b). The effects of ground reflections have been suppressed in both cases.

Some antenna designs make use of these currents to enhance the gain of the monopole; they sometimes have a second set of ground-plane rods further down the mast, tuned to

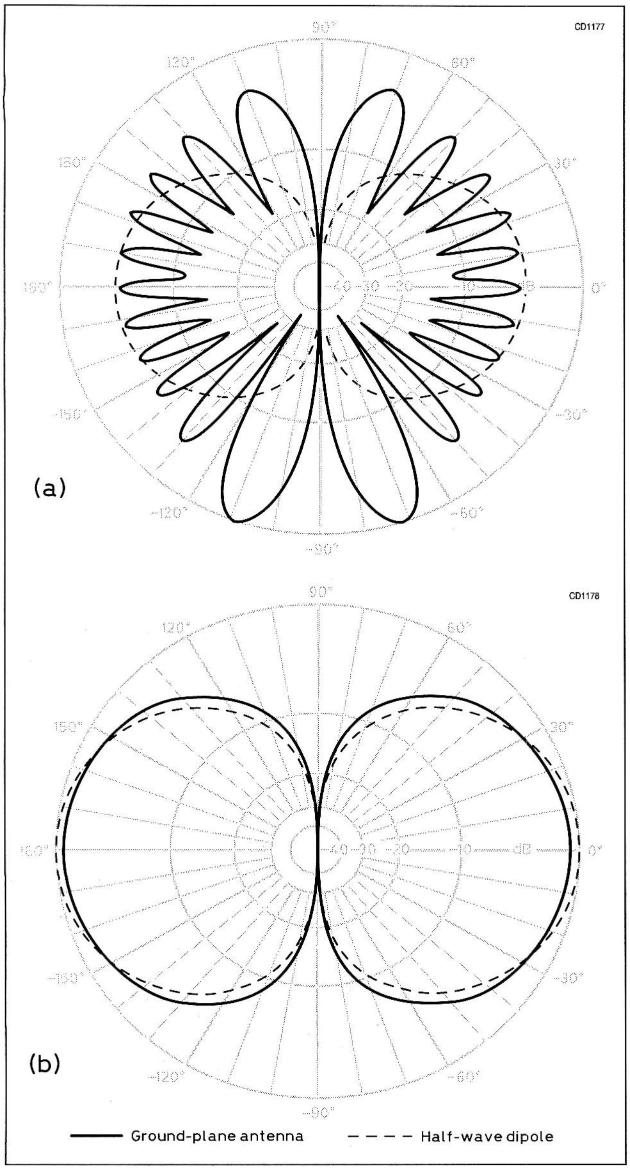


Fig 5.71. Radiation patterns of quarter-wave ground-plane antenna. (a) On top of a 5λ (10m) mast. Note that the main lobes are directed downwards, and would be prone to pick up man-made interference. The pattern of a λ/2 dipole is shown for comparison. (b) In free space. A λ/2 dipole is again shown for comparison

present a high impedance to reduce currents flowing below that point. The mast currents can be reduced a little by using more radials in the ground plane or extending their length to around 0.3λ.

An open-circuited choke sleeve can be more effective than radial wires for mast current control. This technique is used in the skirted antenna described below.

The skirted antenna

The skirted antenna (Fig 5.72) does not require ground-plane radials, and can be mounted in a cylindrical radome for better appearance and lower wind-induced noise. The skirt forms the lower part of a λ/2 dipole and, being λ/4 long, presents a

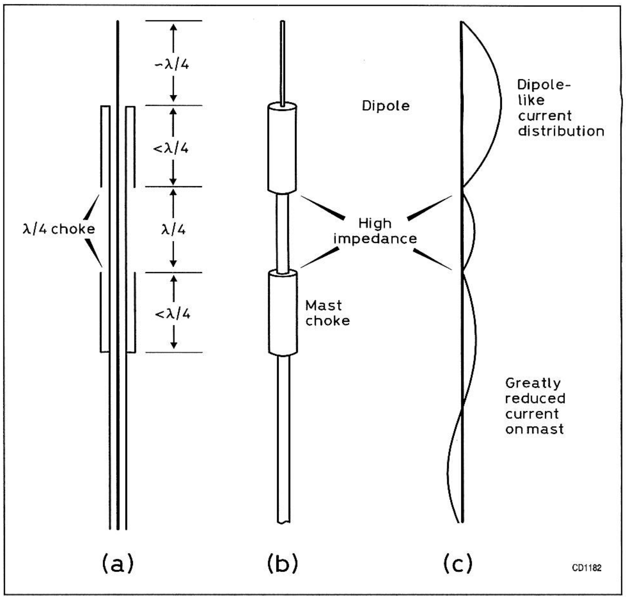


Fig 5.72. Skirted dipole antenna with mast choke. Note that the interior of the choke must be λ/4 in electrical length. Some designs use dielectric loading inside the dipole skirt to shorten its length, which can be used to adjust the impedance of the dipole as a radiator

high impedance at its lower end, reducing unwanted currents on the mast. The current is further reduced by a second choke, with its open, high-impedance end placed λ/4 below the dipole skirt for best effect. The radiation pattern of this antenna closely resembles that of a λ/2 dipole in free space.

The gain sleeve dipole (vertically polarised)

The gain sleeve dipole in Fig 5.73 is derived from the 1.8dBD shunt-feed 5λ/8 mentioned later in the mobile antenna section.

The radiating element B-B is in principle a centre-fed 1λ element but is fed coaxially to make it end fed. Having effectively twice the aperture of the λ/2 dipole, a gain of typically 2.5–3dB is achieved.

Mechanical construction is open to interpretation but a beer can or plastic water-pipe format are two solutions. It should be noted that the mounting point should be at A-A and not on the 0.25λ sleeve.

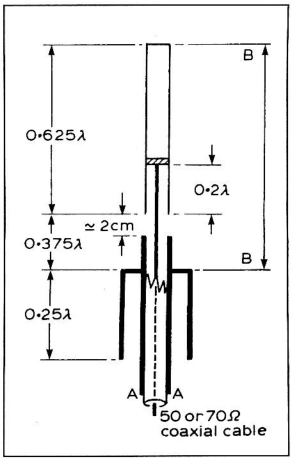


Fig 5.73. Gain sleeve dipole

The discone

This antenna has not found too much favour with amateurs in the past, though frequently used for commercial and military purposes. Unlike many other types this antenna is not only omnidirectional but also has wide-band characteristics. It is capable of covering, say, the 70, 144 and 432MHz bands or 144, 432

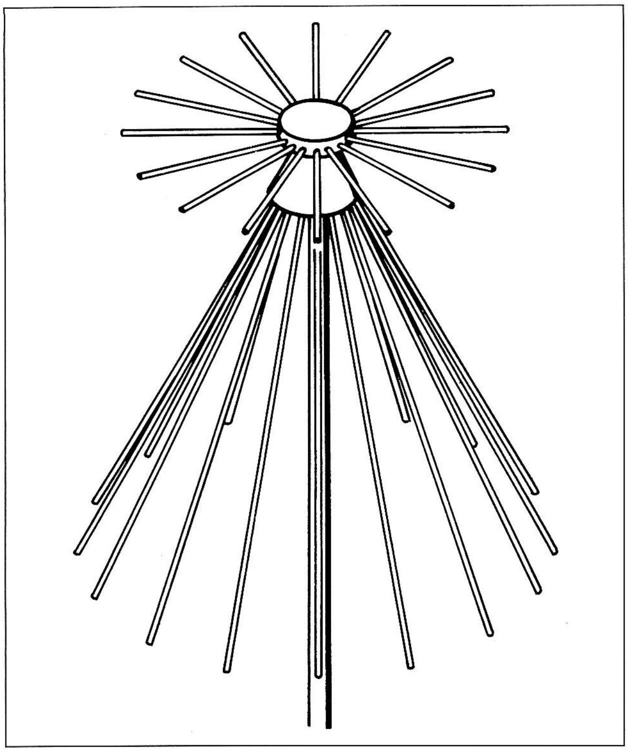


Fig 5.74. General arrangement of skeleton form of discone antenna

and 1296MHz, although there will of course be some variation of the SWR over such a wide range.

Also, since the antenna can operate over roughly a 10-to-one frequency range, it will more readily radiate any harmonics present in the transmitter output. It is therefore important to use a suitable filter to adequately attenuate the harmonic outputs. The radiation angle tends to rise after the first frequency octave.

The discone consists of a disc mounted above a cone, and ideally should be constructed from sheet material. Many amateurs would find this impossible to realise, but with little loss the components may be made of rods or tubes as illustrated in Fig 5.74, with a minimum number of rods of eight or preferably 16. Of course, open mesh may be used as an alternative, bearing in mind the windage increase, and that the current flows radially away from the feedpoint.

The important dimensions are the end diameter of the cone and the spacing of this from the centre of the disc, so that the terminating impedance is correct, eg 50Ω.

The primary parameters are shown in Fig 5.75 with dimensions as follows:

A the length of the cone elements – these are $\lambda/4$ at the lowest operating frequency, or $2952/f$(MHz)in.

B the overall disc diameter – this should be 70% of $\lambda/4$.

C the diameter of the top of the cone – this will be decided to some extent by the diameter of the coaxial cable but for most purposes 0.5in will be suitable.

D the spacing of the centre of the top disc to the cone top – this is 20% of C, or 0.1in for 50Ω.

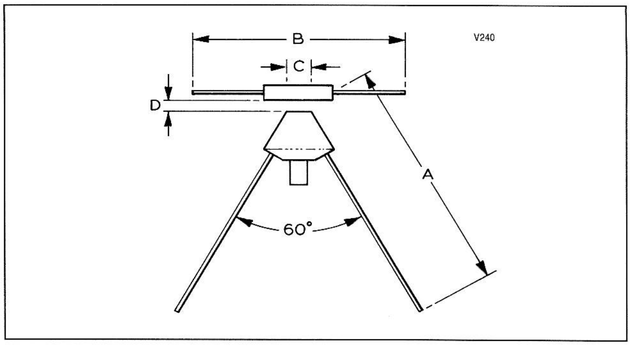


Fig 5.75. Primary dimensions of discone antenna

The detail given in Fig 5.76 of the hub construction will be suitable for any design using a 50Ω cable feed and may be taken as an example. There is likely to be some problem in producing a suitable insulator which may be made of a potting resin or turned from PTFE or other stable low-loss material.

An extension of the discone is the *helicone*. The elements of the conventional discone can be replaced with helical elements working in the normal mode as discussed later.

In its simplest form only eight elements are required for the disc and for the cone. Gain and the radiation pattern is essentially the same for both the discone and helicone but for the helicone the usable bandwidth is reduced to approximately one-third.

Collinear antennas

Communication with mobile stations is best achieved with a vertically polarised omnidirectional antenna, as there is no

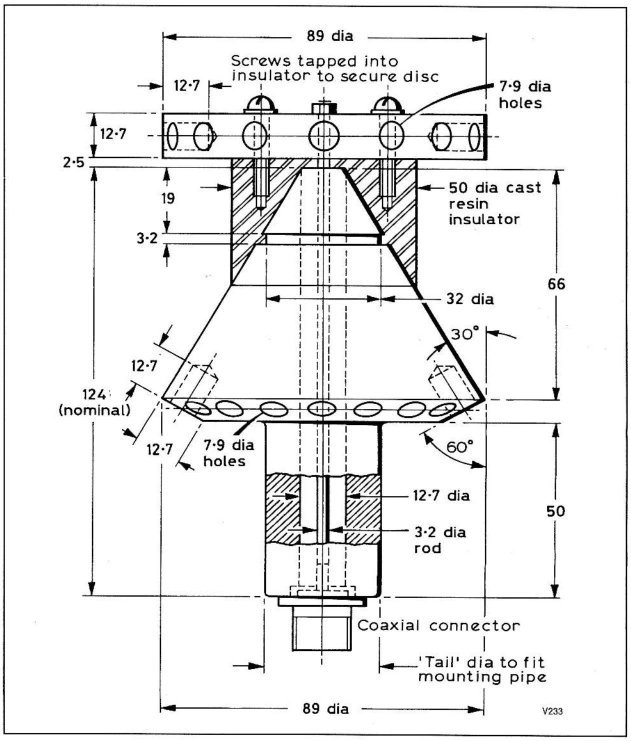


Fig 5.76. Details of a hub assembly

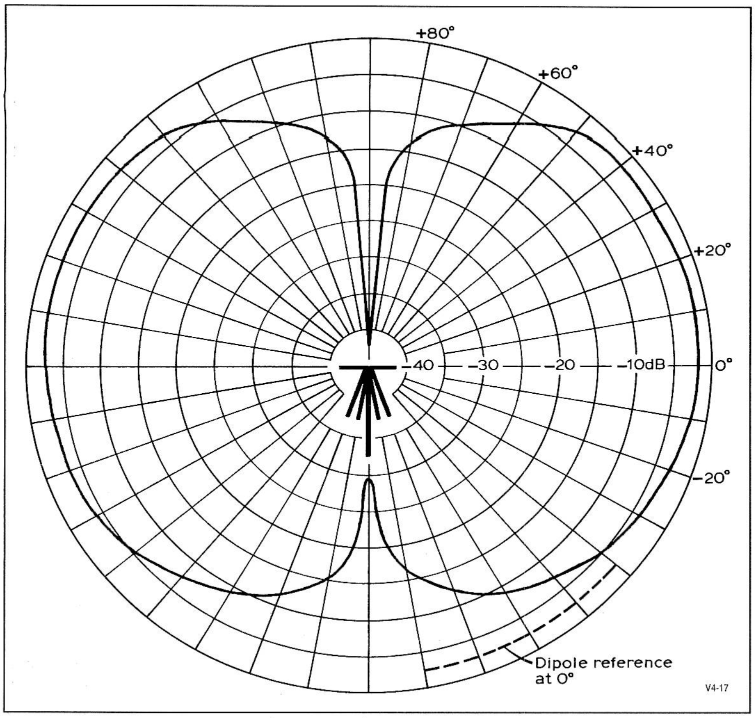


Fig 5.77. Typical discone and helicone decibel radiation pattern over the first 2:1 frequency range. As the frequency increases above 2:1 the pattern tends to rise above the horizontal level until at about 5:1 in frequency the main direction of radiation is above 45° from the horizontal

need to point the antenna in the direction of the mobile. However, a fixed station is not as constrained by mechanical considerations as a mobile, and can thus be fitted with larger, and hence higher-gain, antennas.

This can be achieved by stacking dipoles vertically above one another in a collinear array, and feeding them with cables of equal lengths, as shown for the GB2ER repeater antenna later in this chapter. Another method of achieving gain with simpler feed arrangements is discussed below.

A formula was presented in the section on antenna arrays for calculating the radiation pattern of groups of elements. The radiation pattern of broadside arrays develops nulls as the elements are spaced further apart, so it is desirable to place the elements as close together as practicable, taking into account the capture area and mutual coupling effects also discussed in the section on arrays.

The current on a length of wire several wavelengths long will be distributed as shown in Fig 5.78(a). The wire shown is 2λ long. Radiation at right-angles to the wire will be poor, as the successive half-wavelength current maxima are in opposing phase, and if the currents were equal, there would be perfect cancellation of the radiation in that direction. However, if all the current maxima were in phase, the radiated fields would add, and a high gain could be achieved: Fig 5.78(b).

There are several ways of achieving this phase reversal. The simplest is to insert an anti-resonant network or a non-radiating half-wavelength of transmission line as a phasing section between the λ/2 radiating elements: Fig 5.78(c). The λ/2 transmission line can be realised as a quarter-wavelength of ribbon cable, which can be wound around the insulator between the radiating elements (see the section on mobile antennas).

A more subtle approach uses radiating elements that are a little longer or shorter than λ/2. This helps the feeding arrangements, as it will be remembered that end-feeding a λ/2 dipole is difficult because of its very high impedance. The self-reactance of the longer or shorter dipole is then used in the design of the phasing network between the elements to achieve the desired overall phase shift. The non-radiating transmission line can then often be replaced by a capacitor or an inductor in series with the residual element reactance: Fig 5.79(a) and (b). Again, a transmission line stub can be used to synthesise the required reactance, which may be more convenient or cheaper than a lumped component, especially if significant RF power handling is required: Fig 5.79(c). Sometimes a parallel-tuned

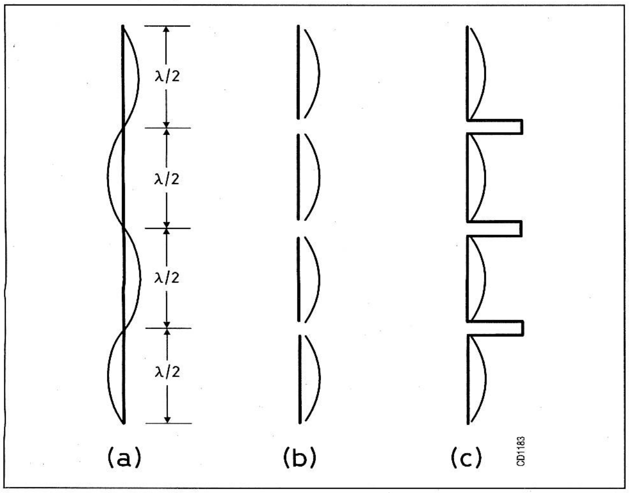


Fig 5.78. Current distributions on a wire and derivation of the collinear antenna

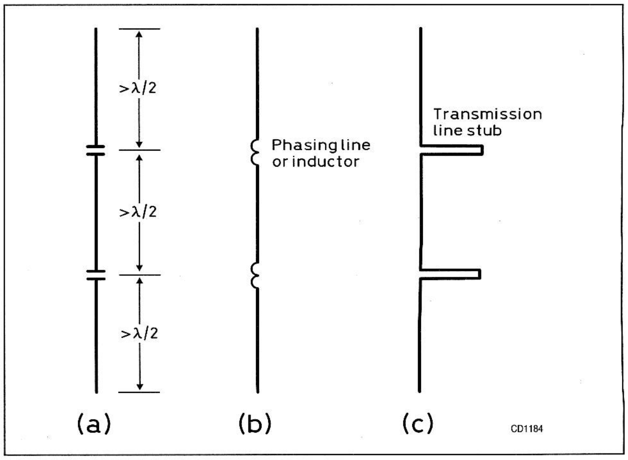


Fig 5.79. Realisation of collinear antennas. The antennas are end-fed

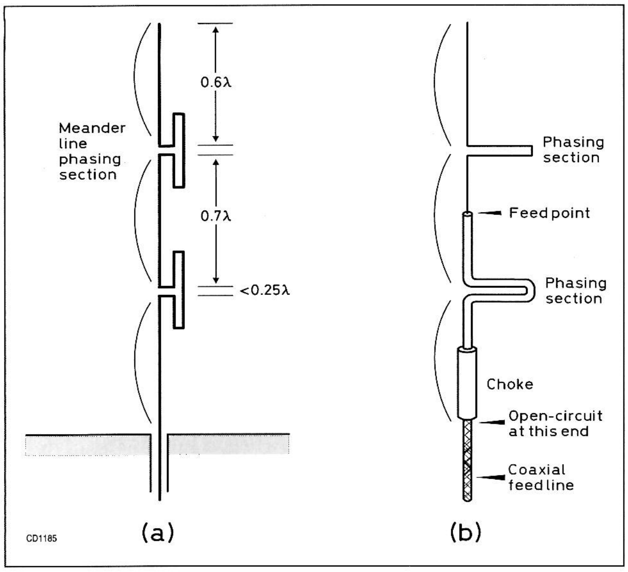


Fig 5.80. Frankin collinear antennas, end-fed and centre-fed

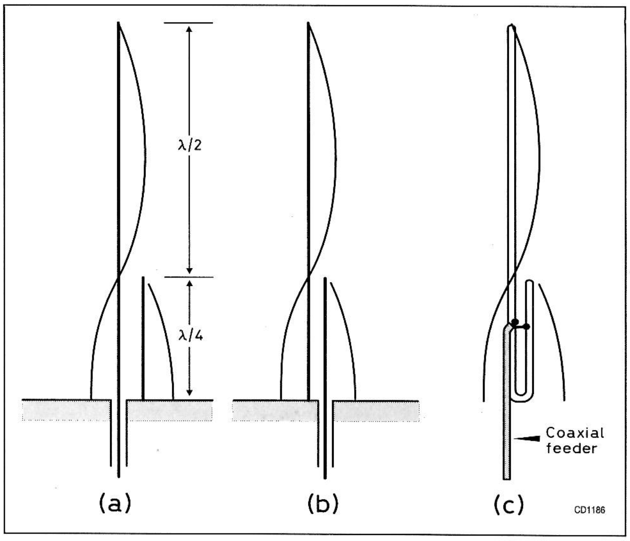


Fig 5.81. The J antenna

circuit is realised as an inductor resonated by the self-capacitance of the insulator separating the radiating elements, and on which it is wound.

A technique devised by Franklin that has been attractive to VHF antenna manufacturers folds parts of the radiating element to provide the phasing section as shown in Fig 5.80(a). Provided that the folded sections are significantly shorter than the radiating elements, the gain is not significantly degraded, although the whole structure is sensitive to capacitive loading by any housing and insulators required. The radiation pattern is frequency sensitive, and the main lobe will squint upwards or downwards as the frequency changes from the nominal. While these folded element designs look attractive for home construction, adjustments to optimise both the radiation pattern and input impedance are very difficult without proper measuring facilities. Poor gain and broken radiation patterns result if the sections are not properly excited and phased.

All these designs are end-fed, which have practical disadvantages with longer, multi-element arrays. If identical sections are used, the end elements carry less current than those close to the feed, reducing the overall efficiency of the antenna. While different length radiators and phasing elements can be used to equalise the current distribution, the design and adjustment is lengthy and definitely requires good radiation pattern measurement facilities. If the array can be centre-fed, any residual phasing errors tend to cancel out and, for a given length, the performance tends to be better because of a more uniform current distribution. Fig 5.80(b) shows one means of achieving centre feeding with a Franklin array. Note the use of the λ/4 choke section at the base of the array, which is essential to prevent current flowing down the outer of the coaxial cable and destroying the performance of the collinear antenna. The practical gain limit of the singly fed collinear antenna is around 10dBi.

Practical collinears in radio amateur use tend to use variations on Fig 5.79. The radiating elements may comprise combinations of lengths up to 5λ/8, with or without ground planes. The presence of a good ground plane increases the gain as the image or reflection effectively doubles the length of the array (see also the section on mobile antennas). However, good results can be achieved with collinears directly mounted on pipe masts, especially if care is taken to control unwanted currents from flowing on the mast.

None of the above considers the practicalities of feeding and matching the antenna. Collinear antennas, by virtue of their operation as end-fed structures, have high feed-point impedances. A good feed arrangement, valid for both ground-plane and mast mounted antennas, is the use of a λ/4 short-circuited transmission line, as described in the section on matching in this chapter.

Fig 5.81(a) shows such an arrangement to end-feed a λ/2 dipole mounted over a ground plane. The matching section should not radiate, and the overall effect is that of a λ/2 radiator raised λ/4 above the ground plane. Either leg of the λ/4 section can be fed, leading to the structure in Fig 5.81(b), which is identical to Fig 5.81(a) in terms of current distribution, and hence radiation performance. The evolution can be taken a stage further by removing the ground plane and feeding either leg of the λ/4 section as in Fig 5.81(c); this is the *J antenna* or *J-pole antenna*, which may use different diameters of tubing for the radiator and stub.

The *Slim Jim antenna* provides an elegant solution for a simple, mechanically robust antenna made from a single piece of tubing as shown in Fig 5.82. This antenna [24] comprises a folded, open-circuit λ/2 radiator above a λ/4 transformer section, and is a derivative of the J antenna. The folded-stub characteristics of the radiator provide some control over the reactive element of the input impedance. The two ends of the tube can be joined by an insulator, eg a piece of stiff plastic tubing, to provide weather proofing and enhanced mechanical rigidity. Either balanced or unbalanced feeds can be used, tapped on to the λ/4 transformer section at the point that provides the best match to the feeder. Coaxial feeders should be strapped or bonded to the λ/4 section to reduce unwanted currents on the outer of the cable. The antenna has a maximum gain of around 2.8dBi in free space, although the main

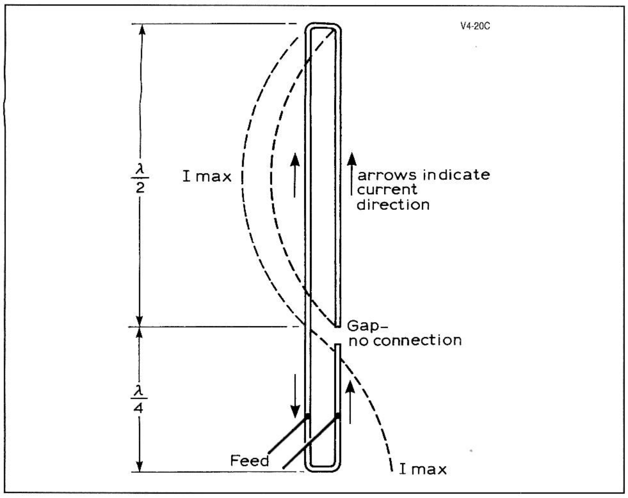


Fig 5.82. The basic Slim Jim, showing direction of current flow and phase reversal in matching stub (*Practical Wireless*)

lobe is tilted up about 10°. The main lobe can be brought to the horizontal by reducing the length of the upper section to about 0.4λ. This reduces the peak gain to around 2.5dBi, and can make the feed impedance capacitive.

Phasing sections and additional elements can be combined to produce a collinear form for the J antenna as shown in Fig 5.83(b). This antenna and that of Fig 5.83(c) have been used

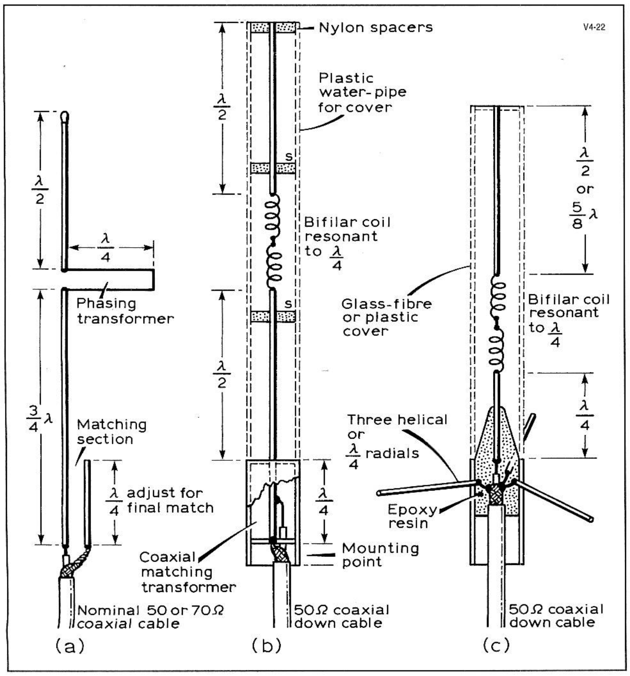


Fig 5.83. A collinear form of the J antenna. (a) The addition of λ/4 sections as suggested by Franklin. (b) Use of a coaxial short-circuit λ/4 transformer to give an unbalanced input. The tapping point in the matching transformer is approximately 0.15λ from the 'earthy' end. (c) A variant of (b) with radials. With both (b) and (c) the λ/4 phasing transformer has been 'wound up' as a bifilar coil (each coil being wound in the opposite hand). While the inductive component is cancelled, the mutual capacitance on the windings makes them physically shorter than λ/4

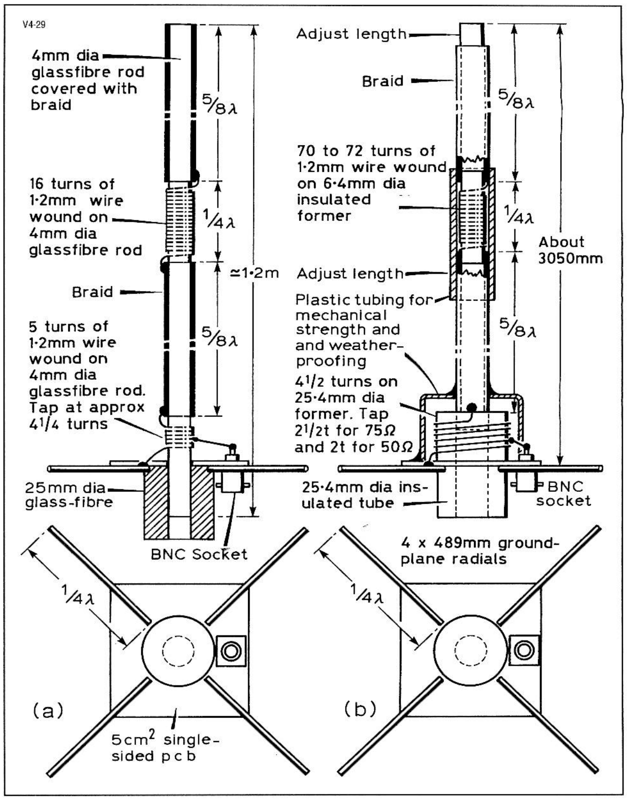


Fig 5.84. (a) A 432MHz collinear. (b) A 144MHz collinear (*UK Southern FM Journal*)

successfully to produce low-angle radiation for the GB3SN 144MHz repeater.

A variation of the techniques described but using coils as with the original Marconi concept is shown in Fig 5.84(a) for 432MHz and Fig 5.84(b) for 144MHz. The expected gain is between 6 and 7dBD.

Materials required for Fig 5.84(a) are as follows:

- One 2.5cm dia 10cm long glassfibre tube
- One 40mm dia 1.2m long glassfibre rod
- Four 20mm dia 20cm long glassfibre rods
- Length of braiding from junk multicore cable
- Length of 18 SWG wire for matching coils
- Approx 5cm square of singled-sided PCB

First, adjust the bottom 5λ/8 element to give minimum SWR; this is done by adjusting the tapping point on the bottom coil (approx 4¼ turns). A fine adjustment can be made by altering the length of the first 5λ/8 element.

Next fix on the centre matching coil and the top element. Please note that to obtain the best results both elements should be approximately 5λ/8 and within reason the same length. A good SWR is obtained by adjusting the centre matching coil (the coil is spread over λ/4).

The matching coil provides the phase change necessary to feed the top element and so adjustment is quite critical. It has been found that if the matching coil has to be 'squeezed up' to obtain a good SWR, then the coil has too many turns. The opposite is true if the coil has to be greater than λ/2 for a good SWR.

To prevent the collinear going off tune once set up, the elements were secured to the centre glassfibre rod and the matching coil taped with self-amalgamating tape. Provided care is taken in setting up, an SWR of close to 1:1 to 1 can be obtained.

Materials required for Fig 5.84(b) are as follows:

- Two ½in dia by 47½in ± ½in, 5λ/8 elements (adjustable)
- Four 19¼in rods for ground plane
- One 1in dia by 30in insulated rod
- One 1in dia insulated tube (a cotton reel can be used instead)
- 18 SWG wire for matching and phasing coils.

The diagram shows extra insulated tubing over the matching and phasing coils to give more mechanical strength and weather-proofing.

Setting up is carried out as follows. First, the length of the bottom 5λ/8 element must be adjusted to give the minimum SWR possible.

Next fix on the phasing coil and the top element which must be the same length as the bottom element. Then obtain the best SWR possible by adjusting the phasing coil.

This coil provides the phase change necessary to feed the top element; it is a length of 18 SWG wire, about 1λ long, coiled up to give 70–72 turns on a 4in former. It was found that the λ/4 spacing between the two elements is more critical than the number of turns. 68 turns gave satisfactory SWR on one version.

Some difficulty may occur in setting up the phasing coil. Before taking too many turns off, go back to the first stage to ensure that the bottom 5λ/8 element is correctly matched. If the bottom element is not correctly matched the collinear will not tune up. Careful adjustments in setting up should produce a SWR of 1:1 to 1.

A technique that has not been discussed but is widely used involves feeding conventional λ/2 dipoles in phase from a single source or adjusting the phase relationship of cable lengths between dipoles. There is a degree of interaction between cables and radiating elements but individual dipoles can be positioned to modify the pattern shape. The example given in Figs 5.85–5.87 is probably the simplest to implement and was devised for the GB3ER 432MHz band repeater.

ANTENNAS FOR MOBILE AND PORTABLE STATIONS

The choice of an antenna for mobile VHF and UHF use is dependent on several factors. As the frequency increases the aperture of the antenna decreases. This means that larger gains are required for UHF than VHF to overcome the loss of aperture as well as the radiation path loss due again to the increase of working frequency.

As the direction of a vehicle, relative to the station to which it is transmitting or receiving, is continually changing there is a need for an omnidirectional antenna system. This will mean that to achieve gain in the horizontal plane, while retaining an omnidirectional pattern, will require considerable reduction of the pattern in the vertical plane. For example an omnidirectional antenna of 6dBD gain will have a typical half-power point (–3dB) of under 30°. The narrow beam or disc that is produced will result in considerable variation in transmitted and received signal strength as the vehicle or antenna tilts or

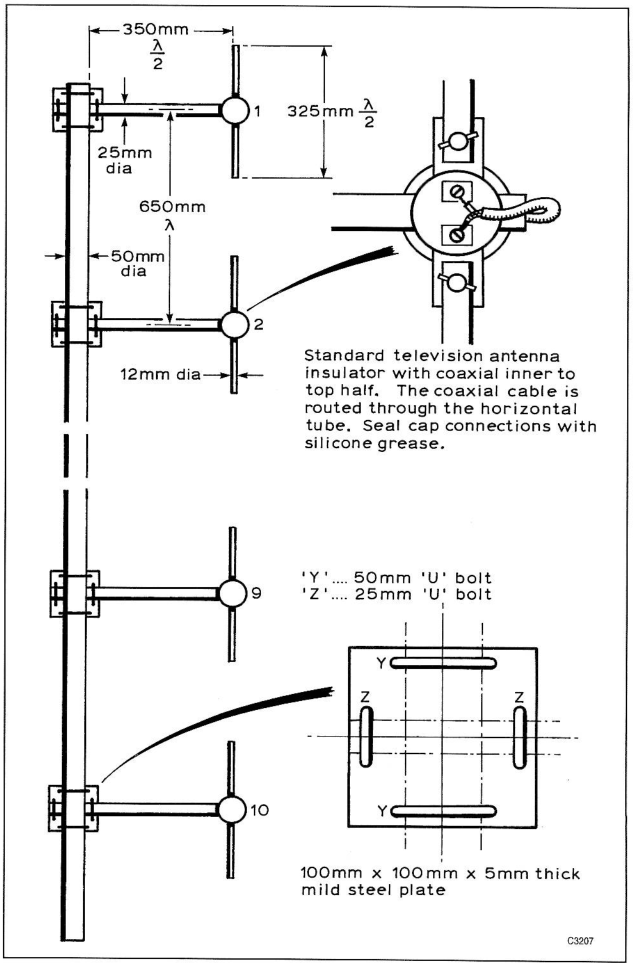


Fig 5.85. Mechanical details of GB3ER collinear

where signals are reflected, as will always be the case, from nearby objects. A compromise has therefore to be arrived at to obtain maximum gain in the best direction which gives minimum disruption of signals when mobile.

The choice of polarisation is not only dependent on compatibility with stations being received and the optimum

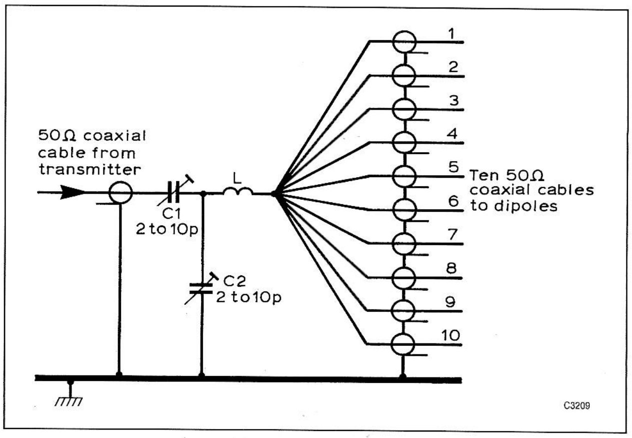


Fig 5.86. Matching unit of GB3ER collinear

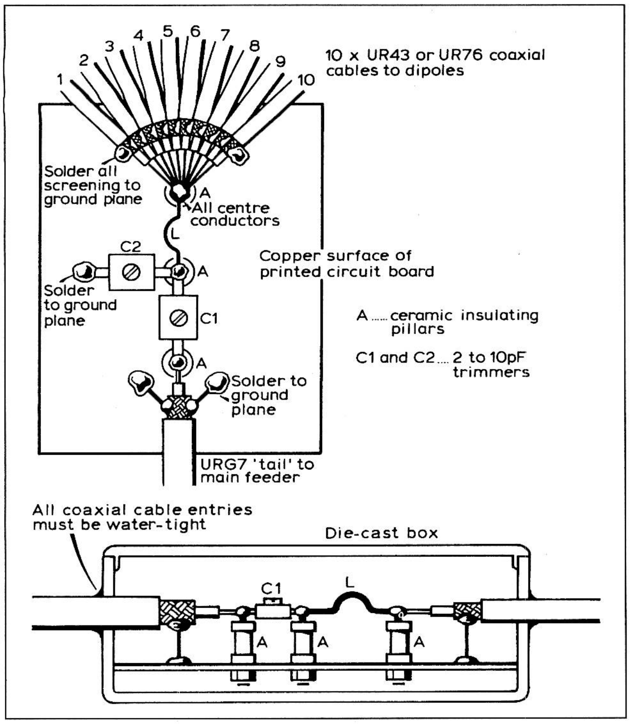


Fig 5.87. Matching unit layout of GB3ER collinear

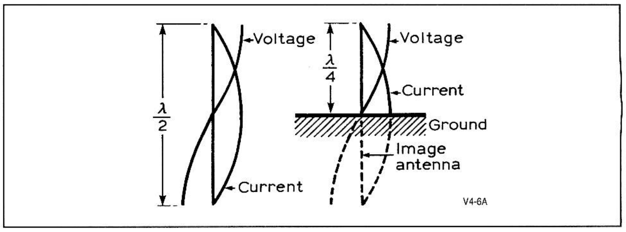


Fig 5.88. The λ/2 antenna and its grounded λ/4 counterpart. The missing λ/4 can be considered to be supplied by the image in ground of good conductivity

polarisation for the propagation path concerned, but the aesthetics and mechanical complexity of the antenna used and its mounting position on the vehicle.

Antennas, particularly when vehicle-mounted, must always be considered as an integral part of the environment in which they are to be used. Radiation patterns quoted by manufacturers can be completely different when an antenna is in use. Increased gain normally means an increase in physical size. This improvement of gain can be lost, with a probable loss of omnidirectivity, if, due to its physical size, the antenna is mounted at a lower point to facilitate access to a garage, for instance. The difference in mounting an antenna on the wing or boot of a car compared with mounting it on the top dead centre of the car roof can lose at least 3dB of gain with the variations of the expected radiation patterns.

There are several antennas in current use which are worth considering. In addition one or two specialised antennas are available or can be readily fabricated by the radio amateur which also merit consideration. Mobile antennas can be considered in three basic groups:

1. Vertically polarised antennas, more often used for FM and repeaters.
2. Horizontally polarised antennas, normally used for SSB transmission.
3. Circularly polarised antennas.

together with a sub-group of low-profile antennas to produce vertical or horizontal polarisation but with physical heights below 0.1λ.

Quarter-wave whip

This is the simplest and most basic mobile antenna. It is derived from the doublet or λ/2 dipole. Marconi, by replacing half of the doublet with a ground plane as shown in Fig 5.88, found that the image of the vertical λ/4 section was 'reflected' in the ground plane, producing an antenna which was substantially the same as the original dipole. The theory of operation showed that if the ground plane was infinitely large and made of a perfectly conducting material, all of the radiation associated with the lower half of the dipole was radiated by the top half giving, in fact, a 3dB improvement over the dipole. In practice the size of the ground plane and its resistive losses modify the pattern and this 3dB is never realised. Figs 5.89 and 5.90 show optimum patterns of a λ/4 whip measured on a ground plane of λ/2 sides and 1λ sides. Although the pattern is raised from the horizontal, on a medium ground plane the loss of horizontal gain is relatively small (20° and 1dB at 0° in Fig 5.89 but 40° and 6dB at 0° in Fig 5.79).

However, as the ground-plane size increases the main lobe continues to rise until the situation of Fig 5.91 occurs. When a radiator is mounted over a ground plane as described, the input impedance is typically halved. So for the λ/4 whip or monopole the input impedance is typically 36Ω + j, that is to say, approximately half the resistance of the dipole but with an additional reactive component.

Considering 50Ω as being the standard cable impedance used at VHF and UHF, this would produce a standing wave at the antenna base of about 1.5 to 1. The simplest way to

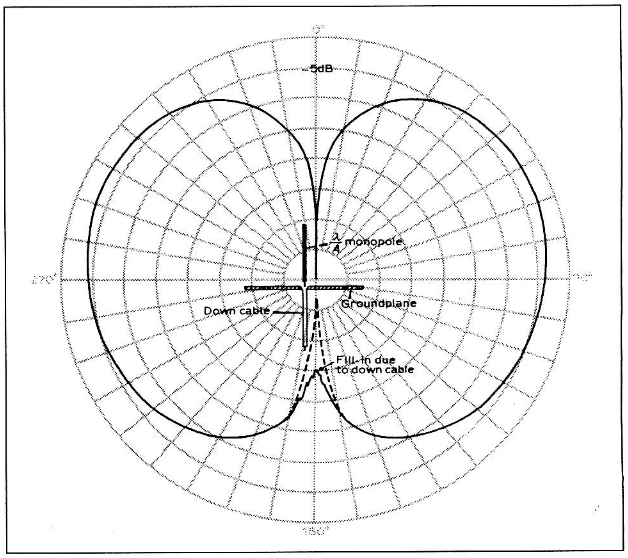


Fig 5.89. Radiation pattern of a λ/4 monopole over a λ/2 square ground plane at 145MHz

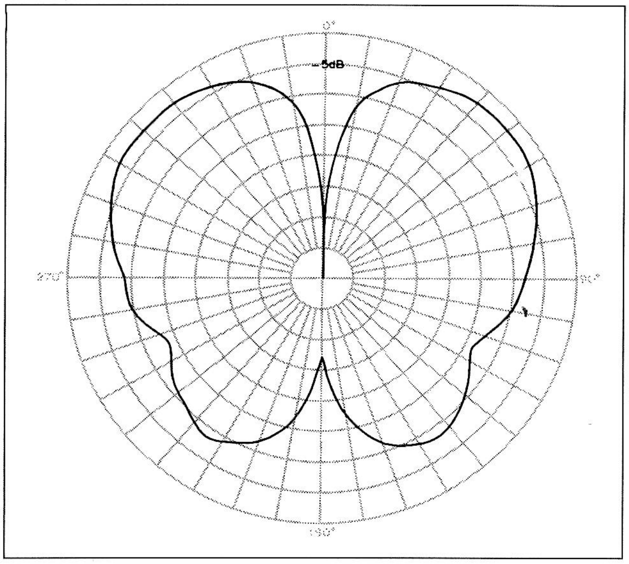


Fig 5.90. Radiation pattern of a λ/4 monopole over a 1λ square ground plane at 145MHz

overcome this mismatch is first to slightly increase the length of the whip to produce an inductive reactance to cancel the capacitive reactance normally obtained. In practice this also raises the resistive value of the whip and a close match can usually be obtained to 50Ω cable. Should a VSWR bridge or similar (of 50Ω characteristic impedance) be used to set up the whip, when a match has been achieved, the length of the cable should be changed and the match re-checked. If there is no change in the meter reading, then the antenna is matched to the cable. If a change does occur then the antenna/cable combination has been matched to the VSWR meter and the whip should be readjusted until changes in cable length have minimal or no effect. It is preferable that the added cable length is not an exact multiple of a λ/2 or λ/4 as this, particularly with a multiple of λ/2, will confuse the results.

The ground-plane effects and aperture size of the λ/4 whip tend to limit its use at VHF and UHF. At UHF the aperture is small and the pattern tends to be raised in the vertical plane due to the large ground-plane area. It is therefore not often used at those frequencies. At VHF, ie 144MHz, the compromise of the λ/4 whip's simplicity and size (about 49cm or 19¼in) often balances with its medium aperture and tendency on some vehicles to have a raised vertical pattern. At 70MHz the physical dimensions are such (about 102cm/40in) that this is the normal limiting factor to the use of the λ/4 whip as opposed to gain devices. The aperture of the antenna at this frequency is compatible with path-loss conditions, and the ground-plane size is such that the radiation angle when roof mounted is fairly low. However, the shape of the radiation pattern can have a loss of 3dB in the omnidirectivity each side of the vehicle.

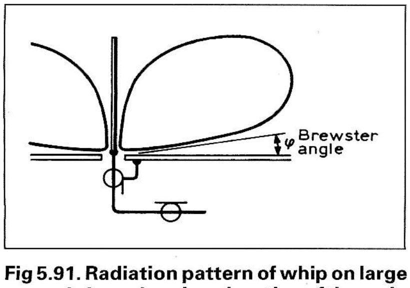


Fig 5.91. Radiation pattern of whip on large ground plane showing elevation of the main lobe

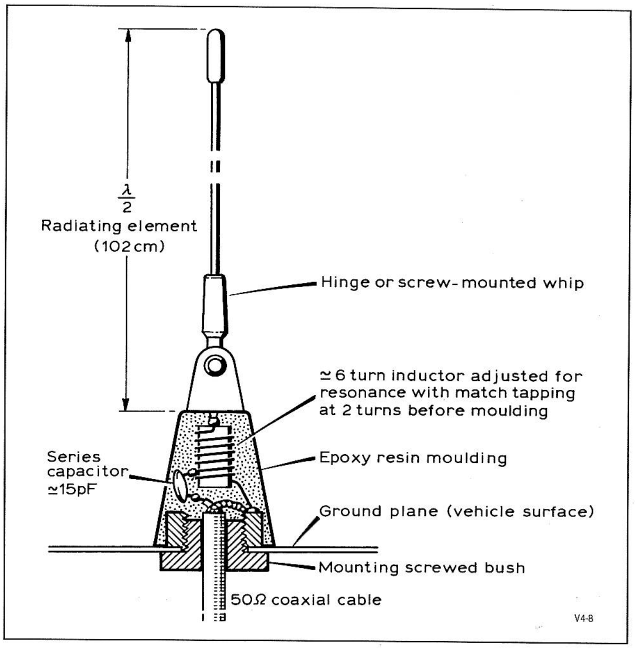


Fig 5.92. A typical home-built λ/2 mobile antenna and mount

The λ/2 and 5λ/8 gain antennas

Using the ground-planing techniques described for the λ/4 whip, gain antennas can be produced. If the λ/2 dipole is extended in length, maximum forward gain (before the pattern divides into several lobes) is obtained when the dipole is about 1.2λ. This becomes the maximum length of 5λ/8 for a ground-plane antenna. A natural extension to the λ/4 whip is the λ/2 whip. However, such a radiator fed against a ground plane has a high input impedance. On the other hand, a 3λ/4 radiator fed against a ground plane has a resistive input of almost exactly 50Ω but is above the optimum length for a reasonable pattern shape.

If the λ/2 whip could be made to look like a 3λ/4 radiator then it would be possible to obtain a 50Ω resistive input. A series coil at the ground-plane end of a λ/2 radiator can be used to resonate it to 3λ/4, but the input is still fairly high impedance and reactive. If, however, the coil is shorted to the ground plane a tapping point up the coil will provide the required impedance and the addition of a non-critical capacitor in series will compensate for the reactive components. Fig 5.92 shows details of such an antenna.

As the aperture of the antenna has been doubled compared with the λ/4 whip, twice the effective radiation is obtained, ie approaching 3dB gain. This assumes however, that there is minimum resistance in the radiating element, ie it must be copper-plated or similar.

The maximum radiator size of 5λ/8 for a single-lobe pattern can also make use of the impedance characteristics of the 3λ/4 radiator. Construction is in fact simpler than the λ/2 antenna. If the radiating element is made 5λ/8 with a series coil equivalent to λ/8 at the ground plane end, an input impedance very close to 50Ω can be obtained. With correct materials a gain close to 4dBD can be achieved by the increase in aperture

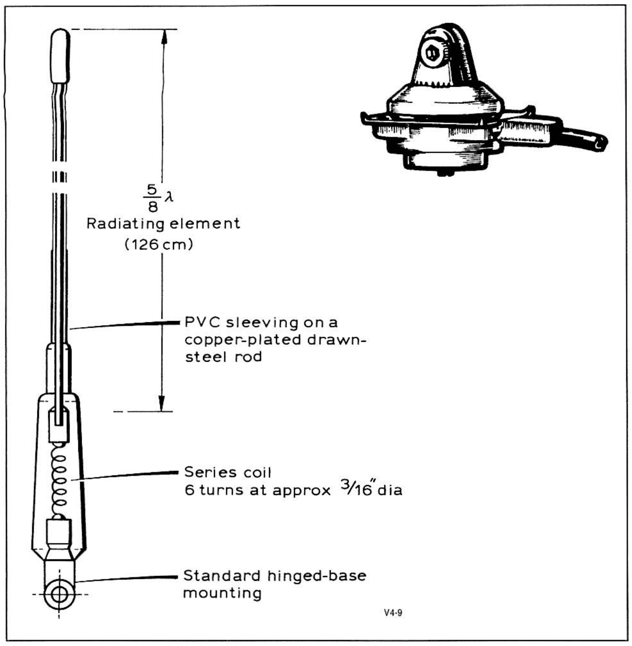


Fig 5.93. A typical commercial 5λ/8 mobile antenna and mount (top right)

over the λ/2 antenna. The radiation pattern is often raised more than that of a λ/2 antenna so the slightly improved gain of the 5λ/8 may not always be realised.

Fig 5.93 gives details of the series 5λ/8 whip. One other advantage of this antenna is that over a wide range of mounting and ground-plane conditions it will self-compensate for impedance and resonance changes. It is preferable for both the λ/2 and 5λ/8 antennas to be on a hinge mount, particularly if roofmounted, to enable folding or 'knock' down with obstructions like trees and garages.

Various gain figures have been given for the 'five-eighth-wave'. Unfortunately not all antennas use optimum materials. As previously stated, the DC resistive losses of the radiator must be a minimum, and in addition the use of a glassfibre rod changes the resonant length because the dielectric material changes the velocity factor by as much as 20%. This means the radiator has to be cut shorter than 5λ/8 with the accompanying loss of aperture.

Incidentally, the series coil with the true 5λ/8 whip must be held rigidly as movement of the coil turns will change the antenna's resonance, giving apparent flutter. With certain transceivers with VSWR-activated transmitter close-down this can produce a situation where the power output of the transmitter is continually being turned down or switched off, producing extremely severe 'flutter'.

Apart from the above reasons for different gain figures, several ground-plane antennas discussed in articles about the 5λ/8 system are in fact discussing antennas which are not truly of this nature. One of these devices worth considering for its own merits is that shown in Fig 5.94. It consists of a 5λ/8 vertical element with a reactive sleeve of 0.2λ at the ground-plane end of the vertical as shown. The gain obtained from this antenna is typically 1.8dBD and, as can be seen, the actual radiating element A-A and therefore its aperture is under that of a λ/2 antenna.

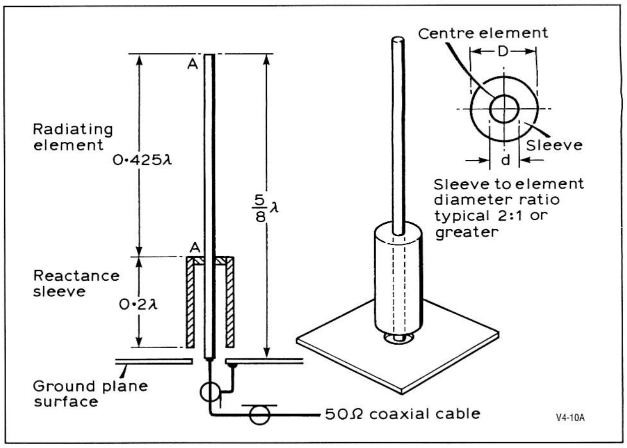


Fig 5.94. The reactance-fed 5λ/8 monopole. Typical gain is +1.8dBD (*Ham Radio*)

Other antennas with similar properties but different in construction are the 'J' and 'Slim Jim'. These were described earlier in this chapter.

7/8λ whip antenna

This is a variant of the 5/8λ antenna and its current distribution is shown in Fig 5.95. This antenna uses the principles of the collinear antennas shown earlier in this chapter, using either a

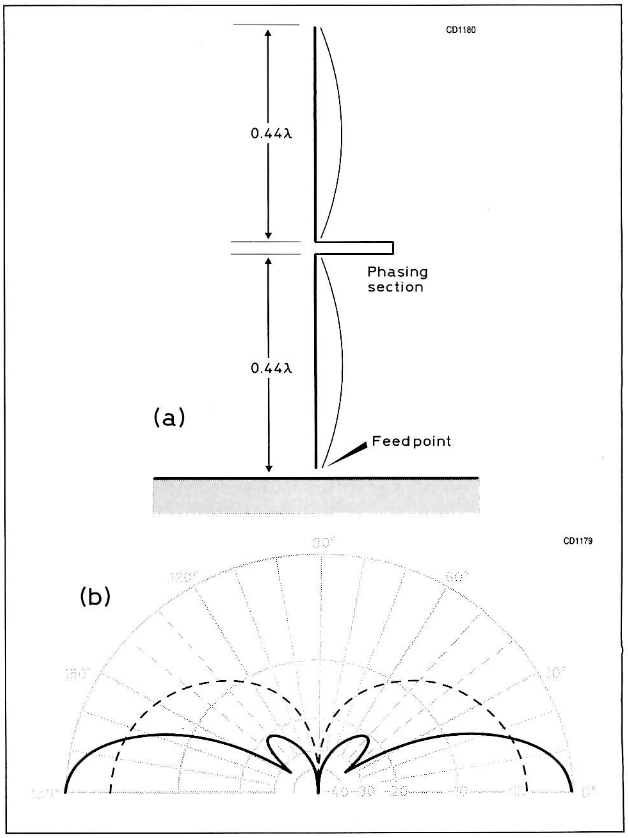


Fig 5.95. 7/8λ whip antenna with theoretical current distribution and radiation pattern on infinite, perfectly conducting ground plane. λ/4 whip pattern shown for comparison

phasing section or a series capacitor to ensure that the radiation from the upper and lower elements is approximately in phase. A stub phasing section, if properly adjusted, can enhance the gain by approximately 0.5dB over the capacitor variant, and can be realised as a short-circuited piece of ribbon cable wrapped round the insulator separating the upper and lower elements. Phasing can also be achieved with a wound section along the lines discussed in the section on collinear antennas. Care should be taken in designing this insulator if the whip is for use on a vehicle, as it must withstand drag of the upper element at speed and the shock loads of the whip snapping from side to side.

For the dimensions shown, the stub length should be approximately 0.245λ. If a capacitor is used, it should present an impedance of around 500Ω at the working frequency (about 2.2pF at 145MHz). The base impedance will be high, between 200 and 400Ω, and capacitive. Base matching arrangements similar to those of the 0.5λ whip are suitable, although the series capacitor should not be necessary. Component values will depend on the materials and construction used, especially around the mounting base.

Table 5.12. λ/2 ring radiator dimensions

	Theoretical	VHF measurement antenna	UHF measurement antenna
Frequency	*f* MHz	145MHz*	433MHz
Diameter *D*	52°	298mm	100mm
Height *H*	8°	39mm	15.5mm
Diameter *d*	nom. 1–2°	15mm 20 SWG strip	10mm 20 SWG strip
Match *M*	5° for 50Ω	28.7mm	9.7mm
Tuning capacitor C	To give capacitive reactance, nominally 250–500Ω	2–5pF	0.5–2pF

* Tuneable 137–148MHz.

Low-profile antennas

An alternative to vertical ground-plane antennas are devices to reduce the physical size of the system. The reduction of physical size normally implies loss of aperture and therefore gain. However, of the antennas discussed in this section, one in fact produces a gain referred to a dipole of +1dB.

The λ/2 ring radiator

Although called a 'ring' radiator, in fact radiation is produced by the slot formed between the horizontal λ/2 ring and the ground plane.

Consider a λ/2 slot in a metal sheet. If the sheet is rolled into a cylinder such that the two ends of the slot come together, an omnidirectional vertically polarised radiator is produced. As with the conventional λ/2 slot an impedance match can be obtained by tapping along from one end. Also, if the slot is just under λ/2 a capacitor across the centre will resonate it to λ/2 again. As with the skeleton slot developed by G2HCG, if the ground-plane sheet at the top of the slot is reduced to produce a ring and the lower ground-plane section is bent into the horizontal plane the low profile of Fig 5.96 is produced. Dimensions in terms of electrical degrees and specific sizes for optimum performances for 144MHz and 432MHz are shown in Table 5.12. Halving the dimension *H* or the loop diameter *D* (with the necessary increase of capacitance and match point to re-tune to frequency) will halve the radiation capability.

The λ/2 ring radiator is a fairly high-*Q* antenna and has therefore a reduced bandwidth compared to a dipole (typically 3% compared to 10% for a monopole). Gain is 1dBD. If the ground plane is completely reduced, as was the top section previously described, a double ring radiator is produced. Both ring radiators lend themselves to discreet fixed antennas.

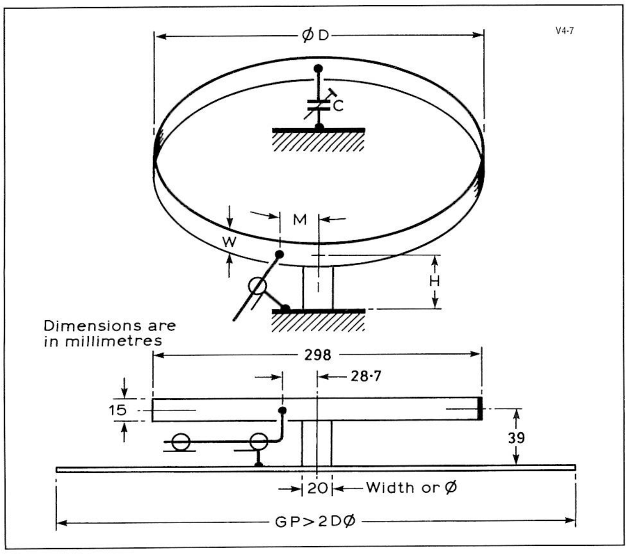


Fig 5.96. A low-profile vehicular antenna with vertical polarisation. Gain is 1dBD, termination 50Ω

Normal-mode helix

Typically less than 0.1λ high, this is described in the next section.

The normal-mode helix antenna

Much has been said for and against what is termed the *normal-mode helix* as used on hand-held transceivers. Unfortunately the method of operation and the results obtainable for this type of antenna have been much misunderstood by amateurs and professionals alike. Most theoretical papers only consider the helical equivalent of the λ/4 whip while most users of this antenna are in fact using the equivalent of a physically reduced 3λ/4 whip.

A helix will work in the normal mode when the diameter and pitch of the helix is less than 0.1λ. When working in this mode the radiation is from the side of the helix, and when the diameter is considerably less than 0.1λ the resultant 'spring' has a radiation pattern similar to a short vertical monopole or whip.

A 3λ/4 whip over a moderate ground plane has a resistive match very close to 50Ω. If this whip is coiled into a helical spring as previously described it will resonate to approximately 50Ω but at a somewhat lower frequency. If the spring

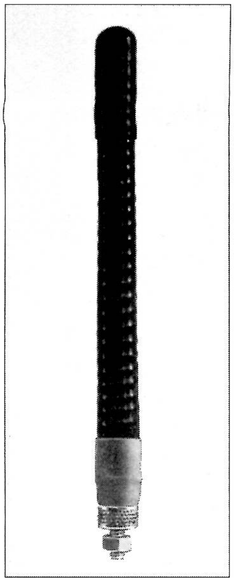


Fig 5.97. A typical commercial helical antenna with screw mounting facility

is trimmed to the original frequency the result will be an antenna of about 0.1λ long matching to approximately 50Ω. The actual wire length tends to be around λ/2 to 5λ/8 long at the working frequency. The capacitance formed between the turns of the spring has 'loaded' the antenna such that it still resonates as a 3λ/4 antenna. This capacitance also tends to modify the matching under various conditions.

Because of its construction, the spring is very reactive off-resonance and this makes it very important that it be resonated for the specific conditions that prevail in its working environment.

Fortunately it is only necessary to change the number of turns to resonate the spring over such diverse conditions as a large ground plane and no ground plane at all. However, the match referred to 50Ω can vary between about 30 and 150Ω at the extremities. Under typical hand-held conditions, however, and depending on the frequency of operation, the spring tends to be fairly close to a 50Ω impedance match. This is shown in Fig 5.98 which also gives an indication of the number of turns required for a typical 9mm diameter helix for 3λ/4 resonance.

An important consideration is that since the helix is a reduced size and aperture antenna two factors arise. First, the radiation resistance is lower than the equivalent linear whip so the choice of a good conducting material is important to minimise resistive losses. A steel spring compared with a brass or copper-plated helical can waste 3dB of power in heating up the spring. The aperture of the helical is a third the physical size of the λ/4 whip and would moreover indicate a loss of 4.77dB. However, results obtainable with copper-plated, Neoprene-sheathed helical antennas, correctly matched to a hand-held transmitter at 145MHz, are at worst –3dB and at best are +1dB compared to the equivalent λ/4 whip (which is –6dB compared to a λ/2 dipole). One thing that will be seen however is that the top of the spring on a hand-held transceiver will often need to be raised to a position corresponding to the top of the equivalent λ/4 whip to receive or transmit the maximum signal strength.

A similar device resonated on to a λ/2 square ground plane could give results 2–3dB below a λ/2 dipole. An alternative arrangement using a bifilar wound helix gives identical results (within 0.2dB) to a λ/2 dipole.

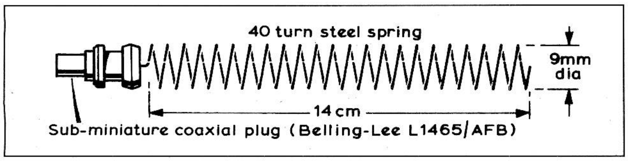


Fig 5.98. Details of a home-made helical whip for 145MHz. A BNC plug could also be used

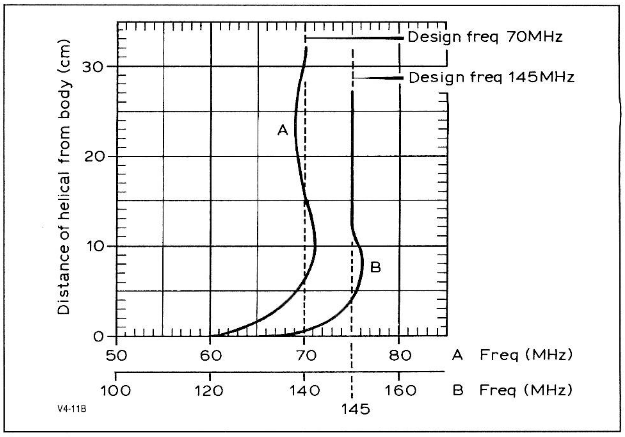


Fig 5.99. Frequency shift of a helical antenna on a typical hand-held transceiver for various distances from the body

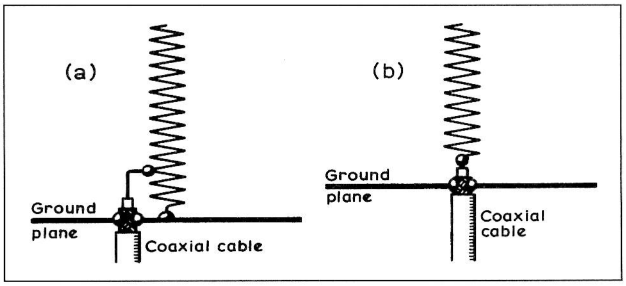


Fig 5.100. Two ways of feeding a helical antenna: (a) shunt feed, (b) series feed

The helical or spring has an interesting and difficult operating characteristic when supported close to the body, particularly at the higher frequencies. Fig 5.99 shows the typical results of a 145MHz or high-band spring and a 70MHz or low-band spring as it is brought closer to the body. The interesting effect which occurs at several centimetres from the body can be seen, where the resonance of the spring, instead of continuing to decrease due to body capacitance, suddenly increases the frequency of resonance. At 2cm and closer the operating frequency suddenly decreased due to body capacitance. Unfortunately this very changeable area occurs at the typical mounting distance of a body-worn transceiver. However, many transceivers are required to be raised to the mouth when transmitting and this puts the antenna back to its best operating conditions.

The normal-mode helical antenna can be vehicle or ground-plane mounted if desired. The height is typically less than 0.1λ and the gain is around to 2–3dB below a dipole. An acceptable match to 50Ω can often be achieved by simply trimming the resonant length. Alternatively, a small inductance or capacitor across the base or a shunt feed as shown in Fig 5.100 will provide the required matching.

The halo and super turnstile

Horizontally polarised antennas for a mobile station become complex and bulky when gain is required. A simple antenna which produces an almost omnidirectional horizontal radiation pattern is the *halo* in its various forms. Basically this is a λ/2 dipole, often gamma matched, which is bent round into a

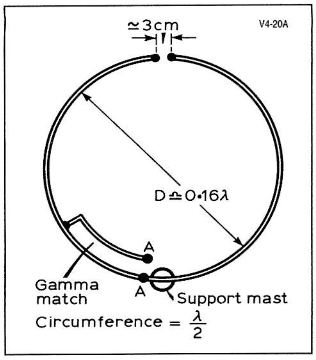


Fig 5.101. Dimensions of the λ/2 halo

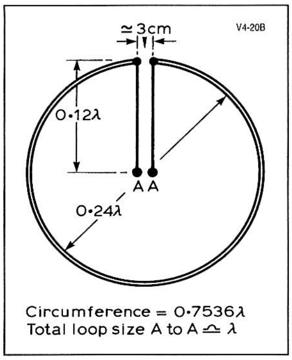


Fig 5.102. Dimensions of the lambda loop

circle or square. As can be seen in Fig 5.101, when correctly resonant the resultant radiation pattern is somewhat offset in favour of the direction of the gap. Best results are obtained when mounted at a minimum height of 70cm, 0.34λ at 144MHz, above the ground plane produced by the vehicle roof.

An extension of the λ/2 halo is the full-wave or *lambda loop.* A 1λ loop is drawn in at one point to the centre to produce both a support and a match transformer, to approximately 50Ω (see Fig 5.102). The addition of a 1:1 Pawsey stub or similar balun (see earlier section on matching) produces a near-omnidirectional pattern with a unity gain relative to the maximum radiation of a dipole. A comparison of the halo and the lambda radiation patterns are shown in Fig 5.103.

Further extension to three loops can be produced to form the *super turnstile* but the complexity and sheer physical size tends to limit this sort of structure to only the most daring mobile radio amateur.

Both the lambda loop and the super turnstile require to be at least 0.34λ above the ground plane surface to work satisfactorily.

Turnstile antennas along the lines of the crossed dipoles shown in the section on antennas for fixed stations were at one time popular for mobile operations, typically mounted on a λ/2 mast above the vehicle bodywork. However, the aerodynamic drag, wind noise, higher vehicle speeds and the danger of injuring pedestrians (the elements were often around eye height) have all discouraged the use of this and other horizontally polarised antennas for mobile use.

ANTENNAS FOR SATELLITE OPERATIONS

For the average radio amateur satellite antennas fall into two groups, both of which are *ground-station antennas,* that is, those on the ground rather than on the satellite itself. The two groups are *steerable,* which enable the passage of the satellite to be tracked across the sky, and *fixed* which, as they in the ideal case have a hemispherical radiation pattern, receive the satellite signals equally in any direction and do not require to track the satellite's passage. The tracking antennas are usually of high gain while the fixed antennas are usually relatively low gain due to the hemispherical coverage required. Fortunately, as signal losses between ground and satellite are low, being mostly line-of-sight path with no obstructions, relatively low-gain antennas of the fixed variety are often acceptable for reception of amateur or weather satellites.

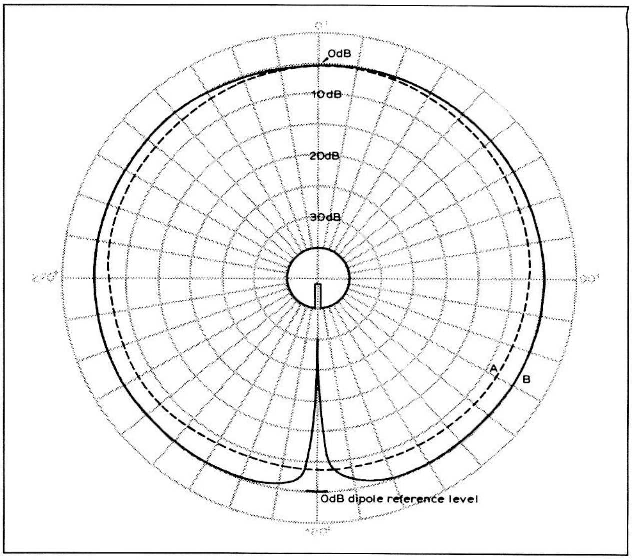


Fig 5.103. Decibel radiation pattern comparison of halo (A) and lambda loop (B) antennas

However, as the satellite tends to rotate, both groups of antennas are normally circularly polarised, right-hand by convention, to compensate for variations in polarisation.

Of the higher-gain tracking antennas, crossed Yagis and the helix are used in the main, with the crossed Yagis probably the easiest to construct and most readily available commercially. The sections on pp5.31 and 5.30 give details of crossed Yagis and the helix respectively.

For fixed or low-gain steerable antennas, several variations of crossed dipoles can be used and also the *volute,* which is a fractional turn four-element helix that can be made to give either directional gain or hemispherical circular polarised coverage (see later). It is worth noting that a conventional single-element helix requires two or more complete 'turns' to obtain circular polarisation.

Crossed dipoles over a ground plane

Fig 5.104 shows a simple arrangement of crossed dipoles above a ground plane. This type of antenna can be scaled for use at 29, 145 or 432MHz. A suggested version for 145MHz is shown in the figure. Mechanical problems may make the reflectors inadvisable in a 29MHz version. The height above ground can be about 2m for 145MHz and 3m for 29MHz. Typical dimensions are:

29MHz	driven elements	(λ/2)	188in	477.5cm
145MHz	driven elements	(λ/2)	38in	96.5cm
	reflectors		40.5in	103cm
	spacing	(0.3λ)	24.5in	62.2cm

The phasing line comprises λ/4 of 72λ coaxial cable, and the matching section λ/4 of 50Ω cable.

When calculating the length of the λ/4 sections, the velocity factor of the cable must be taken into account. Typically this is 0.8 for cellular and semi-airspaced types, and 0.66 for solid dielectric cables, but verification of the correct figure for the cable used should be obtained. As an example, a matching section of RG59/U would be 13in (33cm) in length. To obtain a 50Ω input impedance, the 50Ω transformer section

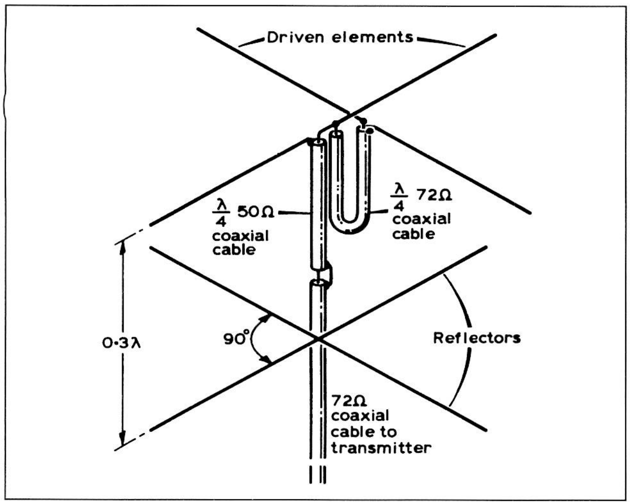


Fig 5.104. A crossed-dipole antenna for 145MHz

should be replaced by two pieces of 95Ω cable, λ/4 long, connected in parallel.

It is preferable to have a 1:1 balun included at each dipole centre to ensure a consistent pattern through 360° of azimuth. Dependent on the spacing between the dipoles and ground plane the radiation pattern can be made to be predominantly to the side for satellites low on the horizon or up for overhead passes. By drooping the dipole elements at 45° and with a spacing of approximately 0.4λ of the dipole mounting boss about the ground plane, a compromise radiation pattern can be achieved that tends to be hemispherical. As horizontal and vertical polarisation is affected differently by ground reflections, low-to-horizon flight paths will not produce circular polarisation. This is due both to ground scatter from the satellite and ground reflections at low levels of incidence at the receiving antenna and its ground plane.

Circular polarisation is normally produced by feeding one dipole 90° out of phase to the second dipole by means of a phasing harness containing an extra λ/4 on one side.

An alternative approach to this method of phasing is to utilise the phase properties of a capacitive or inductive reactance. Suppose, for example, that the length and diameter of the dipoles are made to give a terminal impedance of 70–j70Ω (capacitive). By introducing a series reactance (inductive) of +j70Ω at each terminal of one of the dipoles (Fig 5.105) the terminal impedance of this dipole becomes 70 + j70Ω. With

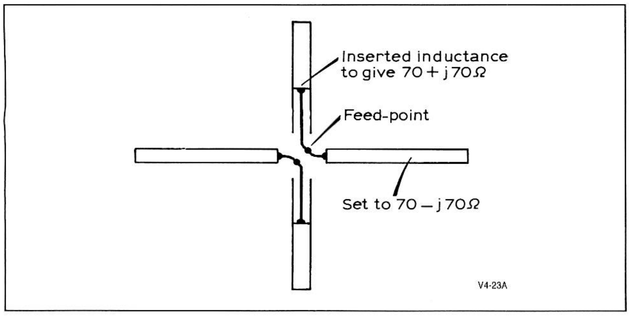


Fig 5.105. Achieving phase quadrature by introducing a reactance in one arm

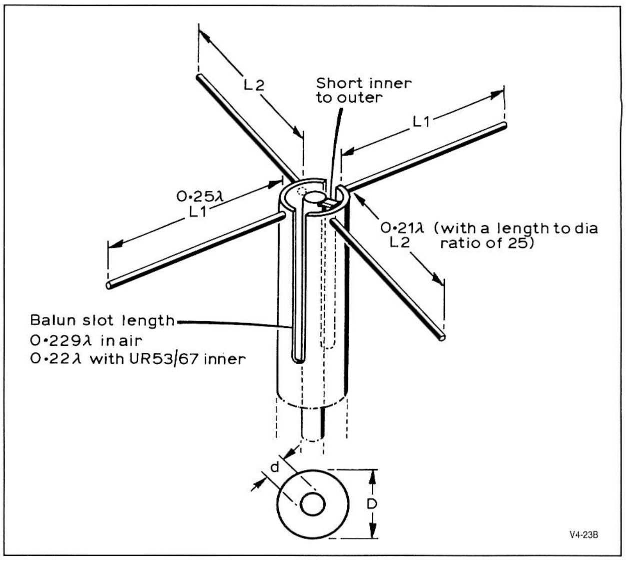


Fig 5.106. A starpole turnstile. *D*/*d* = 1.86 for 75Ω and 1.5 for 50Ω

the two dipoles connected in parallel the current in each dipole is equal in magnitude but, due to the opposite phase differences of 45° in each dipole, a total phase difference of 90° (phase quadrature) is achieved which produces circular polarisation.

The two impedances in parallel become 70 + j0Ω so the addition of a 1:1 balun provides a direct match to a 70Ω coaxial line. If the impedance of the balun is correctly proportioned this match can be to the standard 50Ω coaxial line. Radiating elements can be drooped as previously described to improve the hemispherical coverage. An easier way of introducing the series inductance is simply to make one dipole long, therefore inductive, at the working frequency, and one dipole short, therefore capacitive at the working frequency.

Fig 5.106 shows a working example of the *starpole turnstile* arrangement. The reactive components were chosen as ±25Ω and dimensions were based on the reactive information for dipoles as shown in Fig 5.107.

Volute antennas

The volute can also make use of both phasing line or the reactance method to produce circular polarisation. The number of 'turns' or part turns of the radiating elements combined with their length can be used to produce various radiation patterns. Radiation patterns produced for several combination of turns and resonant lengths are shown in Figs 5.107(a) to (d) with general details of the volute in Figs 5.108 and 5.109 [25, 26]. It must be noted that elements that are multiples of λ/4 have open-circuit ends, while the elements that are multiples of λ/2 can be short-circuited to the mounting structure.

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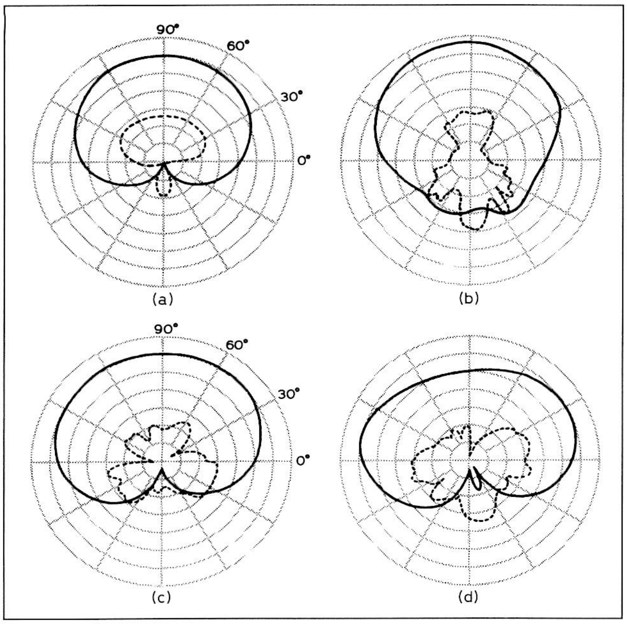

Fig 5.107. Volute radiation patterns. (a) Three-quarter-turn λ/4 volute. (b) Three-quarter-turn λ/2 volute. (c) Three-quarter-turn 3λ/4 volute. (d) Three-quarter-turn 1λ volute (*Microwave Journal*)

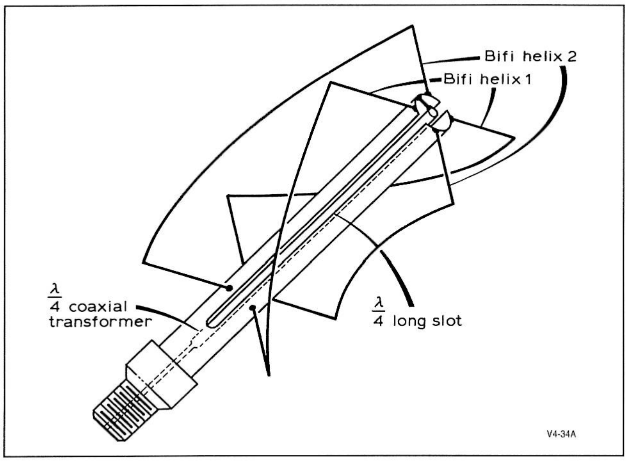

Fig 5.108. A quarter-turn volute with split sheath or slot balun (*Microwave Journal*)

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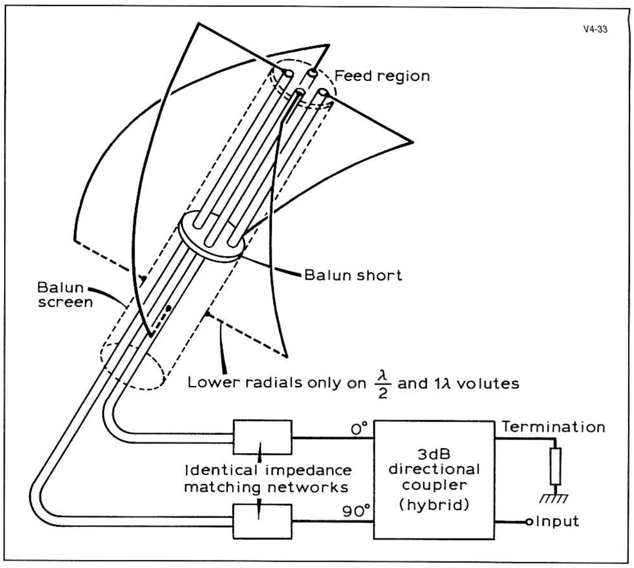

Fig 5.109. The general arrangement using Pawsey stub baluns. A half-hybrid or λ/4 phasing harness as used for the crossed dipoles can be used in place of the directional coupler (*Microwave Journal*)

[11] *Yagi Antenna Design*, J L Lawson, ARRL, 1986.

[12] 'Yagi antennas – principle of operation and optimum design criteria', G Hoch, DL6WU, *VHF Communications* 3/1977.

[13] 'More gain from Yagi antennas', G Hoch, DL6WU, *VHF Communications* 4/1977.

[14] 'Extremely long Yagi antennas', G Hoch, DL6WU, *VHF Communications* 1/1982.

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[16] 'Diplexers for the VHF bands', J Regnault, G4SWX, *Radio Communication* March 1996, pp63, 65 and April 1996, pp61–63.

[17] 'Wind loading', D J Reynolds, G3ZPF, *Radio Communication* April 1988, pp252–255 and May 1988, pp340–341.

[18] 'A quadruple quad antenna – an efficient portable antenna for 2 metres', M Ragaller, DL6DW, *VHF Communications* 2/1971, pp82–84.

[19] *The Antenna Experimenter's Guide*, P Dodd, G3LDO, 2nd edn, RSGB, 1996.

[20] 'Antenna modelling on a PC', I White, G3SEK, *Radio Communication* August 1993, pp39–41.

[21] 'Antenna modelling with ELNEC', J Bazley, G3HCT, *Radio Communication* August 1993, pp41–42.

[22] 'The VHF log-periodic Yagi', M Gibbings, G3FDW, *Radio Communication* July 1994, pp13–17.

[23] 'Helical antennas for 435MHz', J Miller, G3RUH, *Electronics and Wireless World* June 1985, pp43–46.

[24] 'Slim Jim antenna', F C Judd, G2BCX, *Practical Wireless* April 1978, pp899–901.

[25] 'Resonant quadrifilar helix design', C C Kilgus, *Microwave Journal*, Vol 13-12, December 1970, pp49–54.

[26] 'Shaped-conical radiation pattern performance of the backfire quadrifilar helix', C C Kilgus, *IEEE Trans Antennas and Propagation* May 1975, pp392–397.

6 EMC

ELECTROMAGNETIC compatibility is the desirable situation where nearby electronic equipment is not affected by amateur transmissions and does not unduly affect amateur reception. Further information on EMC for radio amateurs is given in references [1] and [2]. This chapter also deals with interference to amateur reception by non-radio equipment, particularly a computer in the shack.

If amateur transmissions affect a neighbour's TV or other electronic equipment, there are three possible causes:

(a) The affected equipment has insufficient immunity to the fundamental frequency of the transmitter.
(b) A harmonic or other unwanted emission from the amateur transmitter is not sufficiently well suppressed.
(c) A non-linear device somewhere else is receiving the amateur signal and re-radiating a harmonic.

The most common reason is (a), while (b) is less likely and (c) is rare. If the affected equipment is a TV, video recorder or some sort of radio receiver then it is important to establish whether the cause is (a) or (b). If the amateur transmitter is producing an unwanted signal then adding a filter at the TV cannot remove it! Conversely, if a TV or other affected equipment has insufficient immunity to the fundamental frequency of the transmitter then the problem cannot be solved by filtering at the transmitter! If the affected equipment is an audio amplifier, wired telephone or other equipment which is not intended to receive radio signals at all, then clearly the problem is not due to harmonics.

Even if harmonics etc are not causing a problem, it is wise to ensure that they are adequately suppressed in case of a visit from the Radiocommunications Agency. It is also important to realise that a VHF or UHF amateur station may be able to generate very high field strengths nearby.

GOOD RADIO HOUSEKEEPING

The Schedule of the *Terms, Provisions and Limitations Booklet BR68*, which accompanies the full UK Amateur Licence 'A' or 'B', lists a maximum output power level of 26dBW (400W) on most bands but this is subject to certain other conditions such as Note (l) which states:

> "In densely populated areas sufficient separation of amateur equipment from surrounding transmitters, receivers and electronic equipment may not be possible to permit the amateur to operate with high power without the high probability of causing interference."

The above statement emphasises the need to plan with EMC in mind where an amateur station is to be installed in a typical urban environment in close proximity to neighbours. The RF from the transmitter should be kept under reasonable control, with the highest possible percentage going where it is wanted (in the direction of the distant station) and as little as possible going into the local environment. This is sometimes known as *good radio housekeeping*. Fortunately, installations designed to achieve this are also likely to minimise the pick-up of locally generated interference.

WARNING – Protective multiple earthing (PME)

Many houses in the UK, particularly those built or wired since the middle 'seventies, are wired on what is known as the *PME system*. In this system the mains earth of the house wiring is bonded to the neutral where the supply enters the building. In the event of certain rare fault conditions it is possible for the earth and neutral conductors all over the house to rise to a voltage significantly above that of the true earth (ie the earth out in the garden). In extreme cases the earth neutral voltage could be the full mains voltage above true earth. For this reason the supply authorities advise certain precautions regarding the bonding of metalwork inside the house.

WHERE A HOUSE IS WIRED ON THE PME SYSTEM, DO NOT CONNECT ANY EXTERNAL (ie radio) EARTHS TO APPARATUS INSIDE THE HOUSE unless suitable precautions are taken. (See reference [1] or [2]).

Antennas

It is always good practice to erect any antenna as far from houses as possible, and as high as practical, subject to planning constraints. If possible, VHF/UHF beams should be located so that most of the power is beamed over the roof tops towards the horizon rather than towards neighbouring buildings or TV antennas. A mast away from the house (and neighbouring houses) is preferable subject to considerations of feeder loss.

Coaxial feeder should be well screened with good-quality woven braid to minimise RF leakage into and out of the cable. When a balanced antenna is fed by coaxial cable, a balun should be used to minimise RF radiation from the braid of the cable. Any remaining RF currents on the braid of the cable can be reduced by passing the latter through some large ferrite beads or clip-on ferrite chokes near the antenna.

Earths

Most VHF or UHF antenna systems do not require an RF earth for EMC reasons but don't forget the need for lightning protection. It should also be borne in mind that when antennas are mounted on an earthed tower, the braid of the coaxial feeder entering the radio shack is likely to be earthed via the antenna to the tower. In such cases, it is important to consider PME (protective multiple earthing) – see the warning panel on this page.

Field strengths

Note (l) in the BR68 booklet mentions excessive field strength from an amateur station. It is therefore useful to be able to calculate the electric field strength generated near a transmitting antenna. The electric or E-field strength at a certain distance is given by equation 6.1 which assumes far-field 'free space' conditions.

$$E = \frac{\sqrt{49.15 P_d}}{d} \qquad (6.1)$$

where E is the electric field strength in volts/metre, d is the distance from the antenna in metres and P_d is the effective radiated power.

It should be noted that P_d is ERP, ie the input power to the antenna multiplied by the gain of the antenna relative to a dipole, not an isotropic radiator.

Fig 6.1 shows the field strength generated by an antenna radiating between 1W and 4kW ERP at distances between 1m and 100m. For example, 100W (20dBW) into an antenna with a gain of 10dBd gives an ERP of 30dBW or 1kW. At a distance of 10m, this would produce a field strength of 22V/m in the direction of maximum radiation. This example shows that a high-power VHF/UHF amateur station can easily generate a field strength which is higher than nearby electronic equipment can reasonably be expected to withstand. Field strength is also mentioned in leaflet RA234 (see below). It may therefore be necessary to avoid using maximum power in certain directions in order to avoid generating excessive field strengths in neighbours' premises. In any case, it is good practice to use only as much power as necessary for a contact. In the case of local contacts, the power necessary may only be one watt or less.

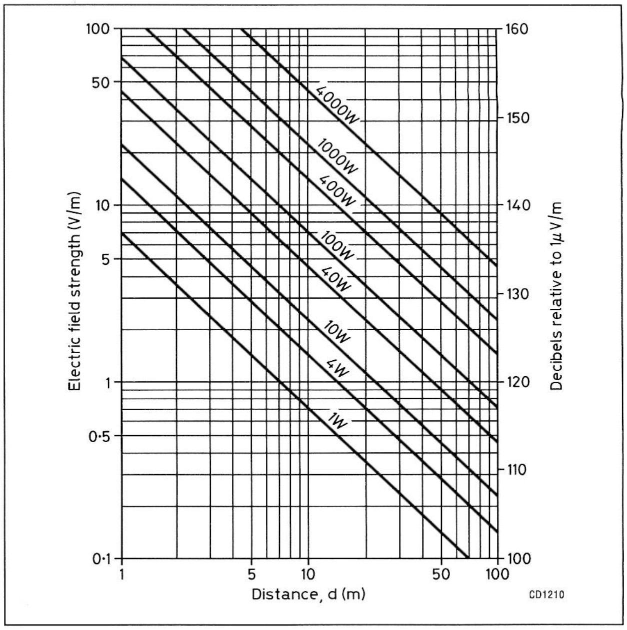


Fig 6.1. Field strength in the direction of maximum radiation from an antenna radiating from 1W to 4kW ERP into free space.

RF IMMUNITY STANDARDS

The UK EMC Regulations (Statutory Instrument 1992 No 2372) came into force on 1 January 1996. Almost all electronic equipment manufactured since that date must meet certain requirements for RF immunity and emissions and must also carry the CE mark to indicate compliance with all applicable European Directives.

This was a major step forward in EMC, as previously there was no requirement for consumer electronic equipment to meet any immunity standard in the UK. Nevertheless, the standards only set a basic level of immunity compared to the possible field strengths from an amateur station. In some cases it is argued that if immunity problems arise when an amateur station generates field strengths in excess of the levels specified in the relevant immunity standard, the amateur should take steps to reduce the field strength. In such cases, it is worth noting the standard, IEC 1000-2-5, *Electromagnetic Compatibility (EMC) – Part 2: Environment – Section 5: Classification of Electromagnetic Environments*. This lists various sources of RF fields (and other types of electromagnetic phenomena) which may exist in various classes of location including residential (urban), residential (rural), commercial etc. In most of these environment classes including residential (urban), the standard states that field strengths of up to 10V/m may be encountered in locations which are at least 20m from the nearest amateur radio transmitter.

Selected information about the RF immunity requirements of some CENELEC harmonised European immunity standards is given below. As standards are revised from time to time, this information may not apply to future editions. For further details, an up-to-date copy of the relevant standard should be consulted. These are often kept by university or college libraries and by learned societies such as the IEE.

EN 55104 : 1995 EMC – immunity requirements for household appliances, tools and similar apparatus

This standard specifies immunity requirements which are generally less demanding than the Generic Standard. It is expected to be replaced with EN 55014-2 : 1996.

EN 55020 : 1995 Immunity of broadcast receivers and associated equipment

This is applicable to television broadcast receivers, sound broadcast receivers and associated equipment such as audio systems. Radio and TV equipment without a connection for an external antenna or without a mains power connection is not required to meet this standard.

The standard specifies various tests for immunity to radiated and conducted signals. For UHF TV receivers, VHF Band II (87.5–108MHz) sound broadcast receivers and audio equipment, the radiated immunity is tested up to 150MHz at a field strength of 125dB(µV/m) (1.78V/m) (carrier) with 80% AM at 1kHz. The modulated signal therefore has a peak envelope voltage (PEV) of 3.2V/m. There are some exclusion bands around IF frequencies and the tuned frequency. Long-wave, medium-wave and short-wave sound broadcast receivers are only tested for immunity to common-mode conducted currents on the antenna terminals (if any) at 26–30MHz.

Draft CISPR 24 Ed. 1 : Information technology equipment – immunity characteristics – Limits and methods of measurement

CISPR 24 will form the basis of EN 55024-1 and requires similar levels of immunity to the 1996 draft Generic standard. Although the title refers to information technologyequipment, this standard also applies to Telecommunications Terminal Equipment (TTE) such as telephones. Until EN 55024-1 is issued, however, a manufacturer can CE mark a telephone by testing it to the Generic Immunity Standard, EN 50082-1 : 1992. Fortunately, the immunity of many CE-marked telephones exceeds this minimum requirement.

EN 50082-1 : 1992 EMC Generic Immunity Standard – Part 1; residential, commercial and light industry

The Generic Standard EN 50082-1 : 1992 can be applied if no product-specific standard exists although it may not be appropriate for all cases. This is a 'watered-down' version of the 1990 draft, prEN 50082-1. The 1992 edition references the IEC 801-3 standard for immunity to radiated fields but only tests from 27–500MHz at 3V/m with an *unmodulated* carrier. Immunity tests below 27MHz are not required and neither are tests for immunity to RF signals picked up on cables.

A new edition of the Generic Standard BS EN 50082-1 : 1996 was published on 1 June 1997. This is a great improvement on the 1992 edition but equipment can still be tested to the weak 1992 standard until 1 July 2001 which is the date of withdrawal. The 1996 draft references the EN 61000 (IEC 1000) series of standards and specifies tests which are broadly similar to the 1990 draft. These include conducted immunity tests on interconnecting cables and mains cables with a 3V signal from 0.15–80MHz and radiated immunity tests at 3V/m from 80–1000MHz. In each case, a 3V or 3V/m carrier is modulated with 80% AM at 1kHz. The PEV (peak envelope voltage) on modulation peaks is therefore 1.8 times higher at 5.4V or 5.4V/m. There is also a new test at 900MHz with 200Hz pulse modulation to simulate GSM cellular telephones.

Section 3 of prEN 50082-1 : 1996 includes the following statements:

> "The immunity requirements have been selected so as to ensure an adequate level of immunity for apparatus at the locations described. The levels do not however cover extreme cases which may occur in any location but with an extremely low probability of occurrence."
>
> "In special cases, situations will arise when the level of disturbance may exceed the levels specified in this standard; for example, a hand-held transmitter used in close proximity to an apparatus. In these instances special mitigation measures may have to be applied."

prETS 300 683 EMC Standard for Short Range Devices

This is an ETSI (European Telecommunications Standards Institute) standard for short-range radio devices which includes specifications for RF immunity. Such devices include baby alarms, cordless household alarm systems etc. Immunity to radiated fields is not tested at 3V/m with 80% AM except within an exclusion band of ±5% of the receive frequency as performance within this range is considered to be a 'spectrum utilisation parameter' rather than an 'EMC parameter'. Narrow-band spurious responses (such as image frequencies) can be declared outside the ±5% exclusion band.

BREAKTHROUGH OF AMATEUR TRANSMISSIONS

Television and video

If breakthrough occurs with a UHF TV set connected directly to an antenna, a suitable high-pass filter should be plugged in at the TV antenna socket. If the problem cannot be cured by filtering the antenna input, it is possible that winding the mains cable through a ferrite core close to the TV set may help. At VHF, however, direct pick-up in the TV set itself may occur and cannot be cured by external filtering. Possible causes include pick-up in the vision IF amplifier circuitry or pick-up in the internal loudspeaker cable. In the case of TV sets with external speakers, pick-up in loudspeaker cables can be tackled in the same way as for audio systems (see below).

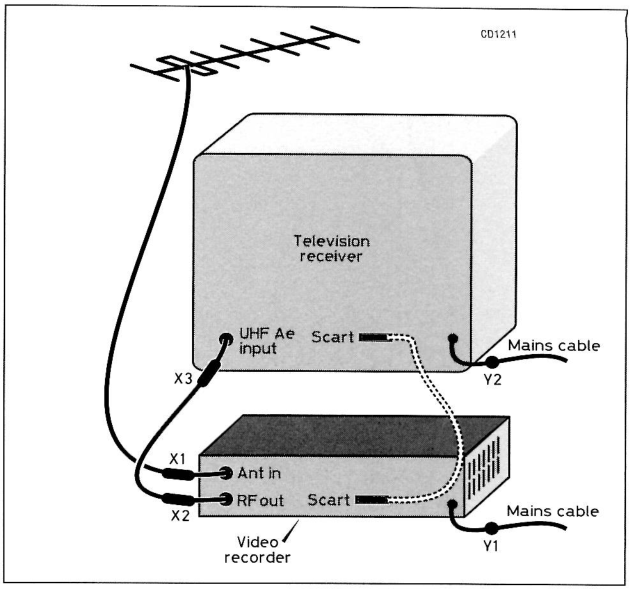


Fig 6.2. Fitting filters to a TV with video recorder

If breakthrough occurs when a TV is connected to a video recorder as in Fig 6.2, the first thing to establish is whether the TV set alone is immune. The addition of a video recorder makes breakthrough more likely because most video recorders contain a VHF/UHF amplifier with a bandwidth of typically 40–860MHz. A high-pass filter should be fitted at X1 and another may be required at X2 or X3. In some cases, winding the mains cable through a ferrite core at Y1 or Y2 may help. If the TV and video recorder both have a SCART connector, the use of a SCART cable (shown dotted) provides a direct 'base-band' audio and video connection on playback which bypasses the UHF modulator in the video recorder and the tuner and IF stages in the TV. This generally improves picture quality on video playback and also reduces the possibility of RF breakthrough. If the tuner or IF stages in the TV are susceptible to breakthrough but the video recorder is immune, it is possible to receive off-air UHF signals without using the tuner and IF stages in the TV. The channel tuned on the video recorder can be viewed on the TV via the SCART connection. The only exception to this is when recording one channel while watching another.

Satellite TV

In a satellite television receiving system, the LNB (low noise block converter) in the dish is unlikely to be affected by amateur signals directly, except on the 10GHz amateur band. The output of the LNB is at the first intermediate frequency covering 950–1750MHz or 950–2050MHz so there is a possibility of 23cm amateur signals causing IF breakthrough.

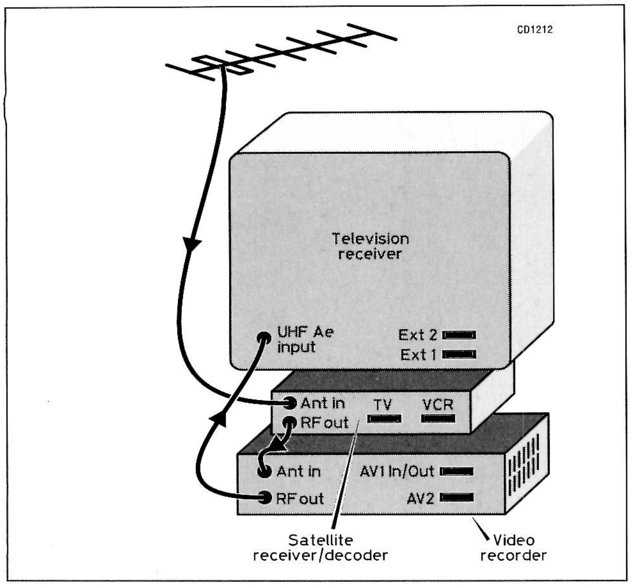


Fig 6.3. TV, satellite receiver and video recorder with UHF connections only

Fortunately, the cable between the LNB and the indoor receiver unit is normally quite well screened with foil and braid which minimises the possibility of IF breakthrough. Nevertheless, common-mode signals can be picked up on the braid of the cable to the dish which may require a common-mode ferrite choke.

The indoor satellite receiver unit normally contains a broad-band amplifier and UHF modulator for use with UHF connections as shown in Fig 6.3. Such a set-up may suffer breakthrough of amateur signals due to the cascaded broad-band amplifiers in the satellite receiver and video recorder.

Fig 6.4 shows a possible set-up which should reduce the chance of RF breakthrough while providing improved picture quality for satellite reception and video playback. With some satellite receivers, video playback via the SCART lead can pass through the satellite receiver to the TV but, if this is not possible, a third SCART cable (shown dotted) is required for video playback.

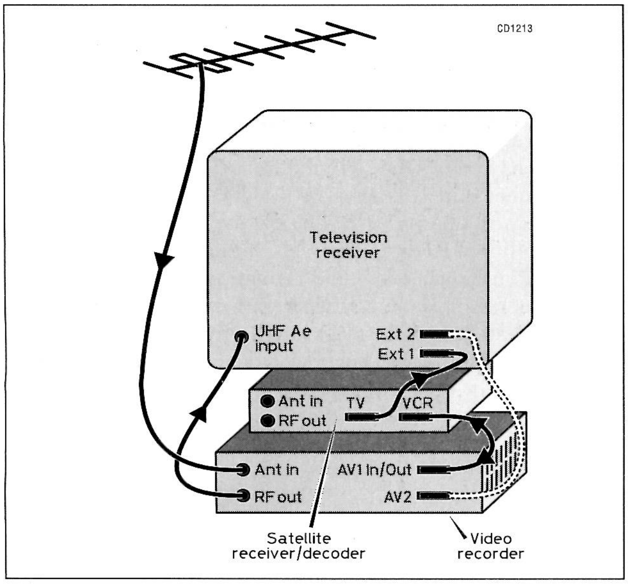


Fig 6.4. TV, satellite receiver and video recorder with SCART connections

TV mast-head preamplifiers

In a fringe area or where a long feeder cable is necessary, a TV masthead preamplifier can give a significant improvement in reception. Nevertheless, mast-head preamplifiers are sometimes used inappropriately in areas of adequate reception in an attempt to compensate for a low-gain antenna and/or poor-quality feeder cable. The use of a mast-head preamplifier increases the chances of breakthrough of out-of-band signals from amateur transmitters or other radio transmitters nearby. Older preamplifiers may be unscreened and have broad bandwidth which can lead to breakthrough problems on any VHF/UHF amateur band. If strong amateur signals are picked up by the TV antenna and cause breakthrough, a filter is required at the input to the TV antenna but unfortunately, this is not easy to install with a mast-head preamp! In such cases, the only solution may be to install a new preamplifier which is CE marked, adequately screened and covers UHF only.

TV distribution amplifiers

Indoor or loft-mounted TV distribution amplifiers are available with two, three, four or more outputs. If any of the amplifier outputs are not used, these should be terminated with a 75Ω load otherwise the gain of the amplifier is increased. Distribution amplifiers are available in two types: broad-band and UHF only. The broad-band type, which typically covers 40–860MHz, can distribute 88–108MHz Band 2 FM broadcast and Band 4/5 UHF TV signals simultaneously. In installations where the Band 2 capability is not used, a UHF high-pass filter may be required at the input to the amplifier. If UHF TV and Band 2 antennas are both connected, a combiner is required and any UHF high-pass filter must be fitted at the UHF input to the combiner. The Band 2 antenna may pick up 50, 70 or 144MHz amateur signals, in which case an 88–108MHz band-pass filter is required at the VHF input to the combiner. Non-CE marked amplifiers in unscreened cases may also be susceptible to picking up signals via the mains cable or directly in the amplifier itself. Such effects may be reduced by fitting the amplifier in a screened box grounded to the coaxial braid or by fitting a ferrite ring to the mains cable.

Fig 6.5 shows a 'worst case' configuration from the EMC point of view. This configuration allows multiple TV sets to select the currently tuned satellite channel, video playback or any terrestrial channel. Band 2 FM signals are also distributed by the same broad-band amplifier and output cables. With three wide-band amplifiers cascaded, the probability of breakthrough is clearly increased. To solve a problem with this type of set-up, it is advisable to start with the satellite receiver feeding the TV alone and to cure any breakthrough problem. This may require a high-pass filter at the input to the satellite receiver. The video recorder should then be connected, followed by the distribution amplifier alone. Finally, the combiner and Band 2 FM antenna should be added. It may be necessary to fit additional filters at the input to the video recorder or distribution amplifier and an attenuator may also be required to reduce the total gain.

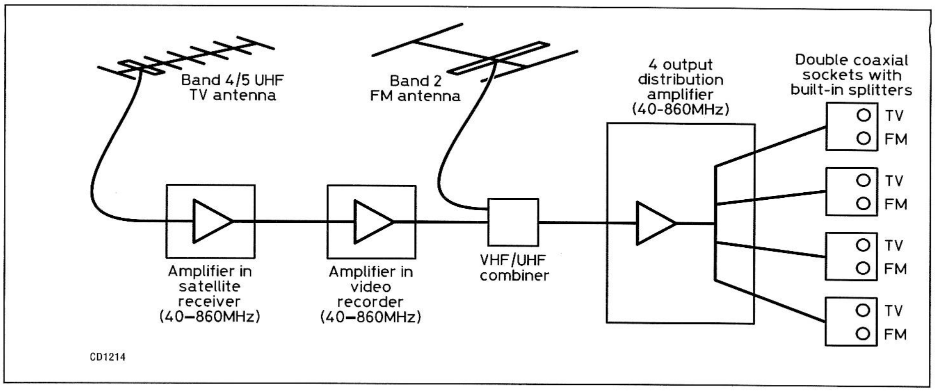


Fig 6.5. 'Worst case' configuration with satellite receiver, video recorder and TV/FM distribution amplifier

Cable TV

The introduction of new cable television (CATV) systems may reduce TVI problems because neighbours who subscribe to the cable system no longer need to use their UHF TV antenna. There are, however, some potential EMC problems due to signals leaking into and out of CATV systems (known to cable TV companies as *ingress* and *egress* respectively). Many systems use harmonically related vision carriers on multiples of 8MHz, for example from 128 to 560MHz or higher with an FM sound carrier 6MHz above each vision carrier. Some systems use a vision carrier at 432MHz with sound on 438MHz. Although some UK cable TV operators avoid the use of 144MHz as a vision carrier, others use it for a leakage detection test signal or for programmes. This may lead to 144MHz breakthrough problems even with low power.

For cable TV systems, the permitted levels of radiated emissions are defined in the Radiocommunications Agency standard MPT 1510. This standard prohibits the use of certain frequencies and specifies low levels of leakage in certain bands including 50–54MHz, 144–146MHz and 432–440MHz [8]. It is anticipated that this standard will eventually be withdrawn and replaced by a harmonised European standard.

FILTERS FOR RADIO AND TV RECEIVERS

The details below relate to filters for reducing breakthrough of 6m, 4m, 2m and 70cm amateur band signals. For details of filter performance on the HF bands and a filter which also rejects the 23cm band, see Appendix 3 of reference [1] or [2]. All the filters listed below are for use in the antenna input of TV and FM radio receivers and allow UHF TV signals to pass through. As with any filter, there is a small loss in the pass band. Only the HPF2 allows both VHF/FM broadcast radio (Band 2, 88–108 MHz) and UHF TV (Bands 4/5) to pass through.

Unwanted pick-up of amateur signals by a UHF TV antenna system can occur in two different ways. The first is where the TV antenna itself picks up VHF or UHF amateur signals which pass along the inner of the coaxial cable and return via the braid in the normal way. This is most likely to occur at 432MHz where a UHF TV antenna still has a moderate gain but it can still be a problem at 144MHz. Pick-up by the UHF TV antenna itself tends to be less significant on 70 and 50MHz, however. The second type of pick-up is where the whole TV antenna with its downlead acts as a receiving antenna, resulting in signals which are on the braid and inner together relative to earth. If such signals cause breakthrough, some sort of 'braid breaker' is required. There are four types, the transformer type such as the BB1 or HPFS, the capacitive type such as the HPF1, the resonant type such as the TNF range and the ferrite common-mode choke.

HPF1 high-pass filter and braid breaker

Pass band: UHF TV.

Stop band: All bands up to and including 144MHz (good performance on 50 and 70MHz, some effect on 144MHz).

'Braid breaking': Capacitive braid breaking at HF, some effect on 50MHz but little effect on 70 and 144MHz.

Remarks: The HPF1 is not stocked separately by RSGB but is included in the RFK1 filter kit (see below).

HPF2 High-pass filter (RSGB order code Filter 2)

Pass band: FM radio broadcast (88–108MHz) up to UHF TV.

Stop band: All HF bands plus limited effect at 50MHz.

'Braid breaking': None.

HPF6 high-pass filter (RSGB order code Filter 8)

Pass band: UHF TV.

Stop band: All bands up to and including 430–440MHz.

'Braid breaking': None.

Remarks: This high-performance six-section filter has a very sharp cut-off below 470MHz and is primarily intended for rejecting the 430–440MHz amateur band. It also offers a high degree of rejection at 144MHz and below.

BB1 braid breaker (RSGB order code Filter 1)

Pass band: All VHF/UHF amateur bands plus FM Band 2 (88–108MHz), and UHF TV.

Stop band: The BB1 does not give any rejection of any VHF/UHF amateur band as it is a 'braid breaker' rather than a filter.

'Braid breaking': A 1:1 transformer type braid breaker. Moderately effective on 50 and 70MHz, limited effect on 144MHz.

Remarks: A BB1 on its own is only likely to be useful where the pick-up is primarily on the braid of the coaxial cable. It can be cascaded with other filters such as HPF2 or HPF6 which do not have any braid-breaking action, although this increases the total pass-band loss.

HPFS high-pass filter (special) (RSGB order code Filter 3)

Pass band: UHF TV.

Stop band: All bands up to and including 144MHz.

'Braid breaking': Includes 1:1 transformer type braid breaker (see BB1 above).

Remarks: The HPFS is a BB1 combined with a high-pass filter. Due to the relatively high pass-band loss, it is not suitable for areas where the TV signal strength is low.

RBF1/70cms notch filter (RSGB order code Filter 5)

Pass band: UHF TV

Stop band: 430–440MHz.

'Braid breaking': None.

Remarks: The RBF1/70cms is pretuned to reject 435MHz although the HPF6 is more effective on the 430–440MHz band. The RBF1/70cms also has a high-pass action with some rejection at 70MHz and below.

TNF2 tuned notch filter range (VHF types)

Type	RSGB order code
TNF2/145	Filter 4, notch tuned to 145MHz
TNF2/70	Filter 7, notch tuned to 70MHz
TNF2/50	Filter 6, notch tuned to 50MHz

Pass band: UHF TV.

Stop band: Only the specified band.

'Braid breaking': Only on the specified band.

Remarks: These filters are designed to provide rejection of one particular amateur band. They have a parallel L-C notch (band-stop) filter in series with the inner conductor and with the braid. The manufacturers state that these notch filters are unsuitable for use with some TV distribution amplifiers.

Ferrite rings and cores

Unwanted breakthrough of RF signals into electronic equipment can occur in two ways. RF signals may be picked up directly inside the affected equipment or they may be conducted in via external cables such as loudspeaker cables, antenna cables and mains cables. If the pick-up is on cables, it can often be reduced or eliminated by winding the affected cable onto a suitable ferrite core. This forms a 'common-mode choke' which presents a high impedance to unwanted RF signals without affecting the wanted signals. A common-mode choke can also be used as a 'braid breaker' for a TV or FM radio coaxial antenna cable. Compared to other types of braid breaker, it has the advantage that it introduces little or no loss to the wanted signal and maintains the integrity of the braid of the coaxial cable.

For the best chance of success, it is important to use a suitable grade of ferrite and a suitable number of turns in order to achieve the highest possible impedance at the frequencies of interest. Ferrite rings available from RSGB are made in the USA by Fair-Rite Corporation in type 43 material. The inside diameter is 22.85mm (0.9in) and the width is 12.7mm (0.5in). They are equivalent to FT140-43. These rings give good results at VHF whereas some other grades of ferrite are poor at these frequencies.

At 50 or 70MHz, an impedance of 3kΩ or more can be obtained using 12 turns on two Fair-Rite grade 43 ring cores wound separately, although in many cases one ring is sufficient. At 144MHz, six turns are recommended on a ring core. Further details of the characteristics of various ferrite rings are given in reference [3]. Stray capacitance between the ends of the winding core is critical at VHF and, in the case of a ring core, this should be minimised by keeping the ends of the winding separated and securing them to the core with cable ties. If the cable is very thick, if it has connectors which cannot easily be removed or if it is not long enough, the best solution is usually to make up a short extension lead by winding a length of the thinnest suitable cable through a ferrite core and then fitting suitable connectors. This also has the advantage that it simply plugs in which is much more satisfactory when dealing with neighbours' equipment.

Normal TV coaxial cable has a minimum bending radius of typically 26mm so it should not be wound tightly through a ferrite ring otherwise it may collapse internally and short-circuit. Instead, a one metre length of miniature 75Ω coaxial cable can be wound onto a ring core or a clip-on core and fitted with coaxial connectors. Suitable cable is available from Maplin Electronics (Stock No XR88V). To ensure that the cable grip in the coaxial plug grips reliably, sleeving or PVC tape should be fitted to the end of the cable to increase its outside diameter to 5–6mm.

Various types of clip-on core are available but the type with 'U' shaped cores is not particularly effective unless four pairs of 'U' cores are stacked together. For best results, the core aperture should have a length which is two or three times its diameter. The split bead type of clip-on ferrite core is available from several sources including Maplin Electronics (BZ34M), Farnell Components (535-904) or RS Components (779-813 or 779-863). With such cores, three turns are recommended for all VHF bands.

Audio systems

A common cause of RF breakthrough in audio systems is RF being picked up in the loudspeaker cables and fed into the power amplifier where it is detected and comes out again as audio. A symptom of this effect is that turning the volume down does not reduce the breakthrough but it disappears if the speakers are unplugged and headphones are used. This effect can often be reduced by fitting suitable ferrite ring cores or clip-on chokes to the loudspeaker cables close to the outputs from the amplifier. RF may also get in to the amplifier inputs, in which case the breakthrough is affected by the setting of the volume control and may affect only one source such as cassette. In hi-fi systems composed of separate units, a plug-in filter may be required for the appropriate audio input.

Telephones

There are large variations in RF immunity between different models of wired telephone. If RF breakthrough occurs on a telephone which is rented from BT, the customer should ask to have it exchanged for another model with better RF immunity. In many cases, breakthrough at VHF is caused by direct pick-up in the telephone itself or in the handset cable so that it cannot be cured by means of a line filter. In any case, at the time of writing, telephone line filters which are available in the UK and are BABT approved are not suitable for VHF use. For example, the BT 'Freelance' RFI filter LJU 10/14A and the BT80A/RF2 filtered junction box contain chokes with a self-resonant frequency of around 1.3MHz for filtering medium-wave broadcast signals and are little use at VHF.

As telephones used on the UK public telephone system require BABT approval, they should not be modified to improve RF immunity and neither should home-constructed telephone line filters be used. There is, however, no objection to winding the cable of the telephone through a ferrite core which may reduce RF breakthrough if this is caused by common-mode RF signals on the cable. In the case of answering machines, fax machines or modems, a ferrite ring may also be required on the power supply cable or other cables.

Low-power devices

Various low-power devices, also known as *short-range devices* (SRDs) are exempt from UK licensing provided they are UK type-approved. Some devices such as vehicle radio keys or baby alarms operate on frequencies in or near amateur bands and can be susceptible to breakthrough.

Vehicle keys

Most radio-controlled car alarms and immobilisers made since mid-1994 operate on a harmonised European frequency of 433.92MHz which is allocated to them on a secondary unprotected basis. Some cars contain superhet receivers whose local oscillator radiates a detectable signal at around 433.275–433.475MHz. These receivers use an IF centre frequency of approximately 500kHz and earlier types have no image rejection so they can be blocked by UK 70cm repeater output frequencies. Some types have a SAW RF bandpass filter to provide some image rejection.

Another type of receiver is the super-regenerative. Some types may have a SAW stabilised oscillator or SAW bandpass filter but those without either have a –6dB bandwidth of typically 6MHz.

In some systems, the radio key only controls central locking whereas in other systems, it also operates the immobiliser. If an amateur receives a complaint about blocking of vehicle radio keys by 70cm transmissions, the first thing to find out is whether there is an alternative way of disarming the immobiliser, for example by entering a code number manually. If there is no alternative to the radio key then it could be argued that the designers of such a system have used an unprotected frequency allocation for an unsuitable purpose.

Baby alarms

Devices which are UK approved to MPT 1336 are allowed up to 10mW ERP from 49.82–49.98MHz although many have an ERP of 1mW or less. This frequency allocation is used for wireless baby alarms, licence-exempt walkie-talkies and cordless headphones. The receivers used in these low-cost devices may have poor rejection of signals in the 50MHz amateur band, and in some cases only a few watts on 50MHz can cause breakthrough up to 100m away [4]. Some baby alarms have a switch to select one of two frequencies but this is unlikely to make much difference to breakthrough from 50MHz amateur signals. A few models have sockets to allow a cable to be connected between the units.

Users of wireless baby alarms should be aware that they are using an unprotected radio service. Models which are also sold in the USA include an FCC statement in the instructions stating that the device must accept any interference received, including interference that may cause undesired operation. The instructions for CE-marked baby alarms may include a statement about possible interference from a nearby transmitter.

Intruder alarms and security lights

If RF triggering of a neighbour's intruder alarm system occurs, the radio amateur is in a strong position technically, particularly if the alarm installation is claimed to meet BS4737 which refers to the "environmental conditions" at the protected premises. As these conditions include "electrical interference", a system which complies with BS4737 should be immune to radio signals from licensed transmitters nearby.

The most likely cause of RF triggering of an intruder alarm is insufficient immunity of PIR (passive infra-red) sensors [5]. These use high-gain operational amplifier circuits whose DC bias conditions may shift slightly when an RF carrier is keyed on or off. Most types use some form of pulse counting and can be set to count two, three or more pulse edges within about 5–10 seconds before sounding the alarm. This also provides some degree of immunity to RF triggering by an FM voice transmission but if the basic RF immunity is insufficient, amateur SSB, CW or packet transmissions will soon exceed the pulse count and trigger the alarm.

It is not advisable to attempt any modifications to PIR sensors in a neighbour's intruder alarm and in any case, improving immunity at VHF may require changes to the PCB layout. Fitting filters or ferrite rings to the cables is unlikely to give a significant improvement in immunity so in most cases, it will be necessary to replace the PIR sensors with a more immune type. If the make and model of the PIR sensor can be identified, it is worth approaching the manufacturer or importer to see whether they have a more immune model which they may be prepared to supply in exchange for the existing PIRs. Even if new PIRs are purchased, these need not be expensive as some low-cost types are available with high RF immunity. Unless the system is a DIY installation, the replacement normally needs to be done by the installer, possibly under a maintenance contract.

PIR sensors used in security lights work on a similar principle to intruder alarm PIRs but generally have lower RF immunity and no pulse counting. On most types, the DC power supply to the electronic circuitry is not isolated from the mains so great care is required if modifications are attempted. Further details on how to improve the RF immunity of PIR security lights is given in reference [6].

SPURIOUS SIGNALS FROM AN AMATEUR TRANSMITTER

The majority of amateur radio EMC problems are caused by the affected equipment having insufficient immunity to the fundamental frequency of transmission but it cannot be assumed that this is true in all cases. If interference occurs due to insufficient suppression of transmitter harmonics or other spurious outputs, it is likely to affect certain specific TV channels or FM broadcast stations. Every receiver within a certain distance is likely to be affected, although this depends on the directional properties of the transmitting and receiving antennas. Spurious emissions may also be generated if an RF power amplifier is unstable or has been incorrectly tuned. In the latter case, frequency halving may occur leading to strong spurious signals at half and 1.5 times the carrier frequency.

Harmonics

Clause 4.(1) of the UK Amateur Radio Licence (A) or (B) *Terms, Provisions and Limitations Booklet BR68* states:

> "The Licensee shall ensure that: (a) the emitted frequency of the apparatus comprised in the station is as stable and as free from Unwanted Emissions as the state of technical development for amateur radio apparatus reasonably permits;"

In the USA, the FCC spectral purity regulations for VHF amateur equipment which reached the market since 1978 require spurious emissions at VHF to be at least 60dB below the level of the carrier (–60dBc) for power levels of 25W or more.

For commercially available amateur radio equipment manufactured or imported into Europe since 1 January 1996, the EMC standard ETS 300 684 specifies levels of emissions and immunity. Section 8.1.3 defines limits for unwanted emissions from the antenna port when the transmitter is active. From 50–1000MHz, the limit is of –36dBm (0.25μW) or –60dBc, whichever is higher.

Even if all harmonics and other unwanted emissions are at least 60dB below the carrier level, this may not be sufficiently low for all situations. In the case of the 50–52MHz band, suppression of the second harmonic at 100–104 MHz is particularly important. For example, even with only 10 nanowatts ERP of second-harmonic power, equation 6.1 shows that at a distance of 10m this would produce a field strength of 70μV/m or 37dB(μV/m). Within the service area of an FM broadcast transmitter, the field strength should be at least 54dB(μV/m) at a height of 10m above ground. Towards the edge of the service area, a spurious signal of 37dB(μV/m) within the passband of the FM broadcast receiver would be more than enough to cause noticeable interference. Clearly, effective low-pass filtering together with a second harmonic trap is required for 50MHz. It is also important to avoid any radiation of second harmonic from sources other than the antenna, for example from a power amplifier with insufficient screening or decoupling.

It is advisable to identify all FM broadcasts between 100 and 104MHz which are intended to serve your area and to check that you can transmit on or near half the frequency without causing interference. If sufficient suppression of the second harmonic is still not achieved after all possible steps have been taken, it will be necessary to avoid transmitting on or near certain frequencies.

In the case of the 144–146MHz band, the fourth harmonic falls at 576–584MHz in UHF TV channels 34 or 35. The fifth harmonic falls at 720–730MHz and could interfere with UHF TV channels 52 or 53. In areas where the above UHF channels are used, additional low-pass or band-pass filtering may be required at the amateur transmitter.

However, it should be noted that, if a sufficiently high level of amateur signal is received by a TV antenna amplifier or the front end of a TV or FM receiver, this can cause harmonics to be generated *within the antenna amplifier or receiver* even though the amateur signal itself may be free of them.

Frequency synthesiser lock-up

When constructing a synthesised transceiver or modifying synthesised ex-PMR equipment, attention should be paid to the lock-up characteristics of the synthesiser, particularly if the transceiver is to be used for packet radio where it is likely to transmit frequently for short periods. With most VHF transceivers, the PLL (phase-locked loop) synthesiser must get into lock on switching from receive to transmit or vice versa. While it is locking, RF drive to the transmit amplifier chain must be inhibited for long enough to avoid a full-power transmitted signal sweeping rapidly across other frequencies inside or outside the amateur band.

In a synthesised transceiver, out-of-lock detection should also be considered. If the VCO (voltage-controlled oscillator) cannot reach the frequency required by the synthesiser IC, the loop fails to lock, resulting in an unstable transmission on an incorrect frequency. There are two reasons why a synthesiser may fail to lock. The first is if the VCO frequency range is incorrectly adjusted and the synthesiser is programmed for a frequency which the VCO cannot reach. The second, which could affect a serially programmed synthesiser, is if the microprocessor fails to program the synthesiser when switching to transmit, leaving the synthesiser programmed to an incorrect frequency. To protect against such error conditions, the out-of-lock condition should be detected and should inhibit transmission. Frequency synthesiser ICs intended for use in transceivers normally have a lock detector output but devices such as the Philips TSA 6057 which are intended for receive-only applications have no lock detector which makes their use in a transceiver inadvisable.

INTERFERENCE TO AMATEUR RECEPTION

In most cases where nearby electronic equipment causes RFI on amateur bands, the equipment complies with any RFI standards which were required at the date of manufacture and the owner is not obliged to take any action unless reception of a protected service UHF TV broadcasting or FM radio is also affected. There is, however, a possibility that a fault has developed, that screening/suppression components have been removed during servicing or that the equipment was not intended for the European market and was imported by the owner.

The first thing to check is whether any interference can be seen on TV or heard on FM radio when using a satisfactory receiving antenna within the intended service area of the broadcast transmitter. If so, the matter can be referred to the local office of the UK Radiocommunications Agency using form RA179 (see below). In such cases, it is worth trying to locate the source so that the unpaid service can be used.

In most cases, only amateur bands are affected and the RFI can only be reduced if the owner of the equipment in question is prepared to co-operate, so a diplomatic approach is recommended. In any case, it is difficult to be certain of the source unless the owner is prepared to co-operate in doing tests. Any RFI reduction should be restricted to measures which can be fitted by the owner without the need for you to touch or dismantle the equipment in question.

It is worth trying to find out details of the make, model number and date of purchase of the equipment so that a complaint can be made directly to the manufacturer or importer. A polite and technically well-informed approach is recommended when dealing with manufacturers. The most effective approach is to phone first to find out the name of the person responsible for EMC then follow up the phone call with a letter, fax or e-mail. It is also worth finding out whether a newer model with reduced RFI is available. In some cases, the manufacturer may be prepared to exchange the equipment in question for a newer model at a reduced price.

In cases where it is not possible to reduce the RFI, cancellation techniques may be used. See for example, reference [7].

RF emission standards

Radio amateurs might wish for nearby electronic equipment to be so well screened and suppressed that it emits no detectable signals in any amateur band but existing RF emission standards fall far short of this ideal. At VHF, emission standards were designed primarily to protect broadcast radio and TV reception with an outdoor antenna at a distance of 10m from the source of the emission. Consequently, they allow levels of RFI which are far higher than radio amateurs would like. Nevertheless the situation has improved for electronic equipment manufactured since 1 January 1996 because previously many types of electronic equipment were not required to meet the relevant emission standards in the UK.

Radiated emissions above 30MHz are covered by BS EN 55 022 which applies to information technology equipment. This not only includes computers but also other equipment containing microprocessors. Various other standards such as the Generic Standard EN 50081-1 are based on EN 55022. The EN 55022 Class 'B' limits are specified as a field strength of 30dB(μV/m) at a distance of 10m over the range 30–230MHz, increasing above 230MHz. This is a very large signal compared to the minimum discernible signal in VHF/UHF amateur bands, but in practice the situation is seldom as bad as it might appear because emissions near the limit are only likely to be found at a few frequencies and in most cases these are not in an amateur band. Further information on RF emission limits in relation to received amateur signal levels can be found in reference [8].

For a VHF/UHF amateur station with a high-gain antenna and low-noise preamp in a quiet rural area, the MDS (minimum discernible signal) for 144MHz SSB corresponds to a field strength of about –30dB(μV/m). In an urban area, the MDS is likely to be higher on certain beam headings due to man-made broad-band noise sources. At certain spot frequencies, there may be significantly stronger narrow-band signals from nearby electronic equipment.

Computer RFI reduction

Although many laptop computers generate little RFI at VHF, other types of computers and associated equipment can be a major source of RFI. The RFI reduction measures described below are primarily intended for desktop PC-type computers used in the radio shack but similar principles can be applied to other types of computer and to digital electronic equipment in general. Clearly, internal modification of a computer or monitor should not be attempted on someone else's computer or one which is still under guarantee. Further details of computer RFI reduction are given in reference [9].

If a computer is used in the radio shack, the antenna and computer should obviously be as far apart as possible. It is also worth checking that the RFI disappears when a dummy load is plugged into the antenna socket, proving that it is not getting into the radio by some other route.

A computer may contain many different clock oscillators, for example for the CPU, graphics controller, disc drives, keyboard, mouse etc. Most of these oscillators are divided down to lower frequencies which can in turn produce many other harmonics. Some frequencies commonly used in digital electronic equipment include 4, 6, 8, 12, 16 and 24MHz, and unfortunately all of these have harmonics at 144.0MHz. Another frequency found in virtually all PC-compatible computers is 14.318MHz (±25kHz or so). This is four times the NTSC colour TV sub-carrier and may be used to synthesise other frequencies such as the clock for the CPU and the pixel clock for the graphics card. For 486 and Pentium processors, a clock frequency of around 30 or 33MHz is commonly used and is multiplied up on the CPU chip to 66, 90, 100, 133 or 166MHz etc.

Reducing RF leakage

Unwanted emissions may escape from a computer or other digital equipment via a number of routes so a step-by-step approach is normally required. It is best to monitor the level of RFI with an indoor antenna 2–3m from the computer, using an SSB receiver with an S-meter if possible. Displaying a graphics screen with a lot of fine detail is recommended as a 'worst case' test even when testing without the monitor. The starting point is to unplug the video lead from the computer, switch off the monitor and unplug the keyboard, mouse and all other interface cables. Any remaining RFI is likely to be a radiated emission due to insufficient screening of the case or a conducted emission via the mains cable. The latter is more likely to affect the HF bands than VHF.

On a CE-marked computer, the case normally has several features to improve screening as shown in Fig 6.6. Arrows 'A' indicate lugs or 'pips' at intervals of about 50mm on the base or cover to ensure good electrical contact at many points. Holes for unused disc drive bays are usually filled with a metal blanking plate ('B'). The wires to LEDs and switches on the front panel come through the metal case ('C') and could cause a slight leak. In practice, however, the shielding of the case may not be the limiting factor so a metal case without these features may be adequate if there is good electrical contact between the lid and the base all around the joints. If not, it may be necessary to add extra fixing screws.

Keyboard

On a PC, the keyboard cable can radiate RFI if the shell of the keyboard connector is not solidly grounded to chassis. Although PC main boards nearly always use four or more layers with power and ground plane, there can still be a small RF potential difference between 'ground' at the shell of the keyboard socket and ground at the back of the case. The fixing hole on the main board nearest the keyboard socket ('D' in Fig 6.6), should be grounded via a metal pillar to minimise common-mode emissions. On CE-marked PCs, the shell of the keyboard DIN socket is normally grounded directly to the back of the case using four spring fingers ('E' in Fig 6.6).

Another way of tackling common-mode emissions is by means of a common-mode choke using a ferrite core on the cable ('F' in Fig 6.6). Most keyboard cables already have one of these moulded on but with only one 'turn' – they introduce a series impedance of only about 100–200Ω. Up to four times as much impedance can be introduced by threading the cable twice through a clip-on ferrite core with 13mm inside diameter.

In some cases, the keyboard itself may radiate a harmonic in an amateur band. As the clock normally uses a ceramic resonator rather than a crystal, it may be possible to move the

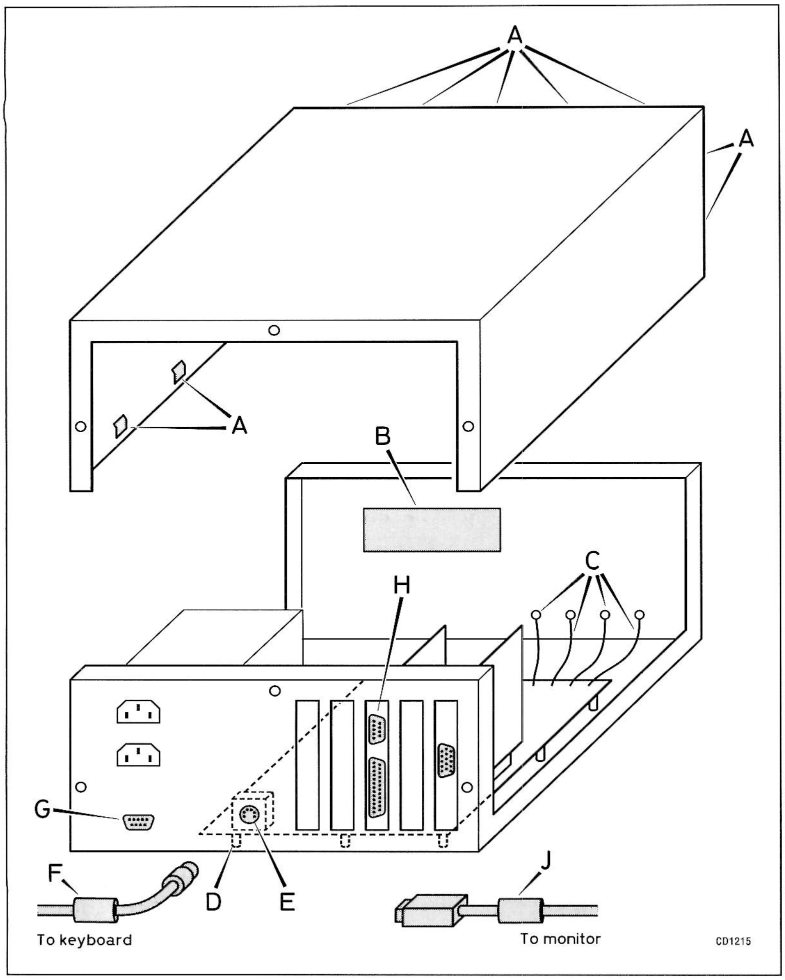


Fig 6.6. A typical PC case showing features which can affect EMC

harmonic out of the band by 'pulling' the frequency of the resonator by increasing its loading capacitors. If this is not successful, it may be necessary to substitute a new keyboard as it is not easy to screen a keyboard effectively.

Serial and parallel I/O ports

Any unscreened serial or parallel interface cable can radiate RFI due to unwanted coupling inside the computer. Where possible, a screened cable should be used with the screen well grounded to the metal connector shell. For a computer mouse, however, unscreened cable is normally used and a clip-on ferrite choke or a filtered connector may be required. A filtered 'D' type connector can be made by soldering a 1nF or 4.7nF ceramic capacitor from each pin to the connector shell. Filtered 'D' connectors or adaptors with built-in feedthrough capacitors are available ready made from suppliers such as RS Components or Farnell Components but are relatively expensive.

A filtered connector on a cable will only be effective if there is good electrical contact between the two halves of the connector shell which requires 'dimples' on the male 'D' type connector. The shell of the plug on the computer must also be well grounded to the case of the computer. This is likely to be true of connector 'G' in Fig 6.6 but connector 'H' may not be well grounded because it is on an expansion card.

For a computer with an internal modem, the unscreened telephone cable can radiate or pick up RF, especially if connected to an overhead telephone line. The only feasible way of filtering the telephone line where it leaves the modem card is by means of a ferrite ring or clip-on choke. Connecting series inductors or shunt capacitors directly to the telephone line is not permissible as it would invalidate BABT approval for the modem.

Graphics card

It is important that the shell of the video connector on a PC graphics card is well grounded to avoid a VHF common-mode emission from the braid of the video cable. The shell of the video output connector is grounded to the mounting bracket but this is only grounded to chassis with one bolt at the top. If possible, the graphics card should be fitted in one of the end slots where grounding to the back panel is better. Some good-quality cases for CE-marked PCs have spring fingers around the expansion slots on the back panel to improve grounding but, if these are not fitted, RF grounding can be improved by bolting or clamping the mounting bracket of the video card to the back panel close to the connector. Another way of reducing RFI from a video cable is to clip on a ferrite choke ('J' in Fig 6.6) if one is not already fitted.

Computer monitors

Following the above steps should lead to a fairly quiet computer at VHF, until the monitor is switched on. RFI from a monitor may have definite peaks at certain frequencies or may be broad-band noise. There are large variations in the amount of VHF RFI emitted by different models of computer monitor. To reduce RFI from a monitor at VHF, it is normally necessary to make internal modifications to circuitry which operates at a high voltage. Such modifications should only be attempted by those with sufficient experience. Note that ferrite is conductive and any ferrite cores should be fitted so as to avoid causing a short-circuit or flashover.

In many computer monitors, the switching frequency for the power supply is synchronised to the line timebase frequency which can make it difficult to distinguish between power supply harmonics, line timebase harmonics and video amplifier harmonics. If turning the brightness and contrast right down give a large reduction in RFI, this could be due to reduced loading on the power supply, reduced loading on the line output stage or reduced output from the video amplifiers. The switch-mode power supply in a monitor may be on an unscreened PCB in a plastic case leading to direct radiation, in which case adding additional mains filtering is unlikely to give much improvement at VHF.

The line output stage can radiate harmonics up to VHF due to the fast switching of the output transistor. As the collector of the line output transistor is usually connected directly to the 'hot' side of the line scan coils, these coils and associated wiring can act as a radiating antenna. The only effective way of reducing VHF harmonics of the line timebase is usually to cut the track to the collector of the line output transistor and put in a series choke consisting of 2–3 turns on an FX1115 ferrite bead. As the added inductance in the collector could alter the operating conditions of this highly stressed device, such modifications should be regarded as experimental. A similar modification can be applied to the main switching transistor in the switch-mode power supply if necessary.

If the video amplifiers are radiating, displaying a screen with a lot of fine detail will produce more RFI than a plain screen of the same brightness. The video output transistors are nearly always mounted on the base if the CRT on a small PCB with a tinplate screen. This screen should be grounded to the chassis by two short lengths of braid. The video cable from the computer should have its braid solidly grounded to chassis where it enters the monitor. Some CE marked monitors have two large ferrite beads on the video cable, one each side of this ground point.

Although it would be possible to coat the whole inside of a monitor's case with nickel RF shielding spray grounded to chassis, this is not recommended for several reasons. First, there are high voltages on the PCB which may flash over to any conductive coating inside the case. Secondly, the conductive paint may find its way through ventilation slots so that it can be touched from outside. This presents a shock hazard if part of the coating inside comes into contact with a high voltage. A third problem is that the coating may not adhere well to certain types of plastic unless a special primer is used. If the coating flakes off, this could cause short circuits.

Even the software set-up of a PC can affect the emissions from a monitor. For example some Cirrus Logic VGA video cards operating in the 800 × 600 resolution modes use a pixel clock frequency of 36.088MHz or 72.176MHz. This produces a second or fourth harmonic at nominally 144.352MHz with sidebands either side. A VGA utility program such as CLMODE can be used to demonstrate the various graphics modes and identify any differences in RFI.

Other RFI sources

There are many possible sources of unwanted signals in the VHF and UHF bands apart from computers and associated equipment. Other domestic products which generally incorporate a microprocessor include intruder alarms and fax machines. Satellite TV receivers and decoders, video recorders and TV sets with NICAM stereo or digital signal processing can also generate RFI in the VHF bands. Many consumer products use a ceramic resonator rather than a crystal in the clock oscillator for the digital circuitry. Harmonics of ceramic resonators tend to drift with temperature and they may even be microphonic where nearby sound or other vibrations cause slight frequency modulation.

Some types of heating thermostat may develop a fault causing them to arc for several seconds or even tens of seconds when the contacts open. This may occur every few minutes and tends to occur more frequently in cold weather. In some cases, the same type of thermostat may have been installed in a number of houses and there could be several arcing thermostats.

Equipment in nearby commercial or industrial premises which may radiate RFI at VHF includes arc welders, computers and computer networks using unscreened twisted pair (UTP) rather than coaxial cable or optical fibre. Some types of fire alarm systems can also be a problem, particularly the 'analogue addressable' type if these are wired with unscreened cable.

A broad-band noise source which can affect all or part of the 70cm band is a super-regenerative receiver on 433.92MHz. With a high-gain 70cm antenna and low-noise preamp, some types within 30–50m can cause a substantial degradation of signal-to-noise ratio. Noisy 433.92MHz super-regenerative receivers are found on some after-market car alarms manufactured in 1994 and 1995. Receivers manufactured from 1996 onwards have to meet ETS 300 220 and are therefore much quieter. Some particularly poor 173MHz garage door openers sold in the late 'eighties also produced high levels of noise on 70cm and up into the UHF TV band.

RFI from noisy super-regenerative receivers can take two forms. In the absence of a signal, they radiate broad-band noise, possibly covering tens of megahertz. When the super-regenerative receiver detects a carrier somewhere near its operating frequency, its emission changes to a number of discrete frequencies spaced at intervals of the quench frequency which may be around 800kHz. These emissions drift and are modulated by signals from other radio services such as radio paging. They can give the misleading impression that a paging transmitter has spurious outputs or that an amateur receiver has spurious responses.

Other radio users

If signals from other radio services such as radio paging or PMR are heard on a receiver tuned to an amateur band, a likely cause is a spurious response in the amateur receiver or overloading of any preamplifier. In particular, some 144MHz amateur transceivers with extended receive coverage can be susceptible to breakthrough of nearby radio paging signals at around 138 or 153MHz. This problem can often be cured by means of a 144–146MHz bandpass filter or a notch filter tuned to the pager frequency [10].

It is possible for radio paging transmitters to develop a fault which produces a number of unstable spurious frequencies either side of the carrier and, if this occurs, the interference typically drifts up or down the amateur band. Note, however, that similar symptoms can be produced on 70cm by some types of super-regenerative garage door receiver nearby reradiating pager signals.

If it is suspected that another radio service is radiating spurious signals in an amateur band, it is advisable to obtain conclusive proof of this before proceeding further.

DEALING WITH NEIGHBOURS

With some amateur radio EMC cases, the technical problem is easy to solve but applying the solution is difficult because of a social problem. If relations deteriorate too far, even a simple matter like getting a plug-in filter fitted could become a major issue. *It is therefore well worth trying to maintain friendly relations even if the neighbour's initial approach is unfriendly.* The neighbour's point of view may be that they have bought a good-quality product which works perfectly well when the radio amateur is not transmitting, so they blame the amateur. The radio amateur's point of view is that he or she is operating within the terms of the amateur licence so the problem is caused by shortcomings in the neighbour's equipment.

To explain your point of view to a neighbour may not be easy and a diplomatic approach is called for. If you take the view that it is not your problem, there is a risk of a much bigger problem later on! Even if your station is 'in the clear' technically, an unco-operative or, worse still, an alienated, neighbour could make life very unpleasant. Some radio amateurs have even resorted to moving house in such a situation.

It is wise to be prepared for the possibility of a breakthrough complaint before it happens. First of all, make sure your own house is in order by solving any EMC problems with your own domestic electronic equipment as far as possible. Being able to show that your TV/video recorder/hi-fi/telephone does not suffer breakthrough when you are transmitting should convince anyone that your transmitter is not at fault. Interference-free radio and TV reception in your own house is also an additional check that any spurious outputs from your transmitter are adequately suppressed. It does not prove this conclusively, however, due to the directional properties of transmitting and receiving antennas.

Solving any breakthrough on your own domestic electronic equipment is also good practice and means that you will probably have a selection of suitable filters or ferrite rings to hand. Even if none are required for your own equipment, it is advisable to keep an 'EMC first aid kit' consisting of at least one suitable TV filter together with a few ferrite rings. It is also worth having at least one RF immune telephone available even if your own telephone is never used while you are transmitting.

If a neighbour reports a problem, this could be your only chance to negotiate so *great diplomacy is necessary*. If there is any doubt about whether your station is the cause of the problem, you could ask the neighbour to keep a written log of dates and times when breakthrough occurs but they may be unwilling to do this so it is worth offering to conduct test transmissions immediately. If possible, the breakthrough should be solved promptly using a filter which you already have.

You are under no obligation to pay for filters for neighbours' TVs etc but in many cases, the neighbour is unwilling to pay, so it is in the interests of good relations to provide a filter on loan for as long it is needed. A small neat label with your name and address makes the point that it remains your property rather than being a gift (which might be taken as an admission of liability). It may also reduce the chance of a TV service engineer taking it away.

REGULATORY ISSUES

The Radiocommunications Agency of the DTI produces a leaflet RA234, *EMC and the Radio Amateur*. This is available from the RA Document Distribution Centre and also via the World Wide Web (http://www.open.gov.uk/radiocom/ra234.htm). RA234 states the following:

> "**What is EMC?**
> EMC, short for electromagnetic compatibility, is the capacity of equipment to function without causing excessive interference and without being unduly affected by emissions from other apparatus.
>
> **Why is EMC important?**
> Amateurs are privileged in being allowed to operate at high power levels in residential areas. This privilege brings responsibility. Interference can be immensely annoying. As a responsible amateur you will naturally take care not to interfere with television and radio reception, for example. Apart from general considerations of good neighbourliness, there are conditions in the Amateur Radio Licence on interference. In addition from 1 January 1996 an EC Directive will impose new EMC standards on virtually all electrical and electronic equipment.
>
> **Does the Directive apply to amateur equipment?**
> Self-built amateur equipment is not covered by the Directive but it will still be necessary when using it to abide by the Licence conditions on interference. Commercially available products will have to comply and carry the CE mark to show compliance.
>
> **What happens if an interference problem arises?**
> If a problem arises, as a first step, the amateur should check that his or her own equipment is not at fault. Poor immunity is often to blame for reception problems and it may be necessary to take steps to improve the immunity of the affected installation. The amateur should co-operate with the neighbour and/or the dealer to identify and resolve the problem. But, if this does not work, the Radiocommunications Agency is likely to become involved.
>
> **What happens then?**
> The Agency is empowered to vary the amateur's permitted power so that the amateur does not cause excessive interference. Before resorting to this, the Agency will take all relevant circumstances into account, including the immunity of the affected installation. In the final analysis, however, the Agency will be guided by the immunity required by the relevant European Standard. If poor immunity is not to blame and other steps to reduce interference have failed, the amateur may be required to take steps to stop the field strength exceeding the level that the relevant European standard requires the affected installation to be able to withstand."

Another useful RA publication is RA323, *Guidelines for Improving Television and Radio Reception*. It consists of 16 pages plus a colour section with photos showing various types of TV interference. It is primarily intended for radio and television dealers, service engineers and antenna installers rather than for the general public. Topics of particular relevance to amateur radio include TV antenna amplifiers, CE marking and effects due to lack of immunity.

Radiocommunications Agency involvement

The following information is believed to be correct at the time of writing (late 1997) but may be subject to change.

If a UK householder experiences a reception problem with UHF television, a video recorder or FM radio, they can refer the matter to the RA using form RA179, *Advice on Television and Radio Reception*. It is useful to keep an up-to-date copy of RA179 in case of a complaint from a neighbour. Copies are available from the Radiocommunications Agency Document Distribution Centre but are no longer available from Post Offices. RA179 is only applicable to domestic complaints. If a businesses is affected, there is a different procedure, details of which are available from the local offices of the RA.

Part A of form RA179 is used when reporting a known or suspected source of interference for possible investigation by the RA. There is no charge for reporting a source but the RA does not visit the complainant to investigate the affected equipment. If a radio amateur is nominated as a source of interference, it is likely that the local officers of the Radiocommunications Agency (previously known as the 'Radio Investigation Service' or 'RIS') would visit the amateur's station. Such a visit may include checking for spurious emissions and ensuring that the station is being operated within the terms of the Amateur Licence. In some cases, the field strength produced by the amateur station could be measured.

Part B of form RA179 is used by a householder to request a visit from the RA to investigate the affected equipment. There is a charge (£45 at the time of writing) which includes the supply of any necessary filters.

Form RA179 states that the paid service is only available for UHF televisions, video recorders or FM radios. It does not cover long-wave and medium-wave radios, satellite TV, cable TV, telephones, fax machines or answering machines. Neither does it cover other equipment not intended to pick up radio such as record players, CD players, tape recorders, electronic keyboards, baby alarms, computers or monitors. PIR security lights and intruder alarm systems are not specifically mentioned but appear to come under the category of 'other equipment'.

REFERENCES

[1] *The Radio Amateur's Guide to EMC*, Robin Page-Jones, G3JWI, RSGB, 1992.

[2] *RSGB Guide to EMC*, Robin Page-Jones, G3JWI, RSGB, to be published 1998.

[3] *RSGB Yearbook*, EMC section.

[4] 49MHz baby monitors item in 'EMC' column, *Radio Communication* June 1996, p74.

[5] Alarm PIR sensors item in 'EMC' column, *Radio Communication* December 1994, pp75–77.

[6] PIR lights item in 'EMC' column, *Radio Communication* April 1994, pp76–77.

[7] 'Two metre interference reduction system', T Day, G3ZYY, *Radio Communication* April 1992, pp48–50.

[8] RF emission standards item in 'EMC' column, *Radio Communication* June 1995, pp76–77.

[9] Computer RFI reduction items in 'EMC' column, *Radio Communication* December 1996, pp77–78, and February 1997, pp 80–81.

[10] 'Intermod – A modern urban problem', E Hare, KA1CV, *QST* August 1996, pp40–43.

7 Data modes

DATA modes can mean a variety of modulation modes, from Morse (which is arguably the simplest form of data transmission using on-off keying) to advanced error correcting (eg packet radio) and direct-sequence (eg spread-spectrum) modes. The commonly used modes are:

1. Data networking, eg AX25 packet and TCP/P.
2. Weak-signal modes, eg Morse, PACTOR, CLOVER etc. It must be said that, with the exception of Morse, these are much less used at present at VHF/UHF than at HF.
3. Digitised speech, image and multimedia. This series of modes are still in their infancy at the time of writing (late 1996) and will not be dealt with further.

DATA NETWORKING

AX.25 packet radio

The overwhelmingly most popular data mode on VHF/UHF at present is packet radio, used with an FM transceiver. Unlike weak-signal modes such as PACTOR, CLOVER etc (see later) as well as visual modes such as fax and SSTV, AX.25 packet uses a amateur protocol derived from X.25, defining the content, format and the handling of packetised data. Information on AX.25 and other amateur packet protocols is extensively covered in other RSGB publications [1, 2] as well as the ARRL Computer Network Conference documents, so only a brief description is given here, concentrating instead on the 'physical layer' and user operation.

HDLC

Packet radio uses high-level data link control (HDLC) to handle the forwarding of error-free frames of data over a communications link. The transmission is based upon a series of 'packets' of data, each packet containing a portion of the transmitted information preceded by routing information and ending with a cyclic redundancy check (CRC), which is a value calculated by the sending station based upon the content of the information in the transmitted packet.

At the receiving end, each packet is automatically checked for correct and valid information content. If a valid packet is received, an 'ACK' (acknowledgement) is sent by the receiving end station to acknowledge that the packet has been received without errors. Otherwise, the transmitting station automatically repeats the transmission of that data packet.

By the use of this protocol, a number of packet radio stations may all use a common frequency, with individual data packets 'interleaving' with others as required. Packet addressing, by callsign or a short 'alias', ensures that the desired receiving station is correctly addressed, with each station ignoring packets not intended for that station's callsign or alias.

TCP/IP

A progression from AX.25 is TCP/IP, also used in other data systems such as the Internet. Like AX.25, TCP/IP (Transmission Control Protocol / Internet Protocol) is similarly used for packet processing, formatting and routing. However, TCP/IP differs from AX25 in that it automatically adapts to system network delays, lengthening transmission delay times accordingly to provide the best overall system throughput performance.

The various protocols in TCP/IP include:

FTP (File Transfer Protocol) – used to exchange data or binary files with another FTP user on the network.

Telnet – allows person-to-person keyboard communication, together with remote log-in facilities to any other computer system on the network.

SMTP (Simple Mail Transfer Protocol) – used to send messages to other TCP/IP users on the network, the message being stored locally with SMTP automatically feeding it to the network for onward forwarding.

POP (Post Office Protocol) – a mail handling facility which is a variant of SMTP, where a remote system is used as a 'post office' for the storage and onward transmission of mail.

PING (Packet INternet Groper) – used to check if a given remote user is on air, and the 'return trip time' for a response.

A packet TNC is used in 'KISS' (ie 'Keep It Simple, Stupid') mode for TCP/IP, with appropriate software running on the PC. This system is *NOS* (Network Operating System), originally developed by Phil Karn, KA9Q – subsequent derivations have included *GRINOS, JNOS, TNOS* and *MFNOS*. Before you can use TCP/IP on the air, you'll need to contact your local TCP/IP co-ordinator for a numeric 'address', which consists of a four-part number, for your station.

TERMINAL NODE CONTROLLER

To handle the AX.25 protocol, as well as providing a platform for TCP/IP, a terminal node controller (TNC) is typically used. This is a self-contained unit comprising a modem and microprocessor-based control system with the operating firmware typically stored in a plug-in erasable programmable read-only memory (EPROM) IC. This allows firmware changes or upgrades to be accomplished as needed.

For the 'user interface', the TNC has an RS-232 serial data connector for connection to a terminal. This may either be a 'dumb terminal', a computer operating in terminal emulation mode, ie running the Windows Terminal program, or a computer running a dedicated packet radio program to give added operating features such as automatic connections, logging,

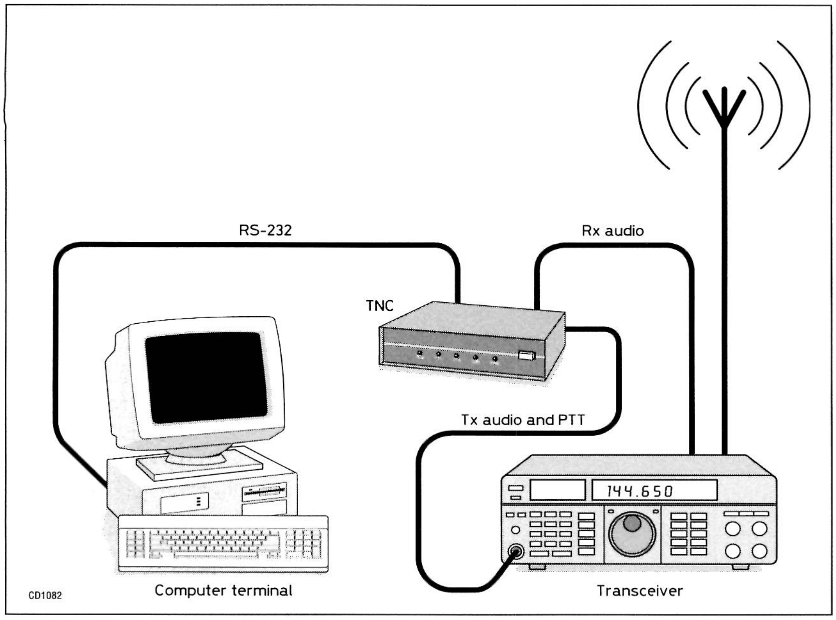


Fig 7.1. Packet TNC connections

mailbox facilities etc. For personal use, many TNCs also have a self-contained message 'mailbox' to provide storage facilities of personal messages to and from the TNC user.

For the radio interface, the TNC connects directly to your VHF/UHF transmitter/receiver although, depending upon the packet radio data rate, differing connections are required. Typical packet data rates used by individual amateurs are 1200 baud and 9600 baud, although 'backbone networking' data rates typically use 9600 baud, 56 kbaud, 64 kbaud, or higher speeds still (eg 2Mbits/s) on the microwave bands. Note that the terminal baud rate you select for use between the TNC and your terminal does *not* affect the radio baud rate, the latter being dependent upon the type and speed of modem used within the TNC.

See Fig 7.1 which shows these connections diagrammatically.

Multiple users

The TNC receives audio from the packet station receiver, and on a given frequency used by a number of packet stations this is usually a sequential combination of many individual packets from differing stations, as well as packets from individual stations intended for a number of other 'connected' stations. Up to 26 different connection 'streams' can be handled by each TNC. In 'monitor' mode, the TNC decodes all received packets, and transfers the decoded information to the RS-232 terminal port for subsequent display and optional processing. However, in communication or 'connected' mode, it still decodes all received packets but typically only transfers to the RS-232 port packets addressed to the callsign or alias the user has manually stored (again via the terminal) into the TNC. It also automatically waits until the frequency is clear before transmitting a packet, to prevent interference to other packet stations on the same frequency. Pseudo-random 'wait' timings prior to transmission give a degree of protection in preventing 'collisions' of packets due to simultaneous transmissions from different stations on a given frequency.

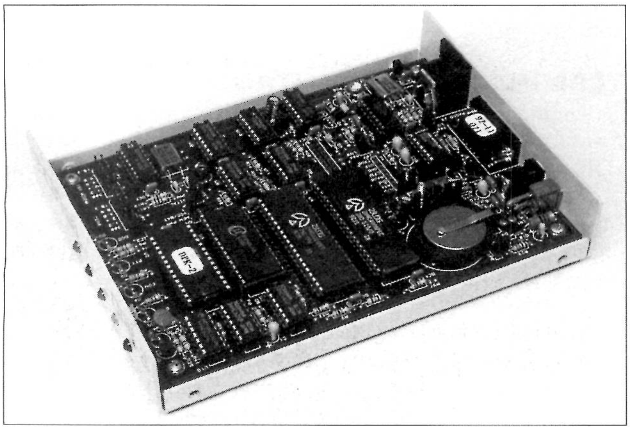


Fig 7.2. Here's what's inside a TNC

HARDWARE

TNC kits of parts are available from groups such as MAXPAK in the UK to allow you to build a TNC-2 'clone', ie a 'generic' TNC which uses a plug-in EPROM common to many other TNCs. The TNC-2 clone is the type required if you intend to substitute the normal plug-in user EPROM with EPROM-based firmware for dedicated network node operation.

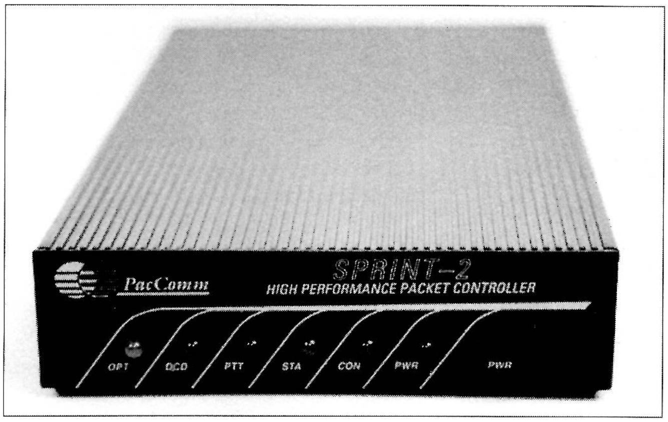
A wide variety of commercial ready-built TNCs are of course available. Many of these are based upon a TNC-2 clone (eg some PacComm and AEA TNCs) whilst others use proprietary firmware to provide additional features (such as some Kantronics TNCs).

TNC connection and operation

The TNC requires a suitable RS-232 connection link to your terminal for operation. Note, however, that some TNCs may employ otherwise unused connections on the 25-way D-type connector for other purposes, eg alternative supply voltage input, test points etc. If in doubt, follow the instructions supplied with your TNC. Tables 7.1 and 7.2 show the most-commonly used connections required for typical TNCs, and Figs 7.6 and 7.7 the serial port connector wiring diagrams.

It is beyond the scope of this chapter to describe the many TNC operational commands due to the wide variety of firmware versions available (see your TNC firmware manual or a dedicated packet radio handbook [1, 2] for these), although TNCs commonly use a pre-defined set of LED indicators on the front panel, which give the operator a degree of information about the status of the TNC at any time. These are:

PWR	Power on
RCV	Off-air signal data being received
XMIT	TNC in transmit mode
CON	TNC in 'Connected' mode
STA	Unacknowledged packets outstanding



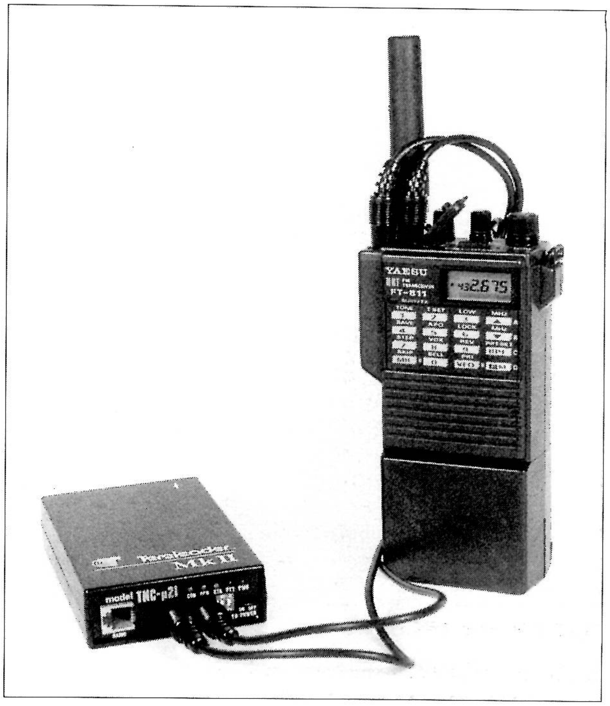


Fig 7.3. Typical factory-built packet TNCs: the DPK-2 (top left), the Sprint-2 (left) and the TNC-µ21 (above)

TRANSCEIVER CONNECTION

For *1200 baud* packet, the TNC transceiver interface may connect simply to the receiver external speaker audio and to the transmit microphone and PTT connections. Unless your TNC has a 'software DCD' incorporated and enabled, ie it has intelligent data detection of packet data to differentiate from receiver squelch noise and other signals, you should ensure that your receiver squelch is suitably adjusted. This is because

Table 7.1. TNC-2 'clone' RS-232 connections (25-pin D type)

Pin	Function	Signal direction
1	Frame Ground (FG)	Common
2	Transmit data (TXD)	PC o/p
3	Receive Data (RXD)	TNC o/p
5	Clear To Send (CTS)	TNC o/p
6	Data Set Ready (DSR)	TNC o/p
7	Signal Ground (SG)	PC o/p
8	Data Carrier Detect (DCD)	TNC o/p
20	Data Terminal Ready (DTR)	PC o/p
22	Ring Indicator (RI)	Not usually connected

Table 7.2. PacComm Tiny-2 connections (nine-pin D type)

Pin	Function	Signal direction
1	Data Carrier Detect (DCD)	TNC o/p
2	Receive Data (RXD)	TNC o/p
3	Transmit data (TXD)	PC o/p
5	Signal Ground (SG)	PC o/p
6	Data Set Ready (DSR)	TNC o/p
7	Request to send (RTS)	PC o/p
8	Clear To Send (CTS)	TNC o/p
9	Ring Indicator (RI)	Not connected
Shell	Frame Ground (FG)	Common

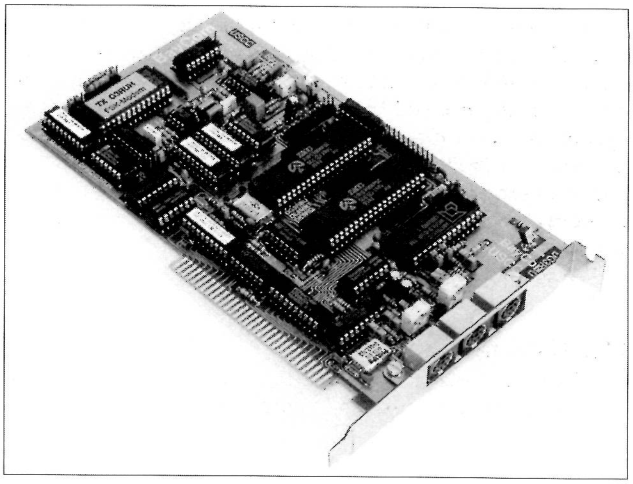


Fig 7.4. A plug-in PC card TNC can be used – this one has modems for 1200 and 9600 baud operation

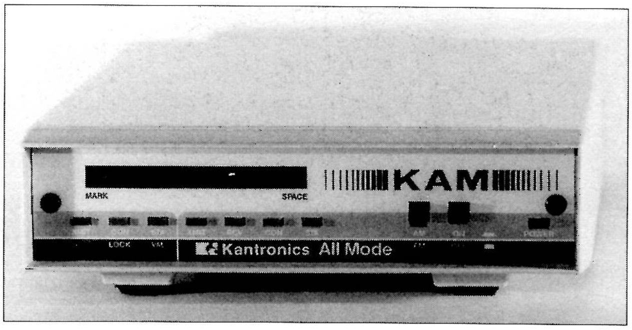


Fig 7.5. The KAM is a popular multimode data controller

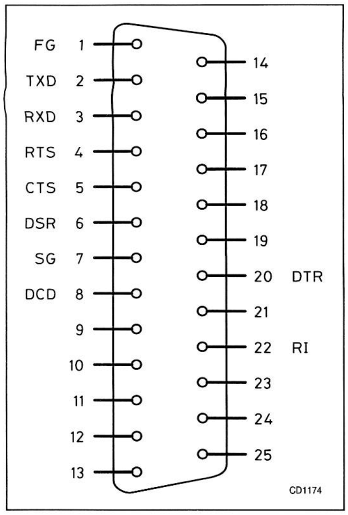


Fig 7.6. IBM PC serial port 25-pin connector

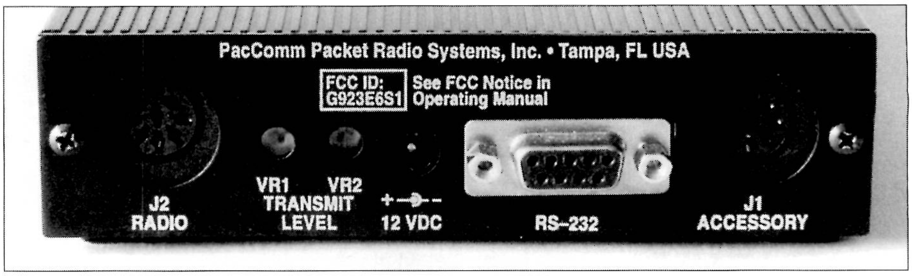


Fig 7.9. Rear-panel connections on the TNC interface to your radio and RS-232 terminal

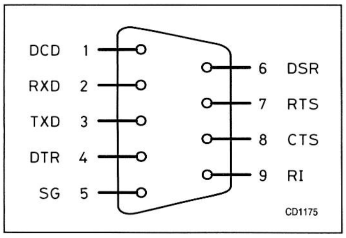


Fig 7.7. IBM PC serial port nine-pin connector

the TNC will not otherwise transmit if the TNC's front panel 'DCD' LED is illuminated, this being controlled by the detection of a received signal by the TNC modem.

9600 baud packet requires connection to the 'flat', ie unprocessed, audio points in your transceiver. On receive, this typically means the receive audio needs to be taken directly from the receiver discriminator, prior to any audio de-emphasis filtering. On transmit, an audio response down to a few hertz, preferably down to DC, is required, together with an essentially 'flat' (ie unprocessed) transmitted frequency response. Some commercially available FM transceivers are fitted with 9600 baud packet data jack connections for this. However, many amateurs also use dedicated low-cost crystal-controlled transceivers, often ex-PMR equipment, for packet radio use, to avoid permanently tying up a high-value commercial transceiver on a given BBS or DX cluster frequency (see the RSGB's *PMR Conversion Handbook* [3] for more details on typical equipment and comprehensive conversion information). Transceiver connections for both 1200 and 9600 baud are shown in Fig 7.8 and a typical FM modulator modification circuit in Fig 7.10.

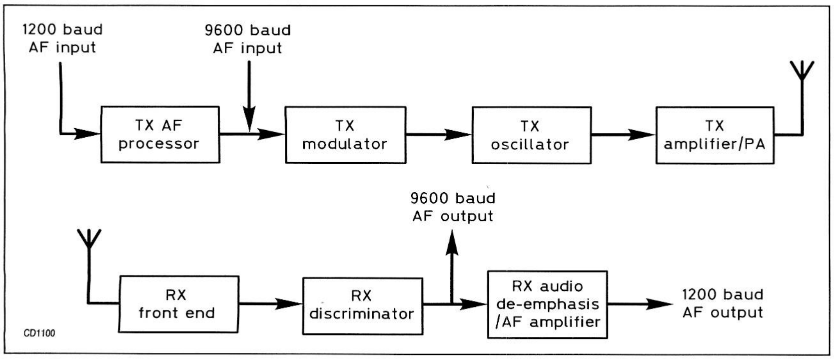


Fig 7.8. 1200 and 9600 baud transceiver connections

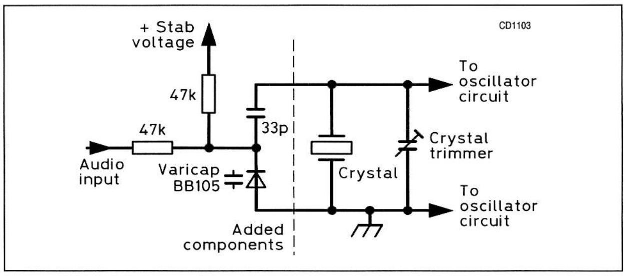


Fig 7.10. Typical 9600 baud direct FM modulator

PACKET DIGIPEATERS AND NODES

For individual one-to-one amateur communication via your terminal you simply issue a 'connect' command to your TNC to directly link to the intended station, who is operational on the same frequency and in communication range of your station. A typical command would be 'CONNECT GB7SMC' (or 'C GB7SMC') if I wished to connect to my local DX packet cluster node.

However, the AX.25 protocol also allows 'digipeating', where a third-party packet station can automatically re-transmit packets by remote command. Thus, 'CONNECT GB7DXW VIA G0SBV' would attempt a link to GB7DXW, using the packet station of G0SBV as an intermediate digital 'relay'. Up to eight intermediate digipeaters may be used for this. Any error-checking here is performed at the 'far end' of the link though, individual digipeaters simply retransmitting the packet information to the next station along. The digipeater facility is an inherent feature in every current packet TNC, although it may be enabled or disabled by the TNC operator as required.

A packet 'node' goes one step further, by employing local error checking and repeat transmission requests. A simple node facility is occasionally also an inherent feature of some TNCs, eg the 'KA-Node' on Kantronics TNCs. One stage further is that of a 'network node', which usually uses dedicated firmware or a PC running appropriate node software such as that written by John Wiseman, G8BPQ.

A network node system has automatic networking abilities, including automatic routing using the best possible transmission quality path between two remote points. A typical network node station arrangement could use multiple transceivers and TNCs, eg operating on 4m, 2m, 70cm and 23cm, with the TNCs locally

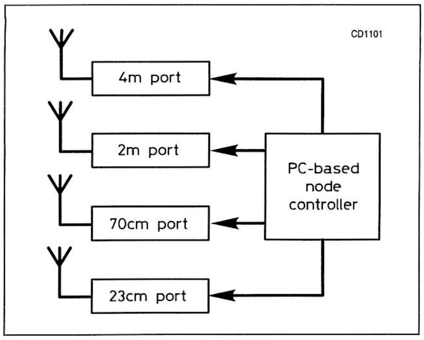


Fig 7.11. Typical network node arrangement

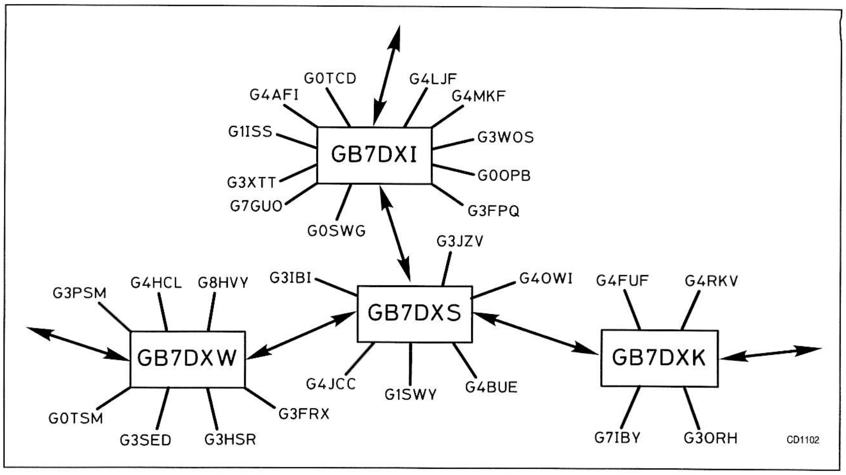


Fig 7.12. Typical DX packet cluster configuration

interconnected either via an RS-232 matrix or with the host computer running appropriate node software (Fig 7.11). In the latter case, the individual TNCs are commanded to run in 'KISS' mode, with the computer handling packet routing.

BULLETIN BOARDS

Although individual contacts are made using packet radio for live 'chats', the most common use of packet radio is by the use of a bulletin board system (BBS). These systems consist of a computer running appropriate BBS software, linked to the packet radio network, usually through individual TNCs and radios at the BBS station.

Two common types of BBSs are in use, the 'normal' BBS and, of greater interest to the DX operator, the DX packet cluster system.

Network BBS

Differing from the 'personal' BBS you may have in your TNC or computer software, which is used for personal messages to and from your station only, a network BBS operates as a linked national and international message storage and forwarding system for third-party messages and bulletins to and from radio amateurs. Some overseas BBS also have facilities for Internet links.

After you have connected to your local BBS, you can then list message titles, either all stored messages or to any given subject, callsign, or whatever, view and download message texts and stored files, as well as uploading files of interest to others, and enter messages addressed to other amateurs. Each BBS is part of the worldwide network, linked via nodes on HF, VHF and UHF, as well as via store-and-forward amateur satellite gateway stations. Thus, you can send and receive messages worldwide, the BBS network routing your messages to the intended recipient. You may read and send general 'bulletins', subject to licensing restrictions, which are intended for general reading, eg to seek help or information on a given subject from other amateur on the packet network.

Each BBS is normally run by an individual, or occasionally by a club, with all running costs being met by him or her and not by any national organisation. This is worth bearing in mind as, due to current UK licensing conditions, the BBS system operator (sysop) usually refrains from directly soliciting donations to pay towards the running costs.

DX packet cluster

A DX packet cluster is a network of interconnected individual DX cluster 'nodes', each of which is located in a given area to serve a local amateur population. Similar in many physical respects to a network BBS, the controlling PC instead runs specialised DX cluster software which has enhancements for DX station activity reporting, propagation information etc. A typical area configuration is shown in Fig 7.12, with local stations each connected to their local DX cluster node, which are in turn linked to each other, and to those further afield (not shown here) either via dedicated RF links or via the national and international AX.25 packet network.

In operation, the users are all part of a shared information resource, where one station can enter an 'announcement', ie a DX 'spot' with DX station details, operation frequency and a short comment, usually when he's heard or worked a station worthy of reporting to others. This announcement is then forwarded throughout the DX cluster system and passed to all other connected stations. A 'filter' is available, where if you wish you can choose to only receive announcement 'spots' relating to certain bands, eg the VHF/UHF bands, you are specifically interested in, or indeed any combination including HF.

The UK is extensively covered by linked DX cluster nodes, which are often also linked to those in other European countries. This system can be an excellent way of obtaining and

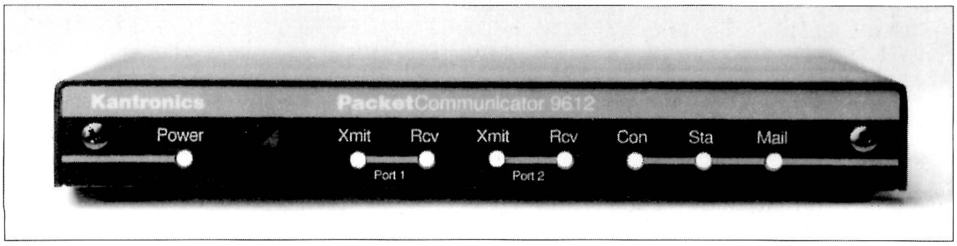


Fig 7.13. This KPC-9612 TNC combines 1200 and 9600 baud packet operation

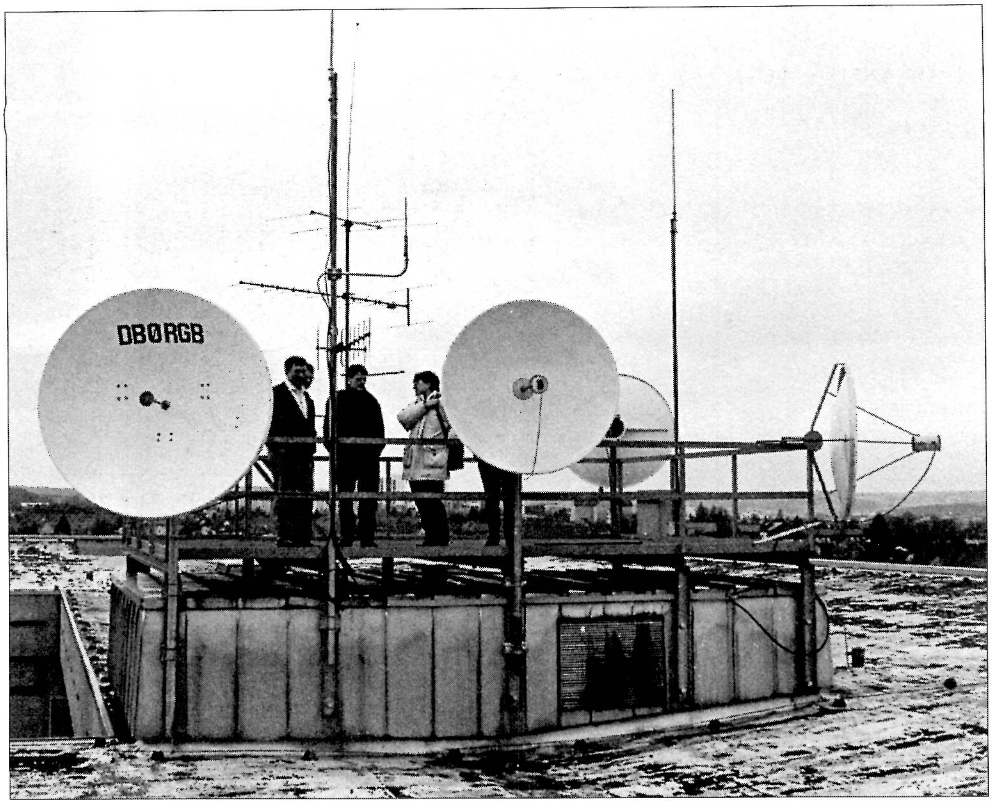


Fig 7.14. The DB0RGB multi-frequency packet radio node site

Typical 2m DX packet cluster 'spots'

144080.0	VE7BQH	3-Nov-1996	0921Z	EME 529	<PA0JMV>
144112.0	DK9IP/P	3-Nov-1996	0918Z	JN48GT	<DF2UU>
144081.0	TM6P	3-Nov-1996	0915Z	JN19pg	<F6HPP>
144015.0	DF0CK	3-Nov-1996	0917Z	cq contest from jo50an	<DG0OPK>
144099.9	F6DJB	3-Nov-1996	0915Z	jn03cl best dx 956 km !	<PA3BAS>
144092.0	G3JRM	3-Nov-1996	0911Z	jo02-jo60 559qsb	<DK0SAX>
144100.0	F5KHG/P	3-Nov-1996	0913Z	ae52b	<PA3BAS>
144067.0	GM0CLN/P	3-Nov-1996	0830Z	io85	<G4WOX>
144096.3	DL0KM/P	3-Nov-1996	0815Z	JO31BC-JN03CF 559 983km!	<F6BOL>
144096.3	F6BOL/65	3-Nov-1996	0812Z	JN03CF-JO31BC 559/559	<DF2JQ>

Typical 6m DX packet cluster 'spots'

50024.0	9H1SIX/B	21-Oct-1996	2128Z	JM75>JN18 559	<F6FLV>
50111.8	9H1AW	21-Oct-1996	2045Z	59 jn08>>jm75	<F1LLS>
50110.0	SP2NJE	21-Oct-1996	2121Z	Es again !	<PA0RDY>
50112.0	9H1AW	21-Oct-1996	2120Z	Alan GW3LDH	<G0OFE>
50112.0	9H1AW	21-Oct-1996	2112Z	jm75>io91 59+	<G4RGK>
50112.0	9H1AW	21-Oct-1996	2053Z	57 >IO83 QSB	<G0JHC>
50112.0	9H1AW	21-Oct-1996	2039Z	JM75>IO92	<G0PQO>
50120.0	F6HTJ	21-Oct-1996	2023Z	JN12 > JO01	<G8RZA>
50112.0	EH1EH	21-Oct-1996	2011Z	IN82 cq cq	<OZ5AGJ>
50110.0	F6HTJ	21-Oct-1996	2009Z	59 JN12	<G0JHC>
50117.0	EH1TA	21-Oct-1996	2010Z	in53>jo32	<PA2TAB>
50125.0	EH7AH	21-Oct-1996	1957Z	IM67-JN67 still 59	<OE2UKL>
50112.0	EH1EH	21-Oct-1996	1935Z	IN82 > JO01	<G8RZA>
50112.0	EH1EH	21-Oct-1996	1925Z	in82>io92	<G4VPD>
50029.5	CT0WW	21-Oct-1996	1920Z	5/5 beacon in61ge	<PA0PAU>
50125.0	EH7AH	21-Oct-1996	1920Z	59 IN JN47	<OE9PTI>
50113.0	EH4EHI	21-Oct-1996	1913Z	59 IM68-IO80 RARE SQUARE	<G4HBA>
50112.8	EH4EHI	21-Oct-1996	1902Z	5/7 im68tv- jo22nw	<PA0PAU>

sharing 'real-time' information on sporadic-E or tropospheric activity with other amateurs, as well as arranging skeds for meteor scatter, EME etc, for example with other European stations.

As with network BBSs, each DX cluster node is usually financed by an individual amateur or local club, which should be borne in mind if you are a regular user of a particular one.

MULTIMEDIA DATA COMMUNICATION

The near-universal use of PCs for amateur data communication together with increasing PC processor speeds has significantly obviated the earlier need for dedicated terminal units for modes such as SSTV. Software-based signal processing within the PC, often using programs written by amateurs, allows the computer to perform the hard work of modulation, demodulation, and signal processing, together with data storage and retrieval facilities.

At the time of writing, amateur freeware and shareware programs are readily available (eg JVFAX, HamComm, MSCAN, EZSSTV etc) for SSTV, fax, CW, RTTY transceive, and PACTOR and packet receive, all using an extremely simple interface which is used between the transceiver and the PC's RS-232 port.

SIMPLE OP-AMP INTERFACE

The circuit given in Fig 7.16 is suitable for use with a variety of readily available programs, and provides a simple, easy-to-build interface for data modes. Received audio from the transceiver is shaped in the

Table 7.3. Typical DX packet cluster commands

BYE	Bye, disconnect from the packet cluster
CONFERENCE	Enter conference mode on the local cluster node
CONFERENCE/FULL	Enter conference mode on the full cluster
DELETE	Delete mail message
DIR	List active mail messages on the local node
DIR/ALL	List all active mail messages on the local node
DIR/BULLETIN	List active messages addressed to 'ALL'
DIR/n	List the last *n* active messages
DIR/NEW	List active messages added since you last used the DIR command
DX	DX spotting info announcement
DX x y z	Announce DX station of callsign *x* on frequency *y* with comment *z*
EXECUTE	Execute your personal command procedure
FINDFILE	Locate file(s) on the system
HELP or ?	Help (displays a short command listing)
HELP x	Display help for command *x*
KILL x	Delete mail message *x*
LIST	List active mail messages
QUIT	Bye, disconnect from the packet cluster
READ n	Read message numbered *n*
REPLY	Reply to the last-read mail message
REPLY/D	Reply to and delete the last-read mail message
SEND	Send a mail message
SEND/P	Send a personal mail message
SEND/NOP	Send a public mail message
SET	Set user-specific parameters
SET/NEED x	Store prefix *x* that you need in the cluster's database
SHOW/C	Display full packet cluster configuration
SHOW/CL	Display shortened packet cluster configuration
SHOW/DX	Show last reported DX spots
SHOW/DX/x y	Show last *x* number of reported spots on band *y*
SHOW/H x	Show heading and distance to country prefix *x*
SHOW/LOC x	Display the latitude and longitude of station callsign *x*
SHOW/M x	Show maximum usable frequency to country prefix *x*
SHOW/QSL x	Show QSL information for station callsign *x*
SHOW/U	Show callsigns of stations connected to local cluster node
SWITCH	Change to alias call
TALK x	Enter 'talk' mode to station with callsign *x*
TALK x y	Send one-line message *y* to station with callsign *x*
TYPE	Display a particular file on the packet cluster
UPDATE	Update a custom database
UPLOAD	Upload a file to the packet cluster
WWV	Log/Announce WWV propagation information
WX	Announce weather conditions

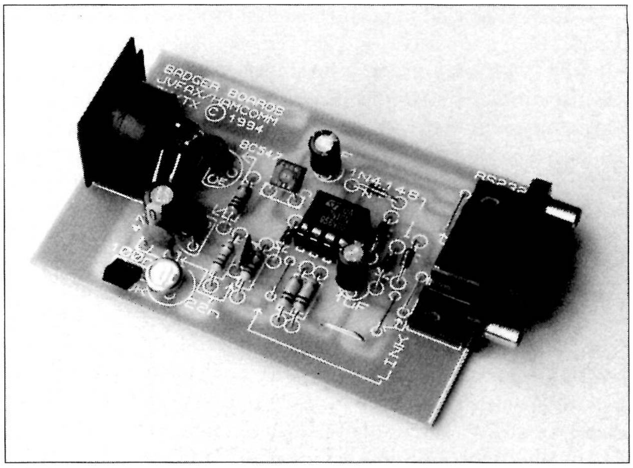


Fig 7.15. A simple one-IC interface can be used for data modes

op-amp to provide a rectangular waveform, and fed directly to one handshake line (DSR) of the PC serial port. For this, the 741 operates as a limiting amplifier with full open-loop amplification. The circuit is powered directly from the PC RS-232 port, where the RTS and DTR lines are used to serve as a power supply to the op-amp, the diodes and capacitors in the interface being used for voltage smoothing. Note that the RS-232 interface boards of some computers may not provide sufficient output here for a standard 741 IC – in these cases the use of a CMOS 741 or the pin-compatible TL071 op-amp may be usefully substituted.

On transmit, the rectangular waveform from the PC's RS-232 TXD line is initially limited by the resistor and back-to-back diodes. This is followed by a passive two-stage low-pass RC filter to filter the resultant square wave, prior to application to the transceiver microphone input, a potentiometer providing a level-setting adjustment. Transmit/receive switching is provided by the RS-232 RTS (Request To Send) line – when this goes high it drives the base of the NPN switching transistor to place the collector/emitter into conduction, the collector driving the transmitter PTT.

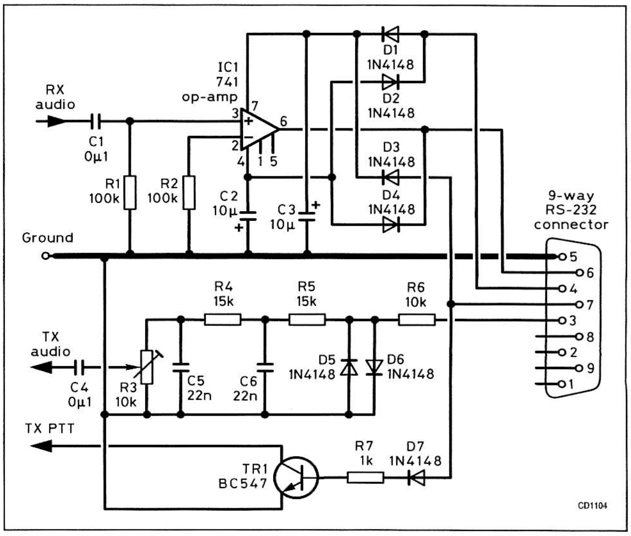


Fig 7.16. Simple SSTV/fax/packet/PACTOR interface

As the PC is used to perform signal processing, the timing is particularly sensitive, and slower computers, such as early 8086/88 XTs, may prove a limitation, although all 'current specification' PCs should be more than adequate. Furthermore, this method in some cases conflicts with some memory managers such as EMM386. (However, most interestingly, not on every machine!) Results obtained with such a 'software demodulation' are from very poor up to excellent.

OTHER INTERFACES

Using the same program but with differing interfaces, the four RS-232 input handshake lines, DCD, RI, CTS and DSR, can be used to provide a 4-bit parallel input port. To expand the data input capability from these 4 lines to 8 bit, a multiplexing technique can also be used by the program if it supports this, where the software initially reads in the four most significant data bits with the RTS line set to 'high'. The RTS line is then toggled to the low state and after a timed delay the lower four bits are read. Alternatively, the PC parallel port may be used to provide greater flexibility, again with alternative interfaces. These are available from commercial suppliers, and construction details for home-made designs are available in specialist amateur publications such as the quarterly journals of the British Amateur Radio Teledata Group (BARTG) and the Remote Imaging Group (RIG).

Packet modems

A popular solution to providing packet radio operation is by use of a BayCom modem for either 1200 baud or 9600 baud packet, together with the use of appropriate PC software such as that also offered by BayCom. The BayCom team is a group of German amateurs, and their modem designs are available either ready-built from commercial suppliers (eg Siskin Electronics and J&P Electronics in the UK) or in kit form from groups such as MAXPAK (Midlands AX.25 Packet group) and the NWPUG (North West Packet Users Group).

The 1200 baud modem is based upon a TCM3105 modem IC which, although reportedly no longer available in the UK, can be obtained from Germany. The interface connects to the PC's RS-232 port, and requires the use of appropriate software to provide the AX25 packet processing – a standard terminal emulator will not suffice. A commercial ready-built 'MiniPak' circuit for this system is available in surface-mount form – this is entirely contained within the shell of an RS-232 connector but kits using discrete components are also readily available.

The PAR96 is a 9600 baud modem, also from BayCom, again using the same BayCom PC software. This is again available either in kit form or ready-built. The circuit design for this unit, including PCB etc, has been published in the German amateur radio press.

PC sound card

At the time of writing, a number of data mode systems are available using PC software together with a standard PC sound card for the audio interface. These currently include CW, SSTV, FAX, AMTOR, RTTY etc as well as for DSP filtering implementations. As technology and PC processing power increases, the use of this type of interface for amateur data modes is likely to grow significantly.

Note, however, that a typical PC modem, designed for

Fig 7.17. 9600 baud modem from BayCom in kit or ready-built form

landline use, is usually not suitable for data-over-radio use, ie for direct connection to an amateur transceiver for packet radio etc. The use of PC-based TCP/IP programs, similar to those used for landline-based Internet connection, is however possible for use with amateur-based TCP/IP radio networks, providing sufficiently high-speed radio links are available. At least one UK TCP/IP group have this facility available for the use of amateurs. A further progression from this is the use of digital speech over TCP/IP.

Here is a quotation from a 1988 article by James Miller, G3RUH, designer of the famous G3RUH 9600 modem used worldwide terrestrially and on amateur satellites: "Who will predict that one day even our voice repeater links will be entirely digital. I will!"

In 1992, the BayCom team demonstrated digital speech via a packet radio store-and-forward system between remote 'speech mailbox' repeaters. In 1996, the Internet was first used with 'RepeaterLink' and 'Iphone' software to provide remote simplex transceive speech operation for amateurs on a number of FM VHF/UHF repeaters around the world. Its use via amateur radio-based TCP/IP networks is certainly possible, again given suitably fast links. We are seeing the gradual merging of data modes with other operation modes, and there may soon be no difference.

WEAK-SIGNAL MODES

These typically use FSK (direct frequency shift keying, or audio frequency shift keying on SSB), and PSK (phase shift keying) of a transmitter.

CLOVER is a proprietary mode, developed in the USA by Hal Communications. It uses PSK with full duplex simulation, where data is transferred between two linked stations, with automatic data link direction changeover. External data compression is used to increase data throughput, and the timing and modulation mode are automatically changed to suit the prevailing propagation conditions.

PACTOR-1 was developed by amateurs in Germany and is an FSK mode. It is increasingly found included as 'standard' on a number of commercially available multimode data controllers, and the hardware design of a PACTOR controller for home construction has been published in the German amateur press. Adaptive Huffman data compression is internally and automatically used, together with variable 100 baud or 200 baud speeds in the controller to improve throughput depending upon the prevailing propagation and signal conditions.

'Memory ARQ' using an internal 8-bit A/D converter is used in the original modem design, which has been adopted by a number of commercial TNCs but not by all, thus offering differing performance based upon the actual modem used. The A/D unit converts received 'packets' into a data stream, and subsequent identical packets (automatically repeated under weak signal conditions) are added to this to eventually 'build up' a valid frame whilst the background noise reduces towards zero. In this way, PACTOR can be successfully used in signal conditions where the wanted data signal is significantly below the level of unwanted noise and interference, even to the point of complete inaudibility by the human ear.

PACTOR-2 is a PSK mode, and is 'backwards-compatible' with PACTOR-1 as it uses the same handshaking protocol. The additional use of DSP techniques provides greater throughput and operation under even weaker signal conditions than PACTOR-1. A combination of Huffman and Pseudo-Markov adaptive automatic software compression is used to further increase throughput. The modulation is automatically changed to suit the prevailing propagation conditions, ie DQPSK, D8-PSK, D16-PSK etc.). It has the ability to employ automatic frequency tracking with fine-adjustment capability, via up/down fine frequency control output lines to your transceiver.

GTOR is a further proprietary mode, developed in the US by Kantronics. It is an FSK mode offering a high data transfer rate given reasonable signal conditions. Software compression is inherent in this mode to improve throughput.

The most common implementation of the above modes in the amateur radio station is by the use of a commercially available multimode data controller, connected to a transceiver with a terminal or PC employed for control. A number of such controllers also feature multiple ports, eg one for weak signal modes as above and often combining CW, RTTY and AMTOR (all used with a transceiver in SSB mode), plus a further port for packet radio (using an FM mode transceiver).

Frequency stability requirements

Unlike other modes, PSK used by CLOVER and PACTOR-2 requires a high degree of frequency stability, with stations 'netted' to typically within 10Hz, although the auto-tracking in PACTOR-2 can cope with up to 100Hz offset drift. This is rarely a problem on HF with current TCXO (temperature compensated crystal oscillator) technology, but additional care must be taken on VHF and UHF to ensure accurate frequency netting.

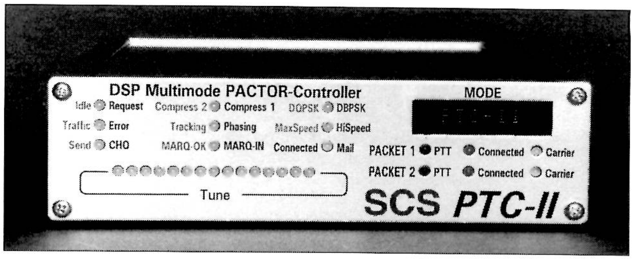


Fig 7.18. The SCS PACTOR-2 DSP Multimode Controller

CONTACTS

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J&P Electronics: Unit 45, Meadowmill Estate, Dixon Street, Kidderminster, DY10 1HH. Tel: 01562 753893.

NWPUG (North West Packet Users Group): W Bateman, Ingleside, Waterloo Whixall, Whitchurch, Shropshire, SY13 2PX.

BARTG (British Amateur Radio Teletype Group): Peter Adams, 464 Whippendell Road, Watford, Herts, WD1 7PT. Tel: 01923 20774 evenings/weekends.

RIG (Remote Imaging Group): Ray Godden, G4GCE, Wayfield Cottage, The Clump, Chorleywood, Herts, WD3 4BG. Tel/fax 01923 720714. Email: 101602.376@compuserve.com.

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8 Amateur television

OF all our senses, sight is probably the most precious to us. Almost every activity we involve ourselves in has a degree of searching, moving and manipulating which require visual cues or feedback at some point. It seems logical that visual communication using television, which has become so much a part of our domestic lives, should also be applied to amateur radio. Listen to any voice conversation on the amateur bands and before long you will hear someone describing a piece of equipment in their shack or asking if anybody knows what a "grey thing with an odd-looking connector in the top corner" is. It's much easier to explain the gadget in your hand when you can hold it up to a camera and let the other person see it for themselves. The adage "a picture is worth a thousand words" is proven true with amateur television (ATV).

The majority of ATV contacts are picture and sound in one direction with 'talkback', usually on 144.750MHz in the UK, in the return direction. Full duplex operation (sound and vision both ways simultaneously) is becoming more popular as activity grows and repeater coverage is extended.

WHAT GETS BROADCAST?

As in the case of voice communication, ATV tends to be unprepared and unrehearsed. Some people use television as an extension of photography – their interest lies in the picture content and production rather than its technical aspects. Others are experimenters, preferring to try new electronic techniques and exotic components. The blend of art and technology works well and provides a wide variety of enjoyable material.

Local shows and events are often tape recorded with camcorders and replayed to an audience over the air – occasionally, depending primarily on location, the transmissions are sent 'live'. Having an ATV station at a public event always attracts attention and is a good way of introducing newcomers to the hobby. The view through a camcorder lens will often make a more rewarding transmission than one from a professional camera team or editing suite.

Of course, many ATV transmissions are simply 'shack shots' with the camera pointing at the operator – this may not always be the prettiest of sights but at least it adds a more personal touch to the contact. There are a few ATVers who prefer not to send camera shots at all and concentrate their efforts on transmitter and video circuitry design – the only evidence on-air of these devotees is an occasional test card transmission.

Like other aspects of amateur radio, TV has its contests and contest groups. The challenge in UK ATV contests is to send a four-digit number over the greatest distance and receive confirmation of its total from the recipient. The numbers themselves are never repeated back in case they are overheard by other competitors but they are entered in the contest log sheets for checking by the adjudicator. Some contests allow slow-scan television (SSTV) on the HF bands, while others only allow normal fast-scan TV on 70cm and above. Either way, striving to exchange pictures as well as voice enhances the competition. Picture quality reports are also exchanged using the 'P' system, where P0 corresponds to an unrecognisable picture through to P5 for a perfect, blemish-free picture.

A CLOSE LOOK AT A TELEVISION SIGNAL

Unlike a voice transmission which only carries a single modulation at audio frequencies, TV signals are a composite of several different component parts. Before getting too deeply involved in transmitter and receiver functions, let's look at exactly what these video components are and their purpose.

The intention is to measure the amount of light falling at each point in the source image and faithfully recreate it on a screen some distance away. Two of the requirements are already defined: a way of measuring the intensity of light and some way of defining its position within the image. If colour is being used, a third signal is needed – this is itself a composite of two *colour difference* signals modulated onto a common carrier. This method of carrying colour information ensures compatibility with monochrome monitors which can simply ignore the colour difference signals and use the intensity (also called *luminance*) signal alone.

In the camera, the source image is focused by a lens and made to illuminate a light-sensitive pick-up. In older cameras this was a sensitised layer which was electrostatically charged – some of the charge was displaced when hit by light. The layer, or *target* as it is correctly named, was also hit by an electron beam which was magnetically or electrostatically scanned to make it sweep side to side and top to bottom over the image area. You may find it useful to think of the scanning process as moving your eyes over these lines of text. The combination of electron beam, fixed charge and light impact made the target voltage change according to the light intensity at the point hit by the beam, and after amplification this was used as the video signal.

More modern cameras use charge-coupled devices (CCDs) to convert the image to an electrical signal. The image is focused onto the CCD which consists of an array of photo-sensitive cells – the charge on the cell depends upon the amount of light falling onto it. To retrieve the image as a usable voltage, the cells are read sequentially, usually as a long shift register (aka *bucket brigade*). The process is analogous to the electron beam scanning in an older camera in that at any point in time the signal from the CCD corresponds to a particular position in the image.

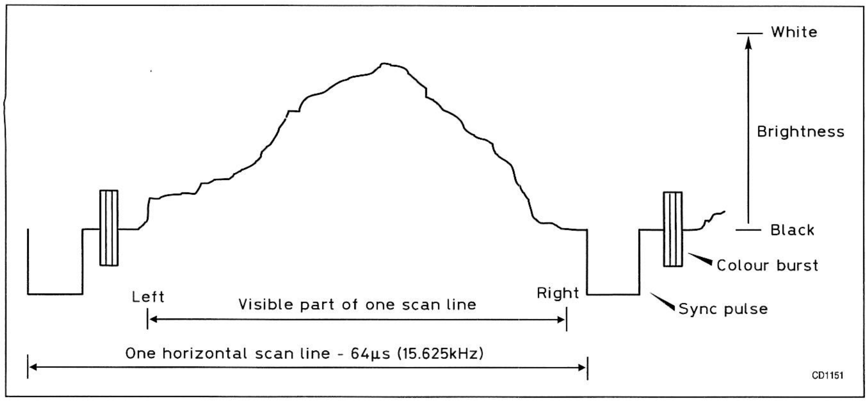

Fig 8.1. Composite video waveform

Once a video signal has been generated, it can be used to modulate a carrier and be sent over the air. At the receiving station, the signal can be demodulated to recover the original voltage. The final step is to use the video signal to change the intensity of a dot of light on the monitor screen.

If the position of the dot and its intensity match that generated in the camera, the image has been successfully reproduced. The process of ensuring the same position within the image is being used at the sending and receiving stations is called *synchronisation* or 'sync' for short. The sync pulses are generated in the camera at the sending end and are included in the transmitted waveform. There are two main types of sync pulse – one occurs just before the start of the horizontal sweep and the other before the vertical sweep – and they can be distinguished by their lengths. Horizontal sync pulses are relatively short compared to the vertical ones, and the difference in length is used in the monitor to decide which sweep should be reset to its start position.

The combined video and sync pulses is known as *composite video* and is shown diagramatically in Fig 8.1. Also shown is a part of the signal called the *colour burst* – this is only present when colour information is present in the picture. The burst is a 10-cycle-long sample of the colour subcarrier oscillator from the camera's colour encoder. It is used to ensure the colour demodulator at the receiving end stays in exact phase lock with the sending one. Inside the monitor, a timing circuit triggered by the horizontal sync pulse opens a gate allowing only the colour burst through. It is then compared to the phase of the monitor's own colour decoder oscillator and any phase errors between them are detected and eliminated. As an extra precaution, the phase of the burst signal is shifted +45 then –45° relative to centre on alternate lines. The monitor averages the phase shift to derive a single central phase for demodulating and uses the instantaneous difference in phase to decide the polarity of one of the colour difference signals. This system is called *phase alternation line* (PAL). The advantage of PAL over the North American NTSC colour system, which does not alternate the burst phase, is that phase errors taken over any two-line average tend to cancel out. Phase errors manifest themselves as shifts in the hue of the colour, for example making flesh tones take on a green or blue tinge. This can be quite important in ATV where poor signal paths are common and can cause considerable distortion to both amplitude and phase. The full process of encoding and decoding colour information is beyond the scope of this text – see the references at the end of this chapter for further reading on this subject.

The final component added to the composite video before transmission is the sound carrier. Audio is first pre-emphasised by increasing the volume of its high frequencies while at the receiving end these higher frequencies are attenuated to bring them back to their original levels. This roll-off of the frequency response also reduces the effect of noise from other sources, thus improving the overall signal-to-noise ratio. The transmission audio is used to frequency modulate a carrier at a nominal 6MHz. This carrier is then added at low level to the picture signal and the total is then used to frequency modulate the transmission carrier. The carrier-in-a-carrier principle is called *intercarrier sound*. In the receiver, the 6MHz is recovered

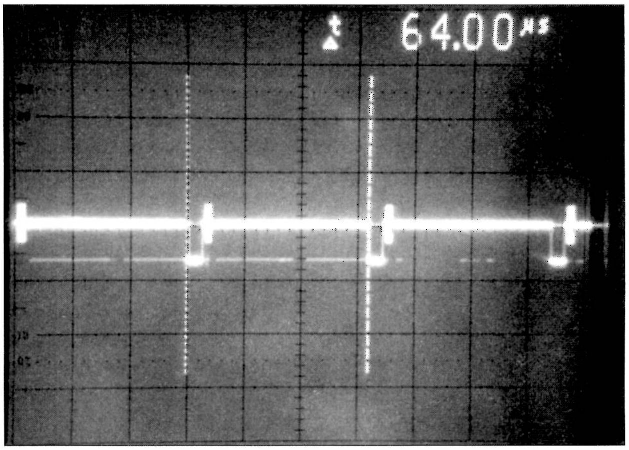

Fig 8.2. An oscilloscope trace of a blank (all-black) picture, showing sync pulses (to the right of the dotted time marker) and the colour burst (to the right of the sync pulse). The time markers are 64µs apart, corresponding exactly to the UK standard 625-line format

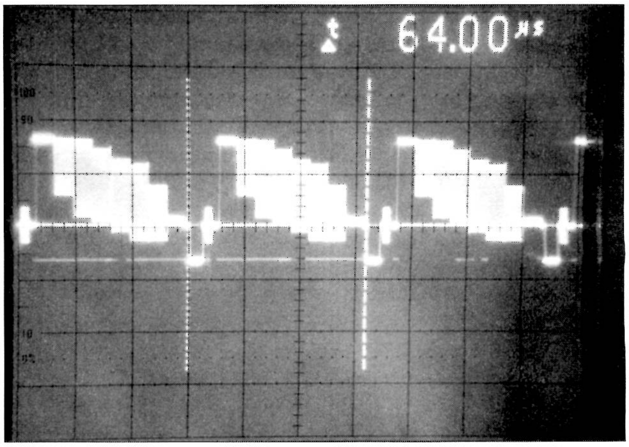

Fig 8.3. An oscilloscope trace of a display of 75% saturated colour bars. The sync pulses and colour burst are still clear

from the demodulated video, filtered and fed to its own amplifier and demodulator.

Fig 8.4. Typical ATV shack

VIDEO EQUIPMENT

The vast majority of ATV operators rely on commercially available cameras and monitors so the complications of generating the composite video source are largely circumnavigated. One area where home-brew video is still the norm is the test card. Most operators use an electronic test card generator to identify themselves. Typically this will have a callsign in large letters, the operator's name and locator square. Many fancy and varied designs can be seen on the air, originating from a variety of electronic designs and sometimes generated by computer.

A typical ATV station consists of a camera, test card generator, a video and sound source switch, a TV and of course a transmitter and receiver. Some operators utilise more advanced gadgets such as video effects and 'wipe' generators or units to display one picture in another. Almost all the commercially available home-video equipment can be applied to ATV use as the transmission standards used are almost the same as those of domestic broadcast television. The narrower bandwidth used on the 70cm band does not permit colour or sound to be sent but on higher frequencies amateur TV sound and picture quality can surpass that of professional broadcasters. It follows that on the 70cm band a low-pass filter must be employed to ensure the transmission bandwidth is restricted to prevent out-of-band radiation from the signal sidebands. Even with the reduced bandwidth, perfectly acceptable pictures can be sent. For those of us who can remember 405-line television, the 70cm band can still give picture quality superior to a good VHF BBC signal!

ATV REPEATERS

Although a great deal of point-to-point operating goes on, particularly during contests, the majority of ATV activity is through repeaters. The directional nature of long Yagi arrays and dishes makes it difficult to find new contacts unless locations and headings are known before hand. Repeaters provide a central target to aim at and their omnidirectional output ensures their reception in the widest possible area. Unlike voice and data repeaters, ATV repeaters are normally operational all the time. Instead of shutting down when not in use, they show a test card (Fig 8.5) or page of descriptive text instead. The presence of a steady constant signal is an enormous help when setting up a receiving system, making it easy to optimise antenna position and receiver tuning. Some repeaters have facilities for reporting the strength of incoming signals, displaying a graphic S-meter to assist with transmitter alignment. The wide separation between repeater input and output frequencies, usually 50MHz or more, also makes it fairly easy to filter out transmitted RF from the receiver input, allowing 'look-through' while sending.

Repeater input and output signals are horizontally polarised, generally using Alford slot antennas, although some allow the selection of a directional antenna to improve reception at the repeater by sending command tones (usually telephone dialling tones) over the sound channel.

ATV TRANSMITTERS

The methods employed depend upon the band being used. On the 70cm band, where space is very limited, transmissions are normally amplitude modulated and ideally will have one of the sidebands reduced in amplitude by filtering – this asymmetrical spectrum is called *vestigial sideband*. Sound

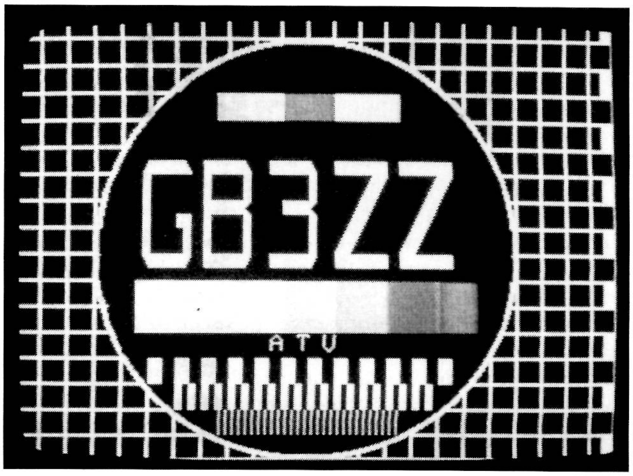

Fig 8.5. Off-air capture of the Bristol 24cm repeater test card over 25km distance

and colour carriers are normally not used on 70cm because they would result in sidebands spreading wider than the band allocation allows. Control of the final amplifier supply voltage is normally used to achieve the amplitude modulation of the carrier. On the higher-frequency bands, FM is the predominant mode of operation, the combined composite video, colour and sound signals being used to directly control the transmission frequency. Some operators have experimented with reduced bandwidth FM in the 70cm band and achieved good results but unfortunately the techniques used are somewhat incompatible with domestic television sets and are therefore not very popular.

ATV construction guidelines

The frequencies present in a video waveform span DC up to about 5MHz and therefore need to be treated rather like a HF band signal. Screening is important, not only to prevent pick-up of magnetic or radio signals but to prevent radiation of the same. A strong signal entering the video signal chain will show as a pattern overlaid on the picture – this may be stationary or random depending on the type of interfering signal. Interference escaping from video circuitry manifests itself as a buzzing sound on nearby radio receivers. The level and harshness of the buzz changes with the picture content.

Always use screened cables to carry video signals – 75Ω cable is generally used rather than the 50Ω type used to carry RF. Unfortunately, the quality of domestic UHF TV feeder cables leaves a lot to be desired and in many cases is unusable. Use a cable with a properly woven braid to ensure signal leakage is minimised. Cables should be correctly fed and terminated with resistive loads – mismatch causes standing wave problems which show as 'ghosts' or repeated images side by side as the signal bounces back and forth along the cable. Very short reflections can cause phase cancellation of certain frequencies which may result in loss of colour or sound subcarriers.

Most ATV stations will have more than one video source on hand, probably at least one camera and a test card generator. Because these are most likely not synchronised with each other, if any cross-coupling occurs between them, the weaker signal will probably appear as a faint image drifting slowly through the dominant one. Care is needed to minimise this breakthrough, particularly in source-switching or mixing units. Standard CMOS signal switch ICs can be used at video frequencies but grounding and decoupling needs to be very efficient to keep the signals apart. Note also that switching between unsynchronised signals will almost certainly cause monitor 'jump' until the new sync pulses are recognised and cross-fading will wreak havoc while two sets of sync pulses appear together.

Unlike audio where a potentiometer can control the volume, video fading requires the reduction of the visible part of the waveform while leaving the sync pulses and colour burst at the same amplitude. Reducing sync level will result in an unlocked picture, while reducing burst level will cause severe colour noise and eventually no colour at all. Several designs for video faders or 'fade to black' units are available from sources listed later. Basically they use the sync pulses to operate a changeover switch – syncs pass straight through while the picture information alone is routed through an attenuator. The two paths are then recombined.

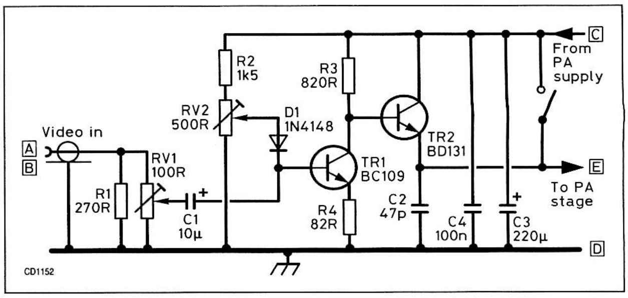

Fig 8.6. 70cm AM modulator. BD131 will require a small heatsink (minimum 10°C/W)

Practical design for an AM modulator

The circuit shown in Fig 8.6 is a simple AM modulator for use in the 70cm band and can be used to convert a low-powered (no more than 3W) conventional voice rig to transmit ATV. Ideally the rig should use CW mode but FM will work just as well if the microphone audio is disconnected. Although a small amount of audio frequency deviation will not affect the picture, it will be receivable on a conventional FM receiver. To use the modulator, the power supply to the PA stage (and possibly driver stage) is redirected through TR2. The switch across TR2, when closed, will restore normal operation so the transmitter can be used as before when not using ATV mode. Before adding the circuit, it will be necessary to locate and disconnect any large-value decoupling capacitors which will almost certainly be present across the PA supply lines. A small capacitance, no more than 470pF, should be left in place to provide a low impedance to RF and to help attenuate any high frequencies in the video signal. Component and PCB layouts are given in Appendix 1.

Adjustment is straightforward. Initially set RV2 to mid-position and tune the rig to the desired frequency – 435.5MHz is a commonly used frequency but avoid moving too close to the band edges as sidebands will start to radiate out of band. With the switch closed, check that the rig is working normally. If all is well, open the switch – the output power should drop and be adjustable by setting RV2. Apply a video signal, preferably of a stationary image or test card, and set RV1 to mid-position. While monitoring on a receiver, adjust RV2 for optimum picture – at one end of its range the sync pulses will

Table 8.1. Component list for 70cm AM ATV modulator

Component	Value	Maplin order code
R1	270R	M270R
R2	1k5	M1K5
R3	820R	M820R
R4	82R	M82R
RV1	100R	UF97F
RV2	470R	UF99H
C1	10µ	AT98G
C2	47p	WX52G
C3	220µ	AT41U
C4	100n	YR75S
D1	1N4148	QL80B
TR1	BC109	QB33L
TR2	BD131	QF03D
Heatsink for TR2	—	JW29G

Note: Heatsink is suitable for rigs up to 1W – a larger heatsink may be needed if the modulator is used with more powerful rigs.

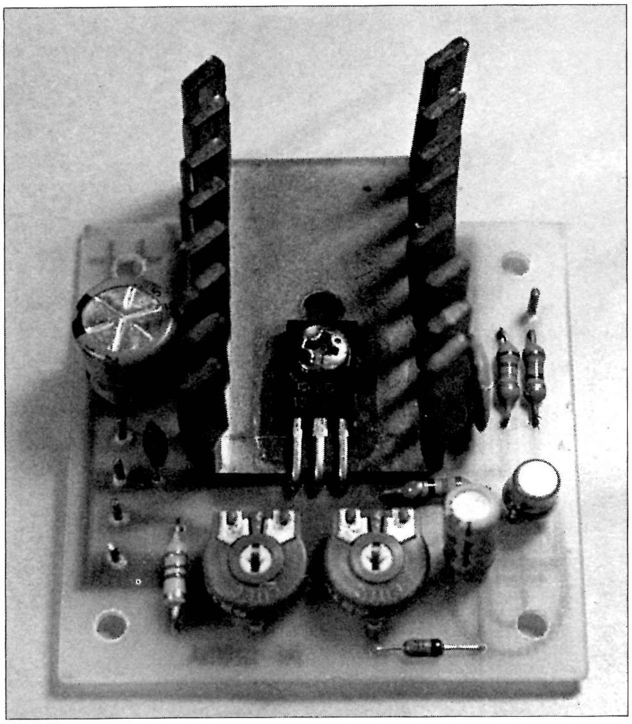

Fig 8.7. Photo of modulator board

'crush', causing the monitor to lose synchronisation. At the other end the brightest parts of the picture will 'wash out', rather like the appearance of an over-exposed photograph. The correct setting is midway between the onset of each symptom. Once the optimum setting is established, adjust RV1 for best contrast – its setting will have some effect on RV2 so it may be necessary to repeat both adjustments until best results are obtained.

If an oscilloscope is available, the circuit in Fig 8.8 will help with the alignment. It is a simple RF pick-up probe and detector. If used, the oscilloscope should display the same video waveform as the one fed into the modulator. This circuit can also be used as a field-strength meter by connecting a millivolt meter instead of the oscilloscope and will work over frequencies from about 10MHz up to about 1GHz.

As with all transmitters, before connecting the antenna, check for spurious emissions and that the signal is confined within the band edges. If the sidebands are wide enough to reach the band edges, a low-pass filter should be fitted in line with the video input – a roll-off starting at about 2.5MHz should be adequate. Remember the filter response must be relatively flat right down to DC to avoid video

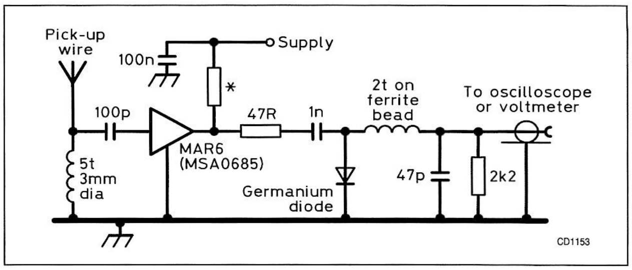

Fig 8.8. Simple RF pick-up probe and detector. * Use 330R for 12V supply – reduce to 270R for 9V battery

distortion. Use a linear amplifier after the rig to boost the power output if necessary – one suitable for SSB should do the job but under no circumstances use one designed for FM only. 70cm ATV transmissions are AM, so linearity is important.

23/24cm (1.3GHz) transmitter designs

Several excellent and inexpensive kits are available for this band, and it is doubtful if one could be built with better quality and lower price than these kits offer. A list of some kit suppliers is given later.

Most designs use a varactor diode (varicap) to tune a VCO at the transmitted frequency. The alternative of using a lower-frequency oscillator and passing it through frequency-multiplying stages is sometimes used but the relatively wide deviation can cause problems, especially as the deviation is multiplied along with the carrier. After the oscillator there is usually a buffer stage and then a modular power amplifier block. Several amplifiers are available at affordable prices. Mitsubishi make one capable of producing 20W when adequately heatsinked. Frequency stability is usually controlled by using a PLL and frequency divider (normally in one IC) – the Plessey SP5060 and SP5070 are popular, containing the phase detector, a divide-by-256 prescaler and the reference oscillator circuits. For example, a crystal cut to

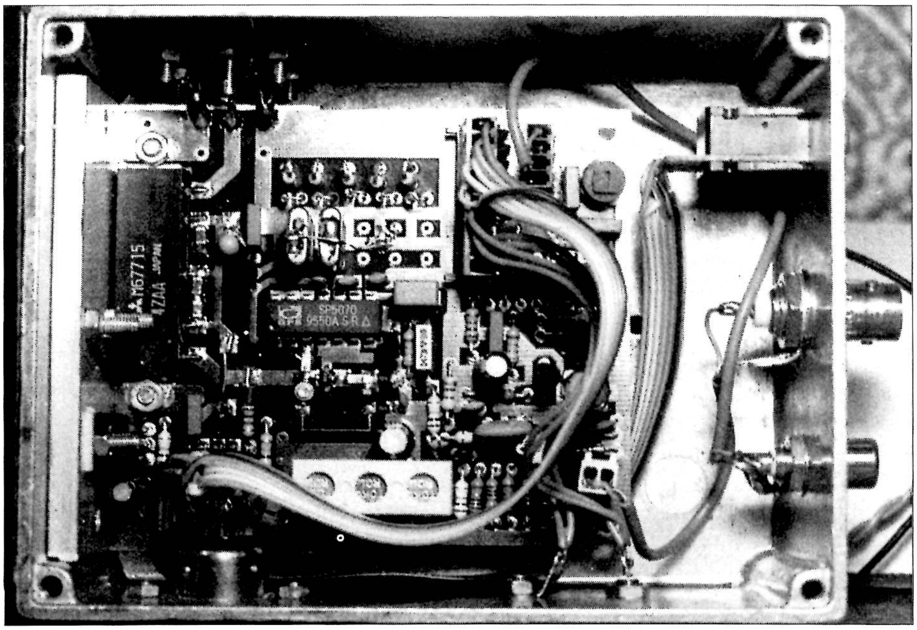

Fig 8.9. 1.5W 24cm ATV transmitter designed by G8OZP

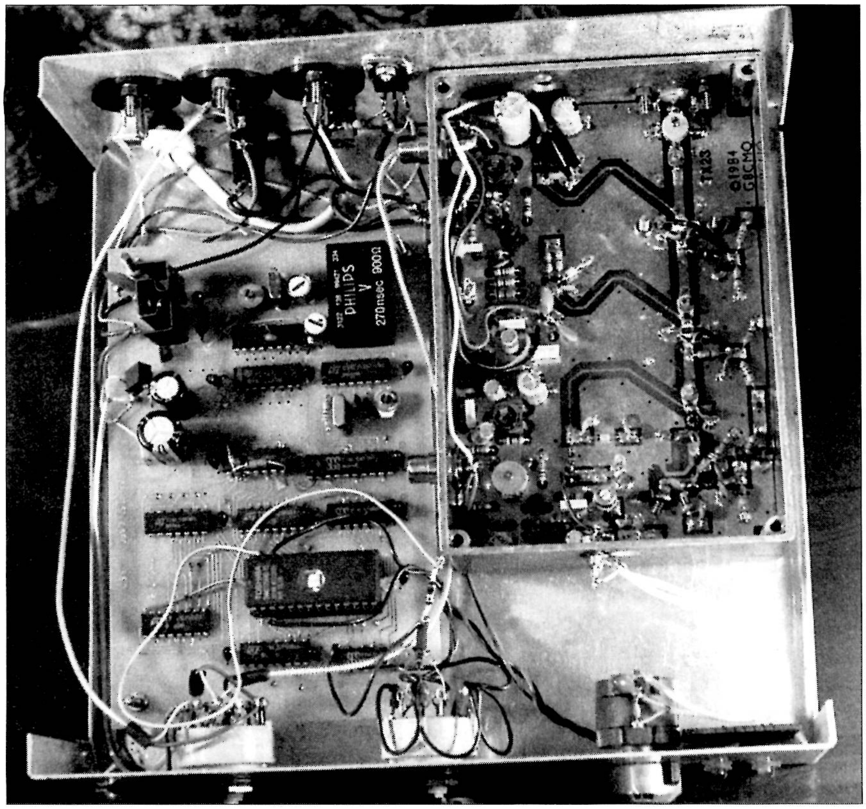

Fig 8.10. A typical 23/24cm transmitter and test-card generator unit at GW6BWX. The transmitter is on the right in the die-cast box and is the GB3VR group design as mentioned in the text under 'Kits'. The PCB to its left is a home-brew testcard generator. The front-panel switches allow either normal operation or link the test-card generator to the transmitter so it can send identification without a camera attached.

4.902MHz will result in a carrier 256 times higher in frequency on the 1255MHz simplex TV frequency.

ATV RECEIVERS

As with transmitters, receiver techniques depend very much on the band being used. On the 70cm band, where AM is the predominant mode, most stations use up-converters to shift the band so it can be received on a normal UHF broadcast TV. Some televisions and VCRs will tune low enough to receive the band without any additional converter or modification but the later-generation synthesised tuners are less generous when it comes to out-of-UHF-band reception. Some manually tuneable TV sets have a resistor in series with the low-voltage side of the tuning control, and shorting it out will usually allow the bottom end of the tuning range to extended down to the ATV frequencies. Domestic terrestrial broadcasts in the UK are AM so the mode used commercially and in 70cm ATV are completely compatible.

On bands above 70cm the mode usually used is FM which makes a separate receiver mandatory. This isn't as bad as it first sounds because domestic satellite broadcasts are FM and surplus receivers are inexpensive and only require minor modification before being usable for ATV. To see what changes are needed to convert a satellite receiver it is first necessary to understand how one works. In essence they are nothing more than a normal superheterodyne receiver with a tuneable IF and an additional intercarrier sound demodulator.

Fig 8.11 shows the structure of a typical system – the satellite signal is typically in the 11GHz (K) band and is down-converted inside the LNB by mixing with a 10GHz local oscillator and filtering to accept the subtractive product. After amplification, this signal is sent down a coaxial cable to the indoor unit which is tuneable across the filter bandwidth, typically 750MHz to 2GHz. Sometimes the LNB local oscillator can be switched to different fixed frequencies so a wider range of input bands can be converted to the same IF. Note that the whole of the 23/24cm band falls within the tuning range of the indoor unit – many receivers even give a direct frequency readout of the IF and therefore the frequency being received.

Satellite units can be used without any modification at all but performance can be significantly enhanced by making a few small changes. First, a warning. When used for satellite reception, the DC supply to the LNB is fed from the receiver via the coaxial cable. If an antenna utilising a looped dipole or balun is directly attached it will short out the supply and may cause damage. There are several ways to avoid this problem – the simplest is to cut the supply feed wire inside the receiver but alternatively a small capacitor can be wired in line with the input socket. If a capacitor is used, make sure it is a type suitable for use at UHF or losses will be incurred.

Probably the most satisfactory solution is to use the feed to power a preamplifier because, as receivers are normally preceded by an LNB which provides 50dB or more gain, they

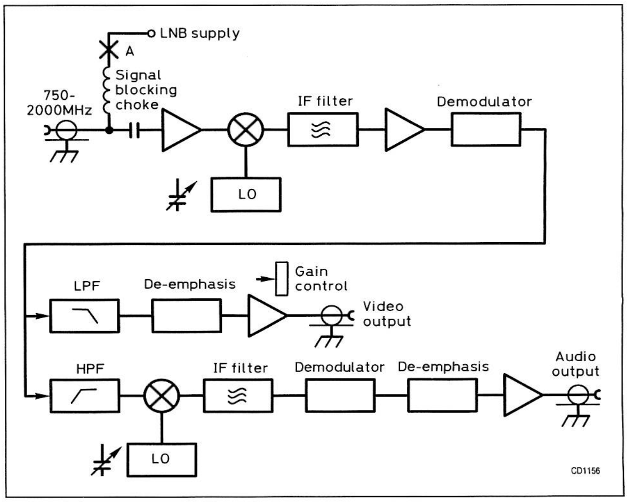

Fig 8.11. Functional blocks of a typical satellite receiver. Cut at 'A' to isolate voltage from input socket if connecting antenna directly

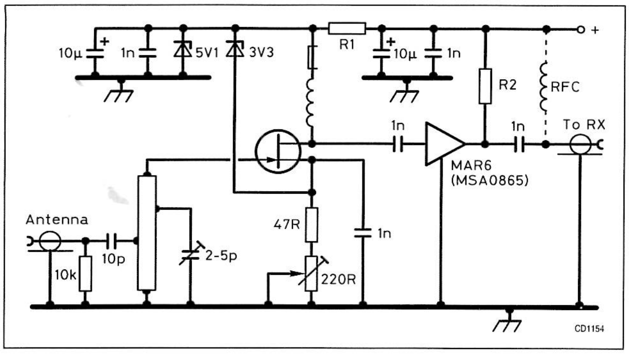

Fig 8.12. High-gain 23/24cm preamp using GaAs FET and MMIC from dismantled LNB. For 12V operation, R1 = 150R (0.5W) and R2 = 330R (0.5W). To power from satellite receiver, R1 = 270R (1W), R2 = 560R (0.5W) and connect RFC (5t on 3mm former) as shown to link receiver supply from coaxial output cable

are relatively insensitive when used alone. Many designs for preamps exist – the single GaAsFET followed by a MMIC seems popular and yields excellent results. If the preamp includes a filter, it should have sufficient bandwidth to allow a 12MHz-wide TV signal through. A narrow-band filter will result in poor sound and colour performance because both of these use the higher-frequency contents of the video signal.

A simple and inexpensive design is illustrated in Fig 8.12. The GaAsFET and MMIC devices can be salvaged from a broken LNB – typically there are three low-noise FETs and two MMICs inside an LNB so they make an excellent source of RF amplifier components, yet are usually thrown away when they break down. A trip to the local TV dealer to ask for discarded LNBs can be very rewarding. Before removing semiconductors from an LNB, mark them to identify the input pin, for once removed it can be difficult to tell their original orientation. Hint: use sharp scissors to cut the PCB around the component before unsoldering, it reduces the heat dissipation into the board and shortens the time the component is heated by the soldering iron.

It should be noted that satellite broadcasts use a much higher modulation index than used in ATV and this will result in a lower-than-expected video voltage from the demodulator. Almost all satellite receivers have an internal video gain control which can usually be advanced to maximum to increase the video level. There may be a penalty to pay for increasing the

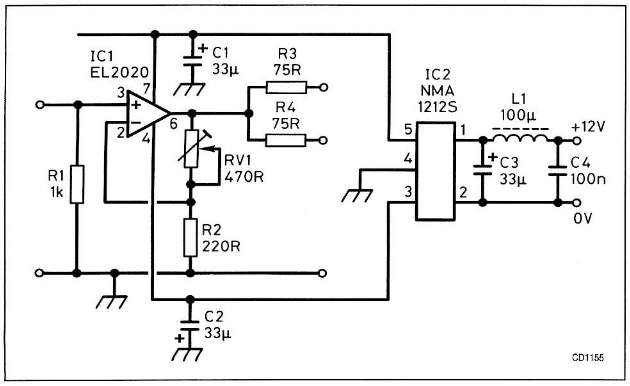

Fig 8.13. Amplifier suitable to raise video level from satellite receiver

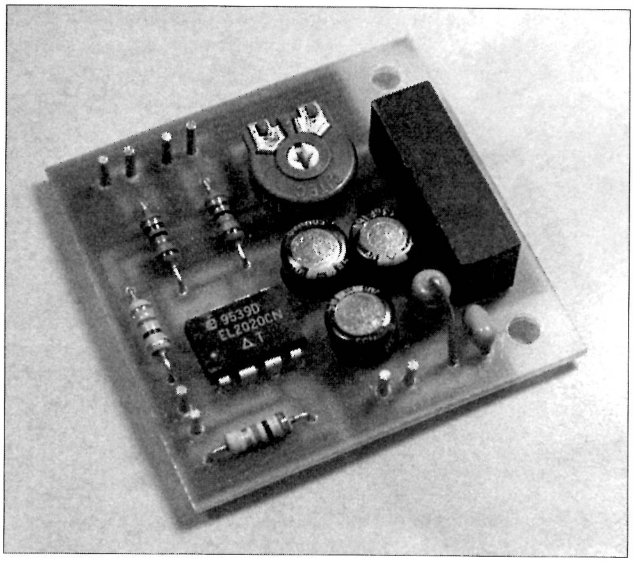

Fig 8.14. Photo of amplifier board

level control – the amplifier stage in some receivers will exhibit poor HF response due to bandwidth reduction as more gain is demanded. If this happens, little can be done with the existing amplifier and another stage will have to be added in series – this can be in-line with the video output socket and external to the receiver if desired.

A suitable external amplifier with two identical output channels is shown in Fig 8.13. Component and PCB layouts are given in Appendix 1. Deriving a split-polarity supply from a DC-DC inverter unit eliminates the need for coupling capacitors, so ensuring a near-perfect LF response. Do not be tempted to use normal op-amp ICs in this application – purpose-designed video amplifier ICs have a wide bandwidth, low phase distortion and are designed to feed 150Ω loads (75Ω in series with the cable and 75Ω terminating load at the far end). Also, note that large-value coupling capacitors should normally be used to carry video signals – the relatively low impedance used and requirement to convey frequencies close to DC makes a low reactance essential. It is also a good idea to connect a low-value ceramic capacitor in parallel with large-value electrolytic capacitors as their internal inductance can reduce their effectiveness at carrying high frequencies. Poor low-frequency response will be seen as a brightness gradient across large areas of dark or light picture and in extreme cases will lead to poor vertical synchronisation or field roll. A poor high-frequency response causes a loss of definition and makes the picture appear smudged.

Table 8.2. Component list for video amplifier

Component	Value	Maplin order code
R1	1k	M1K
R2	220R	M220R
R3, R4	75R	M75R
RV1	470R	UF99H
C1, C2, C3	33µ	AU00A
C4	100n	RA49D
L1	100µ	WH41U
IC1	EL2020	UR06G
IC2	NMA1212S	AH17T

Some older satellite receivers expect fixed sound carrier frequencies, usually 6.5MHz, but modern ones are tuneable usually between 5.5MHz and 8MHz. If the frequency is fixed, the sound IF filter needs to be replaced – they are all a standard size and 6MHz ones, as used in domestic TV sets, are inexpensive and easy to obtain. Obviously, satellite receivers are not ideal for ATV use but they are inexpensive and make a good 'base' system on which to build. Experience shows that the performance of a 'hotted-up' satellite receiver is almost indistinguishable from a purpose-designed ATV receiver.

OPERATING TIPS

To check for local activity, try calling on 144.750MHz – it's used throughout the UK and much of Europe as a 'talkback' frequency so it is likely to be monitored by active ATV stations.

When sending live pictures, try to send sound with them, particularly when talking to the camera. Do this even if you are also using a separate 'voice' transmitter to talk to another station – watching someone's lips moving but not hearing what they are saying is impolite to say the least. Remember there could be people watching who cannot hear the voice transmission.

Try to keep the camera in the same line of view as the monitor – if possible place it immediately above the centre of the monitor screen as a slight downward glance while you watch the picture is hardly noticeable. If the camera and monitor are spaced too far apart, there is a natural tendency to look toward the monitor rather than the lens. At the receiving end this looks like the conversation is with someone out of camera view in the shack instead of the station in contact.

Be aware of the background – avoid bright lights or reflections which can result in flare or a silhouette image. The rules are pretty much the same as with photography – keep the light source behind the camera, not behind the subject.

If the camera is in a position where other members of the family can be seen, make sure they are aware that they can be observed and overheard by a third party. Embarrassing situations when people are unaware of the camera being used are not uncommon!

FURTHER INFORMATION

The BATC (British Amateur Television Club), affiliated to the RSGB, exists for the benefit of ATV enthusiasts. It has about 2000 members and issues a quarterly colour magazine called *CQ-TV*. Membership is currently £12 per year. Details from: The Membership Secretary, 'Grenehurst', Pinewood Road, High Wycombe, Bucks, HP12 4DD.

The BATC also publish several ATV books and offer an extensive library of technical publications to members. They can also provide printed circuit boards and components for many of the projects featured in their magazine. They organise an annual ATV rally and a biannual ATV convention.

KITS

23cm transmitter kits are available from:

Worthing & District Video Repeater Group, GB3VR, c/o 21 St. James Ave, Lancing, Sussex, BN15 0NN. This is a 1W design, priced at £80.

Bob Platts, G8OZP, 220 Rolleston Road, Burton upon Trent, Staffs, DE13 0AY. This is a slightly more powerful design using a PLL and incorporating video and audio filters. It produces 2W and is priced at £125.

23cm antennas are available from:

Severnside Television Group, 18 Linnet Close, Patchway, Bristol, BS12 5RN. They sell three types: a trough reflector design for £19, an 18-element Yagi for £15 and a 38-element Yagi for £26. The 18-element unit can be converted to the 38-element type with a conversion kit also sold by this group.

All the prices quoted are correct at the beginning of 1997 but should be confirmed before ordering. Please send an SAE with enquiries as some of these suppliers are non-profit making organisations, manned by volunteers to raise funds to keep ATV repeaters on the air.

9 Satellite communications

AMATEUR active satellites have been on the scene since the first OSCAR (orbiting satellite carrying amateur radio) launched in 1961. They have been likened to amateur repeaters. They *do* receive a signal on one frequency and transmit it on another but there the similarity ends.

Terrestrial repeaters are at a fixed point and do not need steerable antennas to track them. Most repeaters use FM and most satellites use Morse or SSB. Terrestrial repeaters cost a few thousand pounds to build and support and can be serviced or replaced without much effort. Satellites, on the other hand, cost upwards of £300,000 to design, build and launch, plus a few thousand pounds a year to command into the correct orbit. For the Phase 3 D satellites, it is at present about £2,500,000 (1996 prices).

Orbiting satellites cannot be serviced with new hardware once in orbit and, to date, battery supplies have been provided with solar energy to maintain control and transponder output.

Satellites have a shortish life span of five to 10 years before decay or mishap cause their demise. Generally they operate on a bandpass of a few hundred kilohertz in the internationally allocated Amateur Satellite Service bands using Morse or SSB and providing a worldwide service. A few operate in digital modes.

There are a number of current satellites which are divided into four classes:

1. Those with transponders, receiving on one band and transmitting on another. These are the 'conventional' satellites and use Morse or SSB.
2. Those which receive digital signals, store them and retransmit them later on interrogation. These are the so-called *forward store and retrieve* satellites.
3. Those which contain scientific apparatus and send out data signals for reception only, eg the UOSAT series.
4. Occasionally there are radio amateurs in orbiting spacecraft. Information is usually broadcast by the RSGB with the bands to be used, the times of operation and the modes. These occur every day when the amateur is on board.

Most satellites contain beacons, reception of which is a good indication that the satellite is within range. They always carry information of the condition of the satellite, ie telemetry.

Since there are a large number of active satellites, and they are being added to regularly, no list is given here. Data is available in the *RSGB Yearbook* or, in a more up-to-date form, from AMSAT UK [1]. There are currently (1996) 19 active satellites in orbit.

Table 9.1. Satellite operating modes

Mode	Input ('uplink') (MHz)	Output ('downlink') (MHz)
A	145	29
B	435	145
J	145	435
K	21	29
L	1269	435
S	435	2401
T	21	145

BAND PLANS

Most satellites are built and recommended for use with a low duty cycle. This means that Morse or SSB are the normal modes with recommended use as follows:

Morse:	The lower one-third of the received signal section.
Mixed Morse/SSB:	The middle one-third of the band.
SSB:	The upper one-third of the receive section.

MODES

Modes are the names of the system of input and output frequencies used in each satellite. They are shown in Table 9.1.

A frequency list of all current amateur satellites is available from [1] together with a satellite information package. Each costs £1.94 at present (1996).

ANTENNAS

As will be realised, the best communication is obtained from beam antennas for each band which are fully steerable both in azimuth and elevation.

In some cases, simple dipoles for 145MHz and 29MHz running east-west should enable the satellite to be used for 80% of each pass. This is especially true for LEO (low earth orbiters) devices.

WHERE TO FIND ORBITAL INFORMATION?

The Oscalator and Orbital Calendar are all that is required to find a particular satellite. A computer is not necessary. These are available from [1] at a small charge of £3.00 for *two* months or £18.00 for a year to members of AMSAT-UK. At present (1996), membership costs a minimum of £13.50 per year. This just covers costs and more would be very welcome.

A more complex method using *Kepler elements* [2, 3] is also available from the same source together with tracking software for the IBM PC. Books on the subject are available

Code of practice for satellite users

1. Ensure that your down link is the best possible. No other factor will help more than a really low-noise, high-sensitivity receive system (see Chapter 4). Use the best, lowest-loss, feeders and the shortest runs possible for the up-link and down-link systems. Every decibel counts if the signal has to travel 40,000km *and* back! It should be easy to hear the up-link signal on the down-link without using excessive power (see below).
2. Use the *absolute minimum* of power (ERP is meant here) to make the contact. The return signal should be *no stronger* than the beacon when that is available. No attempt to use the satellite should be made unless the beacon can be heard. Some satellites need as little as 10W ERP. Do not attempt satellite communication unless good power control is possible with the transmitter in use.
3. Observe the amateur code of politeness. Listen before transmitting and be sure that the channel is clear. Join others on the frequency if appropriate. Please do not put out a long CQ call because that is unlikely to be answered.
4. If you are in doubt or requiring help, there are many users who will be glad to help. To find one, listen to the AMSAT net on 144.280MHz in many parts of the UK or the net on 3.780MHz at 10.15 local time on Sunday or 19.00 local time on Monday or Wednesday or ask AMSAT UK [1].

from the RSGB or from AMSAT-UK. An SASE or some IRCs will bring lists and an application form to join AMSAT-UK.

GETTING STARTED

With modern transceivers for the modes to be used, the only other requisite is an antenna system. As mentioned above steerable beams are preferred, but access can be achieved using simple crossed dipoles at least for the lower frequency bands (21, 28 and 144MHz). The need is for a system which has either circular polarisation (see Chapter 5) or has changeable polarisation because polarisation of signals from the satellite are variable by nature and, in any case, may change on transmission through the ionosphere. A start can be made by listening on the beacon frequencies (listed in the data from [1]) which are lower than the downlink frequency by 50–100kHz. Having found the downlink frequency, listen to the traffic and get used to the change of frequency caused by the Doppler effect – the frequency is higher when the satellite is approaching, it falls at its nearest point and is lower when it is going away.

Having become accustomed to the various effects, make an attempt to join a QSO at a suitable break, bearing in mind the suggestions above in the code of practice.

REFERENCES

[1] Contact AMSAT-UK via Ron Broadbent, MBE, G3AAJ, Hon Sec, AMSAT-UK, London, E12 5EQ. Tel: 0181-989 6741 (+44 181-989 6741 for overseas readers). Compuserve ID: 100024,614. Internet at: rbroadbent@ee.surrey.ac.uk.

[2] *Radio Communication Handbook*, ed Dick Biddulph, G8DPS, RSGB, 1994, Chapter 18.

[3] *Space Radio Handbook*, John Branegan, GM4IHJ, RSGB, 1992, Chapter 4.