7 HF Transmitters and Transceivers

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The purpose of a transmitter is to generate RF energy which may be keyed or modulated and thus employed to convey intelligence to one or more receiving stations. This chapter deals with the design of that part of the transmitter which produces the RF signal, while keying, data modulation etc are described separately in other chapters.

Transmitters operating on frequencies between 1.7 and 30MHz only are discussed here; methods of generating frequencies higher than 30MHz are contained in other chapters. Where the frequency-determining oscillators of a combined transmitter and receiver are common to both functions, the equipment is referred to as a transceiver; the design of such equipment operating in the HF spectrum is also included in this chapter.

One of the most important requirements of any transmitter is that the desired frequency of transmission shall be maintained within fine limits to prevent interference with other stations and to ensure that the operator remains within the allowed frequency allocation.

Spurious frequency radiation capable of causing interference with other services, including television and radio broadcasting, must also be avoided. These problems are considered in the chapter on electromagnetic compatibility.

The simplest form of transmitter is a single-stage, self-excited oscillator coupled directly to an antenna system: **Fig 7.1(a)**. Such an elementary arrangement has, however, three serious limitations:

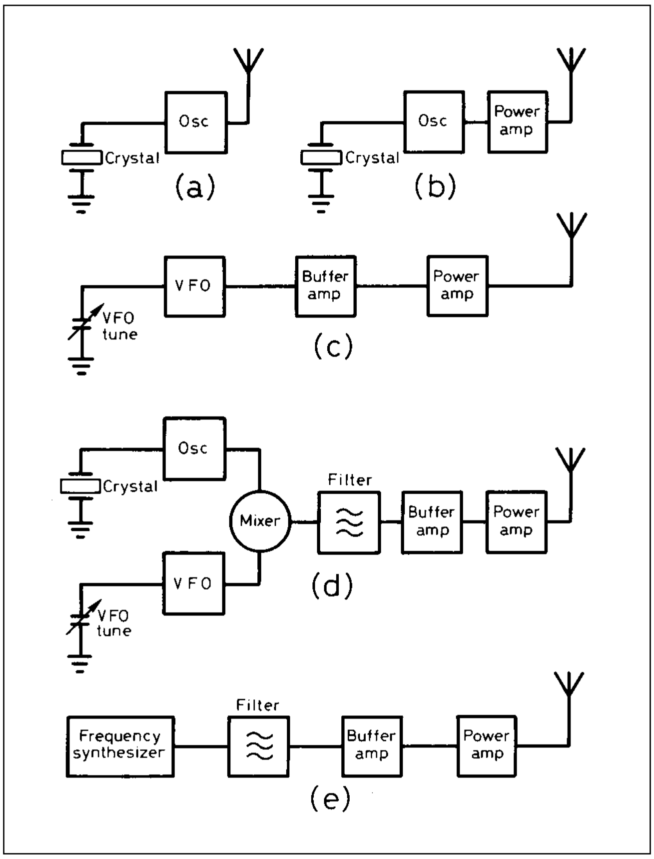


Fig 7.1: Block diagrams of basic transmitter types

- The limited power which is available with adequate frequency stability;
- The possibility of spurious (unwanted) radiation;
- The difficulty of securing satisfactory modulation or keying characteristics.

In order to overcome these difficulties, the oscillator must be called upon to supply only a minimum of power to the following stages. Normally an amplifier is used to provide a constant load on the oscillator: **Fig 7.1(b)**. This will prevent phase shifts caused by variations in the load from adversely affecting the oscillator frequency.

A FET source follower with its characteristic high input impedance makes an ideal buffer after a VFO. In practice, a two-stage buffer amplifier is often used, **Fig 7.1(c)**, incorporating a source follower followed by a Class A amplifier.

Transmitters designed for use on more than one frequency band often employ two or more oscillators; these signals are mixed together to produce a frequency equal to the sum or difference of the original two frequencies, the unwanted frequencies being removed using a suitable filter. This heterodyne transmitter, **Fig 7.1(d)**, is very similar in operation to the superhet receiver. More recently, frequency synthesisers have become popular for the generation of RF energy, **Fig 7.1(e)**, and have almost entirely replaced conventional oscillators in commercial amateur radio equipment. Synthesisers are ideally suited for microprocessor control and multi-frequency coverage (see the chapter on the POC-controlled transceiver). Oscillators are covered in detail in the earlier chapter, Building Blocks: Oscillators.

INTERSTAGE COUPLING

Correct impedance matching between a stage and its load is important if the design power output and efficiency for the stage are to be achieved. The load may be a succeeding stage or an antenna, and correct matching is essential for achieving efficient operation. In order to keep dissipation in the stage within acceptable limits, the load impedance is often higher than that needed for maximum power transfer. Thus a transmitter designed to drive a 50Ω load usually has a much lower source impedance than this. The output power available from a stage can be calculated approximately from the formula:

$$P_{out} = \frac{V_{cc}^2}{2_{ZL}}$$

where ZL is the load impedance in ohms.

Determination of the base input impedance of a stage is difficult without sophisticated test equipment, but if the output of a stage is greater than 2W its input impedance is usually less than 10Ω and possibly even as low as 1 to 2Ω. For this reason some kinds of LC matching do not lend themselves to this application. With the precise input impedance of a stage unknown, adjustable LC networks are often employed as these lend themselves best to matching a wide range of impedances. Additionally, a deliberate mismatch may be introduced by the designer to control power distribution and aid stability.

In the interest of stability it is common practice to use low-Q networks between stages in a solid-state transmitter, but the penalty is poor selectivity and little attenuation of harmonic or

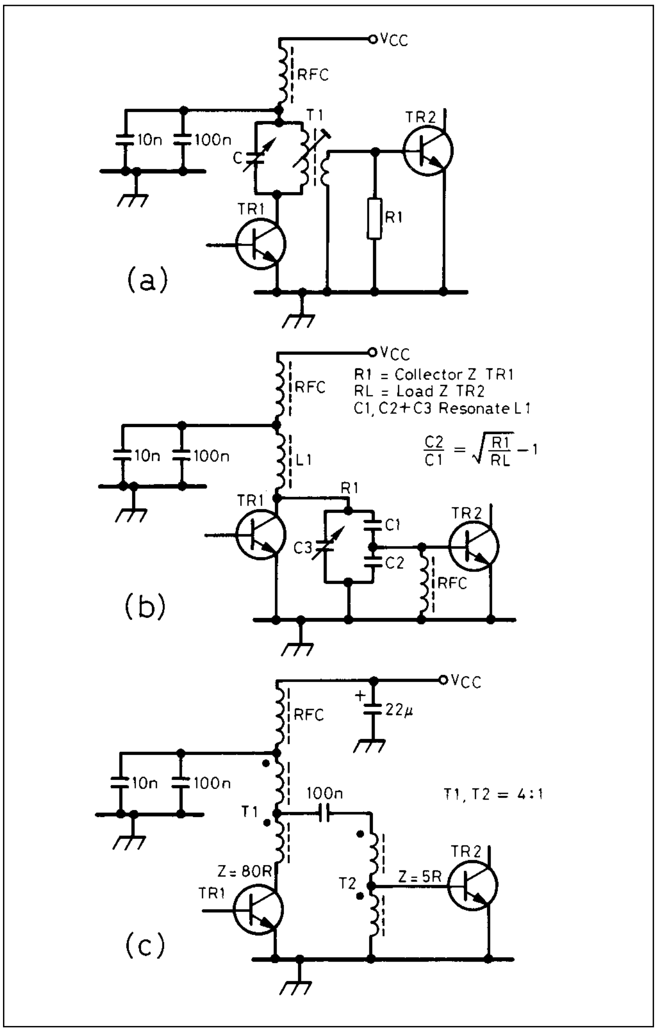


Fig 7.2: Interstage coupling. (a) Transformer coupling; (b) capacitive divider coupling; (c) broad-band transformer matching

spurious energy. Most solid-state amplifiers use loaded Qs of 5 or less compared to Qs of 10 to 15 found in valve circuitry. All calculations must take into account the input and output capacitance of the solid-state devices, which must be included in the network calculations. The best source of information on input and output capacitance of power transistors is the manufacturers' data sheets. Impedances vary considerably with frequency and power level, producing a complex set of curves, but input and output capacitance values are independent of these parameters.

Transformers have always been a popular choice for interstage coupling with solid-state devices: **Fig 7.2(a)**. Toroidal transformers are the most efficient and satisfactory up to 30MHz, the tapping and turns ratio of T1 being arranged to match the collector impedance of TR1 to the base impedance of TR2. In some circumstances only one turn may be required on the secondary winding. A low value of resistor is used to slug this secondary winding to aid stability.

An alternative to inductive coupling is to use a capacitive divider network which resonates L1 as well as providing a suitable impedance tap: **Fig 7.2(b)**. Suitable RF chokes wound on high-permeability (μ = 800) ferrite beads, adequately decoupled, are added to aid stability. This circuit is less prone to VHF parasitics than the one in Fig 7.2(a), especially if C2 has a relatively high value of capacitance.

When the impedance values to be matched are such that specific-ratio broad-band transformers can be used, typically 4:1 and 16:1, one or more fixed-ratio transformers may be cascaded as shown in **Fig 7.2(c)**. This arrangement also exhibits a lack of selectivity due its low Q value, but has the advantage of offering broad-band characteristics. Dots are normally drawn on to transformers to indicate the phasing direction of the windings (all dots start at the same end).

Simple LC networks provide practical solutions to matching impedances between stages in transmitters. It is assumed that normally high output impedances will be matched to lower-value input impedances. However, if the reverse is required, networks can be simply used in reverse to effect the required transformation.

Networks 1 and 2, **Figs 7.3(a) and 7.3b)**, are variations on the L-match and may used fixed or with adjustable inductors and capacitors. Network 3, **Fig 7.3(c)**, is used by many designers because it is capable of matching a wide range of impedances. It is a low-pass T-network and offers harmonic attenuation to a degree determined by the transformation ratio and the total network Q. For stages feeding an antenna, however, additional harmonic suppression will normally be required. The value of Q may vary from 4 to 20 and represents a compromise between bandwidth and attenuation.

Network calculations are shown in **Table 7.1**.

Conventional broad-band transformers are very useful in the construction of solid-state transmitters. They are essentially devices for the transformation of impedances relative to the ratio of the transformer and are not specific impedance devices, eg a solid-state PA with a nominal 50Ω output impedance may employ a 3:1 ratio broad-band output transformer. The impedance ratio is given by the square of the turns ratio and will be 9:1. The impedance seen by the collector is thus:

$$Z_C = \frac{50}{9} = 5.55\Omega$$

The power delivered to the load is then determined by:

$$P_{out} = \frac{V_{CC}^2}{2Z_C}$$

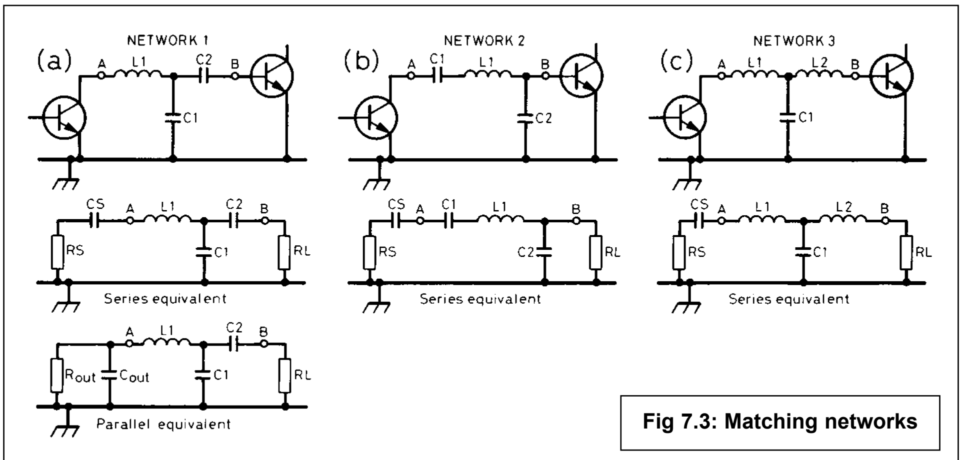


Fig 7.3: Matching networks

Another type of broad-band transformer found in transmitting equipment is the transmission-line transformer. This acts as a conventional transformer at the lower frequencies but, as the frequency increases, the core becomes less 'visible' and the transmission-line properties take over. The calculations are complex, but it is well known that a quarter-wavelength of line exhibits impedance-transformation properties. When a quarter-wave line of impedance Z_0 is terminated with a resistance R_1, a resistance R_2 is seen at the other end of the line.

$$Z_0^2 = R_1 R_2$$

Transmission-line transformers are often constructed from twisted pairs of wires wound onto a ferrite toroidal core having a initial permeability (μ) of at least 800.

This ensures relatively high values of inductance with a small number of turns. It should be borne in mind that with high-power solid-state transmitters some of the impedances to be transformed are very low, often amounting to only a few ohms.

ANODE TANK CIRCUITS

Whilst solid state devices have replaced valves in most amateur equipment, high power valve linear amplifiers are likely to be in use for some considerable time.

Pi Network

The pi-tank (**Fig 7.4**) has been the most popular matching network since the introduction of multiband transmitters in the 1950s. It is easily bandswitched and owes its name to its resemblance to the Greek letter π.

The anode tank circuit must meet the following conditions:

1 The anode circuit of the valve must be presented with the proper resistance in relation to its operating conditions to ensure efficient generation of power.

2. This power must be transferred to the output without appreciable loss.

3 The circuit Q must be sufficient to ensure good flywheel action in order to achieve a close approximation of a sinusoidal RF output voltage. This is especially important in Class C amplifiers where the drive from the valve is in the form of a series of pulses of RF energy.

Tank circuit Q

In order to quantify the ability of a tank circuit to store RF energy (essential for flywheel action), a quality factor Q is defined. Q is the ratio of energy stored to energy lost in the circuit.

$$Q = 2\pi \frac{W_S}{W_L} = \frac{X}{R}$$

where W_S is the energy stored in the tank circuit; W_L is the energy lost to heat and the load; X is the reactance of either the inductor or capacitor in the tank circuit; R is the series resistance.

Since both circulating current and Q are inversely proportional to R, then the circulating current is proportional to Q. By Ohm's Law, the voltage across the tank circuit components must also be proportional to Q.

When the circuit has no load the only resistance contributing to R are the losses in the tank circuit. The unloaded Q_U is given by:

$$Q_U = \frac{X}{R_{loss}}$$

where X is the reactance in circuit and R_{loss} is the sum of the resistance losses in the circuit.

Network 1:

1. Select Q_1
2. $X_{L1} = Q_1 R_S + X_{Cs}$
3. $X_{C2} = Q_L R_L$
4. $X_{C1} = \frac{R_V}{Q_1 - Q_L}$

where:

Loaded Q $= Q_L = \sqrt{\left(\frac{R_S(1+Q_1^2)}{R_L}\right)} - 1 = \sqrt{\frac{R_V}{R_L}} - 1$

$R_V = R_S(1 + Q_1^2)$ = Virtual resistance of network

Network 2:

When $R_S < R_L$

1. Select Q_1
2. $X_{C1} = Q_1 R_S$
3. $X_{C2} = R_L \sqrt{\frac{R_s}{R_L - R_s}}$
4. $X_{L1} = X_{C1} + \left(\frac{R_S R_L}{X_{C2}}\right) + X_{Cs}$

Network 3:

1. Select Q_1
2. $X_{L1} = (R_S Q_1) + X_{Cs}$
3. $X_{L2} = R_L Q_L$
4. $X_{C1} = \frac{R_V}{Q_1 + Q_L}$

where:

$R_V = R_S(1 + Q_1^2)$

$Q_L = \sqrt{\left(\frac{R_V}{R_L}\right)} - 1$ = Loaded Q

Table 7.1: Calculations for the three networks shown in Fig 7.3

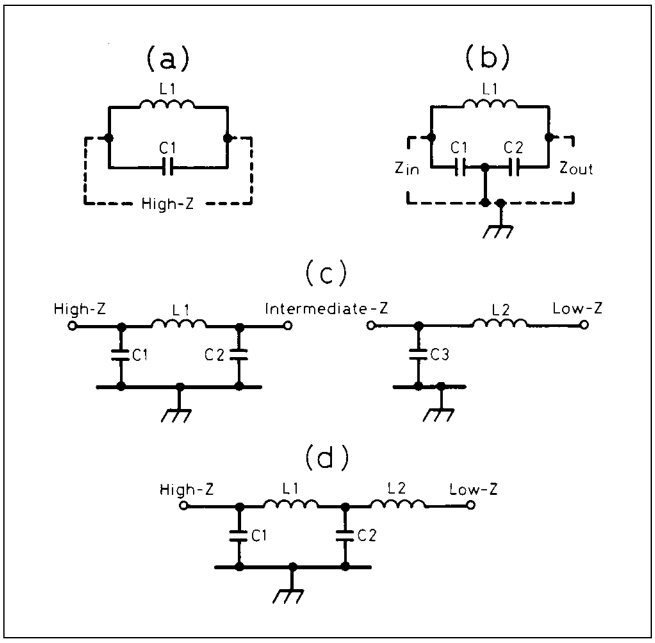


Fig 7.4: Pi-network and derivation. (a) Parallel-tuned tank circuit; (b) parallel-tuned tank with capacitive tap; (c) pi-network and L-network; (d) pi-L network

Table 7.2: Pi-network values for selected anode loads (QL = 12) and load (antenna impedance) of 50 ohms

	MHz	1500	2000	2500	3000	3500	4000	5000	6000	8000
	1.8	708	531	424	354	303	264	229	206	177
	3.5	364	273	218	182	156	136	118	106	91
C1	7.0	182	136	109	91	78	68	59	53	46
(pF)	14.0	91	68	55	46	39	34	30	27	23
	21.0	61	46	36	30	26	23	20	18	15
	28.0	46	34	27	23	20	17	15	13	11
	1.8	3413	2829	2415	2092	1828	1600	1489	1431	1392
	3.5	1755	1455	1242	1076	940	823	766	736	716
C2	7.0	877	728	621	538	470	411	383	368	358
(pF)	14.0	439	364	310	269	235	206	192	184	179
	21.0	293	243	207	179	157	137	128	123	119
	28.0	279	182	155	135	117	103	96	92	90
	1.8	12.81	16.60	20.46	24.21	27.90	31.50	36.09	39.96	46.30
	3.5	6.59	8.57	10.52	12.45	14.35	16.23	18.56	20.55	23.81
L1	7.0	3.29	4.29	5.26	6.22	7.18	8.12	9.28	10.26	11.90
(μH)	14.0	1.64	2.14	2.63	3.11	3.59	4.06	4.64	5.14	5.95
	21.0	1.10	1.43	1.75	2.07	2.39	2.71	3.09	3.43	3.97
	28.0	0.82	1.07	1.32	1.56	1.79	2.03	2.32	2.57	2.98

Table 7.3: Pi-L network values for selected anode loads (Q_L = 12) and load (antenna impedance) of 50 ohms

	MHz	1500	2000	2500	3000	3500	4000	5000	6000	8000
	1.8	784	591	474	397	338	297	238	200	152
	3.5	403	304	244	204	174	153	123	103	78
C1	7.0	188	142	114	94	81	71	57	48	36
(pF)	14.0	93	70	56	47	40	35	29	24	18
	21.0	62	47	38	32	27	23	19	16	12
	28.0	48	36	29	24	21	18	15	13	9
	1.8	2621	2355	2168	2026	1939	1841	1696	1612	1453
	3.5	1348	1211	1115	1042	997	947	872	829	747
C2	7.0	596	534	493	468	444	418	387	368	337
(pF)	14.0	292	264	240	222	215	204	186	172	165
	21.0	191	173	158	146	136	137	125	117	104
	28.0	152	135	127	115	106	107	95	87	86
	1.8	14.047	17.933	21.730	25.466	29.155	32.805	40.011	47.118	61.119
	3.5	7.117	9.086	11.010	12.903	14.772	16.621	20.272	23.873	30.967
L1	7.0	3.900	4.978	6.030	7.070	8.094	9.107	11.108	13.081	16.968
(μH)	14.0	1.984	2.533	3.069	3.597	4.118	4.633	5.651	6.655	8.632
	21.0	1.327	1.694	2.053	2.406	2.755	3.099	3.780	4.452	5.775
	28.0	0.959	1.224	1.483	1.738	1.989	2.238	2.730	3.215	4.171
	1.8	8.917		*The value of L2 remains constant for all values of anode impedance.*						
	3.5	4.518								
L2	7.0	2.476								
(μH)	14.0	1.259								
	21.0	0.843								
	28.0	0.609								

To the tank circuit, a load acts in the same way as circuit losses. Both consume energy but only the circuit losses produce heat energy. When energy is coupled from the tank circuit to the load, the loaded Q (Q_L) is given by:

$$Q_L = \frac{X}{R_{load} + R_{loss}}$$

It follows that if the circuit losses are kept to a minimum the loaded Q value will rise.

Tank efficiency can be calculated from:

$$\text{Tank efficiency (\%)} = \left(1 - \frac{Q_L}{Q_U}\right) \times 100$$

where Q_U is the unloaded Q and Q_L is the loaded Q.

Typically the unloaded Q for a pi-tank circuit will be between 100 and 300 while a value of 12 is accepted as a good compromise for the loaded Q. In order to assist in the design of anode tank circuits for different frequencies, inductance and capacitance values for a pi-network with a loaded Q of 12 are provided in **Table 7.2** for different values of anode load impedance.

Pi-L Network

The pi-L network, **Fig 7.4(d)**, is a combination of the pi-network and the L-network. The pi-network transforms the load resistance to an intermediate impedance, typically several hundred ohms, and the L-network then transforms this intermediate impedance to the output impedance of 50Ω. The output capacitor of the pi-network is in parallel with the input capacitor of the L-network and is combined into one capacitor equal to the sum of the two individual values.

The major advantage of the pi-L network over a pi-network is considerably greater harmonic suppression, making it particularly suitable for high-power linear amplifier applications. A table of values for a pi-L network having a loaded Q of 12 for different values of anode load impedance is given in **Table 7.3**. Both Table 7.2 and Table 7.3 assume that source and load impedances are purely resistive; the values will have to be modified slightly to compensate for any reactance present in the circuits to be matched. Under certain circumstances matching may be compromised by high values of external capacitance, in which case a less-than-ideal value of Q may have to be accepted.

VOICE MODULATION TECHNIQUES

A description of the theory of Amplitude Modulation (AM), Single Sideband (SSB), and Frequency and Phase Modulation (FM and PM) can be found in the Principles chapter.

Although currently still common for broadcasting, AM is now rarely used on the amateur bands. The use of FM and PM is confined to the bands above 30MHz (with the exception of a little activity in the upper part of the 28MHz band), so they are covered in the chapter on VHF/UHF transmitters and receivers. This chapter will deal with the practical aspects of SSB modulation.

An SSB transmitter has a similar architecture to an SSB receiver as shown in **Fig 7.5**. This makes it very suitable for combining into a transceiver, as can be seen later.

BALANCED MODULATORS

Balanced modulators are essentially the same as balanced mixers, balanced demodulators and product detectors; they are tailored to suit different circuit applications especially with respect to the frequencies in use. The balanced modulator is a circuit which mixes or combines a low-frequency (audio) signal with a higher frequency (RF) signal in order to obtain the sum-and-difference frequencies (sidebands); the original RF frequency is considerably attenuated by the anti-phase or balancing action of the circuit.

A *singly* balanced modulator is designed to balance out only one of the input frequencies, either f_1 or f_2, normally the higher frequency. In a *doubly* balanced modulator, both f_1 and f_2 are balanced out, leaving the sum and difference frequencies $f_1 + f_2$ and $f_1 - f_2$. In addition, intermodulation products (IMD) will appear in the form of spurious signals caused by the interaction and mixing of the various signals and their harmonics.

Balanced modulators come in many different forms, employing a wide variety of devices from a simple pair of diodes to complex ICs. In their simplest form they are an adaptation of the bridge circuit, but it should be noted that diodes connected in a modulator circuit are

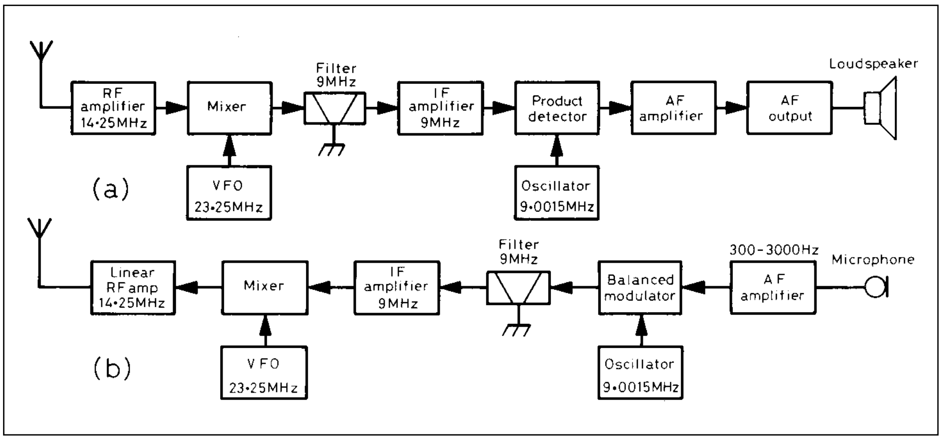


Fig 7.5: (a) Typical single-conversion superhet receiver. (b) Single conversion SSB transmitter

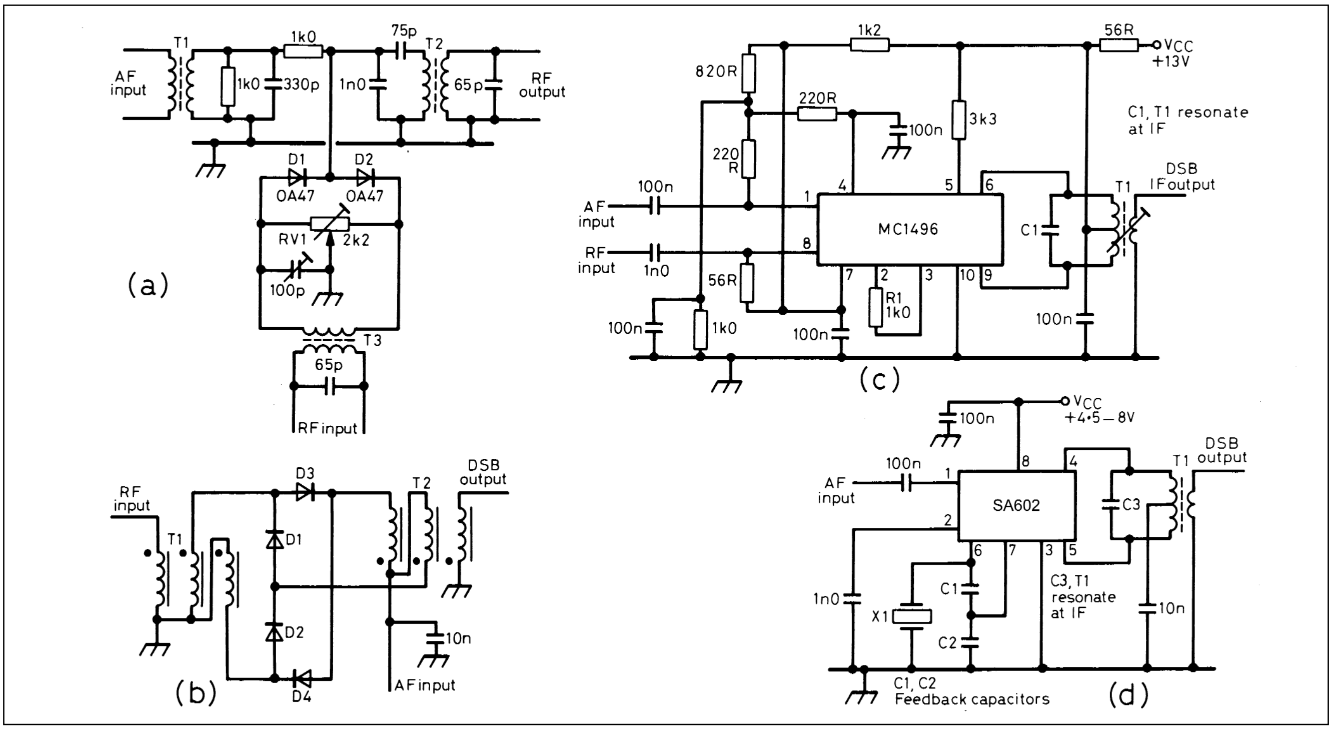


Fig 7.6: Balanced modulators: (a) Singly balanced diode modulator; (b) Doubly balanced diode ring modulator; (c) MC1496 doubly balanced modulator; (d) SA602 doubly balanced modulator with internal oscillator

connected differently to those in a bridge rectifier. Simple diode balanced modulators can provide high performance at low cost. Early designs used point-contact germanium diodes while more recent designs use hot-carrier diodes (HCD). The HCD offers superior performance with lower noise, higher conversion efficiency, higher square law capability, higher breakdown voltage and lower reverse current combined with a lower capacitance. In practice, almost any diode can be used in a balanced modulator circuit, including the ubiquitous 1N914.

In the early days of SSB, simple diode balanced modulators were very popular, easy to adjust and capable of good results. Doubly balanced diode ring modulators have subsequently proved very popular because of their higher performance. However, they incur at least a 6dB signal loss while requiring a high level of oscillator drive. Doubly balanced modulators offer greater isolation between inputs as well as between input and output ports when compared to singly balanced types.

The introduction of integrated circuits resulted in a multitude of ICs suitable for use in balanced modulator and mixer applications. These include the popular Philips SA602 which combines an input amplifier, local oscillator and double balanced modulator in a single package, and the MC1496 double balanced mixer IC from Motorola [1]. The majority of IC mixers are based upon a doubly balanced transistor tree circuit, using six or more transistors on one IC. The major difference between different types of IC lies in the location of resistors which may be either internal or external to the IC.

IC mixers offer conversion gain, lower oscillator drive requirements and high levels of balance, but IMD performance can be inferior to that of diode ring modulators. Some devices, such as the Analog Devices AD831, permit control of the bias current. This allows the designer freedom to improve IMD performance at the expense of power consumption.

Fig 7.6 illustrates a range of practical balanced modulator circuits. The shunt-type diode modulator, Fig 7.6(a), was common in early valve SSB transmitters, offering a superior balance to that achievable with conventional valve circuitry. Fig 7.6(b) shows a simple diode ring balanced modulator capable of very high performance - devices such as the MD108 and SBL1 are derivations of this design. The designs in Figs 7.6(c) and 7.6(d) illustrate the use of IC doubly balanced mixers. The SA602, primarily designed for very-low-power VHF receiver mixer applications, offers simplicity of design and a low external component count. Although most of the devices mentioned are no longer in production they remain popular for homebrew construction and can still be occasionally be found for sale at rallies and on the internet.

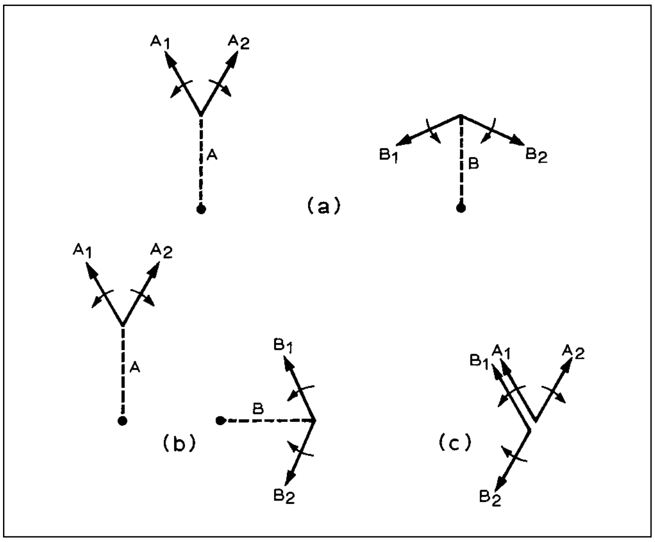


Fig 7.7: Phasing system vectors

GENERATING SSB

The double sideband (DSB) signal generated by the balanced modulator has to be turned into SSB by attenuating one of the sidebands to an acceptable level. A figure of 30-35dB has come to be regarded as the minimum acceptable standard. With care, suppression of 50dB or more is attainable but such high levels of attenuation are of questionable benefit.

The unwanted sideband may be attenuated either by phasing or filtering. The two methods are totally different in conception, and will be discussed in detail.

The Phasing Method

The phasing method of SSB generation can be simply explained with the aid of vector diagrams. Fig 7.7(a) shows two carriers, A and B, of the same frequency and phase, one of which is modulated in a balanced modulator by an audio tone to produce contra-rotating sidebands A1 and A2, and the other modulated by a 90° phase-shifted version of the same audio tone. This produces sidebands B1 and B2 which have a 90° phase relationship with their A counterparts. The carrier vector is shown dotted since the carrier is absent from the output of the balanced modulators. Fig 7.7(b) shows the vector relationship if the carrier B is shifted in phase by 90° and Fig 7.7(c) shows the addition of these two signals. It is evident that sidebands A2 and B2 are in antiphase and therefore cancel whereas A1 and B1 are in phase and are additive. The result is that a single sideband is produced by this process.

A block diagram of a phasing-type transmitter is shown in Fig 7.8 from which will be seen that the output of an RF oscillator is fed into a network in which it is split into two separate components, equal in amplitude but differing in phase by 90°

Similarly, the output of an audio amplifier is split into two components of equal amplitude and 90° phase difference. One RF and one AF component are combined in each of two balanced modulators. The double-sideband, suppressed-carrier energy from the two balanced modulators is fed into a common tank circuit. The relative phases of the sidebands produced by the two balanced modulators are such that one sideband is balanced out while the other is reinforced. The resultant in the common tank circuit is an SSB signal. The main advantages of a phasing exciter are that sideband suppression may be accomplished at the operating frequency and that selection of the upper or lower sideband may be made by reversing the phase of the audio input to one of the

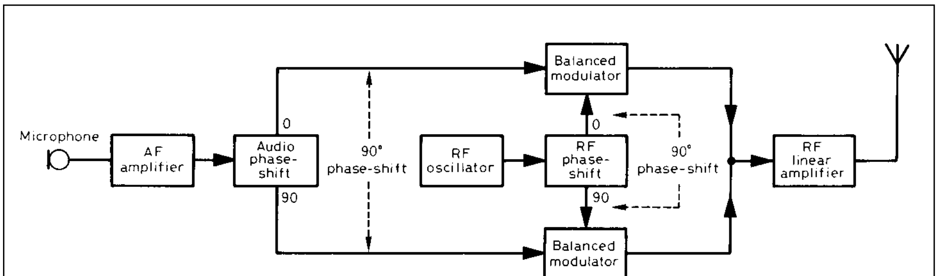


Fig 7.8: Phasing-type exciter

balanced modulators. These facilities are denied to the user of the filter system.

If it were possible to arrange for absolute precision of phase shift in the RF and AF networks, and absolute equality in the amplitude of the outputs, the attenuation of the unwanted sideband would be infinite. In practice, perfection is impossible to achieve, and some degradation of performance is inevitable. Assuming that there is no error in the amplitude adjustment, a phase error of 1° in either the AF or the RF network will reduce the suppression to 40dB, while an error of 2° will produce 35dB, and 3.5° will result in 30dB suppression. If, on the other hand, phase adjustment is exact, a difference of amplitude between the two audio channels will similarly reduce the suppression. A difference between the two voltages of 1% would give 45dB, and 4% 35dB approximately. These figures are not given to discourage the intending constructor, but to stress the need for high precision workmanship and adjustment if a satisfactory phasing-type SSB transmitter is to be produced.

The early amateur phasing transmitters were designed for fundamental-frequency operation, driven directly from an existing VFO tuning the 80m band, and used a low-Q phase-shift network. This low-Q circuit has the ability to maintain the required 90° phase shift over a small frequency range, and this made the network suitable for use at the operating frequency in single-band exciters designed to cover only a portion of the chosen band. The RF phase-shift network is incapable of maintaining the required accuracy of phase shift for operation over ranges of 200kHz or more, and the available sideband suppression deteriorates to a point at which the exciter is virtually radiating a double sideband signal.

For amateur band operation a sideband suppression of 30-35dB and a carrier suppression of 50dB should be considered the minimum acceptable standard. Any operating method that is fundamentally incapable of maintaining this standard should not be used on the amateur bands. For this reason, the fundamental type of phasing unit is not recommended. For acceptable results, the RF phase shift must be operated at a fixed frequency outside the amateur bands. The SSB output from the balanced modulator is then heterodyned to the required bands by means of an external VFO.

Audio phase-shift network

Achieving the audio phase shift necessary for SSB generation in a phasing exciter traditionally required the use of high-tolerance components, often necessitating the use of a commercially made phase-shift network. Such devices can prove more costly than the crystal filter required for a filter-type exciter.

An alternative method of deriving the required 90° phase shift using off-the-shelf values was devised by M J Gingell and is referred to as the polyphase network (**Fig 7.9**). Standard 10% tolerance resistors and capacitors are used in the construction of a six-pole network capable of providing four outputs of equal amplitude, all lagging one another by 90°. The network is designed to phase shift audio signals between 300Hz and 3000Hz, and it is therefore necessary to limit the bandwidth of the audio input using a filter or clipper circuit which can be either active or passive. Audio input to the polyphase network is derived using a simple phase splitter to provide the two phase inputs required by the network. Resistors used in the network are of one common value and Mylar audio-grade capacitors are suitable for the capacitive elements.

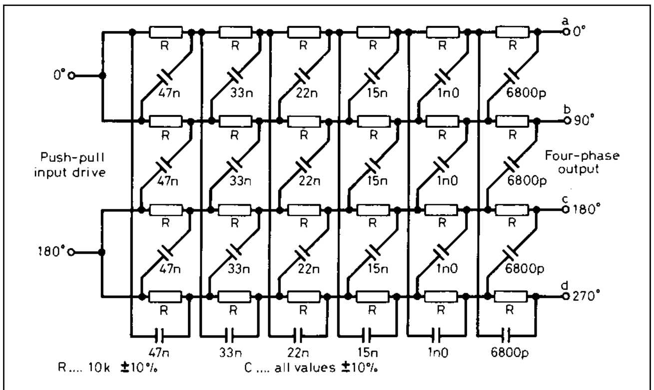


Fig 7.9: Gingell polyphase network

RF phase-shift network

Traditionally the most satisfactory way to produce a 90° RF phase shift was to employ a low-Q network comprising of two loosely coupled tuned circuits which exhibits a combination of inductance, resistance and capacitance.

Fig 7.10(a) illustrates such a network in the anode circuit of a valve amplifier. The primary coils are inductively coupled while the link couplings are connected in series. When both circuits are tuned to resonance there will be exactly 90° phase shift between them. Difficulties occur when the frequency is changed, and the network has to be retuned, restricting the bandwidth to no more than 200kHz.

With the advent of digital ICs, it became relatively easy to obtain the required phase shift by dividing the output of an oscillator using a flip-flop IC. **Fig 7.10(b)** shows the RF circuitry for a 160m phasing exciter, in which the signal from a VFO tuning 7.2-8MHz is divided by four using a 74HC73 (J-K flip-flop), providing a square-wave output between 1.8 and 2MHz.

The VFO is conveniently implemented by one inverter of a 74HC00 (quad NAND-gate), which directly drives the flip-flop producing both 0 and 90 degree outputs. The fundamental square-wave signal will be phased out in the balanced modulator.

Four-way phasing method

The four-way phasing method is an adaptation of the conventional phasing method and can be simply described as a double two-way method. **Fig 7.11** illustrates a four-way phasing exciter. The major requirement for acceptable carrier and sideband suppression is a good audio phase shifter. The polyphase network (Fig 7.9) is ideal and provides the required four output signals at 90° phase intervals. The RF output from the carrier generator must also provide four RF outputs phase-shifted by 90° from one another, and this is achieved by using a dual J-K flip-flop which also divides the input frequency by a factor of four. The phase-shifted AF and RF signals are fed to four modulators, the outputs of which are summed in a tuned adder, resulting in an SSB output signal.

A practical 9MHz four-way SSB generator is illustrated in **Fig 7.12**. The TTL oscillator is operated at 12MHz, providing a 3MHz signal at the output of the flip-flop. As this square wave signal is rich in odd-order harmonics, the tuned adder can be adjusted to tune to the third harmonic in preference to the fundamental signal, and the result is a 9MHz output SSB signal.

One major disadvantage of the digital phase shifter is the necessity to operate the oscillator on four times the output frequency. The technique used in Fig 7.12 represents one solution to the problem, but an alternative would be to heterodyne the output to the desired frequency using a VFO.

The Filter Method

Since the objective is to transmit only a single sideband, it is necessary to select the desired sideband and suppress the unwanted sideband. The relationship between the carrier and

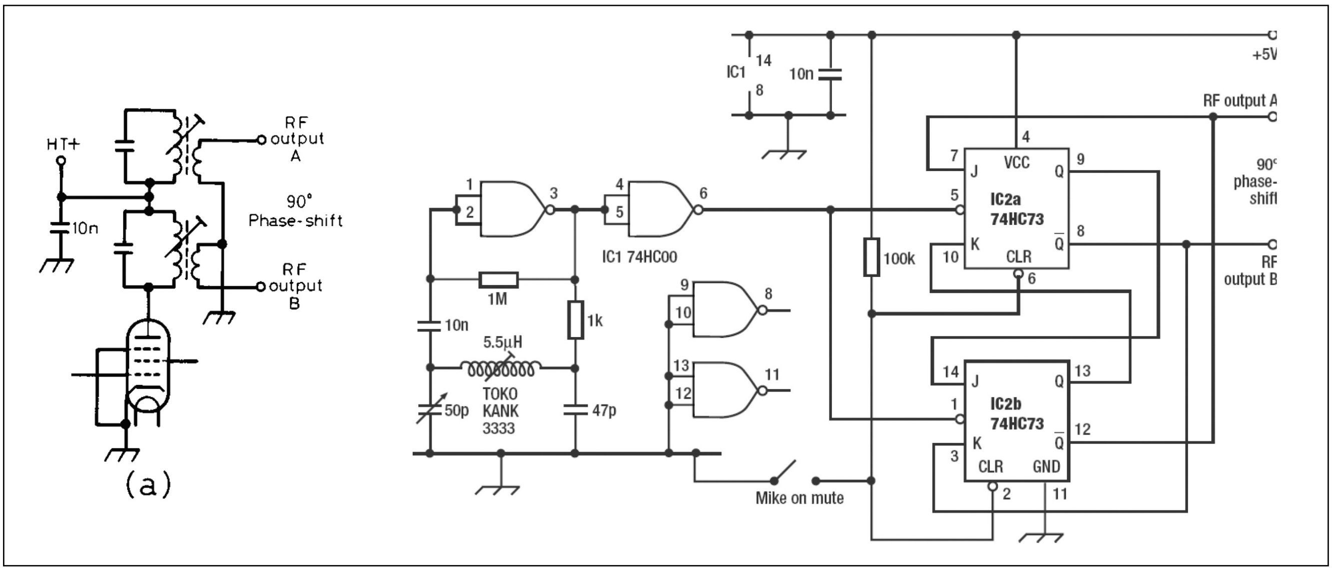


Fig 7.10: Methods of obtaining RF phase shift. (a) Traditional method of obtaining 90° phase shift using loosely coupled tuned circuits. (b) Active RF phase shifter 7.2MHz VFO providing 1.8MHz output with 90° shifts

Fig 7.11: Four-phase SSB generator

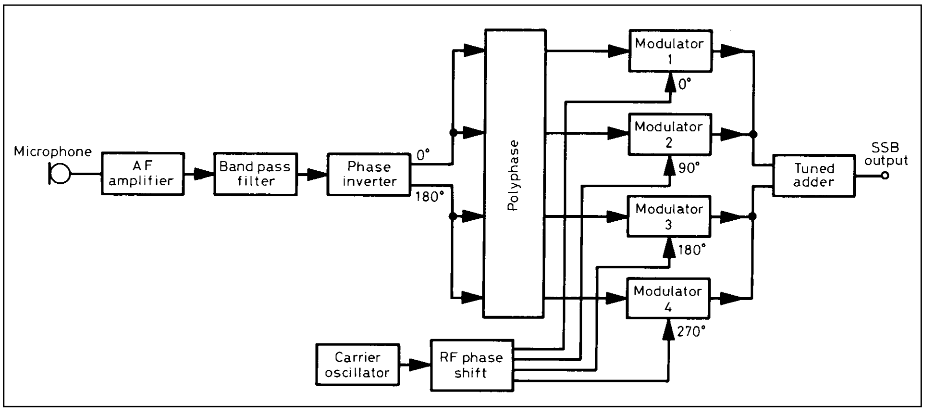
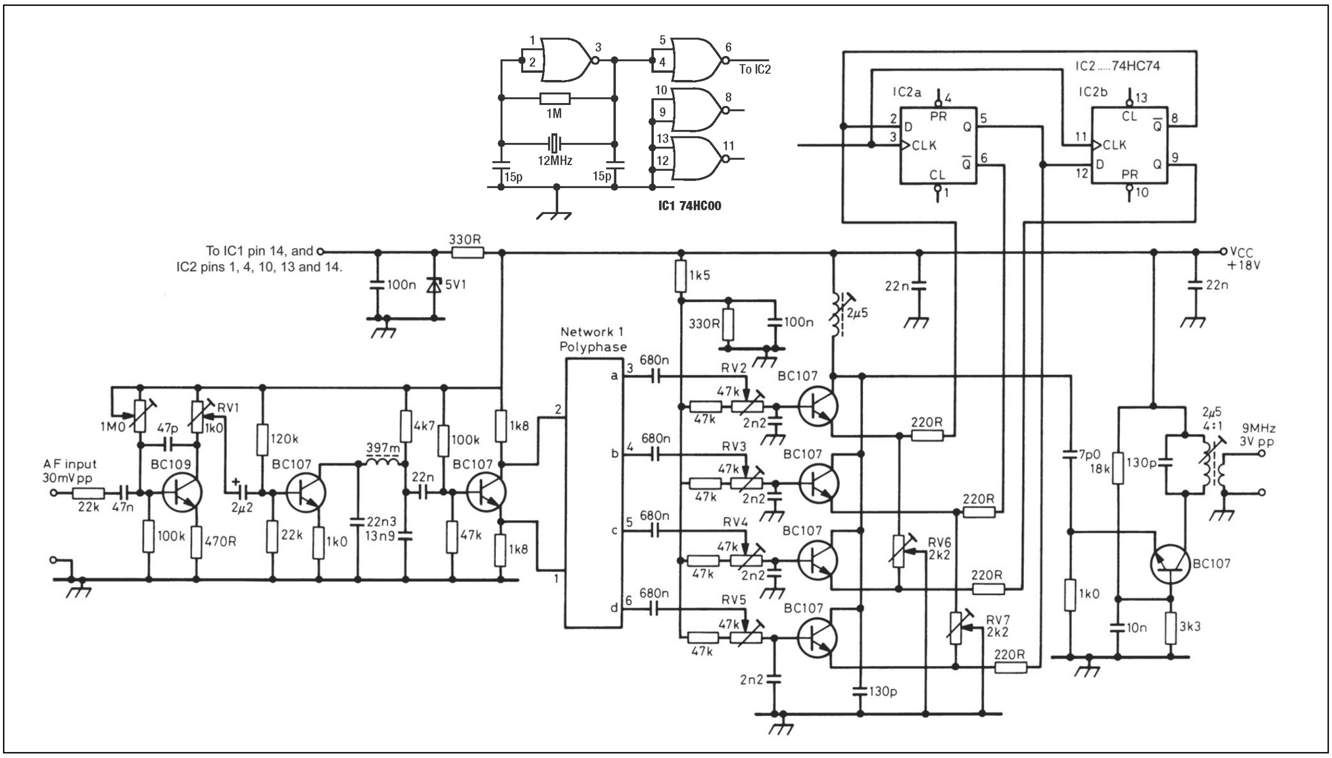


Fig 7.12: 9MHz phasing exciter



sidebands is shown in the Principles chapter. Removing the unwanted sideband by the use of a selective filter has the advantage of simplicity and good stability, and this is therefore the most widely adopted method of generating SSB. The unwanted sideband suppression is determined by the attenuation of the sideband filter, while the stability of this suppression is determined by the stability of the elements used in constructing the filter. High stability can be achieved by using materials that have a very low temperature coefficient. Commonly used materials are quartz, ceramic and metallic plates.

The filter method, because of its proven long-term stability, has become the most popular method used by amateurs. At present three types of selective sideband filters are in common use:

- High-frequency crystal filter
- Low-frequency mechanical filter
- Low-frequency ceramic filter

Crystal filters

The crystal filter is the most widely used type of filter found in SSB transmitters. In a transceiver, one common filter can be used for both transmit and receive functions. Generation of a SSB signal in the 9MHz range permits single-frequency conversion techniques to be employed to cover the entire HF spectrum, and for this reason 9MHz and 10.7MHz filters have virtually dominated the market. A number of other frequencies have also been employed for filters, including 5.2MHz, 3.18MHz and 1.6MHz, the latter mainly for commercial applications.

The principle of operation of a crystal or quartz filter is based upon the piezo-electric effect. When the crystal is excited by an alternating electric current, it mechanically resonates at a frequency dependent upon its physical shape, size and thickness. A crystal will easily pass current at its natural resonant frequency but attenuates signals either side of this frequency (see the chapter on Passive Components).

By cascading a number of crystals having the same, or very closely related, resonant frequencies it is possible to construct a filter having a high degree of attenuation either side of a band of wanted frequencies, typically 40-60dB with a six-pole filter, 60-80dB using an eight-pole filter, and 80-100dB with a 10-pole filter. The characteristic bandwidth of a SSB filter is selected to pass a communications-quality audio spectrum of typically 300-3000Hz. For SSB transmission the best-sounding results can be achieved using a 3kHz wide filter, but for receiver applications a slightly narrower filter is preferable and 2.4kHz has become the accepted compromise. SSB transmissions made using narrower filters have a very restricted audio sound when received. Filter bandwidths are normally quoted at the 6dB and 60dB attenuation levels, and the ratio of the two quoted bandwidths is referred to as the shape factor, 1:1 being the ideal but not realistically achievable. Anything better than 2.5:1 may be regarded as acceptable.

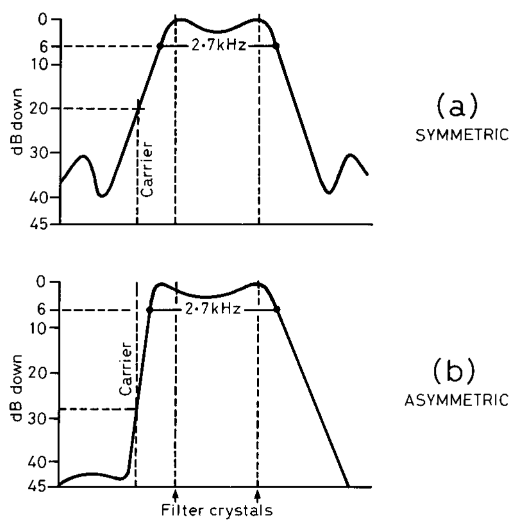


Fig 7.13: Response curves of two- and four-pole symmetric and asymmetric crystal filters

The purpose of the crystal filter is to attenuate the unwanted sideband but it conveniently has a secondary function: to attenuate further the already-suppressed carrier. A balanced modulator seldom attenuates the carrier by more than 40dB, which equates to 1mW of carrier from a 100W SSB transmitter. A further 20dB of carrier suppression is available from the sideband filter, making a carrier suppression of 60dB possible.

The passband of a crystal filter may be symmetrical in shape, Fig 7.13(a), or asymmetric as in Fig 7.13(b). Home-constructed filters are invariably asymmetric to some degree whereas commercially made filters will be designed to fall into either of the two categories. Assuming that we wish to generate a LSB signal with a carrier frequency of 9MHz, a SSB filter will be required to pass the frequency range 8.9975MHz to 9.0MHz, whereas if we wished to change to USB, the filter would be required to pass the frequency range 9.0MHz to 9.0025MHz. At first sight it would appear that two filters are required. In commercial equipment, the use of two filters has been common practice, most probably asymmetric and annotated with the sideband that they are designed to generate and the intended carrier frequency. In amateur radio equipment a much cheaper technique is adopted: it is easier to use a symmetrical filter with a centre frequency of 9.000MHz and move the carrier frequency from one side of the filter to the other in order to change sidebands. Typically 8.9985MHz for USB and 9.0015MHz for LSB are used; note the filter frequency is above the carrier frequency to give USB and below the carrier frequency to give LSB. Asymmetric filters are invariably marked with the carrier frequency and have the advantage of higher attenuation of the unwanted carrier and sideband, due to the steeper characteristic of the filter on the carrier frequency side. This is the primary reason why they are used in commercial applications where they are required to meet a higher specification. One minor disadvantage of the symmetrical filter and switched carrier frequency method is that changing sideband causes a shift in frequency approximately equal to the bandwidth of the filter. This can be compensated for by an equal and opposite movement of the frequency-conversion oscillator.

Radio amateurs have adopted the practice of operating LSB on the low-frequency bands and USB on the high-frequency ones. It is therefore necessary to be able to switch sidebands if operation on all bands is contemplated.

Home construction of crystal filters is only to be recommended if a supply of cheap crystals is available. Fortunately there are now several sources. Clock crystals for microprocessor applications are manufactured in enormous quantities and cost pence rather than pounds. Another source of crystals is those intended for TV colour-burst; the UK and continental frequency of 4.43MHz is the more suitable as the USA colour-burst frequency is in the 80m amateur band. Fig 7.14(a) shows an eight-pole ladder filter designed by G3UUR and constructed using colour-burst crystals. The frequency of individual crystals is found to vary by as much as 200Hz, but by careful selection of crystals it is possible to find those that are on frequency and those that are slightly above or below the nominal frequency.

For optimum results it is recommended that the on-frequency crystals are located at either end of the filter while the centre four crystals should be slightly higher in frequency. **Fig 7.14(b)** shows the frequency response of the G3UUR filter constructed from TV colour-burst crystals.

Commercially available crystal filters are primarily confined to 9MHz and 10.7MHz types, though the sources seem to come and go. Numerous filters are available new, and surplus filters can often be found at bargain prices. **Table 7.4** lists the KVG range of filters - alternative filters are listed in **Table 7.5**.

Mechanical filters

The mechanical filter was developed by the Collins Radio Company for low-frequency applications in the range 60-500kHz. The F455 FA-21 was designed specifically for the amateur radio market with a nominal centre frequency of 455kHz, a 6dB bandwidth of 2.1kHz and a 60dB bandwidth of 5.3kHz, providing a shape factor of just over 2:1.

The mechanical filter is made up from a number of metal discs joined by coupling rods. The discs are excited by magnetostrictive transducers employing polarised biasing magnets and must not be used in circuits where DC is present. The input and output transducers are identical and are balanced to ground so the filter can be used in both directions. Mechanical filters have always been expensive but provide exceptional performance, and examples of the Collins filters can often be found in surplus equipment. Japanese Kokusai filters were available for a number of years and offered a more cost-effective alternative. Mechanical filters are now seldom used in new equipment.

Ceramic filters

Ceramic filters have been developed for broadcast radio applications; they are cheap, small and available in a wide range of frequencies and bandwidths. Narrow-bandwidth ceramic filters with a nominal centre frequency of 455kHz have been manufactured for use in SSB receiver IF applications and provide a level of performance making them ideally suited for use in SSB transmitters. Bandwidths of 2.4 and 3kHz are available with shape factors of 2.5:1 and having attenuation in excess of 90dB (Table 7.5).

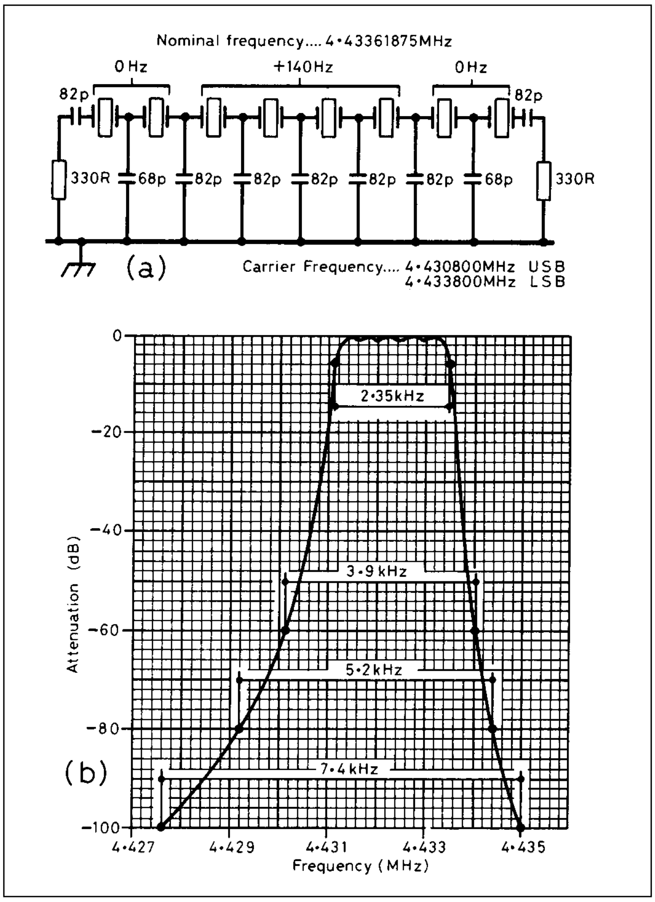


Fig 7.14: (a) Ladder crystal filter using 4.43MHz TV crystals. (b) Ladder crystal filter using P129 specification colour-TV 4.43MHz crystals. It provides a performance comparable with the ladder filter in the Atlas 180 and 215 transceivers. Insertion loss 4-5dB, shape factor (6/60dB) 1.66. Note that the rate of attenuation on the low-frequency side of the response is as good as an eight-crystal lattice design; on the HF side it is better

Filter type	XF-9A	XF-9B	XF-9C	XF-9D	XF-9E	XF-9M
Application	SSB TX	SSB TX/RX	AM	AM	FM	CW
Number of poles	5	8	8	8	8	4
6dB bandwidth (kHz)	2.5	2.4	3.75	5.0	12.0	0.5
Passband ripple (dB)	<1	<2	<2	<2	<2	<1
Insertion loss (dB)	<3	<3.5	<3.5	<3.5	<3	<5
Termination	500 /30pF	500 /30pF	500 /30pF	500 /30pF	1200 /30pF	500 /30pF
Shape factor	1.7 (6-50dB)	1.8 (6-60dB)	1.8 (6-60dB)	1.8 (6-60dB)	1.8 (6-60dB)	2.5 (6-40dB)
		2.2 (6-80dB)	2.2 (6-80dB)	2.2 (6-80dB)	2.2 (6-80dB)	4.4 (6-60dB)
Ultimate attenuation (dB)	>45	>100	>100	>100	>90	>90

Table 7.4: KVG 9MHz crystal filters for SSB, AM, FM and CW applications

Filter type	90H2.4B	10M02DS	CFS455J	CFJ455K5	CFJ455K14	QC1246AX
Centre frequency (MHz)	9.0000	10.7000	0.455	0.455	0.455	9.0000
Number of poles	8	8	Ceramic	Ceramic	Ceramic	8
6dB bandwidth (kHz)	2.4	2.2	3	2.4	2.2	2.5
60dB bandwidth (kHz)	4.3	5	9 (80dB)	4.5	4.5	4.3
Insertion loss (dB)	<3.5	<4	8	6	6	<3
Termination	500 /30pF	600 /20pF	2k	2k	2k	500 30pF
Ultimate attenuation (dB)	>100	-	>60	>60	60	>90
Manufacturer/UK supplier	IQD	Cirkit	Cirkit	Bonex	Bonex	SEI

Table 7.5: Popular SSB filters

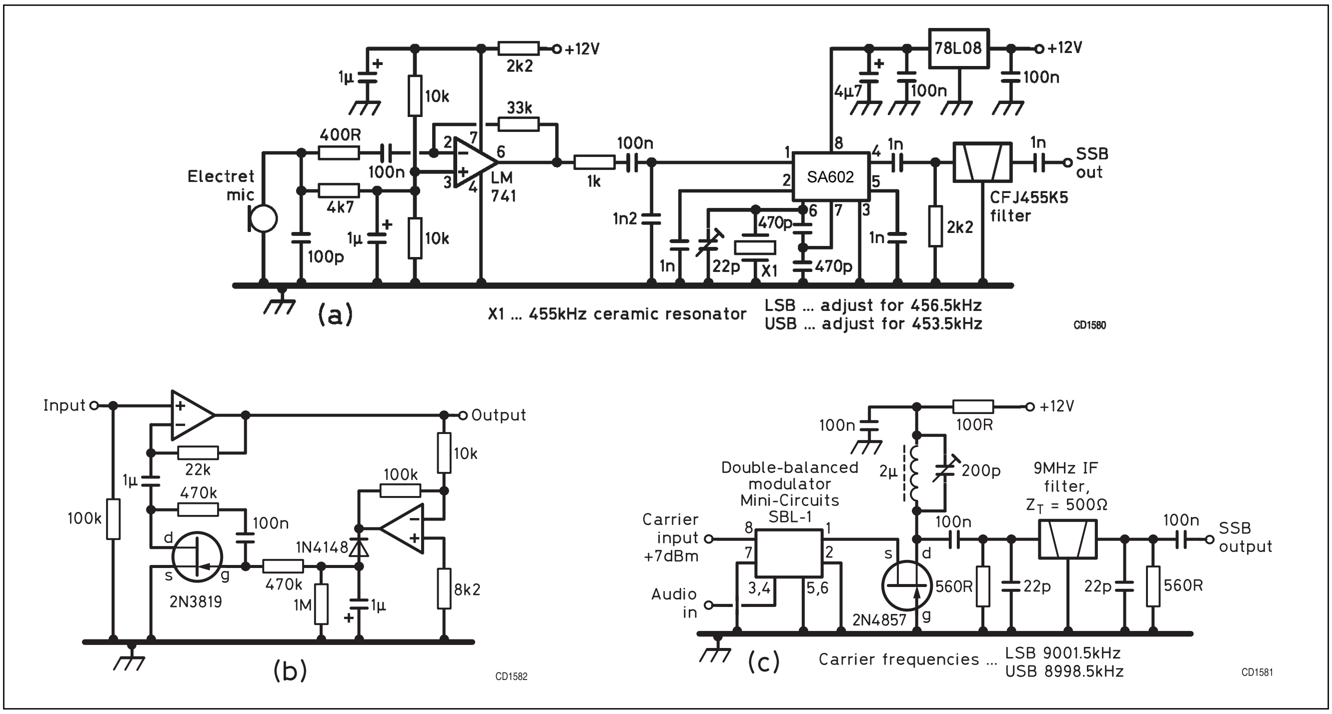


Fig 7.15: (a) 455kHz SSB generator. (b) Example of an audio AGC circuit. (c) 9MHz SSB generator

SSB filter exciter

The SSB filter exciter is inherently simpler than the phasing-type exciter. Fig 7.15(a) illustrates a 455kHz SSB generator requiring just two ICs and using a ceramic filter. The SA602 doubly balanced modulator is provided with a 455kHz RF input signal from its own internal oscillator. This can use a 455kHz crystal or the cheaper ceramic resonator, and the latter can be pulled in frequency to generate either a LSB or USB carrier. A DSB signal at the output of the balanced modulator is fed directly to the ceramic filter for removal of the unwanted sideband, resulting in a 455kHz SSB output signal. This can be heterodyned directly to the lower-frequency amateur bands or via a second higher IF to the HF bands. The circuit performance can be improved by the addition of an audio AGC circuit, such as the one shown in Fig 7.15(b). Fig 7.15(c) illustrates a design for a 9MHz SSB generator. Microphone audio would need to pass through a suitable audio amplifier before being applied to the balanced modulator at pins 3 and 4. Output from the 9MHz SSB exciter can be heterodyned directly to any of the LF and HF amateur bands.

The only adjustment required in either circuit Fig 7.15(a) or (c) is to adjust the oscillator to the correct frequency in relation to the filter; this frequency is normally located 20dB down the filter response curve. Quite often the oscillator can simply be adjusted for the best audio response at the receiver.

POWER AMPLIFIERS

The RF power amplifier is normally considered to be that part of a transmitter which provides RF energy to the antenna. It may be a single valve or transistor, or a composite design embodying numerous devices to take low-level signals to the final output power level. RF amplifiers are also discussed in detail in the Building Blocks chapter.

Push-pull valve amplifiers, Fig 7.16(a), were popular until the 1950s and offered a number of advantages over single-ended output stages as in Fig 7.16(b). The inherent balance obtained when two similar valves having almost identical characteristics are operated in push-pull results in improved stability, while even-order harmonics are phased out in the common tank circuit. One major shortcoming of the push-pull valve amplifier is that switching of the output tank circuit for operation on more than one frequency band is exceedingly difficult due to the high RF voltages present.

During the 1960s, push-pull amplifiers were superseded by the single-ended output stage, often comprising two valves in parallel, Fig 7.16(c), coupled to the antenna via a pi-output network and low-impedance coaxial cable. The changes in design were brought about by two factors. First, the rapid expansion in TV broadcasting introduced problems of harmonically related TVI. This demanded greater attenuation of odd-order harmonics than was possible with the conventional tank circuit used in push-pull amplifiers. Second, there was a trend towards the development of smaller, self-contained transmitters capable of multiband operation, ultimately culminating in the transceiver concept which has almost totally replaced the separate transmitter and receiver in amateur radio stations.

The pi-output tank circuit differs from the conventional parallel-tuned tank circuit in that it uses two variable capacitors in series: Figs 7.4(a) and 7.4(b). The junction of the two capacitors is grounded and the network is isolated from DC using a high-voltage blocking capacitor. The input capacitor has a low value, while the output capacitor has a considerably higher value. This provides high-to-low impedance transformation across the network, enabling the relatively high anode impedance of the power amplifier to be matched to an output impedance of typically 75 or 50Ω. The pi-tank circuit performs the function of a low-pass filter and provides better attenuation of harmonics than a link-coupled tank circuit. The low-impedance output is easily band-switched over the entire HF spectrum by shorting out turns, and facilitates direct connection to a dipole-type antenna, an external low-pass filter for greater reduction of harmonics or an antenna matching unit for connection to a variety of antennas.

The Class C non-linear power amplifiers commonly used in CW and AM transmitters produced high levels of harmonics, making compatibility with VHF TV transmissions exceedingly difficult.

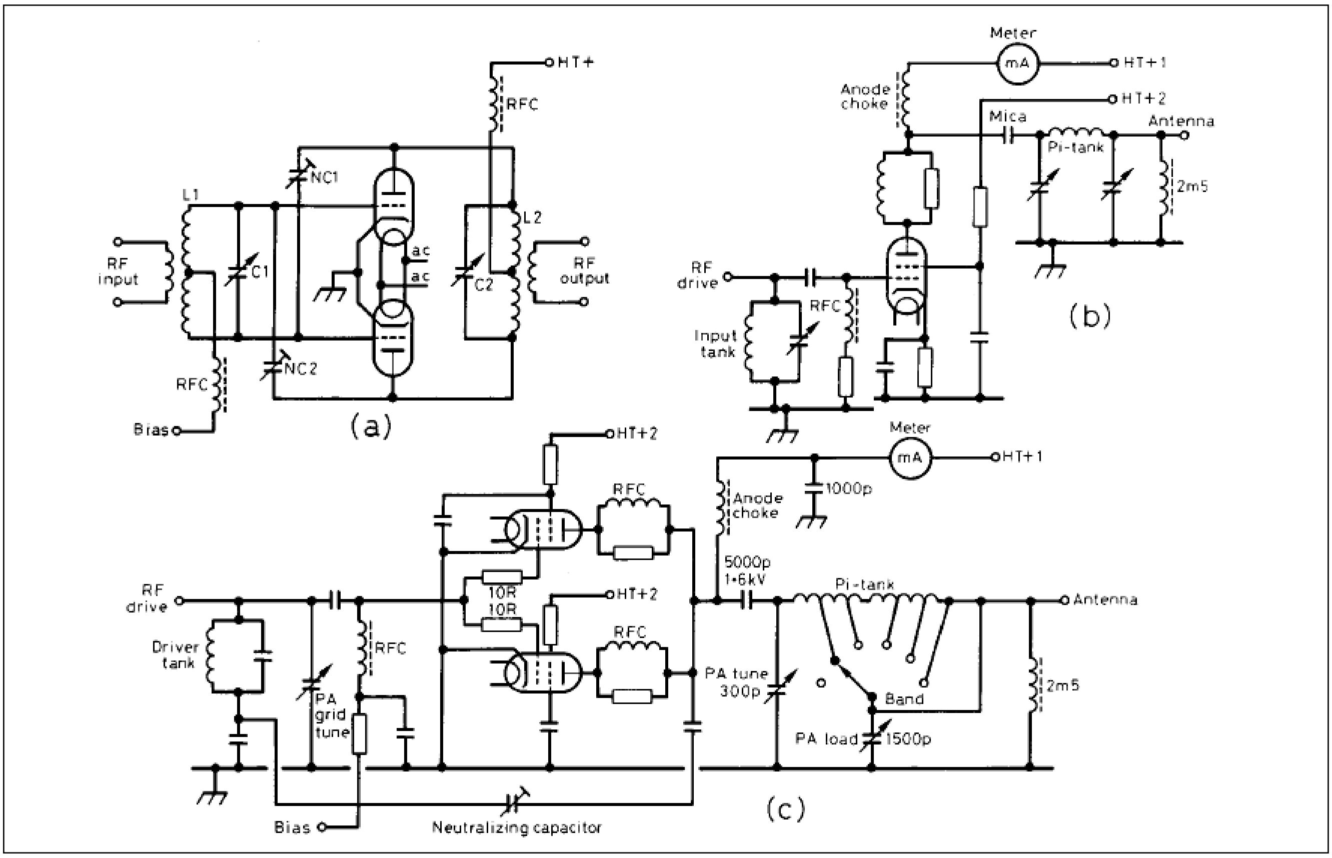


Fig 7.16: Valve power amplifiers. (a) Push-pull amplifier; (b) Single-ended Class-C amplifier; (c) Parallel-pair RF amplifier

Fortunately the introduction of amateur SSB occurred at much the same time as the rapid expansion in television operating hours, and the introduction of 'linear' Class AB amplifiers, essential for the amplification of SSB signals, greatly reduced the problems of TVI with TV frequencies that were often exact multiples of the 14, 21 and 28MHz amateur bands. The parallel-pair valve power amplifier became the standard output stage for amateur band transmitters and is still in common use in many external linear amplifiers and a few amateur radio transceivers .

Solid-state power amplifiers first appeared in the late 1970s and were capable of up to 100W RF output. Initially the reliability was poor but improved rapidly, and within 10 years virtually all commercially made amateur radio transmitters were equipped with a 100W solid-state PA unit. Transistors for high-power RF amplification are specially designed for the purpose and differ considerably internally from low-frequency switching transistors. Initial attempts by amateurs to use devices not specifically designed for RF amplification resulted in a mixture of success and failure, probably giving rise to initial claims that solid-state power amplifiers were less reliable than their valve counterparts.

Due to the very low output impedances of solid-state bipolar devices, it is almost impossible to match the RF output from a solid-state power amplifier to a resonant tank circuit with its characteristic high impedance. As a result, low-Q matching circuits (Fig 7.2 and 7.3) are used to transform the very low (typically 1 or 2Ω) output impedance encountered at the collector of an RF power amplifier to the now-standard output impedance of 50Ω.

As the matching circuits are not resonant, they are broad-band in nature, making it possible to operate the amplifier over a wide range of frequencies with no need for any form of tuning. Unfortunately, the broad-band amplifier also amplifies harmonics and other unwanted products, making it essential to use a low-pass filter immediately after the amplifier in order to achieve an adequate level of spectral purity. It is also essential that the amplifier itself should not contribute to the production of unwanted products. For this reason solid-state amplifier designs have reverted to the balanced, push-pull mode of operation, with its inherent suppression of even-order harmonics. As band switching and tuning of the amplifier is neither possible or necessary, construction of the amplifier is relatively simple. Typically, HF broad-band amplifiers will provide an output over the entire HF spectrum from 1.8 to 30MHz.

For powers in excess of 100W, valves are still popular and are likely to remain in use for some time to come on the grounds of cost, simplicity and superior linearity.

Power amplifiers can be categorised into two basic types, valve and solid-state, and then further into sub-groups based upon the output power. Low-power amplifiers can be regarded as 10W or less, including QRP (usually regarded as less than 5W output power), medium power up to 100W, and high power in excess of 100W.

The class of operation of a power amplifier is largely determined by its function. Class C amplifiers are commonly used for CW, AM and FM transmissions because of their high efficiency. Class C is also a pre-requisite for successful high-level modulation in an AM transmitter. Due to the non-linear operation of Class C amplifiers, the harmonic content is high and must be adequately filtered to minimise interference.

Single sideband transmission demands linear amplification if distortion is to be avoided, and amplifiers may be operated in either Class A or B. Class A operation is inefficient and normally confined to driver stages, the high standing current necessary to achieve this class of operation usually being unacceptable in amplifiers of any appreciable power rating. Most linear amplifiers designed for SSB are operated in between Class A and B, in what are known as Classes AB1 and AB2, in order to achieve a compromise between efficiency and linearity.

Solid-state Versus Valve Amplifiers

The standard RF power output for the majority of commercially produced amateur radio transmitters/transceivers is 100W, and at this power level solid-state amplifiers offer the following:

- Compact design.
- Simpler power supplies requiring only one voltage (normally 13.8V) for the entire equipment.
- Broad-band, no tune-up operation, permitting ease of operation.
- Long life with no gradual deterioration due to loss of emission.
- Ease of manufacture and reduced cost.

There are of course some disadvantages with solid-state amplifiers and it is for this reason that valves have not disappeared entirely. However, valve amplifiers have considerably more complex power supply requirements and great care must be taken with the high voltages involved if home construction is contemplated. Although solid state amplifiers only require relatively low voltages, they require very high currents for the generation of any appreciable power, placing demands upon the devices and their associated power supplies. Heatsinking and voltage stabilisation become very important.

While the construction of solid-state amplifiers up to 1kW is feasible using modern devices operated from a high-current 50V power supply, valve designs offer a simpler and more cost-effective alternative. However, the development of VMOS devices with much higher input and output impedances is beginning to bridge the gap. Quite possibly within a few years VMOS devices will offer a cheaper alternative to the valve power amplifier at the kilowatt level.

Impedance Matching

All types of power amplifier have an internal impedance made up from a combination of the internal resistance, which dissipates power in the form of heat, and reactance. As both source and load impedances are fixed values and not liable to change, it is necessary to employ some form of impedance transformation or matching in order to obtain the maximum efficiency from an amplifier. The power may be expressed as:

$$P_{input} = P_{output} + P_{dissipated}$$

where P_{input} is the DC input power to the stage; P_{output} is the RF power delivered to the load and $P_{dissipated}$ is the power absorbed in the source resistance and dissipated as heat.

$$\text{Efficiency} = \frac{P_{output}}{P_{input}} \times 100\%$$

When the source resistance is equal to the load resistance, the current through either will be equal as they are in series, with the result that 50% of the power will be dissipated by the source and 50% will be supplied to the load. The object of a power amplifier is to provide maximum power to the load. Design of a power amplifier must also take into account the maximum dissipation of the output device as specified by the manufacturer. An optimum load resistance is selected to ensure maximum output from a power amplifier while not exceeding the amplifying device's power dissipation. Efficiency increases as the load resistance to source resistance ratio increases and vice versa. The optimum load resistance is determined by the device's current transfer characteristics and for a solid-state device is given by:

$$R_L = \frac{V_{cc}^2}{2P_{out}}$$

Valves have more complex current transfer characteristics which differ for different classes of operation; the optimum load resistance is proportional to the ratio of the DC anode voltage to the DC anode current divided by a constant which varies from 1.3 in Class A to approximately 2 in Class C.

$$R_L = \frac{V_a}{KI_a}$$

The output from a RF power amplifier is usually connected to an antenna system of different impedance, so a matching network must be employed. Two methods are commonly used: pi-tank circuit matching for valve circuits and transformer matching for solid-state amplifiers. The variable nature of a pi-tank circuit permits matching over a wide range of impedances, whereas the fixed nature of a matching transformer is dependent upon a nominal load impedance of typically 50Ω, and it is therefore almost essential to employ some form of antenna matching unit between the output and the final load impedance. Matching networks serve to equalise load and source resistances while providing inductance and capacitance to cancel any reactance.

Valve Power Amplifiers

Valve power amplifiers are commonly found in the output stages of older amateur transmitters. Usually two valves will be operated in parallel, providing twice the output power possible with a single valve. The output stage may be preceded by a valve or solid-state driver stage.

The valve amplifier is capable of high gain when operated in the tuned input, tuned output configuration often referred to as TPTG (tuned plate tuned grid). Two 6146 valves are capable of producing in excess of 100W RF output with as little as 500mW of RF drive signal at the input. Operation of two valves in parallel increases the inter-electrode capacitances by a factor of two and ultimately affects the upper frequency operating limit. Fig 7.16(c) illustrates a typical output stage found in amateur transceivers.

Neutralisation

The anode-to-grid capacitance of a valve provides a path for RF signals to feed back energy from the anode to the grid. If this is sufficiently high, oscillation will occur. This can be overcome by feeding back a similar level of signal, but of opposite phase, thus cancelling the internal feedback. This process is referred to as neutralisation.

Once set, the neutralisation should not require further adjustment unless the internal capacitance of the valve changes or the valve has to be replaced.

While there are numerous ways of achieving neutralisation, the most common method still in use is series-capacitance neutralisation. A low-value ceramic variable capacitor is connected from the anode circuit to the earthy end of the grid input circuit as in Fig 7.16(c).

The simplest way to adjust neutralisation is firstly to tune up the amplifier into a dummy load and then to remove all the high voltage supplies from the valves, leaving the filaments powered. Connect a sensitive RF voltmeter across the dummy load and drive the input of the amplifier with rated power. When the neutralising capacitor is adjusted, the RF voltage at the output will dip to a minimum at the optimum tuning point. This method is safe and accurate. An alternative is to reverse the amplifier connections, that is apply the drive power to the antenna port and measure the RF voltage at the grid network and similarly tune for a dip. Neutralising capacitors should be adjusted with non-metallic trimming tools as metal tools will interfere with the adjustment.

Parasitic oscillation

It is not uncommon for a power amplifier to oscillate at some frequency other than one in the operating range of the amplifier. This can usually be detected by erratic tuning characteristics and a reduction in efficiency. The parasitics are often caused by the resonance of the connecting leads in the amplifier circuit with the circuit capacitance. To overcome problems at the design stage it is common to place low-value resistors in series with the grid, and low-value RF chokes, often wound on a resistor body, in series with the anode circuit. Ferrite beads may also be strategically placed in the circuit to damp out any tendency to oscillate at VHF.

HIGH-POWER AMPLIFICATION

Output powers in excess of 100W are invariably achieved using an add-on linear amplifier. In view of the high drive power available, the amplifier can be operated at considerably lower gain, with the advantages of improved stability and no requirement for neutralisation. Input circuits are usually passive, with valves operated in either passive-grid or grounded-grid modes. The grounded-grid amplifier has a cathode impedance ideally suited to matching the pi-output circuit of a valve exciter. Pi-input networks are usually employed to provide the optimum 50-ohm match for use with solid-state exciters. Passive grid amplifiers often employ a grid resistor of 200 to 300Ω, which is suitable for connecting to a valve exciter but will require a matching network such as a 4:1 auto-transformer for connection to a solid-state exciter.

Output matching

There are only two output circuits in common use in valve power amplifiers - the pi-output network is by far the most common and suitable values for a range of anode impedances are provided in Table 7.2. More recently, and especially for applications in high-power linear amplifiers, an adaptation of the circuit has appeared called the pi-L output network, and here the conventional pi-network has been combined with an L-network to provide a matching network with a considerable improvement in attenuation of unwanted products. The simple addition of one extra inductor to the circuit provides a considerable improvement in performance. Suitable values for a pi-L network are given in Table 7.3 for a range of anode impedances.

Valve amplifiers employing pi-output networks require a suitable RF choke to isolate the anode of the power amplifier from the high voltage power supply. This choke must be capable of carrying the anode current as well as the high anode voltages likely to be encountered in such an amplifier. The anode choke must not have any resonances within the operating range of the amplifier or it will overheat with quite spectacular results. Chokes are often wound in sections to reduce the capacitance between turns and may employ sections in varying diameters. Ready-made

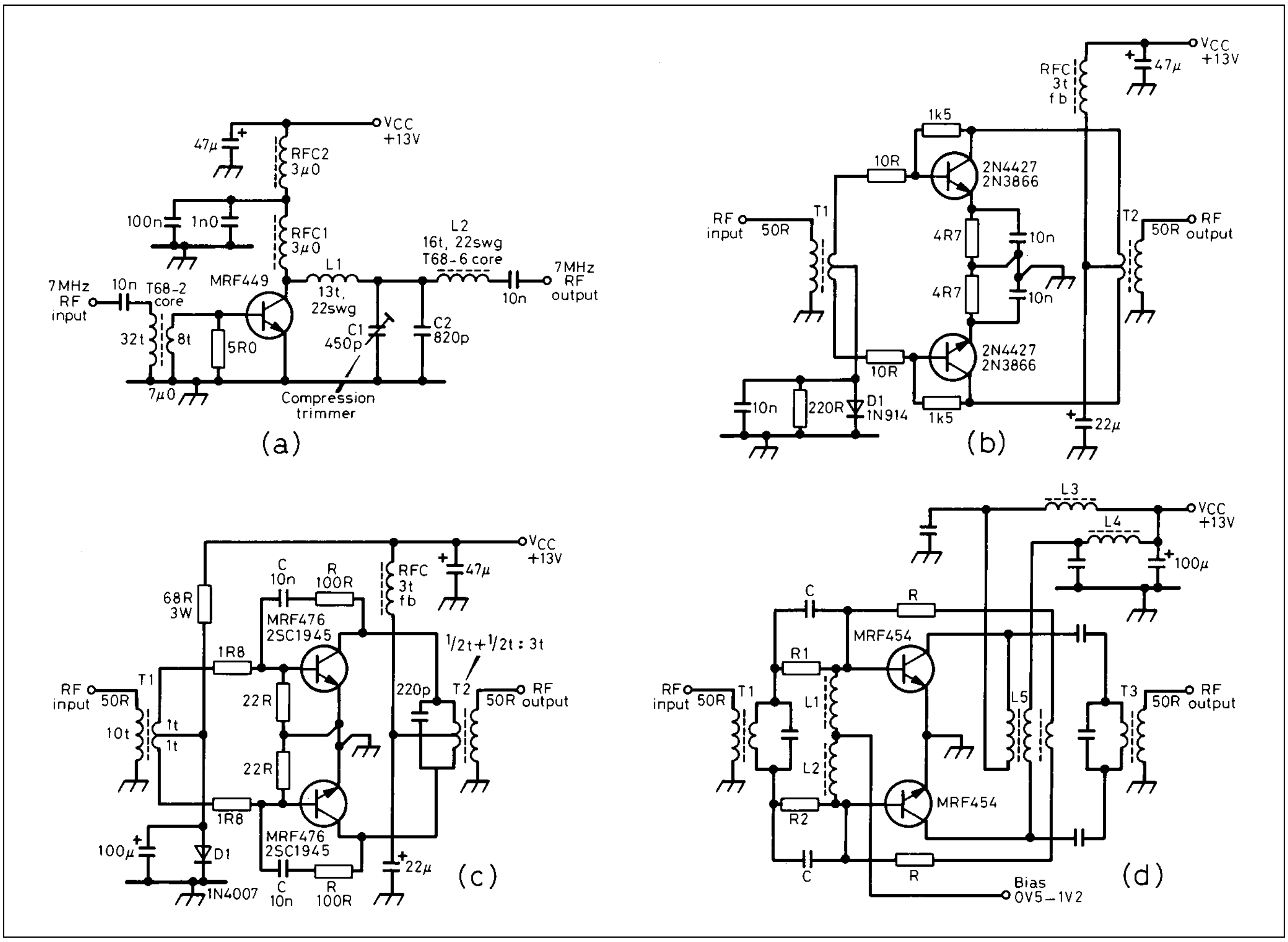


Fig 7.17: Solid-state power amplifiers. (a) Single-ended Class C amplifier (15W); (b) low-power amplifier driver; (c) low-power push-pull amplifier (10-25W); (d) medium-power push-pull amplifier (100W)

chokes are available for high-power operation from a number of sources on the internet. A typical example is the very reasonably priced amateur bands transmitter choke rated at 4kV (part number RFC-3) offered by RF Parts Company in the USA [2].

Solid-state Power Amplifiers

Solid-state power amplifiers are invariably designed for 50Ω input and output impedances, and they employ broad-band transformers to effect the correct matching to the devices. While single-ended amplifiers are often shown in test circuits, Unless they are tuned (**Fig 7.17(a)**), their use is not recommended for the following reasons:

(a) higher levels of second harmonic may be present than when using push-pull. These are difficult to attenuate using standard low-pass filters;

(b) multiband operation is more difficult to achieve due to the more complex filter requirements.

The broad-band nature of a solid-state amplifier with no requirement for bandswitching enables push-pull designs to be used, taking advantage of their improved balance and natural suppression of even-order harmonics (though care should be taken to reduce odd-order harmonics). The gain of a solid-state broad-band amplifier is considerably lower than that of a valve power amplifier (typically 10dB) and may necessitate cascading a number of stages in order to achieve the desired power level. This is easily achieved using common input and output impedances.

The reduced gain has the advantage of aiding stability, but the gain rises rapidly with a reduction in frequency and demands some form of frequency-compensated gain reduction. The latter is achieved using negative feedback with a series combination of R and C: **Figs 7.17(b) and (c)**. Good decoupling of the supply down to audio frequencies is essential. The use of VHF power transistors is not recommended in the HF spectrum as instability can result even when high levels of negative feedback are employed.

Solid-state amplifiers can be operated in Class C for use in CW and FM transmitters. However, their use for AM transmission should be treated very carefully because device ratings must be capable of sustaining double the collector voltage and current on modulation peaks, ie four times the power of the carrier. Additional safety margins must be included to allow for high RF voltages generated by a mismatched load.

For AM transmission it is recommended that the signal be generated at a low level and then amplified using a linear amplifier. Again, allowance should be made for the continuous carrier and the power on modulation peaks.

Output filters

Solid-state amplifiers must not be operated into an antenna without some form of harmonic filtering. The most common design is the pi-section filter, comprising typically of a double pi-section (five-element) and in some cases a triple pi-section (seven-element) filter. Common designs are based upon the Butterworth and Chebyshev filters and derivations of them. The purpose of the low-pass filter is to pass all frequencies below the cut off frequency (f0), normally located just above the upper band edge, while providing a high level of attenuation to all frequencies above the cut-off frequency. Different filter designs provide differing attenuation characteristics versus frequency, and it is desirable to achieve a high level of attenuation by at least three times the cut-off frequency in order to attenuate the third harmonic.

The second harmonic, which should be considerably lower in value due to the balancing action of the PA, will be further attenuated by the filter which should have achieved approximately 50% of its ultimate attenuation.

Elliptic filters are designed to have tuned notches which can provide higher levels of attenuation at selected frequencies such as 2f and 3f. Filters are discussed in detail in the chapters on Building Blocks and General Data.

The desire to achieve high levels of signal purity may tempt constructors to place additional low-pass filters between cascaded broad-band amplifiers, but this practice will almost certainly result in spurious VHF oscillations. These occur when the input circuit resonates at the same frequency as the output circuit, ie when the filter acts as a short-circuit between the amplifier input and output circuits. Low-pass filters should only be employed at the output end of an amplifier chain. If it is essential to add an external amplifier to an exciter which already incorporates a low-pass filter, it is important either to use a resistive matching pad between the exciter and the amplifier, or modify the output low-pass filter to ensure that it has a different characteristic to the input filter. If a capacitive input filter is used in the exciter, an inductive input filter should be employed at the output of the linear amplifier. The resulting parasitics caused by the misuse of filters may not be apparent without the use of a spectrum analyser, and the only noticeable affect may be a rough-sounding signal and warm low-pass filters.

Amplifier matching

Broad-band transformers used in solid-state amplifiers consist of a small number of turns wound on a stacked high-permeability ferrite core. The secondary winding may be wound through the primary winding which may be constructed from either brass tube or copper braid. The grade of ferrite is very important and will normally have an initial permeability of at least 800 (Fairite 43 grade). Too low a permeability will result in poor low-frequency performance and low efficiency. Some designs use conventional centre-tapped transformers for input and output matching (Fig 7.17(b) and (c)); these transformers carry the full DC bias and PA currents.

Other designs, **Fig 7.17(d)**, include phasing transformers to supply the collector current while the output transformer is blocked to DC by series capacitors. The latter arrangement provides an improvement in IMD performance of several decibels but is often omitted in commercial amateur radio equipment on grounds of cost.

Amplifier protection

The unreliability of early solid-state PAs was largely due to a lack of suitable protection circuitry. ALC (automatic level control) has been used for controlling the output of valve amplifiers for many

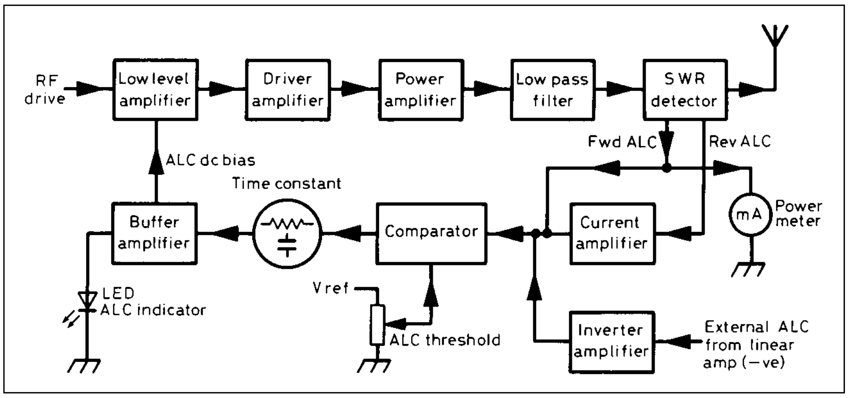


Fig 7.18: ALC system incorporating forward and reverse protection

years - by sampling the grid current in a valve PA it is possible to provide a bias that can be fed back to the exciter to reduce the drive level. A similar system is used for solid-state amplifiers but it is usually derived by sampling some of the RF output present in a SWR bridge circuit.

ALC is very similar to the AGC system found in a receiver. One disadvantage of sampling the RF output signal is that excessive ALC levels will cause severe clipping of the RF signal with an associated degradation of the IMD performance of the amplifier. For optimum performance the ALC system should only just be operating.

One of the major differences between solid-state and valve amplifying devices is that the maximum voltage ratings of solid-state devices are low and cannot be exceeded without disastrous consequences.

Reverse ALC is provided to overcome this problem and works in parallel with the conventional or forward ALC system. High RF voltages appearing at the PA collectors are attributable to operating into mismatched loads which can conveniently be detected using a SWR bridge.

The reverse or reflected voltage can be sampled using the SWR bridge and amplified and used to reduce the exciter drive by a much greater level than that used with the forward ALC. The output power is cut back to a level which then prevents the high RF voltages being generated and so protects the solid-state devices. Some RF output devices are fitted internally with zener diodes to prevent the maximum collector voltage being exceeded. **Fig 7.18** illustrates a typical ALC protection system.

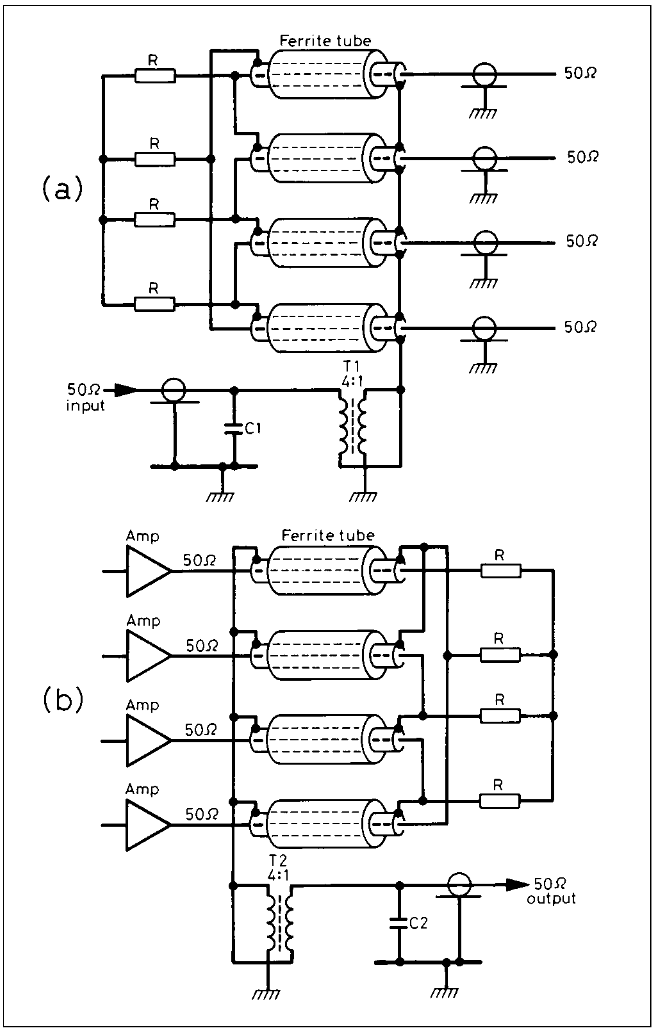


Fig 7.19: Combining multiple power amplifiers. (a) Four port power divider. (b) Four-port output combiner

Heatsinking for solid-state amplifiers

The importance of heatsinking for solid-state amplifiers cannot be overstressed - on no account should any power amplifier be allowed to operate without a heatsink, even for a short period of time.

Low-power amplifiers can often be mounted directly to an aluminium chassis if there is adequate metalwork to dissipate the heat, but a purpose-made heatsink should ideally be included for power levels in excess of 10W. For SSB operation the heatsinking requirements are less stringent than for CW or data operation due to the much lower average power dissipation. Nevertheless, an adequate safety margin should be provided to allow for long periods of key-down operation.

Solid-state power amplifiers are of the order of 50% efficient, and therefore a 100W output amplifier will also have to dissipate 100W of heat. It can be seen that for very-high-power operation heatsinking becomes a major problem due to the compact nature of solid-state amplifiers. The use of copper spreaders is advisable for powers above 200W. The amplifier is bolted directly to a sheet of copper at least 6mm (0.25in) thick, which in turn is bolted directly to the aluminium heatsink. The use of air blowers to circulate air across the heatsink should also be considered.

The actual mounting of power devices requires considerable care; the surface should ideally be milled to a flatness of ±0.012mm (0.5 thousandths of an inch). For this reason, die-cast boxes must not be used for high-power amplifiers as the conductivity is poor and the flatness is nowhere near to being acceptable. Heatsinking compounds are also essential for aiding the rapid conduction of heat away from the device. Motorola Application Note AN1041/D [1] provides guidance in the mounting of power devices in the 200-600W range.

Power dividers and combiners

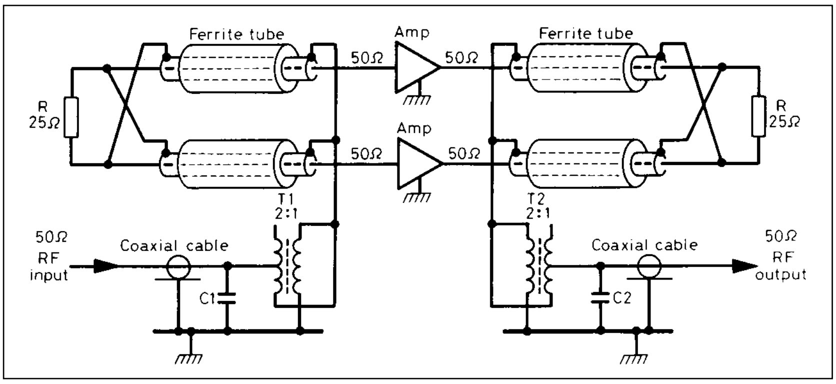
It is possible to increase the power output of a solid-state amplifier by combining the outputs of a number of smaller amplifiers. For instance it is practical to combine the outputs of four 100W amplifiers to produce 400W, or two 300W amplifiers to produce 600W.

Initially the drive signal is split using a power divider and then fed to a number of amplifiers that are effectively operated in parallel. The outputs of the amplifiers are summed together in a power combiner that is virtually the reverse of the divider circuit. **Fig 7.19** illustrates a four-way divider and combiner.

The purpose of the power divider is to divide the input power into four equal sources, providing an amount of isolation between each. The outputs are designed for 50Ω impedance, which sets the common input impedance to 12.5. A 4:1 step-down transformer provides a match to the 50Ω output of the driver amplifier. The phase shift between the input and output ports must be zero and this is achieved by using 1:1 balun transformers.

These are loaded with ferrite tubes to provide the desired low-frequency response without resort to increasing the physical length. In this type of transformer, the currents cancel, making it possible to employ high-permeability ferrite and relatively short lengths of transmission line. In an ideally balanced situation, no power will be dissipated in the magnetic cores and the line loss will be low. The minimum inductance of the input transformer should be 16μH at 2MHz - a lower value will degrade the isolation characteristics between the output ports and this is important in the event of a change in input VSWR to one of the amplifiers. It is unlikely that the splitter will be subjected to an open-

Fig 7.20: Two-port divider and combiner



or short-circuit load at the amplifier input, due to the base-frequency compensation networks in the amplifier modules. The purpose of the balancing resistors R is to dissipate any excess power if the VSWR rises.

The value of R is determined by the number of 50Ω sources assumed unbalanced at any one time. Except for a two-port divider, the resistor values can be calculated for an odd or even number of ports as:

$$R = \left(\frac{R_L - R_{IP}}{n + 1} \right) n$$

where R_L is the impedance of output ports (50Ω); R_{IP} is the impedance of input port (12.5Ω); n is the number of correctly terminated output ports.

Although the resistor values are not critical for the input divider, the same formula applies to the output combiner where mismatches have a larger effect on the total power output and linearity.

The power divider employs ferrite sleeves having a μ of 2500 and uses 1.2in lengths of RG-196 coaxial cable; the inductance is approximately 10μH. The input transformer is wound on a 63-grade ferrite toroid with RG-188 miniature coaxial cable. Seven turns are wound bifilar and the ends are connected inner to outer, outer to inner at both ends to form a 4:1 transformer.

The output combiner is a reverse of the input divider and performs in a similar manner. The ballast resistors R must be capable of dissipating large amounts of heat in the event that one of the sources becomes disabled, and for this reason they must be mounted on a heatsink and be non-inductive. For one source disabled in a four-port system, the heat dissipated will be approximately 15% of the total output power, and if phase differences occur between the sources, it may rise substantially. The resistors are not essential to the operation of the divider/combiner; their function is to provide a reduced output level in the event of an individual amplifier failure. If they are not included, failure of any one amplifier will result in zero output from the amplifier combination. The output transformer must be capable of carrying the combined output power and must have sufficient cross-sectional area. The output transformer is likely to run very warm during operation. High-frequency compensating capacitors C1 and C2 may be fitted to equalise the gain distribution of the amplifier but they are not always necessary.

A two-port divider combiner is illustrated in **Fig 7.20**; operation is principally the same as in the four-port case but the input and output transformers are tapped to provide a 2:1 ratio. Detailed constructional notes of two- and four-port dividers and combiners are given in Motorola Application Notes AN749 and AN758 respectively [1]

G4JNT DIRECT AUDIO UP-CONVERTER

I/Q or quadrature upconverters are ideally suited for translating the audio from a soundcard directly up to RF. Image cancelling or I/Q upconversion is possibly by making use of the two stereo channels from a soundcard carrying the I and Q audio signals - without the need for a separate audio phasing network. Alternatively, for narrow band datamodes, a simple phasing network working over no more than a few hundred Hz bandwidth is quite straightforward, and useable where stereo I/Q output have not been provided by the software author.

The design shown in **Fig 7.21** is a direct-from-audio upconverter suitable for the LF to low HF bands. The only part not shown is the actual RF source itself which could be a crystal oscillator, a VFO or a frequency synthesizer such as a DDS.

The upconverter design is based around fast CMOS bus switches as used in the Softrock and several similar designs of Software Defined Radios for direct conversion to audio. By turning the mixer round a high performance image cancelling direct upconverter can result.

At the left hand side of the circuit, the RF input is buffered by a pair of logic gates biased as a high gain limiting amplifier then applied to a ring counter made from a pair of flip flops. One oddity of this divider making it different from a conventional binary counter is that the outputs count 0, 2, 3, 1 rather than 0, 1, 2, 3. This changes the order of the connections needed to the two poles of the CMOS switch. A ring counter is used as it is easier to ensure both output drives are exactly synchronous than it is with a conventional counter.

The output filter is essential. As the RF drive takes the form of a square wave, all odd order harmonics are present, and the relative power of these rolls off following a 1/F2 law. So the third harmonic is only 10.LOG(1/9) or approximately 10dB down, the fifth at -14dB and so on. The values shown give a cut off of 505kHz. This is followed by a simple broadband amplifier raising the output level to around +6dBm maximum. The filter component values can be scaled for other frequency ranges.

The I/Q drive is generated from a single audio input with an all-pass network made from a pair of op-amps. Each channel of the all-pass network maintains a constant gain of unity whatever the input frequency, with a phase shift over the audio band depending on the CR product connected to each op-amp positive input. This goes from 0 degrees at DC, 90 degrees from input to output at F = 1/2.π.C.R and 180 degrees at high frequencies. By choosing a pair of differing CR time constants for each channel, the difference in phase between the two outputs can be kept sufficiently close to 90 degrees over a limited frequency band as shown in **Fig 7.22**.

The simple arrangement can only generate I/Q signals with a sufficiently accurate 90 degree phase difference over a limited range of frequencies. As the phase shift degrades from the ideal quadrature values the sideband isolation falls with its value, in dB, given by 20 * LOG (TAN(ØI - ØQ)) and assumes the amplitudes of I and Q are identical; if this is not so, sideband rejection is further degraded. At 90 degrees it would yield an infinite value (perfect rejection) but at 89 degrees phase shift (1 degree error) an isolation of 35dB results. At 5 degree of error sideband isolation becomes a rather poor 21dB. **Fig 7.23** shows the resulting sideband rejection that can be achieved. Only single I and Q channel signals are supplied and we need differential drive, so a second pair of op-amps are used as unity gain buffers to generate minus-I and minus-Q.

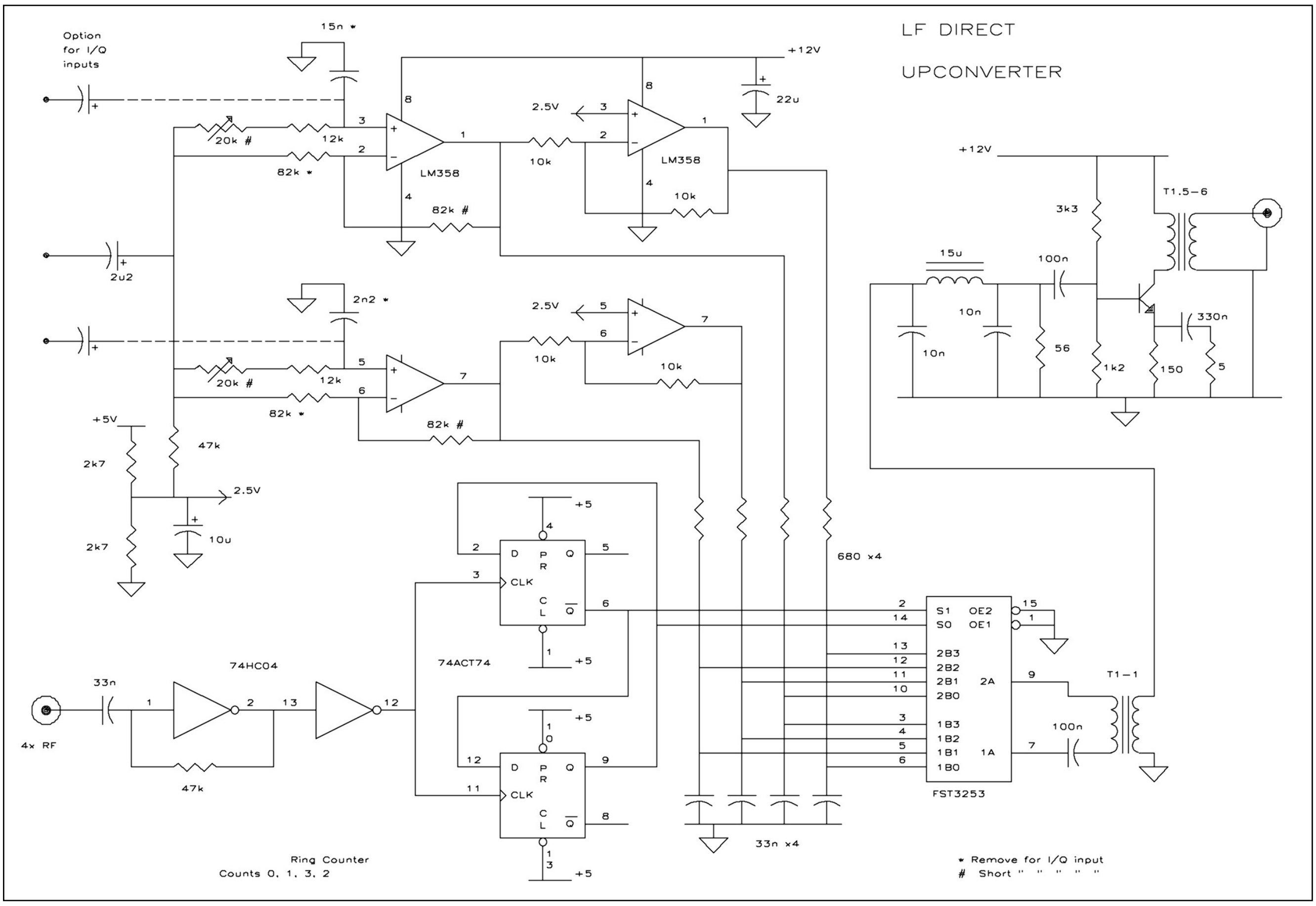


Fig 7.21: Circuit diagram of an upconverter for generating RF direct from soundcard outputs

If I and Q signals can be generated by software instead of having to be synthesized by analogue networks, the connections shown by the dotted lines (in Fig 7.21) can be used to configure the first two pair of op-amps as buffers rather than all-pass networks. Several components need to be shorted out or removed to do this. They are marked with an asterisk on the schematic diagram.

Setting up involves little more than adjusting the I/Q networks for optimum sideband rejection and checking / trimming carrier rejection. As an absolute minimum a receiver that can be tuned either side of the output band of interest to look at sideband rejection is adequate, with a suitable clean sinusoidal drive signal. Ideally the receiver should have a spectral display several kHz wide - any soundcard based SDR will do admirably.

For the source, it is helpful if the drive can be mixed with white noise to show sideband rejection over the whole output band simultaneously. A suitable test signal can come from another SSB receiver tuned to a weak carrier. The audio out from this contains reasonably flat noise typically from 300Hz to 2700Hz and a tone. First display its audio output by feeding directly into the soundcard input while running spectral analysis software such as Spectran [3]. Tune to any single carrier and adjust input attenuation / signal pickup until this sits a few tens of decibels above the background noise. Ensure things stay reasonably constant and that you can change the tone frequency with the tuning knob, and the noise stays flat. Transfer the audio so it becomes the drive to the upconverter, with a way of controlling the level from zero to maximum - such as by adjusting the receiver's volume control.

Apply a four-times RF source to the up-converter RF input buffer, and connect the RF output via suitable attenuation to a test receiver tuned to the wanted frequency. With the audio drive set to minimum some carrier leakage should be observed. Slowly increase audio drive and observe the upconverted RF increase in level by several tens of decibels above the carrier leakage which should remain constant. Audio sidebands either

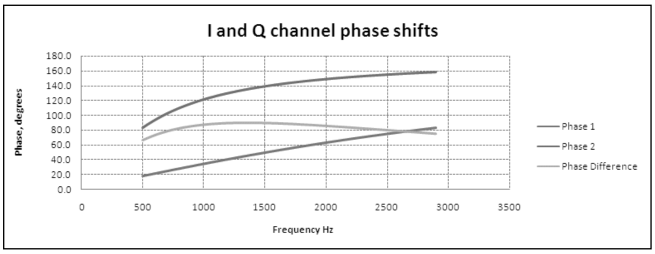


Fig 7.22: Phase response from the all-pass network in Fig 7.21

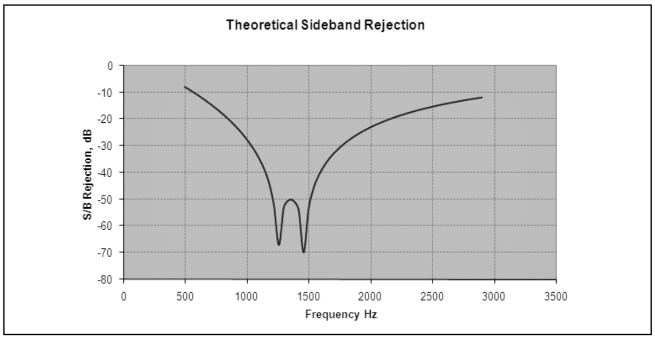


Fig 7.23: Theoretical sideband rejection from the all-pass network in Fig 7.21

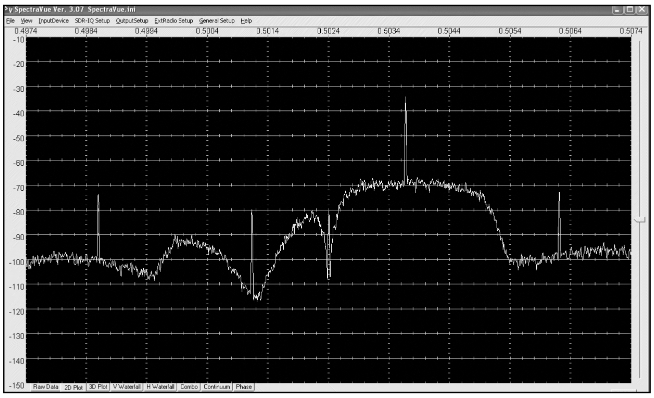


Fig 7.24: Measured plot of the up-converter output when driven by a mixture of white noise and a tone. For this plot the RF input was set at 2.0096MHz (4 * 502.4kHz) with the output from the up-converter fed via a 20dB attenuator directly to my SDR-IQ antenna input. The SDR-IQ was tuned to 502.4kHz centre frequency and 10kHz span. Leakage from the carrier appears exactly in the middle of the display with the upconverted sidebands on either side. The audio drive was generated by tuning my IC-746 to a weak carrier while monitoring the audio output using *Spectran*. The RF was attenuated until the carrier sat about 35dB above noise on the spectral display then the audio drive was transferred to the up-converter input

side of the carrier will be seen, but a section of the audio band, one sideband, should be noticeably different from the other. It may be the upper or lower sideband - it doesn't matter which at this stage. Fig 7.24 shows the sort of plot that can be expected with a tone plus noise drive.

Adjust the two input preset resistors to optimise the rejection. 40dB - 50dB can be achieved in a narrow band, with 35 - 40dB over a range of a few hundred hertz. If it is impossible to reach a particularly high rejection at even a single point, then it is likely that the amplitudes of the I and Q channel are not matched. With 1% resistors used throughout, little trimming should be needed. RF carrier rejection of 50dB ought to be possible, and there is a bit of scope for optimising this by adjusting the inverting buffer gain. But note that this will then affect I/Q balance and it could get very time consuming trying to get the ultimate all round performance. The pragmatic solution is to match resistors using a DVM to around 0.3 - 0.5% then leave well alone, just accepting the result. To change between upper and lower sideband, swap over I and Q drive signals.

TRANSCEIVERS

Separate transmitter and receiver combinations housed in one cabinet may be referred to as a 'transceiver'. This is not strictly correct as a transceiver is a combined transmitter and receiver where specific parts of the circuit are common to both functions. Specifically, the oscillators and frequency-determining components are common and effectively synchronise both the transmitter and receiver to exactly the same frequency. This synchronisation is a pre-requisite for SSB operation and transceivers owe their existence to the development of SSB transmission. While many early attempts at SSB generation used the phasing method, the similarity of the filter-type SSB generator circuit to a superhet SSB receiver circuit (Fig 7.5) makes interconnection of the two circuits an obvious development (Fig 7.25). True transceive operation is possible by simply using common oscillators, but it is also advantageous to use a common SSB filter in the IF amplifier stages, which provides similar audio characteristics on both transmit and receive as well as providing a considerable saving in cost.

Initially low-frequency SSB generation necessitated double-conversion designs, often employing a tuneable second IF and a crystal-controlled oscillator for frequency conversion to the desired amateur bands. This technique was superseded by single-conversion designs using a high-frequency IF in the order of 9MHz, with a heterodyne-type local oscillator consisting of a medium-frequency VFO and a range of high-frequency crystal oscillators. With the advent of the phase-locked loop (PLL) synthesiser and the trend towards wide-band equipment, modern transceivers typically up-convert the HF spectrum to a first IF in the region of 40-70MHz. This is then mixed with a synthesised local oscillator and converted down to a working IF in the order of 9MHz. Often a third IF in the order of 455kHz will be employed, giving a total of three frequency conversions.

In order to simplify the construction of equipment, designers have attempted to combine as many parts of the transceiver circuit as possible. Front-end filtering can be cumbersome and requires a number of filters for successful operation. Fig 7.26 illustrates the typical filtering requirements in a 14MHz transceiver. Traditionally, the receiver band-pass filter and even the transmit band-pass filter would have employed a variable tuning capacitor, often referred to as a preselector, to provide optimum selectivity and sensitivity when correctly peaked. To arrange for a number of filters to tune and track with one another requires careful design and considerable care in alignment, especially in multiband equipment. The introduction of the solid-state PA with its wide-band characteristics has lead to the development of wide-band filters possessing a flat response across an entire amateur band. By necessity these filters are of low Q and consequently must have more sections or elements if they are to exhibit any degree of out-of-band selectivity. By employing low-Q, multi-section, band-pass filters it is possible to eliminate one filter between the receiver RF amplifier and the receive mixer. The gain of the RF amplifier should be kept as low as possible and in most cases can be eliminated entirely for use below 21MHz. By providing the band-pass filters with low input and output impedances, typically

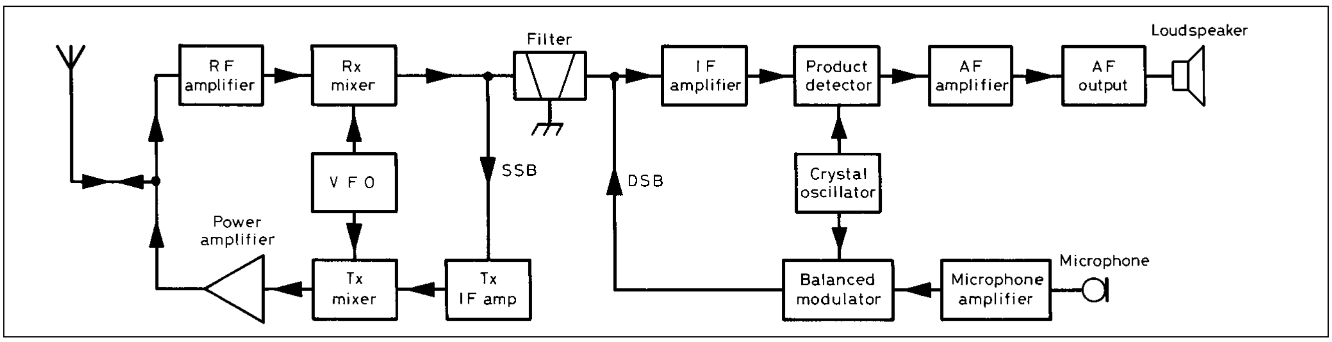


Fig 7.25: SSB transceiver block diagram

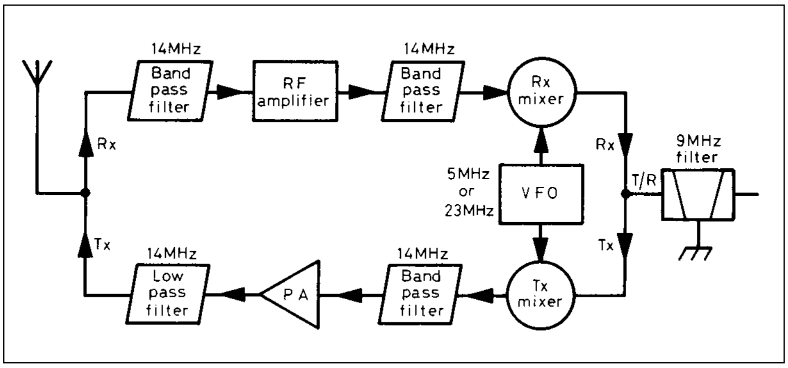


Fig 7.26: HF transceiver radio frequency filtering

50Ω, switching of the filters can simplified to the extent that one filter can be used in both the transmit and receive paths, thus reducing the band-pass filter requirement to one per band.

Fig 7.27(a) illustrates a typical band-pass filter configuration for amateur band use, and suitable component values are listed in **Table 7.6**. A transmit low-pass filter is essential for attenuation of all unwanted products and harmonics amplified by the broad-band amplifier chain. Suitable values for a typical Chebyshev filter, **Fig 7.27(b)**, suitable for use at the output of a power amplifier chain providing up to 100W output, are given in **Table 7.7**. One disadvantage of the standard low pass filter designs is the limited attenuation in the stop band and how quickly the attenuation increases for increasing frequency. The Butterworth filter is the worst, the Chebyshev far superior and the so-called elliptic filter (Cauer-Chebyshev) **Fig 7.27(d)** the best for the same number of poles.

The addition of two capacitors placed in parallel with the two filter inductors results in the circuits L1, C4 and L2, C5 being resonant at approximately two and three times the input frequency respectively to provide peaks of attenuation at the second and third harmonic frequencies. Where possible, the use of elliptic filters is recommended - typical values for amateur band use are given in **Table 7.8**.

For an external solid-state amplifier being driven by an exciter that already contains a capacitive-input low-pass filter such as that in Fig 7.23(b), the inductive input design, **Fig 7.27(c)**, may be necessary at the final amplifier output.

The combination of the two different types of filter should eliminate parasitic oscillations which will almost certainly occur if two filters with similar characteristics are employed. Values for the inductive input filter are given in **Table 7.9** for use at powers up to 300W.

The development of HF synthesisers has led to the development of HF transceivers providing general-coverage facilities and requiring yet further changes in the design of band-pass filters for transceiver front-ends. There is a finite limit to the bandwidth that can be achieved using conventional parallel-tuned circuit filters. For general-coverage operation from 1 to 30MHz, approximately 30-40 filters would be required, and this is obviously not a practical proposition. While returning to the mechanically tuned filter might reduce the total number of filters required, the complexity of electronic band-changing would be formidable. By combining the characteristics of both low-pass and high-pass filters, **Figs 7.27(d) and (f)**, the simple action of cascading two such filters will result in a band-pass filter, **Fig 7.28(a)**, having a bandwidth equal to the difference between the two filter cut-off frequencies. The limiting bandwidth will be one octave, ie the highest frequency is double the lowest frequency. Unless electronically tuned using varicap diodes, filter bandwidth is restricted to slightly less than one octave. Typical filters in a general-coverage HF transceiver might cover the following bands:

(a) 1.5-2.5MHz
(b) 2.3-4.0MHz
(c) 3.9-7.5MHz
(d) 7.4-14.5MHz
(e) 14.0-26.0MHz
(f) 20.0-32.0MHz

It can be seen that six filters will permit operation on all the HF amateur bands as well as providing general coverage of all

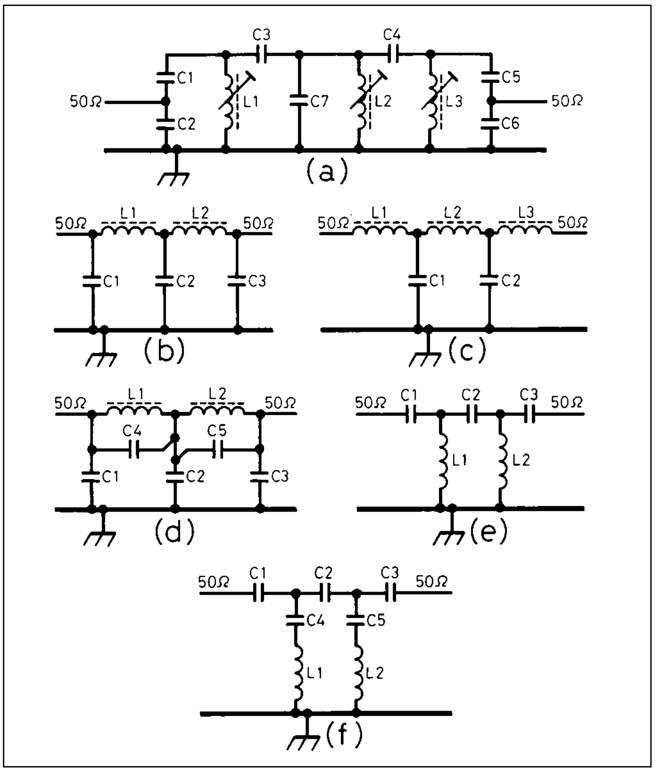


Fig 7.27: Band-pass, low-pass and high-pass filters. (a) Band-pass filter; (b) Chebyshev low-pass filter capacitive input/output; (c) Chebyshev low-pass filter inductive input/output; (d) elliptic function low-pass filter; (e) Chebyshev high-pass filter; (f) elliptic high-pass filter

Band (m)	L1, L3 (µH)	L2 (µH)	L-type	C1, C5 (pF)	C2, C6 (pF)	C3, C4 (pF)	C7 (pF)
160	8.0	8.0	27t KANK3335R	1800	2700	180	750
80	5.8	5.8	KANK3334R	390	1800	47	270
40	2.8	2.8	KXNK4173AO	220	1000	10	150
30	1.3	1.3	KANK3335R	220	1000	10	180
20	1.2	1.2	KANK3335R	120	560	4.7	100
17	0.29	-	Toko S18 Blue	220	750	8.2	-
15	0.29	-	Toko S18 Blue	180	560	6.8	-
12	0.29	-	Toko S18 Blue	100	560	2.7	-
10	0.29	-	Toko S18 Blue	82	390	2.7	-

All capacitors are polystyrene except those less than 10pF which are ceramic. L2 and C7 are not used on the 10-17m bands.

Table 7.6: Band-pass filter of Fig 7.23(a). 50 ohms nominal input/output impedance

Band (m)	L1, L2	Core	C1, C3 (pF)	C2 (pF)
160	31t/24swg	T50-2	1200	2500
80	22t/20swg	T50-2	820	1500
40	18t/20swg	T50-6	360	680
30/20	12t/20swg	T50-6	220	360
17/15	10t/20swg	T50-6	100	220
12/10	9t/20swg	T50-6	75	160

All capacitors are silver mica or polystyrene - for 100W use 300VDC wkg; for <50W use 63VDC wkg. Cores are Micrometals Inc T50-2 Red or T50-6 Yellow from Amidon.

Table 7.7. Chebyshev low-pass filter of Fig 7.23(b)

Band (m)	L1	L2	Core	C1 (pF)	C2 (pF)	C3 (pF)	C4 (pF)	C5 (pF)
160	28t/22SWG	25t/22SWG	T68-2	1200	2200	1000	180	470
80	22t/22SWG	20t/22SWG	T50-2	680	1200	560	90	250
40	18t/20SWG	16t/20SWG	T50-6	390	680	330	33	100
30/20	12t/20SWG	11t/20SWG	T50-6	180	330	150	27	75
17/15	10t/20SWG	9t/20SWG	T50-6	120	220	100	12	33
12/10	8t/20SWG	7t/20SWG	T50-6	82	150	68	12	39

Capacitors 300VDC wkg silver mica up to 200W. All cores Micrometals Inc from Amidon.

Table 7.8: Elliptic low-pass filter of Fig 7.23(d)

Band (m)	L1, L3	L2	Core	C1, C2 (pF)
160	8.1μH/24t	11.4μH	T106-2	1700
80	4.1μH/17t	5.8μH/21t	T106-2	860
40	2.3μH/13t	3.2μH/15t	T106-2	470
30/20	1.18μH/10t	1.65μH/12t	T106-6	240
17/15	0.79μH/8t	1.11μH/10t	T106-6	160
12/10	0.57μH/7t	0.8μH/8t	T106-6	120

This filter is for use with a high-power external amplifier, when capacitive input is fitted to exciter. Capacitors silver mica: 350VDC wkg up to 200W; 750VDC wkg above 300W. Use heaviest possible wire gauge for inductors. All cores Micrometals Inc from Amidon.

Table 7.9: Chebyshev low-pass filter inductive input of Fig 7.27(c)

the in-between frequencies. Hybrid low/high-pass filters invariably use fixed-value components and require no alignment, thus simplifying construction. The transmit low-pass filter may also be left in circuit on receive in order to enhance the high-frequency rejection; it has no effect on low-frequency signals. The use of separate high-pass filters in the receiver input circuit prior to the band-pass filter serves to eliminate low-frequency broadcast signals. Typically, a high-pass filter of the multi-pole elliptic type, Fig 7.28(b), having a cut-off of 1.7MHz, is fitted to most commercial amateur band equipment.

TRANSVERTERS

Transverters are transmit/receive converters that permit equipment to be operated on frequencies not covered by that equipment. Traditionally HF equipment was transverted to the VHF/UHF bands but, with the increase in availability of 144MHz SSB equipment, down-conversion to the HF bands has become popular. Fig 7.29 shows the schematic of a typical 144/14MHz transverter providing HF operation with a VHF transceiver.

A transverter takes the output from a transmitter, attenuated to an acceptable level, heterodynes it with a crystal-controlled oscillator to the desired frequency and then amplifies it to the required level. The receive signal is converted by the same process in reverse to provide transceive capabilities on the new frequency band. The techniques employed in transverters are the same as those used in comparable frequency transmitting and receiving equipment. Where possible it is desirable to provide low-level RF output signals for transverting rather than having to attenuate the high-level output from a transmitter with its associated heatsinking requirements.

Occasionally transverters may be employed from HF to HF in order to include one of the 'WARC bands' on an older transceiver, or to provide 160m band facilities where they have been omitted. In some cases it may prove simpler to add an additional frequency band to existing equipment in preference to using a transverter. Transverters have also been built to convert an HF transceiver to operate on the 136kHz band.

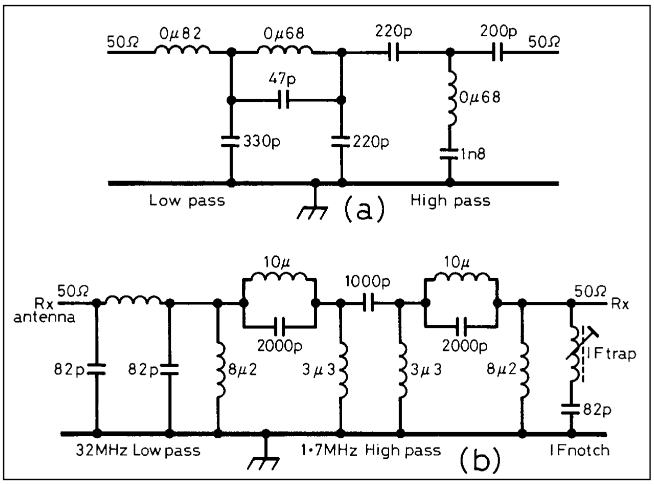


Fig 7.28: Multifunction filters. (a) Band-pass filter using high/low pass filters; (b) composite receiver input filter

PRACTICAL TRANSMITTER DESIGNS

QRP + QSK - A Novel Transceiver with Full Break-in

This design by Peter Asquith, G4ENA, originally appeared in *Radio Communication* [4].

Introduction

The late 20th century saw significant advances in semiconductor development. One such area was that of digital devices. Their speed steadily improved to the point which permitted them to be used in low band transceiver designs. One attractive feature of these components is their relatively low cost.

The QSK QRP Transceiver (Figs 7.30 and 7.31) employs several digital components which, together with simple analogue circuits, provide a small, high-performance and low-cost rig. Many features have been incorporated in the design to make construction and operation simple.

One novel feature of this transceiver is the switching PA stage. The output transistor is a tiny IRFD110 power MOSFET. This device has a very low 'ON' state resistance which means that

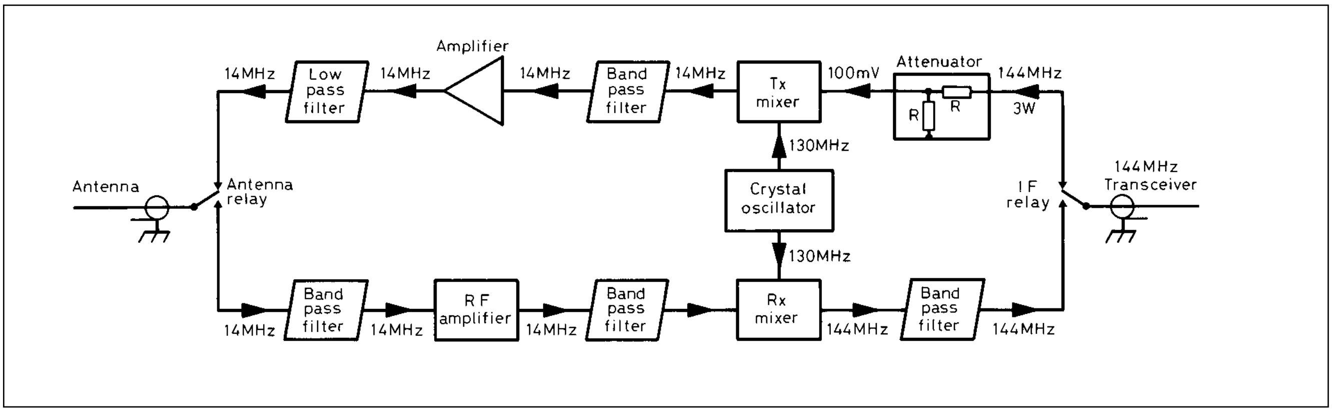


Fig 7.29: 144MHz to 14MHz transverter

Fig 7.30: The G4ENA transceiver

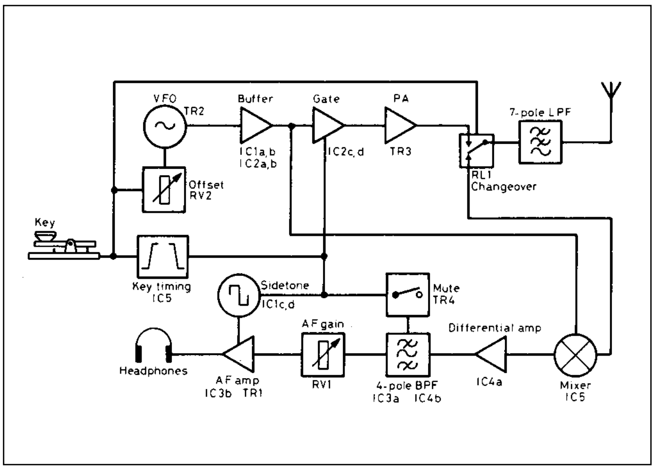


Fig 7.31: Block diagram of the G4ENA transceiver

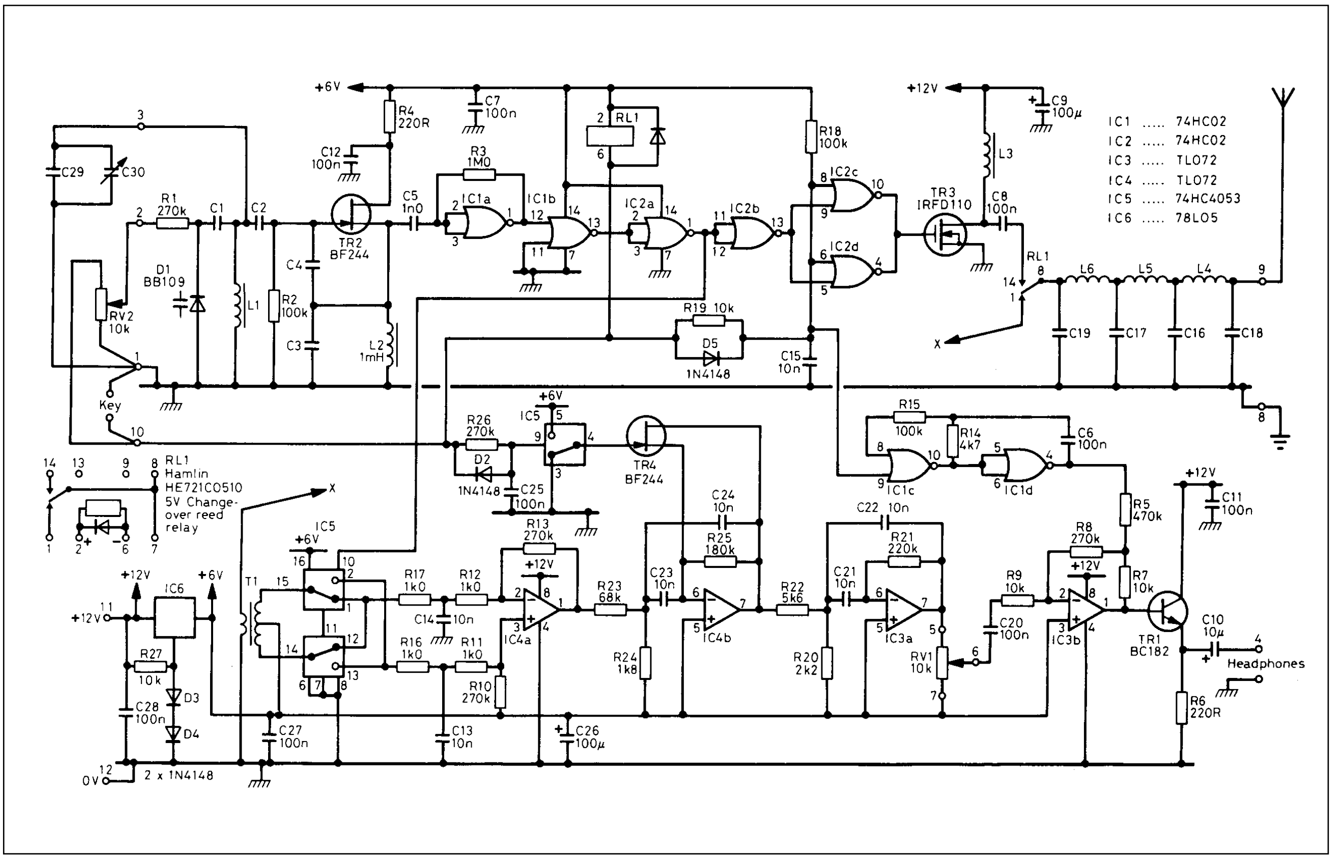


Fig 7.32: The switching PA stage requires a seven-pole filter, as shown in the circuit diagram above

very little power is dissipated in the package, hence no additional heatsinking is required. However, using this concept does mean that good harmonic filtering must be used.

The HC-type logic devices used in the rig are suitable for operation on the 160m and 80m bands. The transmitter efficiency on 40m is poor and could cause overheating problems. Future advances in component design should raise the top operating frequency limit.

Circuit description

VFO: The circuit diagram of the transceiver is shown in **Fig 7.32**. TR2 is used in a Colpitts configuration to provide the oscillator for both receive and transmit. The varicap diode D1 is switched via RV2/R1 by the key to offset the receive signal by up to 2kHz, such that when transmitting, the output will appear in the passband of modern transceivers operating in the USB mode. C1 controls the RIT range and C29/30 the band coverage. IC1a, IC1b and IC2a buffer the VFO. It is important that the mark/space ratio of the square wave at IC1b is about 50:50. Small variations will affect output power.

Transmitter: When the key is operated, RL1 will switch and, after a short delay provided by R19/C15, IC2c and IC2d will gate the buffered VFO to the output FET, TR3. TR3 operates in switch mode and is therefore very efficient. The seven-pole low-pass filter after the changeover relay removes unwanted harmonics, which are better than -40dB relative to the output.

Receiver: The VFO signal is taken from IC2a to control two changeover analogue switches in IC5, so forming a commutating mixer and providing direct conversion to audio of the incoming stations. IC4a is a low-noise, high-gain, differential amplifier whose output feeds the four-pole CW filter, IC4a and IC3a, before driving the volume attenuator, RV1. IC3b amplifies and TR1 buffers the audio to drive headphones or a small speaker. When the key is down TR4 mutes the receiver, and audio oscillator IC1a and b injects a sidetone into the audio output stage, IC3b. The value of R5 sets the sidetone level.

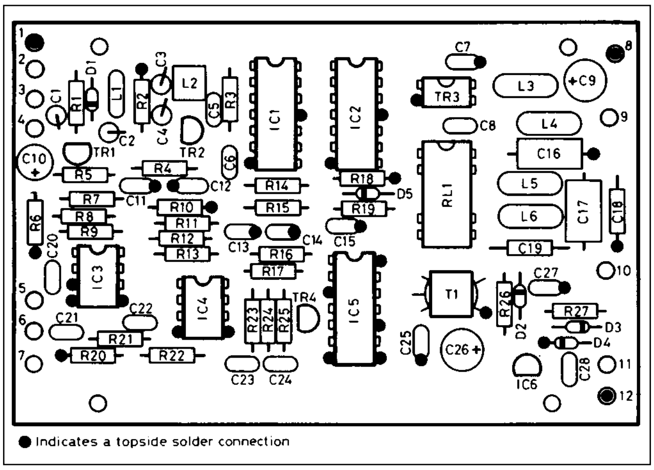


Fig 7.33: A neat component layout results in a compact unit suitable for portable operation (see also Fig 7.34)

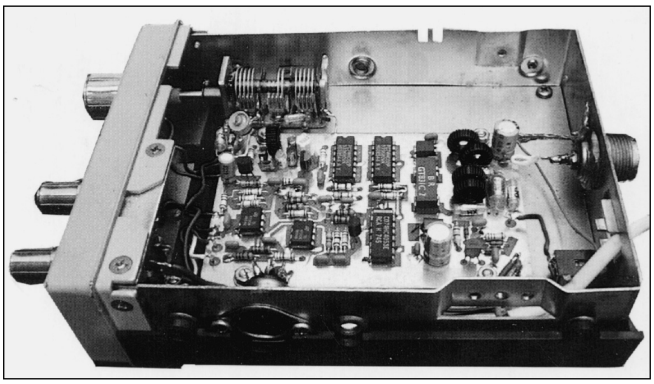


Fig 7.34: Inside view of the G4ENA transceiver

Constructional notes

The component layout is shown in **Figs 7.33 and 7.34**, and the Components List is in **Table 7.10**.

1. Check that all top-side solder connections are made.
2. Wind turns onto toroids tightly and fix to PCB with a spot of glue.
3. Do not use IC sockets. Observe anti-static handling precautions for all ICs and FETs.
4. All VFO components should be earthed close to VFO.
5. Fit a 1A fuse in the supply line.
6. Component suppliers: Bonex, Cirkit, Farnell Electronic Components, JAB Electronic Components etc.

Test and calibration

Before TR3 is fitted, the transceiver must be fully operational and calibrated. Prior to switch-on, undertake a full visual inspection for unsoldered joints and solder splashes. Proceed as follows.

1. Connect external components C29/30, RV1/2, headphones and power supply.
2. Switch on power supply. Current is approx 50mA.
3. Check +6V supply, terminal pin 7. Voltage is approx 6.3V.
4. Select values for C29 to bring oscillator frequency to CW portion of band (1.81-1.86MHz/3.50-3.58MHz). Coverage should be set to fall inside the band limits of 1.81/3.50MHz.
5. Connect an antenna or signal generator to terminal pin 9 and monitor the received signal on headphones. Tuning through the signal will test the response of the CW filter which will peak at about 500Hz.
6. Connect key and check operation of sidetone and antenna changeover relay. Sidetone level can be changed by selecting value of R5.
7. Monitor output of IC2c and IC2d (TR3 gate drive) and check correct operation. A logic low should be present with key-up, and on key-down the VFO frequency will appear. This point can be monitored with an oscilloscope or by listening on a receiver with a short antenna connected to IC2c or IC2d.
8. When all checks are complete fit TR3 (important! - static-sensitive device) and connect the transceiver through a power meter to a dummy load. On key-down the output power should be at least 5W for +12V supply, rising to 8W for 13.8V supply. Note: should the oscillator stop when the key is pressed it will instantly destroy TR3. Switch off power when selecting VFO components.
9. Connect antenna and call CQ. When a station replies note the position of the RIT control. The average receive offset should be used when replying to a CQ call.

On air

The QSK (full break-in) concept of the rig is very exciting in use. The side tone is not a pure sine wave and is easily heard if there is an interfering beat note of the same frequency. One important note is to remember to tune the receiver into a station from the high-frequency side so that when replying your signal falls within his passband.

RESISTORS (*All fixed resistors 0.25W 2%)*

R1, 8, 10, 13, 26	270k	R14	4k7
R2, 15, 18	100k	R20	2k2
R3	1M	R21	220k
R4, 6	220R	R22	5k6
R5	470k	R23	68k
R7, 9, 19, 27	10k	R24	1k8
R11, 12, 16, 17	1k	R25	180k
RV1, 2	10k lin		

CAPACITORS (**Select on test component)*

Ref	Type	Pitch	Value (80m)	Value (160m)
C1	Ceramic plate 9	2.54	4p7	15p
C2	Polystyrene	-	47p	100p
C3, 4	Polystyrene	-	220p	470p
C5	Ceramic monolithic	2.54	1n	1n
C6, 7, 8, 11, 12, 20, 25, 27, 28	Ceramic monolithic	2.54	100n	100n
C9, 26	Aluminium radial 16V	2.5	100	100µ
C10	Aluminium radial 16V	2.0	10µ	10µ
C13, 14, 15, 21, 22, 23, 24	Ceramic monolithic	2.54	10n 10%	10n 10%
C16, 17	Polystyrene	-	1n5	2n7
C18, 19	Polystyrene	-	470p	1n
C29*	Polystyrene	-	470p	820p
C30*	Air-spaced VFO	-	25p	75p

INDUCTORS

Ref	Type	80m	160m
L1	T37-2 (Amidon)	31t 27SWG (0.4mm)	41t 30SWG (0.315mm)
L2	7BS (Toko)	1mH	1mH
L3	T37-2	2.2µH 23t 27SWG	4.5µH 33t 30SWG
L4, 6	T37-2	2.9µH 26t 27SWG	5.45µH 36t 30SWG
L5	T37-2	4.0µH 31t 27SWG	6.9µH 41t 30SWG
T1	Balun	2t primary, 5+5t secondary, 0.2mm 36SWG (28-43002402)	

SEMICONDUCTORS

D1	BB109	IC1, 2	74HC02*
D2, 3, 4, 5	1N4148	IC3, 4	TL072*
TR1	BC182 (not 'L')	IC5	74HC4053*
TR2, 4	BF244	IC6	78L05
TR3	IRFD110*		

**Static-sensitive devices*

MISCELLANEOUS

RL1	5V change-over reed relay, Hamlin, HE721C0510; PED/Electrol, 17708131551-RA30441051
PCB	Boards and parts are available from JAB Electronic Components [5]

Table 7.10: G4ENA transceiver components list

Both the 160 and 80m versions have proved very successful on-air. During the 1990 Low Power Contest the 80m model was operated into a half-wave dipole and powered from a small nicad battery pack. This simple arrangement produced the highest 80m single band score! Its small size and high efficiency makes this rig ideal for portable operation. A 600mAh battery will give several hours of QRP pleasure - no problem hiding away a complete station in the holiday suitcase!

A QRP Transceiver for 1.8MHz

This design by S E Hunt, G3TXQ, originally appeared in *Radio Communication* [6].

Introduction

This transceiver was developed as part of a 1.8MHz portable station, the other components being a QRP ATU, a battery-pack and a 200ft kite-supported antenna. It would be a good constructional project for the new licensee or for anyone whose station lacks 1.8MHz coverage. The 2W output level may seem a little low, but it results in low battery drain and is adequate to give many 1.8MHz contacts.

The designer makes no claim for circuit originality. Much of the design was adapted from other published circuitry; however, he does claim that the design is repeatable - six transceivers have been built to this circuit and have worked first time. Repeatability is achieved by extensive use of negative feedback; this leads to lower gain-per-stage (and therefore the need for more stages) but makes performance largely independent of transistor parameter variations.

Circuit description

The transceiver circuit (**Fig 7.35**) comprises a direct-conversion receiver together with a double-sideband (DSB) transmitter. This approach results in much simpler equipment than a superhet design, and is capable of surprisingly good performance, particularly if care is taken over the mixer circuitry.

During reception, signals are routed through the band-pass filter (L1, L2 and C25-C31) to a double-balanced mixer, M1, where they are translated down to baseband. It is vital for the mixer to be terminated properly over a wide range of frequencies, and this is achieved by a diplexer comprising R34, RFC2 and C32-R34. Unwanted RF products from the mixer, rejected by RFC2, pass through C32 to the 47-ohm terminating resistor R34. The wanted audio products pass through RFC2 and C34 to a common-base amplifier stage which is biased such that it presents a 50-ohm load impedance. The supply rail for this stage comes via an emitter-follower, TR5, which has a long time-constant (4s) RC circuit across its base. This helps to prevent any hum on the 12V rail reaching TR6 and being amplified by IC3.

The voltage gain of the common-base stage (about x20) is controlled by R37 which also determines the source resistance for the following low-pass filter (L3, L4 and C39-C43). This filter is a Chebyshev design and it determines the overall selectivity of the receiver. The filter is followed by a single 741 op-amp stage which give adequate gain for headphone listening; however, an LM380 audio output stage can easily be added if you require loudspeaker operation.

On transmit, audio signals from the microphone are amplified in IC1 and IC2, and routed to the double balanced mixer where they are heterodyned up into the 1.8MHz band as a double-sideband, suppressed-carrier signal. Capacitors C56 and C57 cause some high-frequency roll-off of the audio signal and thereby restrict the transmitted bandwidth.

A 6dB attenuator (R12-R14) provides a good 50Ω termination for the mixer. The DSB signal is amplified by two broad-band feedback amplifiers, TR2 and TR3, each having a gain of 15dB. TR3 is biased to a higher standing current to keep distortion products low.

The PA stage is a single-ended design by VE5FP [7]. The inclusion of unbypassed emitter resistors R30-R32 establishes the gain of the PA and also helps to prevent thermal runaway by stabilising the bias point. Additional RF negative feedback is provided by the shunt feedback resistor R29. The designer chose to run the PA at a moderately high standing current

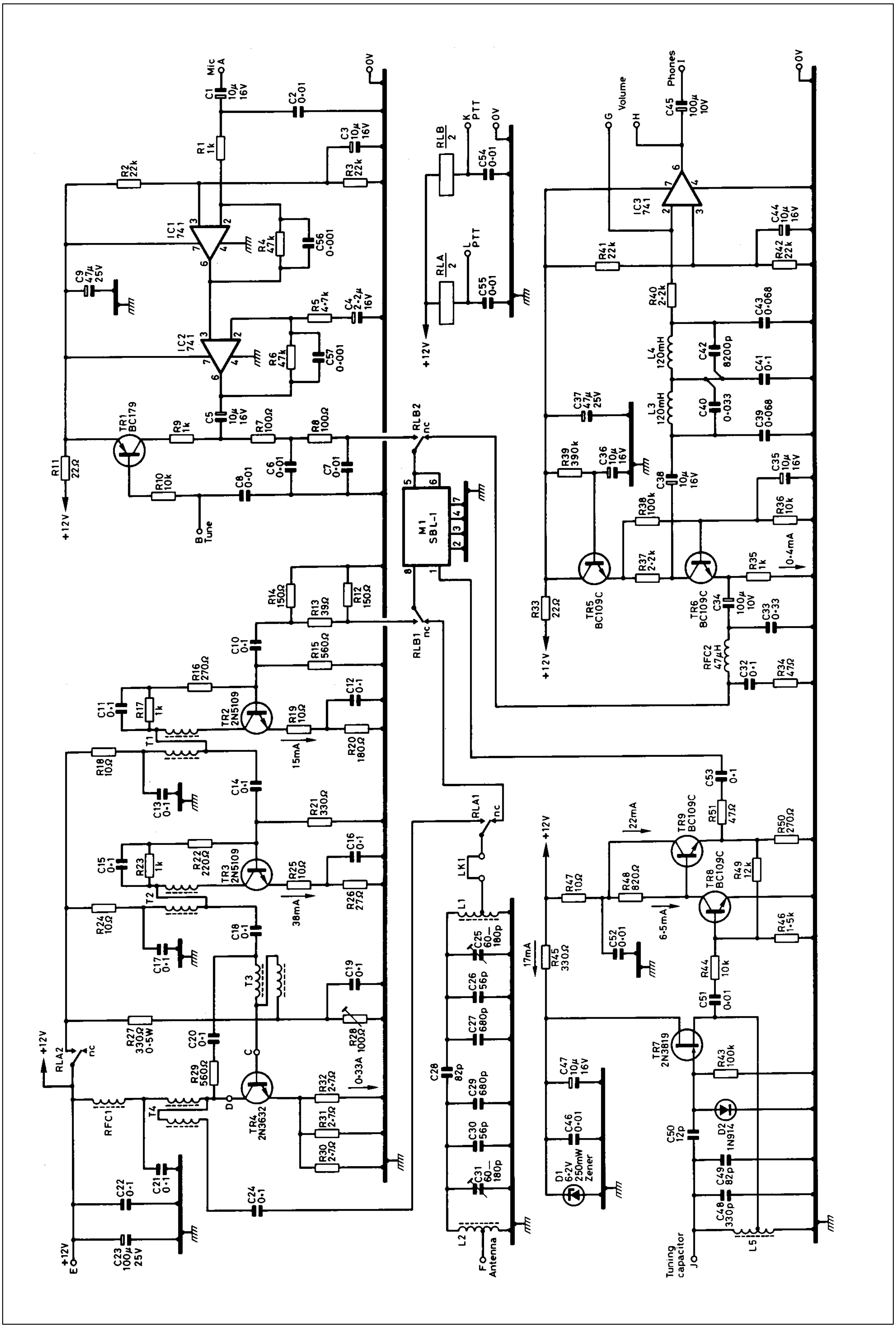


Fig 7.35: Circuit of the G3TXQ transceiver

Fig 7.36: Front view of G3TXQ QRP transceiver. On this prototype the TUNE switch is labelled TEST

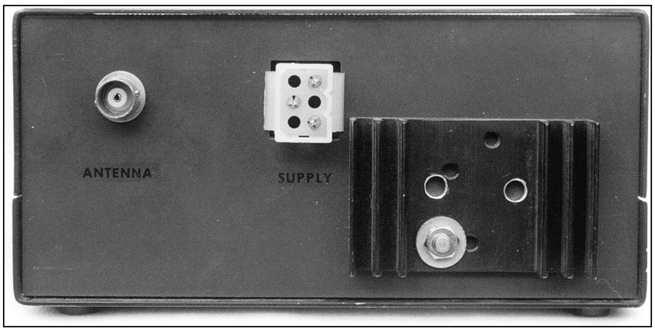


Fig 7.37: Rear view of G3TXQ QRP transceiver

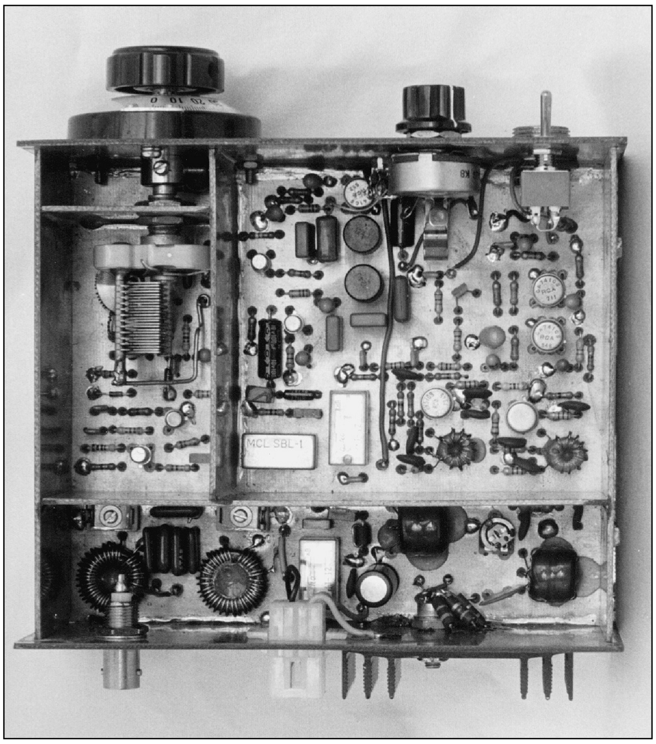


Fig 7.38: The top view of the G3TXQ low power 1.8MHz transceiver

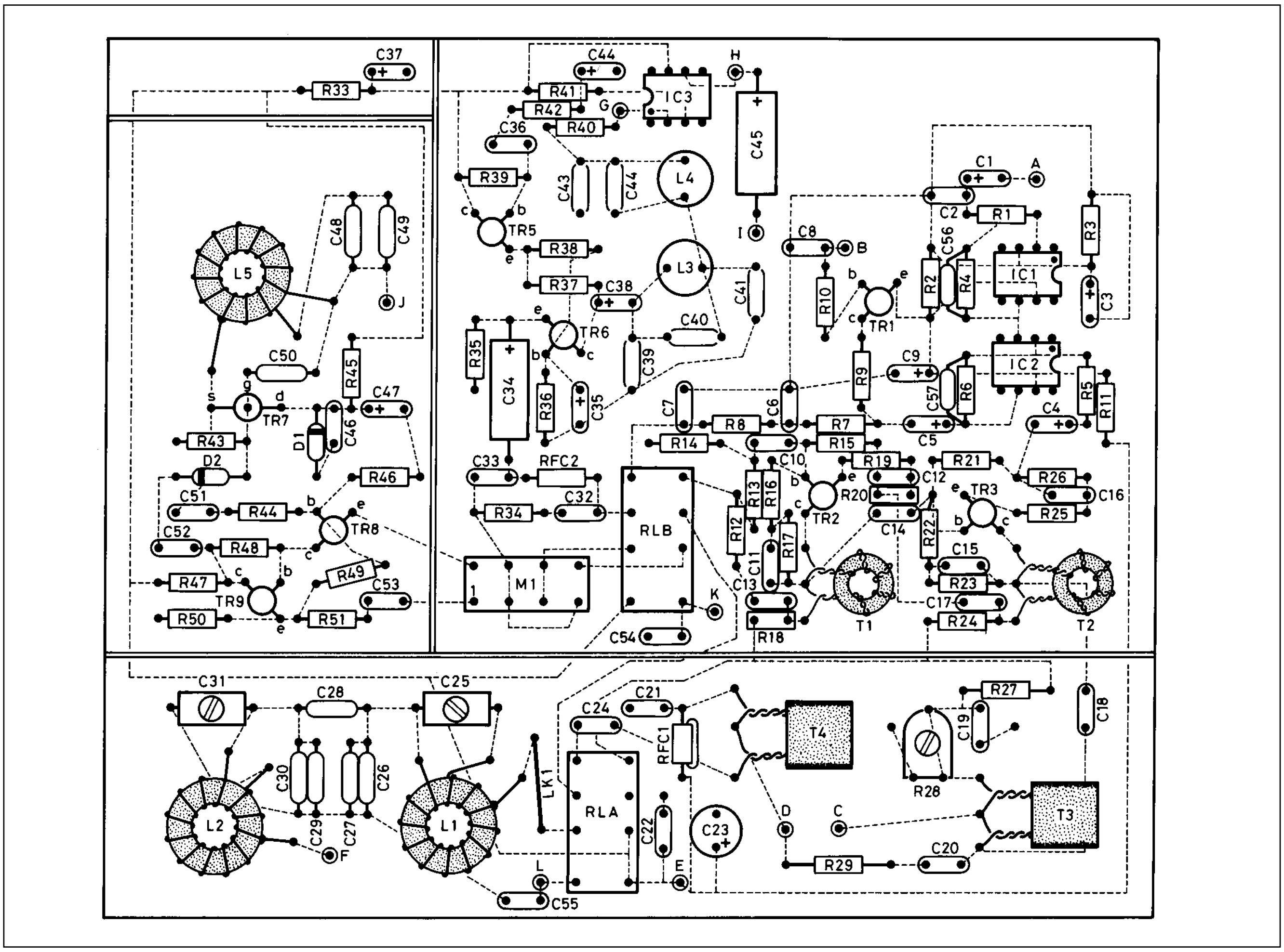


Fig 7.40: Component layout for the G3TXQ transceiver. The PCB details (Fig 7.39) can be found in Appendix B

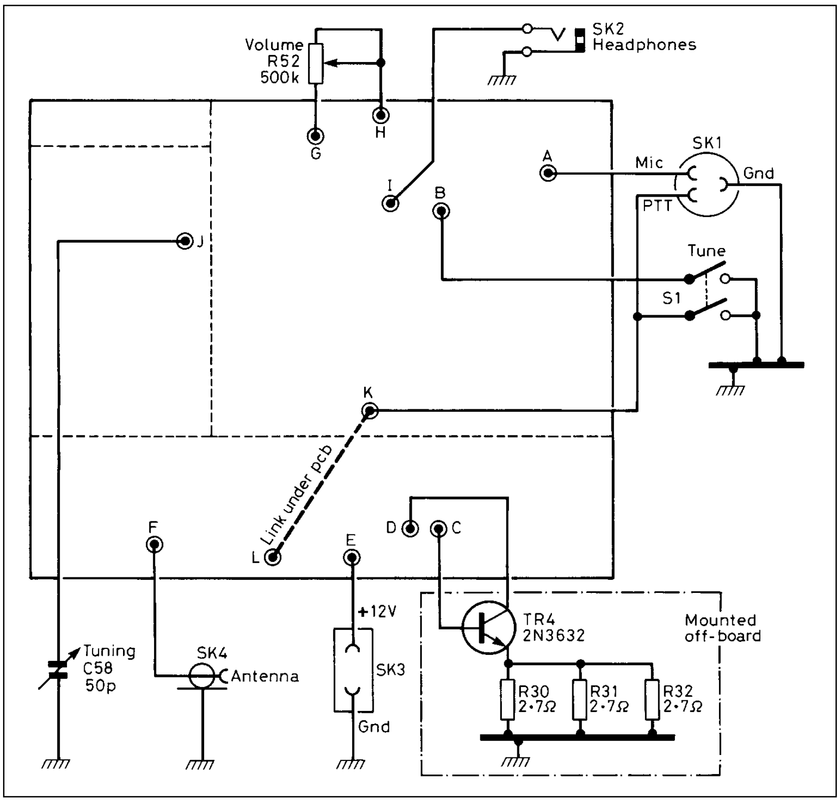


Fig 7.41: Wiring diagram for the G3TXQ transceiver

(330mA) in order to reduce distortion products, thinking that at some stage he might use the transceiver as a 'driver' for a 10-15W linear amplifier.

The PA output (about 2W PEP) is routed through the band-pass filter to the antenna. The designer used a 2N3632 transistor in the PA because he happened to have one in the junk-box; the slightly less expensive 2N3375 would probably perform just all well. VE5FP used a 2N5590 transistor but this would need different mounting arrangements.

At the heart of the transceiver is a Hartley VFO comprising TR7 and associated components. The supply of this stage is stabilised at 6.2V by zener diode D1 and decoupled by C46 and C47. It is important for best stability that the Type 6 core material is used for L5 as this has the lowest temperature coefficient. Output from the VFO is taken from the low impedance tap on L5.

The VFO buffer is a feedback amplifier comprising TR8 and TR9. The input impedance of this buffer is well-defined by R44 and presents little loading of the VFO. Its gain is set by the ratio R49/R44 and R51 has been included to define the source resistance of this stage at approximately 50Ω. Changeover between transmit and receive is accomplished by two DPDT relays which are energised when the PTT lines are grounded.

A CW signal for tuning purposes can be generated by grounding the TUNE pin - this switches on TR1, which in turn unbalances the mixer, allowing carrier to leak through to the driver and PA stages.

Construction

The transceiver, pictured in **Figs 7.36-7.38**,is constructed on a single 6 by 5in PCB. The artwork, component layout and wiring diagram are shown in **Figs 7.39** (see Appendix B)**, 7.40 and 7.41** respectively. The PCB is double-sided - the top (component) surface being a continuous ground plane of unetched copper. The Components List is shown in **Table 7.11**.

R1, 9, 17,		C1, 3, 5,		R52	500k log pot
23, 35	1k	35, 36, 38,		L1, 2	37t on T68-2 core tapped at 7t from ground
R2, 3,		44, 47	10µ 16V tant bead	L3, 4	120mH (eg Cirkit 34-12402)
41, 42	22k	C2, 6, 7,		L5	57t on T68-6 core tapped at 14t from ground
R4, 6	47k	8, 46, 51,		RFC1	2t on small ferrite bead
R5	4k7	52, 54, 55	0.01µ ceramic	RFC2	47µH choke
R7, 8	100R	C4	2µ2 16V tant bead	T1, 2	10t twisted wire on 10mm OD ferrite toroidal core
R10, 36, 44	10k	C9, 37	47µ 25V tant bead		Al = 1µH/t (eg SEI type MM622). See fig.
R11, 33	22R	C10-22, 24,		T3, 4	4t twisted wire on two 2-hole ferrite cores. Al = 4 H/t
R12, 14	150R	32, 53	0.1µ ceramic		(eg Mullard FX2754). See fig.
R13	39R	C23	100µ 25V elect	TR1	BC179
R15, 29	560R	C25, 31	60-180p trimmer	TR2, 3	2N5109 or 2N3866
R16, 50	270R		(Cirkit 06-18006)	TR4	2N3632 (see text)
R18, 19, 24,		C26, 30	56p silver mica	TR5, 6, 8, 9	BC109C
25, 47	10R	C27, 29	680p silver mica	TR7	2N3819
R20	180R	C28	82p silver mica	D1	6.2V 250mW zener
R21, 45	330R	C33	0.33µ	D2	1N914
R22	220R	C34, 45	100µ 10V elect	IC1, 2, 3	741 op-amp
R26	27R	C39, 43	0.068µ	M1	Mini-circuits SBL-1 double-balanced mixer
R27	330R 0.5W	C40	0.033µ	RLA, B	DPDT 12V relay (eg RS Electromail 346-845)
R28	100R preset	C41	0.1µ polystyrene	SK1	Microphone socket
R30, 31, 32	2R7	C42	8200p silver mica	SK2	Headphone socket
R34, 51	47R	C48	330p silver mica	SK3	DC power socket (eg Maplin YX34M)
R37, 40	2k2	C49	82p silver mica	SK4	Antenna socket
R38, 43	100k	C50	12p silver mica	S1	DPDT toggle switch
R39	390k	C56, 57	0.001µ ceramic	Slow-motion drive for C58 (eg Maplin RX40T)	
R46	1k5	C58	50p air-spaced	Heatsink approx 1.5 by 2in	
R48	820R		variable, SLC law	Knob for R52	
R49	12k		(Maplin FF45Y)		

Table 7.11: 1.8MHz QRP transceiver components list

Without the facility to plate-through holes, some care needs to be taken that components are grounded correctly. Where a component lead is not grounded, a small area of copper must be removed from the ground plane, using a spot-cutter or a small twist drill. Where a component lead needs to be grounded, the copper should not be removed and the lead should be soldered to the ground plane as well as to the pad on the underside. This is easy to achieve with axial-lead components (resistors, diodes etc) but can be difficult with radial-lead components. In most cases the PCB layout overcomes this by tracking radial leads to ground via nearby resistor leads. A careful look at the circuit diagram as each component is loaded soon shows what is needed. Remember to put in a wire link between pins L and K, and in position LK1. Screened cable was used for connecting pins G and H to the volume control - connect the outer to pin H.

There are no PCB pads for C56 and C57, so these capacitors should be soldered directly across R4 and R6 respectively. TR4 must be adequately heatsinked as it dissipates almost 4W even under no-drive conditions. TR4 was bolted through the rear panel to a 1.5 by 2.5in finned heatsink. Resistors R30-R32 are soldered directly between the emitter of TR4 and the ground plane.

It is important that the VFO coil L5 is mechanically stable. Ensure that it is wound tightly and fixed rigidly to the PCB; the coil was 'sandwiched' between two Perspex discs and bolted through the discs to the PCB (**Fig 7.42**). Also, be sure to use rigid heavy-gauge wire for connecting to C58.

The designer used a 6:1 vernier slow-motion drive which, with the limited tuning range of 100kHz, provides acceptable bandspread; the 0-100 vernier scale (0 = 1.900MHz, 100 = 2.000MHz) gives a surprisingly accurate read-out of frequency, the worst-case error being 1kHz across the tuning range.

The broad-band transformers, T1-T4, are wound by twisting together two lengths of 22SWG enamelled copper wire. The twisted pair is then either wound on a ferrite toroidal core (T1 and T2), or wound through ferrite double-holed cores (T3 and T4). Identify the start and finish of each winding using an ohmmeter - connect the start of one wire to the finish of the other to form the centre tap (see **Figs 7.43 and 7.44** for more details). All transformers and the band-pass filter coils were secured to the PCB with adhesive. The designer made all of the transceiver, other than the top and bottom panels, by soldering together double-sided PCB materials.

It is vital to have a good screen between the PA and the VFO, otherwise the transmitter will frequency modulate badly. Two-inch high screens were used around the PA and VFO area, and a screen was included at the front of the VFO compartment on which to mount C58. If you use lower screens you may need to put a lid over the VFO; cut a tightly fitting piece of PCB material and bolt it in position to four nuts soldered into the corners of the VFO compartment.

Alignment

Check the PCB thoroughly for correct placement of components and absence of solder bridges.

Turn the volume control fully counter-clockwise, the TUNE switch to the off position and R28 fully counter-clockwise. Connect the transceiver to a 12V supply and switch on. Check that the current drawn from the supply is about 50mA.

Check the frequency of the VFO either by using a frequency counter connected to the source of TR7, or by monitoring the VFO on another receiver. With C58 set to mid-position, the frequency should be about 1.95MHz; if it is very different, you can adjust L5 slightly by spreading or squeezing together the turns. Alternatively, major adjustments can be made by substituting alternative values for C49. Check that the range of the VFO is about 1.9 to 2.0MHz. Plug in a pair of headphones and slowly advance the volume control; you should hear receiver noise (a hissing sound). If you have a signal generator, set it to 1.95MHz and connect it to the antenna socket; if not, you will have to connect the transceiver to an antenna and make the next adjustment using an off-air received signal. Tune to a signal at 1.95MHz and alternately adjust C25 and C31 for a peak in its level.

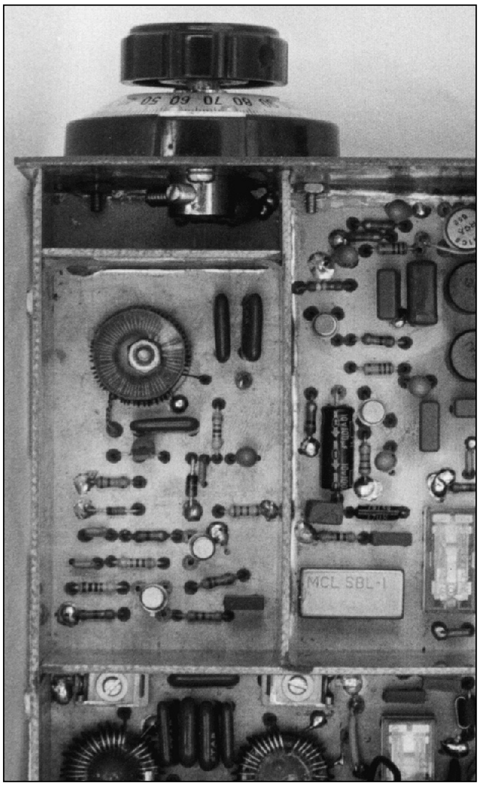


Fig 7.42: Detail of top view with C58 removed to show mounting arrangements

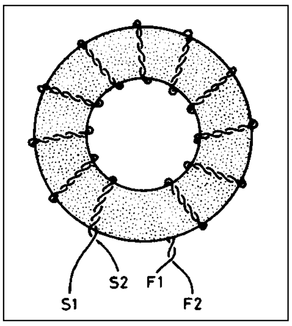


Fig 7.43: Winding details of T1 and T2. Connect S2 and F1 to form the centre tap. Note that the two wires are twisted together before winding. S1, F1: start and finish respectively of winding 1. S2, F2: start and finish respectively of winding 2. Core: 10mm OD ferrite toroid

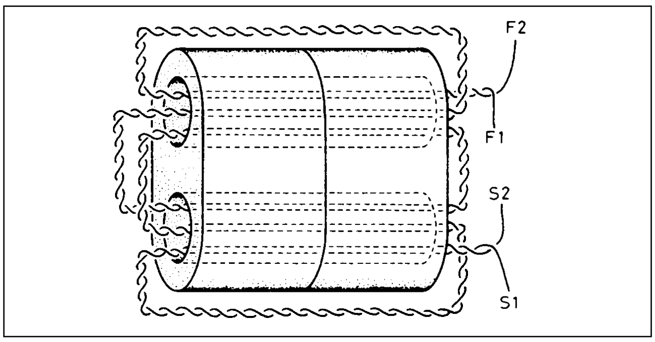


Fig 7.44: Winding details of T3 and T4. Connect S2 and F1 to form the centre tap. Note that the two wires are twisted together before winding. S1, F1: start and finish respectively of winding 1. S2, F2: start and finish respectively of winding 2. Core: two 2-hole cores stacked end-to-end

	Emitter	Base	Collector	Note
TR1	12.2	11.6	11.8	Tune switch operated
TR2	2.85	3.6	12	Transmit
TR3	1.4	2.15	11.6	Transmit
TR4	0.3	1	12.2	Transmit
TR5	11.2	11.8	12.2	
TR6	0.4	1	10.3	
TR8	0	0.65	6.75	
TR9	6	6.75	12	

Table 7.12: Bipolar transistor DC voltages (with 12.2V supply)

	Source	Gate	Drain
TR7	0	0	6.2

Table 7.13: FET DC voltages (with 12.2V supply)

Circuit node	AC voltage	Notes
TR7 source	2.6V p/p	1.8MHz RF
TR9 emitter	2.6V p/p	1.8MHz RF
Mic input	4mV p/p	Transmit audio
IC1 pin 6	200mV p/p	Transmit audio
IC2 pin 6	2.2V p/p	Transmit DSB RF
TR2 base	200mV p/p	Transmit DSB RF
TR4 collector	15V p/p	Transmit DSB RF
Ant (50)	30V p/p	Transmit DSB RF

Table 7:14: AC voltages

Connect the transceiver to a 50Ω power meter or through an SWR bridge to a 50Ω load. Plug in a low-impedance microphone and operate the PTT switch. Note the current drawn from the supply - it should be about 200mA. Slowly turn R28 clockwise and note that the supply current increases; adjust R28 until the supply current has increased by 330mA. Release the PTT switch and operate the TUNE switch; the power meter should indicate between 1 and 2W. At this stage, final adjustments can be made to C25 and C31. Swing the VFO from end to end of its range and note the variation in output power. The desired response is a slight peak in power at either end of the VFO range with a slip dip at mid-range. It should be possible to achieve by successive adjustments to C25 and C31. For those of you lucky enough to have access to a spectrum analyser and tracking generator, LK1 was included to allow isolation of the band-pass filter.

If you have any problems, refer to **Tables 7.12 to 7.14** which show typical AC and DC voltages around the circuit. If necessary, you can tailor the gain of IC2 to suit the sensitivity of your microphone by changing the value of R5.

Final thoughts

In retrospect it would have been useful to have included the low-pass filter (L3, L4, C39-C43) in the transmit audio path to restrict the bandwidth further. Normally the roll-off achieved by C57 / C56 combined with the low output power means that you are unlikely to cause problems for adjacent contacts. But when using a 200ft vertical antenna during portable operation, the transceiver puts out a potent signal and a bandwidth reduction would then be more 'neighbourly'.

A CW facility could be added fairly easily using the TUNE pin as a keying point. You would need to add RIT (receive independent tune) facilities - probably by placing a varactor diode between TR7 source and ground. You might also consider changing to a band-pass audio filter rather than a low-pass audio filter in the receiver.

The transceiver can be adapted for other bands by changing the VFO components and the band-pass filter components - all other circuitry is broad-band. You will need to worry more about VFO stability as you increase frequency, and you may find the gain of the buffer falls - you can overcome this by decreasing the value of R44. The noise figure of the receiver is adequate for operation on the lower frequency bands but on 14MHz and above a preamplifier will probably be needed. Those who enjoy experimentation might try changing the VFO to a VXO, adding a preamplifier to the receiver, and seeing if operation on 50MHz is possible!

Finally, it has been interesting to note that, despite theory, with careful tuning it is quite possible to resolve DSB signals on the direct-conversion receiver.

The FOXX2 Transceiver

The FOXX2 is based on a design by George Burt, GM3OXX, and modified by Rev George Dobbs, G3RJV. This article originally appeared in *SPRAT* [8] and shows just how simple a CW transceiver can be.

Introduction

The FOXX transceiver is an elegant little circuit which uses the same transistor for the transmit power amplifier and the receive mixer. It is capable of transceiver operation on several bands and generates around 1W of RF power out. The original FOXX circuit has been revised with a few design changes.

Circuit design

The circuit diagram is shown in **Fig 7.45**. TR1 is a VXO (variable crystal oscillator) stage. The feedback loop formed by the crystal and the trimmer capacitor (C1) tunes the circuit to the desired frequency. C1 provides a small amount of frequency shift. The output is coupled to a power amplifier stage. This stage is unusual in that a PNP transistor is used with the emitter connected to the positive supply and the output taken from the collector load, which goes to ground. The output of the transmitter may be adjusted by a

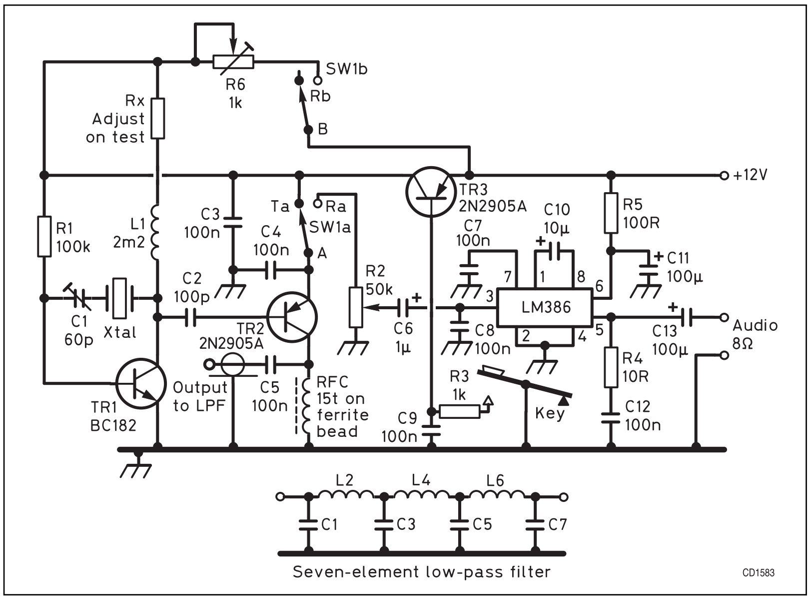


Fig 7.45: FOXX2 circuit diagram

Band (MHz)	C1, 7 (pF)	C3, 5 (pF)	L2, 6 (turns)	L4 (turns)	Core	Wire (SWG)
3.5	470	1200	25	27	T37-2	28
7.0	270	680	19	21	T37-6	26
10.1	270	560	19	20	T37-6	26
14.0	180	390	16	17	T37-6	24

Table 7.15: FOXX2 low-pass filter values

resistor (Rx - a few hundred ohms) to around 1W. TR2 should be fitted with a clip-on heatsink. TR3, another PNP transistor, allows the transmitter to be keyed with respect to ground. TR3 and TR2 are both 2N2905A PNP switching transistors.

The low-pass filter is a seven-element circuit based on the circuit and constants described by W3NQN. The transmit-receive function is performed by a double-pole, double-throw switch, SW1A and SW1B. The receive position has two functions. It bypasses the keying transistor, TR3, to ensure that the oscillator TR1 remains on during the receive position to provide the local oscillator. It also switches the supply line away from the power amplifier TR3 and connects the latter to the audio amplifier. In this position TR3 functions as a diode mixer, mixing the signals from the antenna which appear at the emitter and the signal from TR1.

The audio amplifier is an LM386 working in maximum gain mode. The supply for the LM386 is taken directly from the 12V supply line which means it is on during both transmit and receive functions. This has the advantage of providing a rudimentary sidetone to monitor the keying. 'Sidetone' is an over-statement because all it does is produce clicks in time with the keying.

A preset potentiometer is added in series with TR1 supply on receive. This is a very simple form of RIT (receiver incremental tuning). If the supply voltage to TR1 is reduced enough, it shifts the frequency of the oscillations. Assuming the value of Rx to be in the order of a few hundred ohms (just to reduce the drive from TR1 a little on transmit), a 1kΩ preset at R6 can be set to shift the frequency by around 700-800Hz, giving a comfortable offset for CW reception. Values for the seven-element low pass filter are shown in **Table 7.15**. The wire gauge is not critical but wind the coil so as to comfortably fill the core over about three-quarters of its full circumference. The PCB layout is shown in **Fig 7.46** (in Appendix B), the component layout is in **Fig 7.47**, and the Components List is in **Table 7.16**.

The Epiphyte-2

This section is based on articles by Derry Spittle, VE7QK [9] and Rev George Dobbs, G3RJV [10]. The Epiphyte is a remarkable little transceiver which has introduced many QRP operators to the pleasure of building their own SSB equipment. It was designed by Derry Spittle, VE7QK, who needed reliable radio communication when journeying into wilderness areas in British Columbia. Beyond the range of VHF repeaters, simple battery-operated HF equipment offers the only practical means of communication. The objective was to build a small transceiver capable

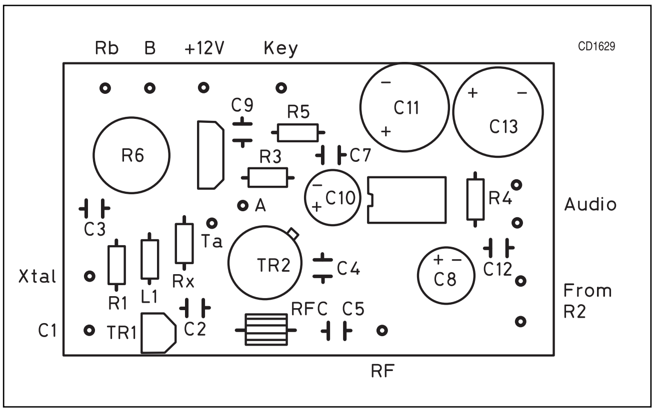


Fig 7.47: FOXX2 component layout (See Appendix B for PCB)

R1	100k	C3-5, 7-9, 12	100n
R2	50k	C6	1µ
R3	1k	C10	10µ
R4	10R	C11, 13	100µ
R5	100R	L1	2.2mH
R6	1k	RFC	15t on ferrite bead
Rx	See text	TR1	BC182
C1	60p trimmer	TR2, 3	2N2905A
C2	100p	IC1	LM386

Table 7.16. FOXX2 component list

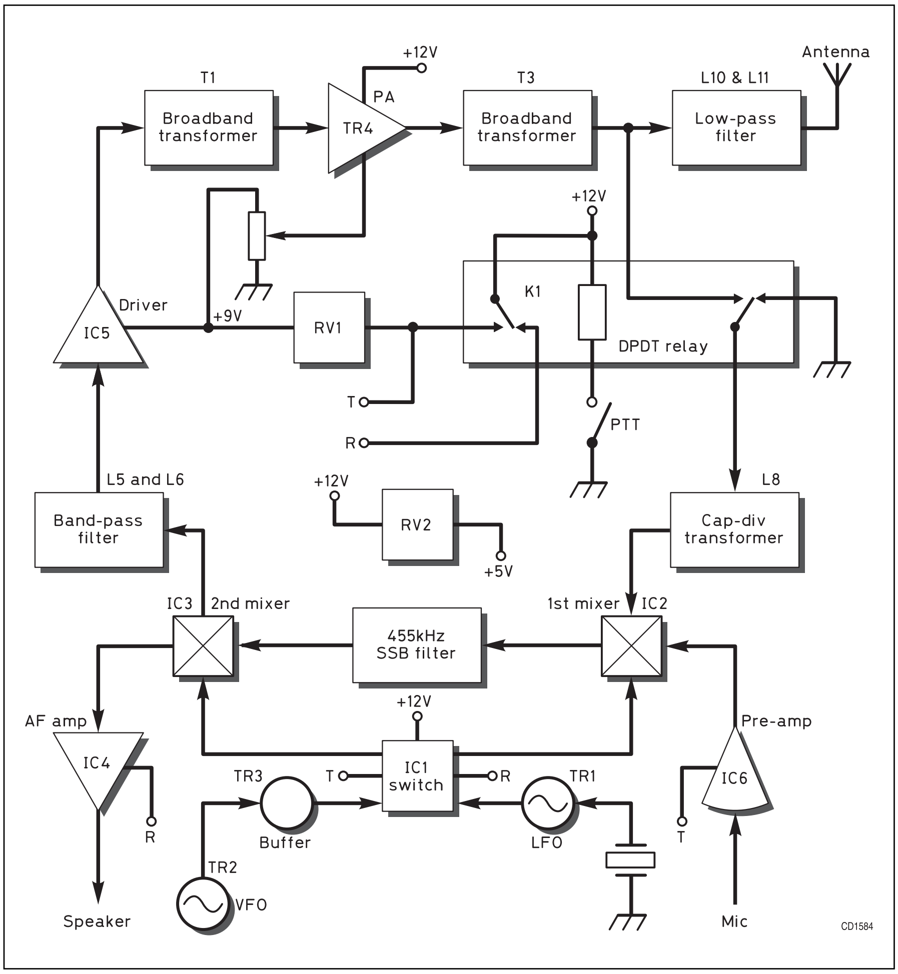


Fig 7.48: Block diagram of Epiphyte-2

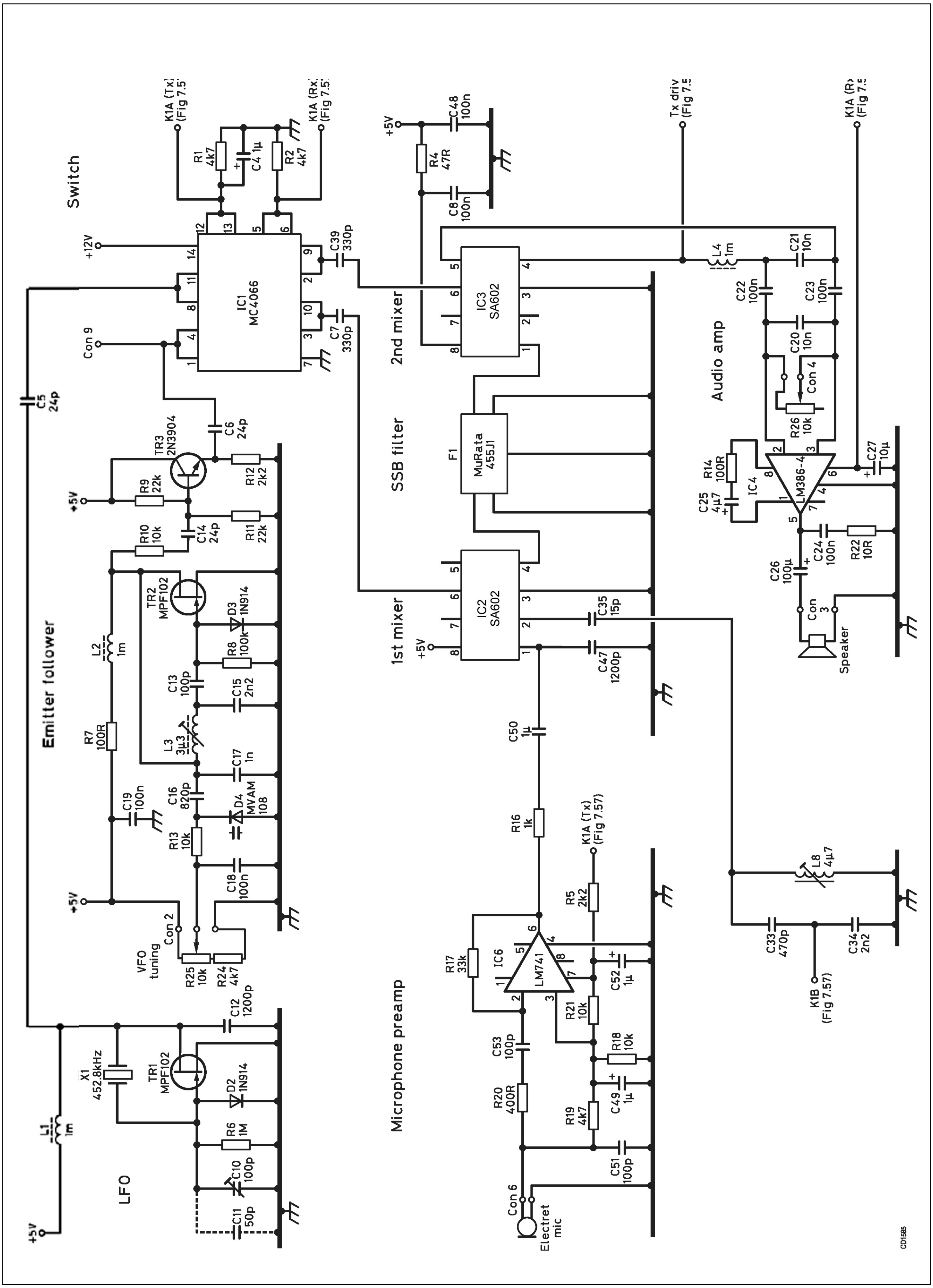


Fig 7.49: Circuit of Epipyhte-2 receiver and SSB generator

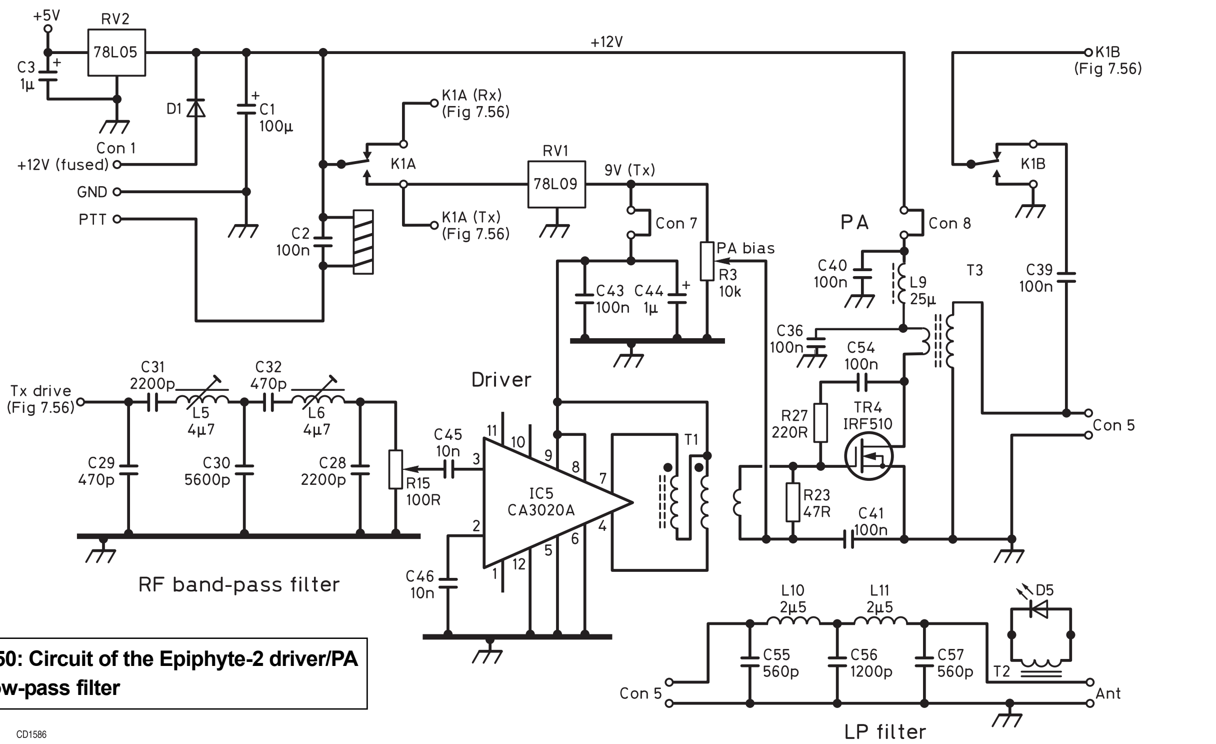


Fig 7.50: Circuit of the Epiphyte-2 driver/PA and low-pass filter

R1, 2, 19, 24	4k7	C27	10µ tantalum
R3	10k lin (multi-turn trim pot)	C29, 32, 33	470p axial polystyrene
R4, 23	47R	C30	5600p axial polystyrene
R5, 12	2k2	C35	15p NPO
R6	1M	C55, 57	560p axial polystyrene
R7, 14	100R	C56	1200p axial polystyrene
R8	100k	L1, 2, 4	1mH choke
R9, 11	22k	L3	3.3µH (Toko BTKANS9445)
R10, 13, 18, 21	10k	L5, 6, 8	4.7µH (Toko 154AN-T1005)
R15	100R lin (multi-turn trim pot)	L9	RFC (7t on Amidon FB-43-801)
R16	1k	L10, 11	2.5µH (25t on Amidon T-32-2)
R17	33k	T1	4:1 broad-band transformer (5t bifilar on FB-43-
R20	400R	801	with 3t link)
R22	10R	T2	2t on Amidon FB-43-2401
R25	10k (precision 10 turn pot)	T3	binocular broadband transformer
R26	10k log		(2t pri, 5t sec on Amidon BM-43-202)
R27	270R	IC1	MC14066
RV1	78L09	IC2, 3	SA602
RV2	78L05	IC4	LM386-4
C1, 26	100µ electrolytic	IC5	CA3020A
C3, 4, 44, 49,		IC6	LM741
52	1µ tantalum	D1	polarity protection diode, eg 1N4001
C50	1µ non-polar ceramic	D2, 3	1N914
C5, 6, 14	24p NPO	D5	LED
C7, 9	330p ceramic	D4	MVAM108
C2, 8, 18, 19,		TR1, 2	MPF102
22, 24, 36,		TR3	2N3904
39-43, 48,		TR4	IRF510
53, 56	100n monolithic ceramic	F1	455kHz SSB filter (MuRata 455J1)
C15, 28, 31, 34	2200p axial polystyrene	K1	Miniature DPDT relay
C10	100p trimmer	X1	455kHz ceramic resonator
C11	50p NPO	Two 3-pin, four 2-pin polarised Molex terminals	
C12, 47	1200p ceramic	Two metering jumpers and terminals	
C13, 51	100p NPO	One 1-pin test point	
C32	470p NPO	Heatsink for TR4	
C16	820p axial polystyrene	Four 8-pin, one 14-pin, one 16-pin IC sockets	
C17	1n axial polystyrene	Four ¼in metal stand-offs	
C20, 21, 45, 46	10n	Printed circuit boards and a kit of parts for the project are available	
C25	4µ7	from JAB Electronic Components [3]	

Table 7.17: Epiphyte-2 component list

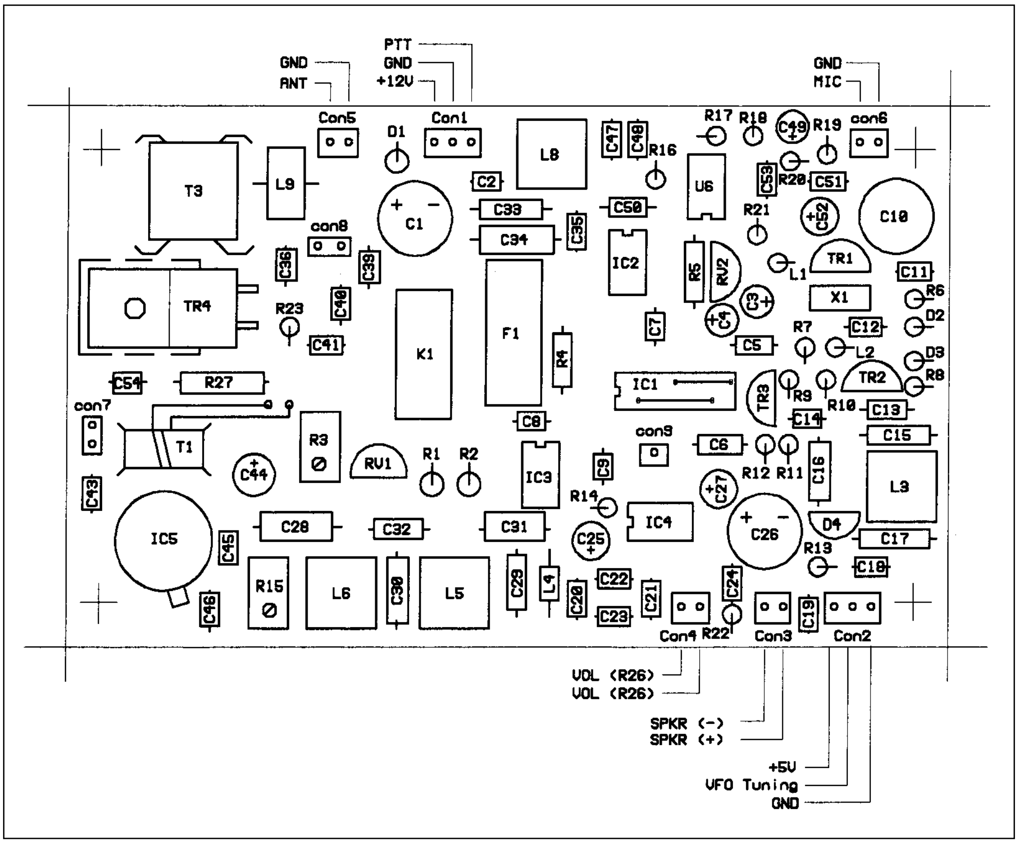


Fig 7.51: Component layout of the Epiphyte-2

of providing effective voice communication with the British Columbia Public Service Net on 3729kHz from anywhere in the province. The Epiphyte began as a project of the QRP Club of British Columbia and the design was first published in 1994 by the G-QRP Club and the QRP Club of Northern California. The Epiphyte-2 now includes many modifications and suggestions since made by their members.

Circuit description

The block diagram of the EP-2 is shown in **Fig 7.48**. The circuit is based around a pair of SA602 double-balanced mixers (IC2, IC3), and a MuRata miniature 455kHz ceramic SSB filter (F1). IC1 switches the LF and HF oscillators, permitting the same mixers to be used for both transmitting and receiving. On transmit, the modulated signal passes though a band-pass filter to remove the image, before being applied to the driver (IC5). Transmit/receive switching is accomplished with a DPDT relay (K1) and IC1. When receiving, B+ is disconnected from the microphone preamplifier (IC6) and the driver (IC5), and forward bias is removed from the PA (TR4). When transmitting, B+ is removed from the audio amplifier (IC6) and the RF input to the first mixer is disconnected from the low-pass filter and grounded. At the same time, the relay provides switching voltages to IC1. The antenna remains connected to the low-pass filter and PA at all times. The circuit is shown in **Figs 7.49 and 7.50. Table 7.17** shows the Components List.

The VFO (TR2) is a varactor-tuned Vackar circuit using a Toko 3.3μH variable coil. This is buffered by an emitter follower (TR3). The LF oscillator (TR1) uses a 455kHz ceramic resonator adjusted to 452.8kHz. An MPF102 FET is used for both oscillators.

The RF bandpass filter was modelled in a series-tuned configuration to give a reasonably flat response over 200kHz and a sharp roll-off on the high-frequency side for rejection of the image frequency. It is designed for an input impedance of 1500Ω (to match the SA602 mixer) and to terminate in a 100Ω resistive load to ensure stability in the driver. It uses a pair of Toko 4.7μH coils and the fixed capacitors are standard values.

The driver stage (IC5) is a CA3020A differential amplifier. Operating from a 9V supply, this stage has a power gain of 60dB and is capable of 500mW output. The broad-band transformer (T1) has a bifilar-wound primary link-coupled to a 47Ω resistive load (R23) at the gate of TR4. The power amplifier (TR4) is an IRF510 MOSFET with an RF output of 5W PEP. A broad-band transformer (T3) matches the output to a conventional 50Ω low-pass filter.

The receiver RF input from the low-pass filter is a capacitance-divider matching circuit to the first mixer and is tuned to the centre of the phone band. The AF amplifier (IC4) is a LM386 with a balanced input. The receiver is quite sensitive and with a dipole or inverted-V antenna at 25ft it provides adequate speaker volume from all but the weakest signals. Band noise is usually the limiting factor.

R19 provides the polarising voltage for an electret microphone (two-terminal type) and should be omitted if a dynamic microphone is used. The speech amplifier (IC6) is an LM741. The value of R20 should match the impedance of the microphone and the stage gain may be set by adjusting the value of R17.

The component layout is shown in **Fig 7.51** and the PCB layout is in **Fig 7.52** (in Appendix B). A ground loop present in earlier versions (eg **Fig 7.53**) has now been eliminated.

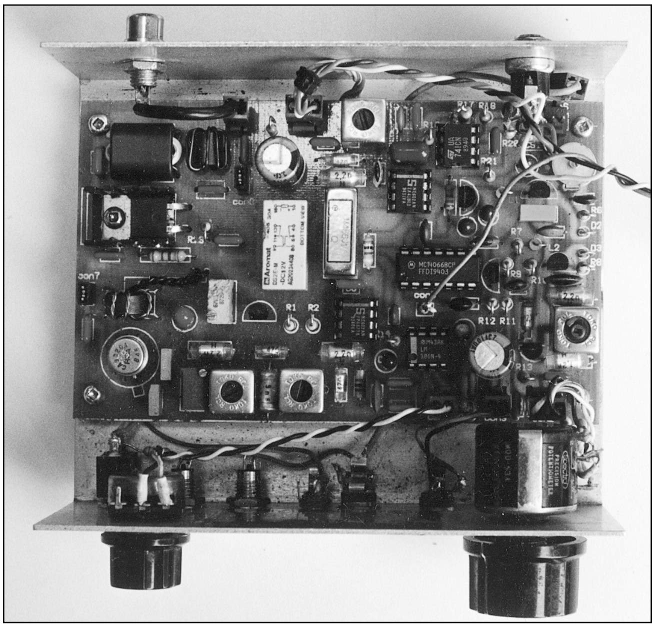


Fig 7.53: An assembled Epiphyte-2 which uses a earlier version of the PCB

Assembly

Assembly is fairly straightforward. Ensure that the Toko coils (L3, 5, 6, 8), SSB filter (F1), ceramic resonator (X1) and trimmer capacitor (C1) fit the PCB and enlarge the holes if necessary. Install the CA3020A (IC5) first as it is easier to align the 12 pins without the other components in place. The tab is over pin 12. Be sure to solder in the two bare jumper wires on top of the PCB before installing the socket for IC1. Some fairly large value polystyrene capacitors are specified and their physical size should be ascertained before ordering if they are to fit comfortably on the board.

Alternatively, NP0/C0G ceramic capacitors may be substituted. Cut off the centre pin (drain) before mounting the IRF510 (TR4) and heatsink with a 4-40 machine screw, nut and star washers. Output from TR4 is taken from the tab. Remove or cut-off unused terminals from the relay socket. Finally, don't bother soldering the three unconnected pins on the Toko coils to the ground plane - you may need to remove them one day!

Alignment and testing

This must be carried out with the single-sided PCB fastened to a ground plane with four metal stand-offs.

1. Remove both metering jumpers. Install the relay (K1) and PTT switch if not built into the microphone. Remove all socketed ICs and connect to a 12-14V FUSED supply. Verify that RV2 is delivering 5V and that RV1 is delivering 9V on transmit. With an RF probe, check that both oscillators are functioning. Install IC1 and verify that the VFO is switching between transmit and receive.
2. Set the LFO to 453kHz with the trimmer (C10). Monitor the ninth harmonic at 4.075MHz on a communications receiver. It may be necessary to change the value of the padder (C11).
3. Set R24 to mid position and adjust L3 until the frequency at the 'test point' (Con9) measures approximately 4.2MHz. To avoid having the ferrite core protrude above the case of L3, screw it completely into the coil and tune it upwards (anticlockwise). The core will remain firmly in position without further 'fixing'. R24 limits the overall frequency coverage and, at the same time, ensures that the varactor diode is always biased positive. The value of R24 may be changed to alter the bandspread.
4. Install IC4 and connect the antenna, speaker and volume control (R26). The leads to R26 should be kept as short as possible. Test the receiver and adjust L8.
5. The RF voltage at pin 6 on IC2 and IC3 should read 100 to 150mV RMS. If necessary, change the value of C5 and/or C6 to adjust.
6. Set the RF drive control (R15) to minimum. The transmit idling current in the driver (IC5) should measure 25mA +10% at Con 7. If not then TR1 or IC5 has probably been installed incorrectly.
7. Adjust R3 to set the transmit idling current in the power amplifier (TR4) to read 50mA at Con8. Install IC6, microphone, both metering jumpers and a 50-ohm dummy load at the antenna terminal. Advance the RF driver (R15) until RF voltage appears across the dummy load while modulating with a constant tone (whistle!). Adjust the band-pass filter (L5, L6) to maximise, but do not 'stagger tune' them. Continue increasing the drive until it peaks at around 16V RMS with modulation. The driver current should rise to 60 or 70mA and the PA current to around 600mA with a continuous heavy tone. The 'average' current with normal speech modulation will be considerably less. Monitoring your own signal with phones on a receiver will disclose any serious problem before testing on the air with a local amateur.

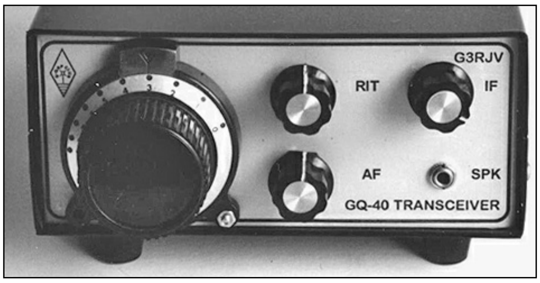


Fig 7.54: GQ-40 exterior view

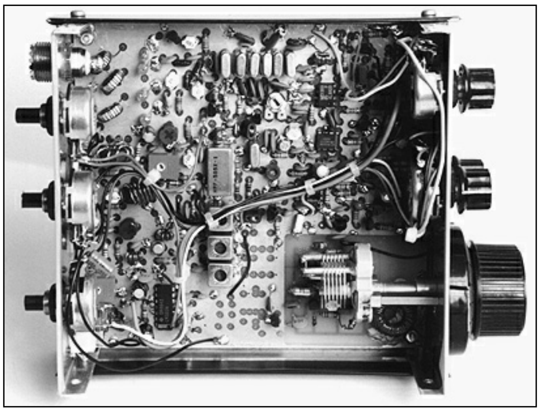


Fig 7.55: GQ-40 interior view

The GQ-40 CW Transceiver

This section describes the GQ-40 (**Figs 7.54 and 7.55**) a high-performance, single-band CW transceiver by GW8ELR which first appeared in *SPRAT* [11].

The main Components List is shown in **Table 7.18**, and component values are provided for 7, 10, and 14MHz versions (**Tables 7.19, 7.20 and 21**). **Table 7.22** shows transformer winding details.

General

The receiver is a conventional superhet design at an IF of 4.4MHz, with a high-dynamic-range, double-balanced, mixer, a six-pole 500Hz crystal IF ladder filter and high-power audio stage. The transmitter is of the mixer type, utilising the receive mixer in a bi-directional mode. The driver and PA are MOSFETs of the inexpensive commercial switching type; adjustable gate bias allows user selection of operating class. The PA is run in push-pull to give a low harmonic content and when in Class AB1 an output of 7W is typical. Power output in normal operation is fully variable via a power control (drive) potentiometer.

Full QSK is achieved by electronic timing of the antenna pin diodes, positive supply lines and IF gain control. Conventional rectifier diodes 1N4007 are used for the antenna changeover system as these are inexpensive and have a similar doping profile to more expensive PIN diodes. Insertion loss is less than 0.1dB at 10MHz with an IP3 of +50dBm when biased at 5mA minimum.

Frequency control is by a Colpitts VFO with a high-quality variable capacitor. Due to the high frequency required for the local oscillator on 14MHz, the VFO is pre-mixed with a crystal oscillator.

Receiver front-end

Signals from the antenna are routed through the LPF L5-7/C74-77 to the diode changeover system (**Fig 7.56**). D5/6 are biased from the permanent 12V line when TR6 is switched on by the receive 12V line. Bias is regulated by R26/25 with RFC2-4 and C55, 56 keeping RF from entering the DC supply. D7 prevents signal loss during receive through T6. Signals now enter the three-pole input filter formed by C1-7/L1-3 (Fig 7.52). The input filter is a low-loss

R1, 41	47k
R2, 4	180R
R3, 5	10R
R6	560R
R7, 9, 24, 51, 52, 65	4k7
R8, 19, 27, 45, 46, 50, 53-56	10k
R15,16	68k
R17, 20, 23, 25, 26, 28	1k
R18	220R
R21	5R6
R22	220k
R29, 42, 44, 47-49, 57, 58, 67	100k
R31	3k
R32	27R
R33, 35, 36	22k
R34	2k2
R10, 38-40	not used
R59	390R
R63	680R
R64	12R
RFC2-5	1mH
VR1	1k preset
VR2, 3	10k preset
RV1, 2	10k log
RV3	22k log switched
RV4	100k lin
RV5, 6	22k lin
C14, 20, 28	100p
C8, 12, 23-27, 29, 34, 35, 40, 42, 45, 55-60, 62, 66, 68,79, 80, 83, 84, 86-91, 100, 101, 110	100n
C9-11, 43, 61, 65, 71, 72	1n
C13, 15, 16, 18, 19, 21, 31	180p
C17	220p
C13a,15a, 9a, 21a	22p
C14a, 20a	4p7
C16a, 18a	8p2
C17a	18p
C22, 32, 38, 39, 49, 53, 63, 64, 69, 70, 73, 78, 97, 111	10n
C30, 48	47p
C33	1µ
C37, 67	10µ
C41	4n7
C44	47µ
C46, 47, 81	100µ
C50	150p
C51	560p
C52	68p
C82	4µ7
C85	22µ tantalum
C92	6p8
TC1, 2	60p (brown)
TC3	6-10p (blue)
T1-3, 6, 7	37KX830 (matt black)
T4, 5, 10	BLN43002402 (two hole balun core)
T8, 9	56-61001101 (matt black)
X1-8	4.4336MHz
RFC1	100µH (Toko 7BS)
RFC2-5	1mH (Toko 7BS)
RFC6,7	15µH (Toko 8RBSH)
RFC8	10µH (Toko 7BS)
D1, 2, 4, 9, 10, 11	1N4148
D3, 8	BA243
D5-7	1N4007
ZD1	BZY88C4V7
ZD3 [varicap D12]	BB105
TR1, 2, 5, 16	J310
TR3, 12	2N3904
TR4	2N2222
TR6	PN2222/MPS3392
TR7	MFE201
TR8	VN66AFD
TR9, 10	IRF510
TR11	2N3906
TR13, 14	BD140
TR15	BS170
TR17	2N3866
IC1	HPF-505X/SBL-1
IC2	MC1350P
IC3, 9	SA612/602
IC4	LF351
IC5	LM380
IC6	4093
IC7, 8	78L08
IC10	78L05
IC11	78L06

Table 7.18: GQ transceiver parts list - all bands

R61	not fitted
R37	820R
C1, 4, 6	100p
C2, 7	1n
C3, 5	3p9
C74, 77	47p + 220p polystyrene 63V
C75, 76	680p polystyrene
C93	47p + 47p + 33p
C94	470p polystyrene
C95, 96	1800p polystyrene
C99	82p
C102-109	not fitted - install bypass link
L1-3	Toko KANK 3334 (yellow)
L4	40t 28SWG on T68-6 or T50-6 (yellow)
L5, 7	21t 26SWG on T37-6 (yellow)
L6	24t 26SWG on T37-6 (yellow)
RFC9	1mH
TC4	not fitted

VFO frequency 2566-2666kHz for 7000-7100kHz coverage

Table 7.19: GQ transceiver - additional parts for 40m version

R61	not fitted
R37	820R
C1, 6	47p
C2, 7	470p
C5	fit wire link
C74, 77	270p polystyrene
C75, 76	560p polystyrene
C93	33p + 39p NP0
C94	220p + 220p polystyrene
C95	180p polystyrene
C96	220p + 220p polystyrene
C99	82p
C102-109	not fitted - install bypass link
C104	47p
TC4	not fitted
L1	not fitted
L2,3	Toko KANK 3334 (yellow)
L4	32t 28SWG on T68-6 or T50-6 (yellow)
L5,7	19t 26SWG on T37-6 (yellow)
L6	20t 26SWG on T37-6 (yellow)
RFC9	1mH

VFO frequency 5666-5766kHz for 10,100-10,200kHz coverage. Adjust VC1S series cap to bandspread

Table 7.20: GQ transceiver parts list - additional parts for 30m version

R61	100R
R37	4k7
C1, 6	120p
C2, 7	1n
C5	4p7
C3	fit wire link
C4	not used
C74, 77	180p polystyrene
C75, 76	390p polystyrene
C93	33p + 39p NP0
C94	220p + 220p polystyrene
C95	180p polystyrene
C96	220p + 220p polystyrene
C98, 102	10n
C99	82p
C103	not used
C104	47p
TC4	18p fixed
C105	56p
C106	100n
C107	1p8
C108	56p
C109	10n
X9	24.00MHz
L1	not fitted
L2, 3	Toko KANK 3335 (pink)
L4	32T 28SWG on T68-6 or T50-6 (yellow)
L5, 7	16T 26SWG on T37-6 (yellow)
L6	17T 26SWG on T37-6 (yellow)
L8, 9	1.2µH Toko KANK 3335 (pink)
RFC9	1mH
IC9	SA612/602

VFO frequency 5566-5466kHz (LO = 24   VFO) for 14000-14100kHz coverage

Table 7.21: GQ transceiver - additional parts for 20m version

T1	K37X830	3t IC1		13t TR2	32SWG
T2	-	15t TR2	10t C13		32SWG
T3	-	6t C21	24t IC2		32SWG
T4	BLN 43002402	2t TR5	Ct D3	6t 12VT	32SWG
T5	-	2t TR1	Ct C11	6t 12VP	32SWG
T6	K37X830	4t D8		15t TR7	32SWG
T7	-	14t TR7	3t R31		32SWG
T8	59-61001101	12t PA		12t TR8	bifilar 28SWG
T9	-	12T			trifilar 28SWG
T10	BLN 43002402	6t R66		2t R3	32SWG

Table 7.22: GQ transceiver transformer winding details

Butterworth design with a bandwidth of 300kHz. Capacitive dividers C1/2 and C6/7 match the characteristic filter impedance to 50Ω. The filter output feeds IC1, a double-balanced hot-carrier mixer, type SBL-1/HPF-505 etc. LO drive for the mixer is provided from a J310 buffer amplifier, and to ensure a correct termination a 50-ohm T pad is used. The mixer is terminated by a simple active termination which uses a step-up transformer to a J310 FET amplifier with a gate resistor. Whilst this is not as good as the more common diplexer arrangement, it requires no setting up.

IF and crystal filter

The filter (**Fig 7.57**) is a six-pole ladder design with a centre frequency of 4.433MHz. The unit uses high-quality, high-volume crystals designed for TV colour burst timing. The filter has a 3dB bandwidth of 500Hz and a Gaussian shape by use of Butterworth design constants. T2, 3 match the input and output impedance of the IF amplifier and product detector. IC2, a MC1350P, provides up to 60dB gain at the IF frequency. During transmit the amplifier is muted by applying bias to the AGC control, pin 5. This bias voltage is switched by TR3 controlled from the T3 line.

A front-panel IF gain control is fed to the AGC control input via D1 and external AGC may also be applied via D2. The exclusion of an AGC system was a deliberate design policy. Many of the

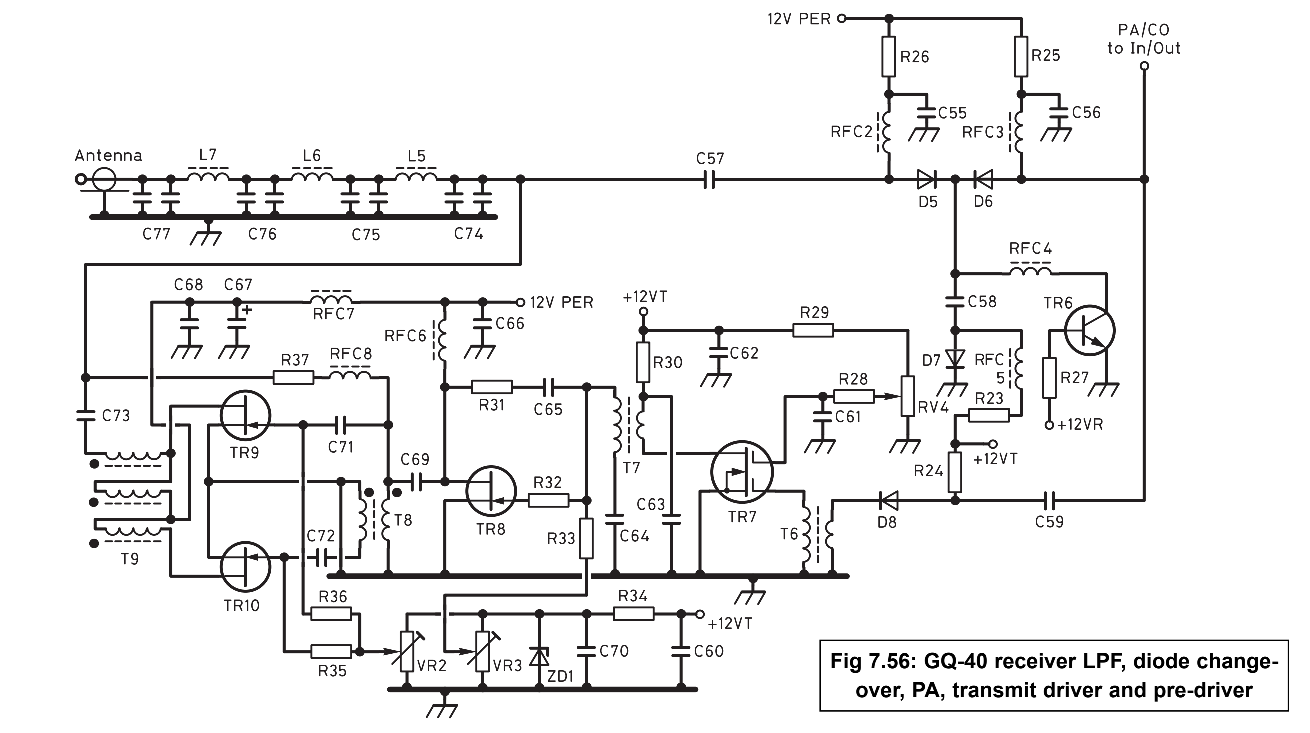


Fig 7.56: GQ-40 receiver LPF, diode change-over, PA, transmit driver and pre-driver

audio-derived AGC systems found in current QRP transceiver designs are far from satisfactory and the inclusion of AGC is often counter-productive to the reception of weak CW signals. If you must add an AGC system, add a good one.

The amplified IF frequency is converted to audio frequency in IC3, an SA602 balanced mixer. LO (BFO) injection for the conversion is provided by an on-chip crystal oscillator. The BFO frequency is adjusted by TC2. The SA602 requires a 6 volt power supply and this is provided from IC11, a three-terminal regulator.

AF preamp and amplifier

A low-noise LM351 preamplifier, IC4, with some frequency tailoring drives an LM380 audio output amplifier (Fig 7.57). The output is capable of about 1W of audio, and has a Zobell network to ensure stability.

Transmit carrier oscillator, buffer and mixer

TR4 is the transmit oscillator (Fig 7.56) and is crystal controlled at 4.433MHz. TC1 allows the frequency to be offset to match the receive BFO audio note. The oscillator is keyed directly at the keying rate via the key line.

TR5 is a J310 buffer amplifier. This feeds the mixer via D3 which is biased on during transmit, via the 12VT line. The output is coupled to the mixer transformer T1 via C11, a DC blocking capacitor. The transmit signal is mixed in IC1 to the output frequency and routed back through the band-pass filter to the pre-driver, TR7.

Transmit pre-driver and driver

When in transmit, D8 is biased on by the 12VT line, coupling the transmit signal to TR7 by T6 (see Fig 7.56). D7 is also biased on to provide an RF ground between D5/6 to improve the input/output receive bypass isolation. C58 provides a DC block to prevent D5/6 being switched on.

The pre-driver is a dual-gate MOSFET; gate 2 is used as a control element by varying its voltage by RV4, a panel-mounted potentiometer. This stage is transformer coupled to the next by T7, a broad-band matching transformer. C64 on the secondary provides an RF ground but provides DC isolation.

The driver, TR8, a VN66 FET, is biased as a Class A amplifier. Feedback is applied to the driver to ensure stability; this is controlled by R3. Bias adjustment is by VR3. The bias voltage is stabilised by ZD1 and supplied from the 12VT line. As the amplifier has no bias during the receive period, TR8 is connected to the 12V permanent line.

Power amplifier

TR9, 10 are run as a push-pull amplifier (Fig 7.56) to maintain stability. R37/RFC8 provide feedback. Push-pull drive for the amplifier is provided from a phase-splitter transformer T8. The amplifier is provided with bias to allow the operating class to be varied. The best balance between output power and efficiency will be in Class B. VR2 is the bias control potentiometer and is adjusted for the required quiescent current.

T9, a trifilar transformer matches the transistors' output impedance to the load presented by the output filter. LPF L5-7 and C74-77 form a seven-element Chebyshev low-pass filter. This has a low SWR at the operating frequency, with the cut-off just above the maximum bandwidth upper frequency. The filter is used bi-directionally to cascade with the bandpass filter when in receive mode.

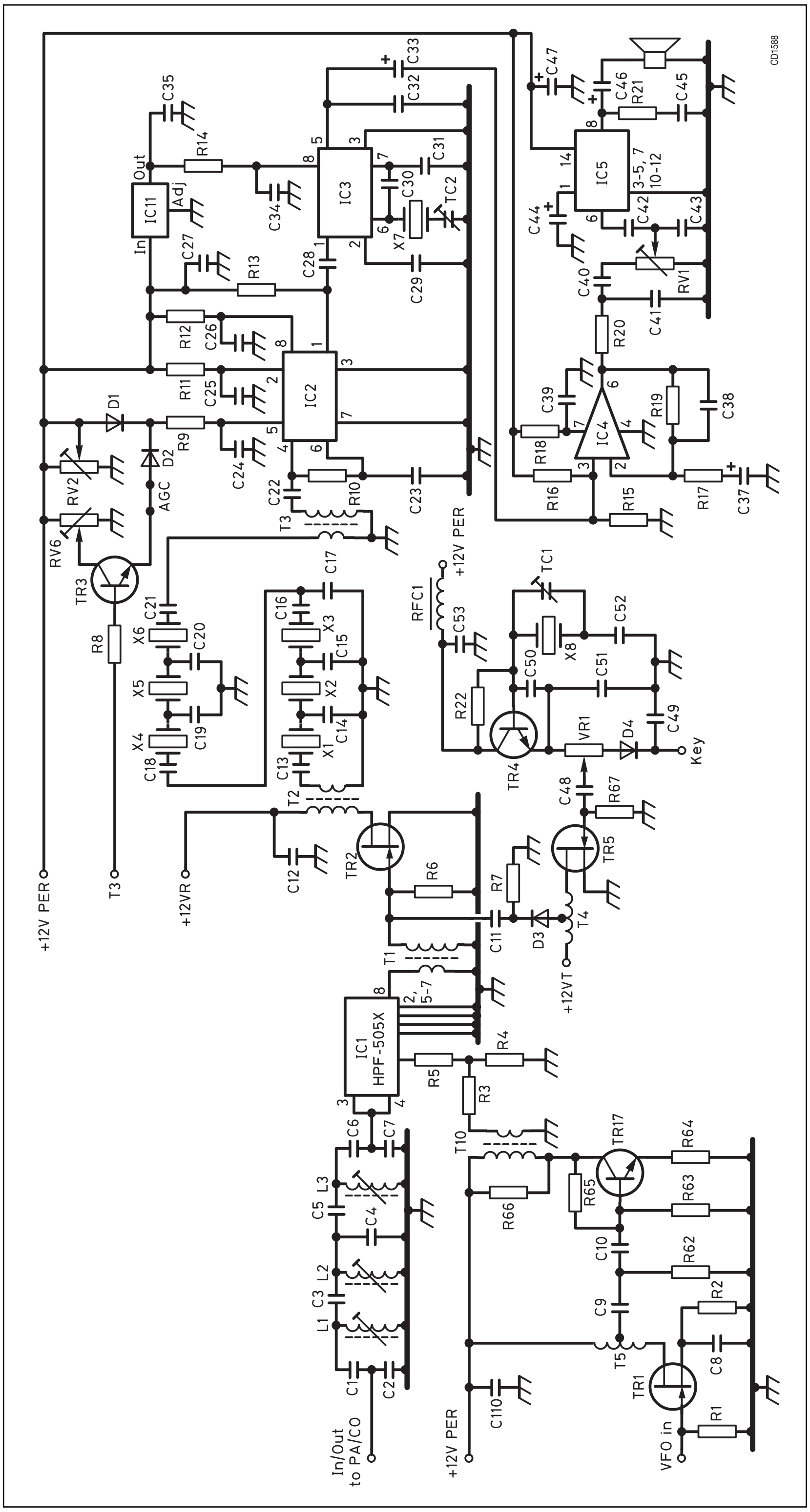


Fig 7.57: GQ-40 crystal filter, CO, buffer and mixer

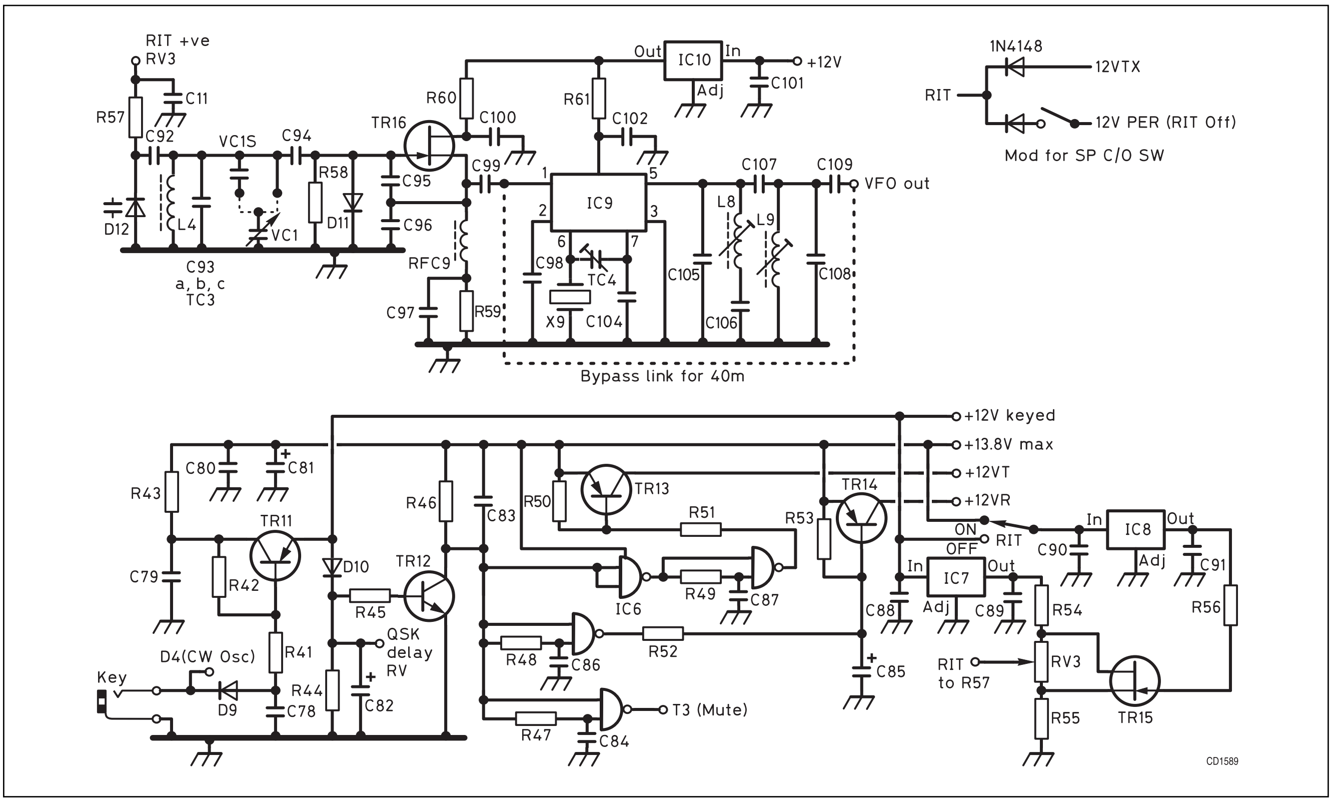


Fig 7.58: VFO, pre-mix and voltage control circuits for the GQ-40 CW transceiver

VFO and pre-mix

The VFO is a standard Colpitts oscillator (**Fig 7.58**). The voltage to the oscillator is stabilised by a three-terminal regulator at 5V. C93 is the main fixed capacitance. This is made up from three values with NP0 dielectric, and a small 10pF trimmer for calibration.

The inductance is wound on type 6 material which is a good compromise between thermal stability and the winding size. Receiver incremental tuning is provided by a variable capacitance diode. To maintain a reasonable voltage swing on the diode, its effective capacitance is reduced by series capacitor C92.

Output from the oscillator is coupled by C99, either to the pre-mixer, IC9, or direct to the main mixer buffer amplifier TR1. When the transceiver model requires a LO frequency greater than 6MHz, IC9 pre-mixer is employed. This is a SA602A used as a crystal oscillator/mixer. The LO pre-mixer output is filtered by a two-pole bandpass filter, C105-109/L8-L9.

LO buffer and drive amplifier

TR1, a J310 FET, buffers either the pre-mix or VFO frequency dependent on the model (Fig 7.57). Its output is transformer coupled via T5 to a Class A driver, TR17. This is a 2N3866, and the amplifier is designed as a 50-ohm line driver feeding the mixer IC1 T-pad.

Voltage and control switching

The 12VT and 12VR power supply lines are switched by PNP power transistors controlled from a time-and-delay circuit (Fig 7.58). The timing switch IC6 is a quad NAND Schmitt trigger package, gate timing being adjusted with small ceramic capacitors C84-87. Timing is initiated by TR11 keyer which directly keys the CW oscillator and the delay generator TR12. TR12 will hold on, dependent on the charge across C82. The charge is adjustable via RV5 which is mounted on the rear panel.

RIT voltage is supplied through a divider network controlled by RV3, a front-panel potentiometer. Voltage to the divider is stabilised at 8V by a three-terminal regulator IC7. During transmit, or when the RIT is switched off, gate voltage is applied to TR15 from IC8. This turns on the FET to short RV3 and centre the voltage equivalent to the mid-position of RV3.

Test and alignment

The minimum requirements for test and alignment are a high-impedance multimeter, power meter, 50Ω dummy load and a general-coverage receiver. The power meter can be a simple peak diode detector across the dummy load used in conjunction with the multimeter. If available, a signal generator or in-band test oscillator are very useful, as is a frequency counter and oscilloscope.

Before starting, carry out a resistance check across the 13.8V input. This should be about 200Ω. If all is OK, connect the multimeter in series with the DC+ lead on its highest amperage range, and switch on. With no signal present and minimum volume, expect the current to be 100-200mA. If the current is appreciably more, switch off and check for faults. Once this test is complete, remove the meter and connect normally to the supply. Turn up the volume control and check for some background noise.

Next test and set up the VFO. Check the VFO output frequency using a counter or a general-coverage receiver with a short antenna near the VFO. Adjust TC3 for the LF band edge. If you are unable to correct the frequency with TC3, then spread or compress the turns on L4. If still outside the range, remove or add turns to L1, or add capacitance using 5mm NP0 discs at C93a, C93b, C93c. Next unmesh VC1 and note the HF frequency. If the coverage is more than required, reconnect VC1 through a fixed capacitor to the VC1 pin. VC1 is now in series with the capacitor and its effective capacitance swing will be reduced dependent on the value. Check the HF and LF limits again and readjust TC3 as necessary. If necessary alter the fixed value and repeat the checks. Once you are happy with the coverage, cement the turns on L4, then cement it to the board using Balsa or plastic adhesive.

Once the VFO is up and running correctly, any temperature drift may be corrected. The capacitance of C93 can be made up with three capacitors using the A and B pads. If required, it should be possible to correct long-term drift by mixing dielectric types.

If a VFO crystal mixer is installed, check for the correct output frequency. If an oscilloscope is available, check the output at R5 or use a general-coverage receiver. Peak the output by adjusting L8/L9.

Attach an antenna to the input pin and adjust the cores of L1-3 for maximum received signal strength - do not use a screwdriver for adjustment as the core is very brittle. If you do not have the correct tool, an old plastic knitting needle with the end filed is ideal. Tune to a test signal and adjust the BFO frequency using TC2. Adjust for the best filter response and opposite sideband suppression.

Test the transmitter next. Connect a 50-ohm dummy load and a power meter or diode detector and multimeter to the antenna output. Fit a temporary heatsink to TR8-10; a clip-on TO220 type is ideal (the tab is part of the drain so be careful that during the checks it is not grounded). Set VR2/3 to the earthy end of their travel. (If you have connected RV4, this must also be at the earth end of its range.) Set VR1 wiper to maximum (the TR5 end of its travel). Connect a multimeter on its 'amps' range in the 13.8V supply line, and key a dash length by grounding the KEY pin. If all is well, the transceiver should change to transmit and hold for a short time. With a multimeter on its DC voltage range, check at R27 that the 12VR line switches correctly. Re-check that 4.7V appears across ZD1 on transmit.

Current consumption will be around 200mA providing there is no drive from TR7. If the consumption is excessive, remove D8 to isolate the driver and PA, and re-check. Possible faults include reversed diodes, ICs or transistors, incorrect values on decoupling or bias resistors, or shorted or incorrectly installed transformers. Once any errors have been corrected, reconnect D8.

Adjust TC1 to obtain a 700Hz beat note from the sidetone. If the audio level is excessive, check that the mute voltage to the IF amp of 8V plus appears at the junction of R9/D1.

Attach a multimeter on its current range in series with the DC supply line. Key the transmitter and set TR8 quiescent current by adjusting VR3 so that the meter current rises by 200mA (ie TR8 standing current is 200mA). Now adjust VR2 so that the current increases by a further 100mA (this sets TR9/10 quiescent to 100mA). The adjustments should be smooth with no power output on the meter. Jumps on the current or the power meter would indicate instability at some level of drive. In this case, remove D8 to isolate the drive and re-check. If there is no improvement, check that T9 is correctly wired. Also check the antenna isolation: confirm that TR6 is switching and that D7 is biased on by the 12VT line. Finally, attach RV4 and adjust it to give 2-3V on the gate of TR7. Key the transmitter and check for power output. This completes the tests.

GQ transceiver PCB

The PCB, a complete kit of parts and a manual containing the latest revisions can be obtained from Hands Electronics, Tegryn, Llanfyrnach, Dyfed SA35 0BL. Tel 01239 698427.

MEDIUM AND HIGH-POWER SOLID-STATE HF LINEAR AMPLIFIERS

The construction of a solid-state linear amplifier now represents a realistic alternative to valve designs even up to the normal maximum UK power level of 400W. The construction of solid-state amplifiers has often been discouraged in the amateur press with claims that such projects can only be made with the use of expensive test equipment, and in particular, a spectrum analyser. The 'simple spectrum analyser' described by Roger Blackwell, G4PMK, in [12] is ideal for the purpose, and fully justifies the small amount of money necessary for its construction.

Number	Power out (W)	Supply voltage (V)
AN762	140	12-14
EB63	140	12-14
EB27A	300	28
AN758	300-1200	50
EB104	600	50

Table 7.23: Motorola RF amplifiers

Motorola Applications Notes AN762, AN758 and EB104, available from Communications Concepts Inc [13] and on the internet [1, 14], provide the designers of sold-state linear amplifiers with outputs in the range 140W to over 1kW.The application notes (AN) and engineering bulletins (EB) describe the construction and operation of suitable amplifiers using Motorola devices, and include printed circuit foil information, making construction relatively easy. It is interesting to note that the PA units fitted to virtually all of the currently available commercial amateur radio equipment are based upon these designs by Motorola, with only a few individual differences to the original design.

With the availability of suitable application notes it is perhaps surprising that few amateurs seem to have embarked upon such a project. High prices for solid-state power devices, combined with a lack of faith in their reliability, may well account for the reluctance to construct solid-state linear amplifiers. Suitable components have been readily available in the USA for some time: Communications Concepts Inc of Xenia, Ohio [13] have offered a complete range of kits and components for the Motorola designs.

The kits include the PCB, solid-state devices, and all the components and the various ferrite transformers already wound, so that all the constructor has to do is solder them together. A large heatsink is necessary and is not supplied.

Home construction of PCBs for an amplifier is possible from the foil patterns available in the application notes, but it should be borne in mind that the thickness of copper on much of the laminate available to amateurs is unknown and inadequate for the high current requirements of a linear amplifier. A ready-made PCB is therefore a very sensible purchase and the kit of parts as supplied by CCI represents a very-cost-effective way of building any of the Motorola designs.

Whilst these HF linear designs by Motorola have now been in existence for more than two decades, they remain a useful standard and a very viable design for home construction. The Motorola Semiconductors business was split up in around the year 2000 and the specified devices may be harder to find today, but the general principles apply and in many cases substitutions are available. A matched pair of MRF454 is available in the US from CCI [13]. A suitable substitution for the MRF454 is the SD1487 and a matched pair is available from Birketts in the UK [15].

Choosing an Amplifier

Perhaps the most useful amplifier for the radio amateur is described in AN762. It operates from a 13V supply and is capable of providing up to 160W output with only 5W of drive. This design is the basis for the majority of commercial 100W PA units and lends itself to both mobile and fixed station operation using readily available power supplies. Table 7.23 illustrates a number of alternative designs.

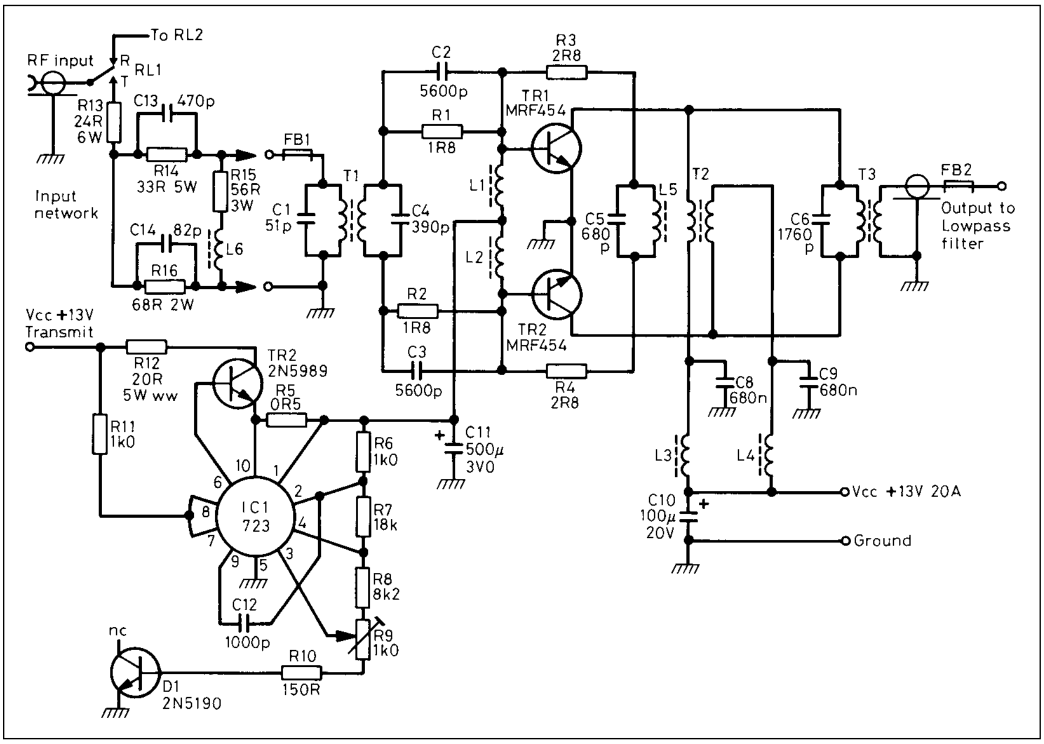


Fig 7.59: The HF SSB 140-300W linear amplifier circuit

AN758 describes the construction of a single 300W output amplifier operating from a 50V supply and further describes a method of using power combiners to sum the outputs of a number of similar units to provide power outputs of 600 and 1200W respectively. Two such units would be ideal for a full-power linear amplifier for UK use, giving an output comparable to the FL2100 type of commercial valve amplifier. It is recommended that any prospective constructor reads the relevant applications note before embarking upon the purchase and construction of such an amplifier.

After constructing a number of low-power transceivers it was decided to commence the construction of the AN762 amplifier. The EB63 design is similar, but uses a slightly simpler bias circuit.

AN762 describes three amplifier variations: 100W, 140W and 180W using MRF453, MRF454 and MRF421 devices respectively, and the middle-of-the road MRF454 140W variant was chosen (**Fig 7.59**).

Constructing the Amplifier

Construction of the amplifier is very straightforward, especially to anyone who has already built up a solid-state amplifier. The PCB drawings (**Fig 7.60**, see Appendix B) and layouts are very good and a copy of the relevant application note was included

Fig 7.61: The HF linear amplifier

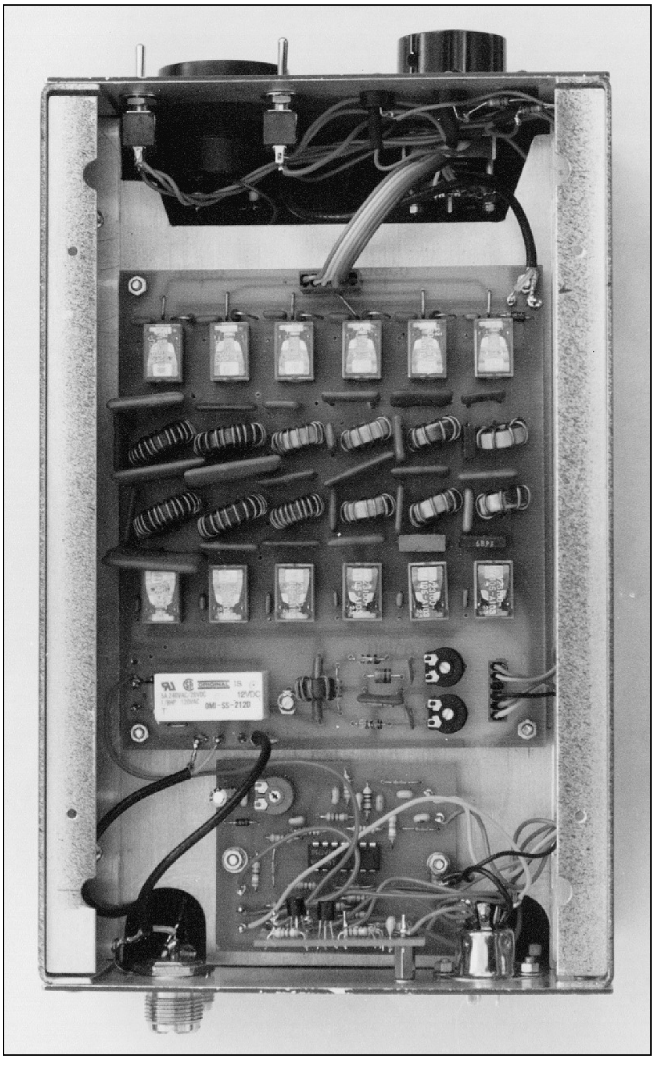


Fig 7.62: Low-pass filter and ALC circuits

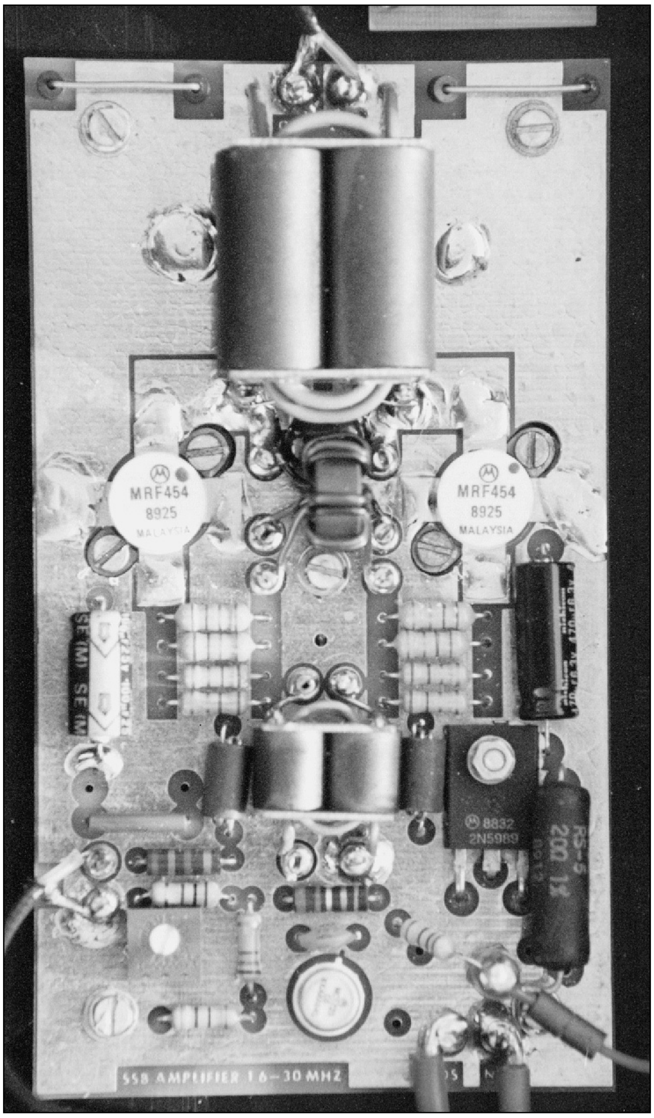


Fig 7.63: Circuit board of the linear amplifier

with the kit. **Figs 7.61, 7.62 and 7.63** show how the amplifier was constructed. A Components List is in **Table 7.24**.

Fixing the transformers to the PCB posed a slight problem: on previous amplifiers that the author has built they were soldered directly to the PCB, but not this one. Approximately 20 turret tags (not supplied) are required: they are riveted to the board and soldered on both sides.

A modification is required if you wish to switch the PA bias supply to control the T/R switching: the bias supply must be brought out to a separate terminal rather than being connected to the main supply line. This is achieved by fitting a stand-off insulator to the PCB at a convenient point; the bias supply components are then soldered directly to this stand-off rather than directly to the PCB.

A number of small ceramic chip capacitors have to be soldered directly to the track on the underside of the PCB; it should be borne in mind when doing this that the clearance between the PCB and heatsink is slightly less than 1/8in and the capacitors must not touch the heatsink.

It is advisable to mount the PCB to the heatsink before attempting to make any solder connections to it, as it is necessary to mark the mounting holes accurately, ideally drilling and tapping them either 6BA or 3mm. The power transistor mounting is very critical in order to avoid stress on the ceramic casing

HF LINEAR AMPLIFIER	
C1	51p chip
C2, 3	5600p chip
C4	390p chip
C5	680p chip
C6 (C7)	1760p (2 x 470p chips plus 820p silver mica in parallel)
C8, 9	0.68μ chip
C10	100μ 20V
C11	500μ 3V
C12	1000p disc
C13	470p silver mica
C14	82p silver mica
R1, 2	2 x 3.6R in parallel
R3, 4	2 x 5.6R in parallel
R5	0.5R
R6	1k
R7	18k
R8	8k2
R9	1k trimpot
R10	150R
R11	1k
R12	20R 5W WW
R13	24R carbon 6W*
R14	33R carbon 5W*
R15	56R carbon 3W*
R16	68R carbon 2W*
* Make up from several higher values in parallel.	
L1, 2	VK200 19/4B choke (6-hole ferrite beads)
L3, 4	Fairite beads x 2 (2673021801) on 16SWG wire
L5	1t through T2
L6	0.82μH (T50-6)
T1	2 x Fairite beads 0.375in x 0.2in 0.4in, 3:1 turns
T2	6t 18SWG ferrite 57-9322 toroid
T3	2 x 57-3238 ferrite cores (7d grade) 4:1 turns
FB1, 2	Fairite 26-43006301 cores
RL1	OUD type
TR1, 2	MRF454
TR3	2N5989
D1	2N5190
IC1	723 regulator
LOW-PASS FILTER	
C1	1200p
C2, 16	180p
C3	2200p
C4	470p
C5	1000p
C6, 13	680p
C7	90p
C9	250p
C10	560p
C11	390p
C12, 24	33p
C14	100p
C15, 18	330p
C17	27p
C19	75p
C20, 28	150p
C21	120p
C22, 27	12p
C23	220p
C25	100p
C26	82p
C29	39p
C30	68p
C31, 32	10n ceramic
C33	10p trimmer
C34	220p silver mica
Note: C1-C30 silver mica 350VDC	
L1	28t 22SWG T68-2
L2	25t 22SWG T68-2
L3	22t 22SWG T50-2
L4	20t 22SWG T50-2
L5	18t 20SWG T50-6
L6	16t 20SWG T50-6
L7	12t 20SWG T50-6
L8	11t 20SWG T50-6
L9	10t 20SWG T50-6
L10	9t 20SWG T50-6
L11	8t 20SWG T50-6
L12	7t 20SWG T50-6
T1	18t bifilar T50-43 pri: 1t
R1	68R
R2, 3	22k trimpot
R4	1k8
L13	1mH RFC
RL2	OM1 type
RL3-14	2A SPCO PCB mtg, 6V coil
D1, 2	OA91 or OA47
SWITCHING AND ALC	
R1, 17, 20	10k
R2	4k7
R3-7	1k
R8, 9	33k
R10, 13, 21, 22	1M
R11	3k3
R12, 14, 18	47k
R15	47k trimpot
R16	390R
R19	220k
R23	8k2
TR1	BC212
TR2	BC640
IC1	LM3900
C1, 14-26	100n
C2-5, 7, 9-11, 13	10n
C6	0.22μ
C8	10 x 16V
C12	1μ 16V
D1-3	LEDs
D4, 5, 7-18	1N914
D6	10V zener
S1	1-pole 6-way
S2, 3	SPCO
Meter	500μA or similar

Table 7.24: HF linear amplifier components list

of the devices. The devices mount directly onto the heatsink and should ideally be fitted by drilling and tapping it. The PCB is raised above the heatsink on stand-offs made from either 6BA or 3mm nuts, so that the tabs on the transistors are flush with the PCB - they must not be bent up or down. When the PCB and transistors have been mounted to the heatsink correctly, they may be removed for the board to be assembled. The transistors should not be soldered in at this stage. The nuts to be used as the stand-offs can be soldered to the PCB, if required, to make refitting the latter to the heatsink a little simpler, but ensure the alignment of the spacers is concentric with the holes.

Assembly should commence with the addition of the turret tags and stand-offs. Then the ceramic chip capacitors should be added under the PCB. D1, which is really a transistor, is also mounted under the PCB; only the emitter and base are connected - the collector lead is cut off and left floating. This transistor is mounted on a mica washer and forms a central stand-off when the PCB is finally screwed down to the heatsink. The mounting screw passes through the device which must be carefully aligned with the hole in the PCB. A number of holes on the PCB are plated through and connect the upper and lower ground planes together. The upper-side components can be mounted starting with the resistors and capacitors, and finally the transformers can be soldered directly to the turret tags. Soldering should be to a high standard as some of the junctions will be carrying up to 10A or more.

When the board is complete it should be checked at least twice for errors and any long leads removed from the underside to ensure clearance from the heatsink. Mount the PCB to the heatsink and tighten it down. Now apply heatsink compound and mount the power transistors which should fit flush with the upper surface of the PCB. Tighten them down by hand, ensuring that there is no stress on the ceramic cases. If any of the connections need to be slightly trimmed to fit, cut them with metal cutters. Ensure the collector tab is in the correct place. Once the transistors fit correctly, they can be removed again and very lightly tinned. The PCB should also be lightly tinned. The transistors can now be refitted and tightened down. Now they can be soldered in but, once in, they are very difficult to remove, so take great care at this stage. The amplifier board is now complete.

Construction of the amplifier takes very little time but requires considerable care to avoid damage to the output devices; the metal work may take a little longer. A large heatsink is required for 140W and an even larger one for 300W. It is recommended that the higher-power amplifiers are mounted onto a sheet of 6mm (0.25in) copper which is in turn bolted to the main heatsink. (Note: on no account should the transistors be mounted onto a diecast box due to surface imperfections and poor thermal conductivity.) Blowing may be necessary at the higher powers or if a less-than-adequate heatsink is used.

Setting Up and Testing the Amplifier

There is only one adjustment on the amplifier, making setting up relatively simple. Before connecting any power supplies, check and re-check the board for any possible errors. The first job is to test the bias supply. This must always be done before connecting the collector supply to the amplifier as a fault here could destroy the devices instantly. With +13V connected to the bias supply only, it should be possible to vary the base bias from approximately 0.5V to 0.9V. Set it to the lowest setting, ie 0.5V. Disconnect the bias supply from the 13V line. When conducting any tests on the amplifier always ensure that it is correctly terminated in a 50-ohm resistive dummy load. Apply +13V to the amplifier and observe the collector current on a suitable meter. It should not exceed a few milliamps - if it does something is wrong, so stop and check everything. Assuming that your amplifier only draws 3 to 4mA, connect the bias supply to the +13V supply and observe an increase in current, partly caused by the bias supply itself and also by the increased standing current in the output devices. The current can be checked individually in each of the output devices by unsoldering the wire links L3 and L4 on the PCB. Set the bias to 100mA per device by adjusting R9. The current should be approximately the same in each device: if it is not it could indicate a fault in either device or the bias circuitry to it. Increase the standing current to ensure that it rises smoothly before returning it to 100mA per device. Once the total standing current is set to 100 + 100 = 200mA, the amplifier is ready for operation.

With a power meter in series with the dummy load, apply a drive signal to the input, steadily increasing the level. The output should increase smoothly to a maximum of about 160W. It will go to 200W but will exceed the device specification. Observing the output on a spectrum analyser should reveal the primary signal, together with its second, third and higher harmonics. Check that there are no other outputs. Removal of the input signal should cause the disappearance of the other signals displayed. It is helpful during initial setting up to monitor the current drawn by the amplifier. At full output, efficiency should be in the order of 50%, perhaps lowering slightly at the

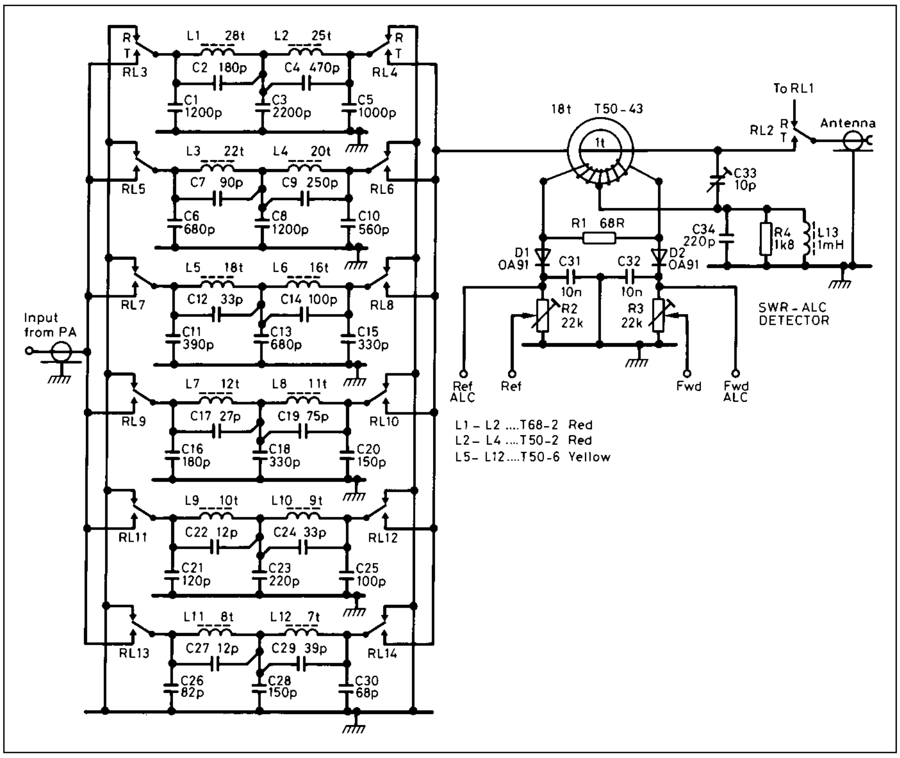


Fig 7.64: The amplifier is followed by six switched filters and a detector for SWR and ALC circuitry

upper and lower frequency limits and increasing a little somewhere in the 20MHz range. The maximum current likely to be drawn by the 140W amplifier is in the order of 24A.

Putting the Amplifier to Use

Building and setting up the amplifier is undoubtedly a simple operation, and may lead one into a false sense of security. Before the amplifier can be used it must have a low-pass filter added to the output to remove the harmonics generated. For single-band operation only one filter would be required, but for operation on the HF amateur bands a range of filters is required with typically six switched filters covering the range 2 to 30MHz. For most applications a five-pole Chebyshev filter will provide all the rejection required, but the majority of commercial designs now use the elliptic type of filter providing peaks of rejection centred around the second and third harmonics. Such filters can be tuned to maximise the rejection at specific frequencies. An elliptic function filter was decided upon as it only requires two extra components over and above the standard Chebyshev design and setting up is not critical.

The construction of a suitable low-pass filter (**Fig 7.64**) may take the form of the inductors and capacitors mounted around a suitable wafer switch, or they may be mounted on a PCB and switched in and out of circuit using small low-profile relays. This makes lead lengths shorter and minimises stray paths across the filter. Unused filters may be grounded easily using relays permitting only one filter path to be open at a time. The relays need only to be able to carry the output current; they are not required to switch it. 2A contacts are suitable in the 100-140W range.

Micrometals Inc cores ensure the duplication of suitable inductors while silver mica capacitors should be used to tune the filters. The voltage working of the capacitors should be scaled to suit the power level being used; ideally, 350V working should be used in the 100-150W range and, for powers in the region of 400-600W, 750V working capacitors should be used, the latter being available from CCI.

The antenna change-over relay may be situated at either end of the low-pass filter. If it is intended to use the filter on receive then the relay will be placed between the amplifier and the filter, but if the amplifier is an add-on unit then it may not be necessary to use the filter on receive and the relay may be located at the output end of the filter. The filter performance is enhanced if it is mounted in a screened box with all DC leads suitably decoupled.

SWR Protection and ALC

One of the major shortcomings of early solid-state amplifiers was PA failure resulting from such abuse as overdriving, short-circuited output, open-circuited output and other situations causing a high SWR. A high SWR destroys transistors either by exceeding the collector-base breakdown voltage for the device or through overcurrent and dissipation.

ALC (automatic level control) serves two functions in a modern-day transmitter: it controls the output power to prevent overdriving and distortion and can be combined with a SWR detector to reduce the power if a high SWR is detected, which reduces the voltages that can appear across the output device and so protects it.

A conventional SWR detector provides indication of power (forward) and SWR (reflected power) which can be amplified and compared with a reference. If the forward power exceeds the preset reference an ALC voltage is fed back to the exciter to reduce the drive and hence hold the power at the preset level. A high SWR will produce a signal that is amplified more than the forward signal and will reach the reference level more quickly, again causing a reduction in the drive level.

The circuit (**Fig 7.65**) has been designed to work in conjunction with the ALC system installed in the G3TSO modular transceiver [16] and produces a positive-going output voltage. The LM3900 IC used to generate the ALC voltage contains two unused current-sensing op-amps which have been used as a meter buffer with a sample-and-hold circuit providing a power meter with almost a peak reading capability. In practice it reads about 85% of the peak power compared to the 25% measured on a typical SWR meter.

T/R Control

There are many ways of controlling the T/R function of an amplifier. It was decided to make this one operate from the PTT line but unfortunately direct connection resulted in a hang-up when relays in the main transceiver remained activated after the PTT line was released. The buffer circuit comprising TR4 and TR5 simply switch the input and output relays from receive to transmit and provide a PA bias supply on transmit.

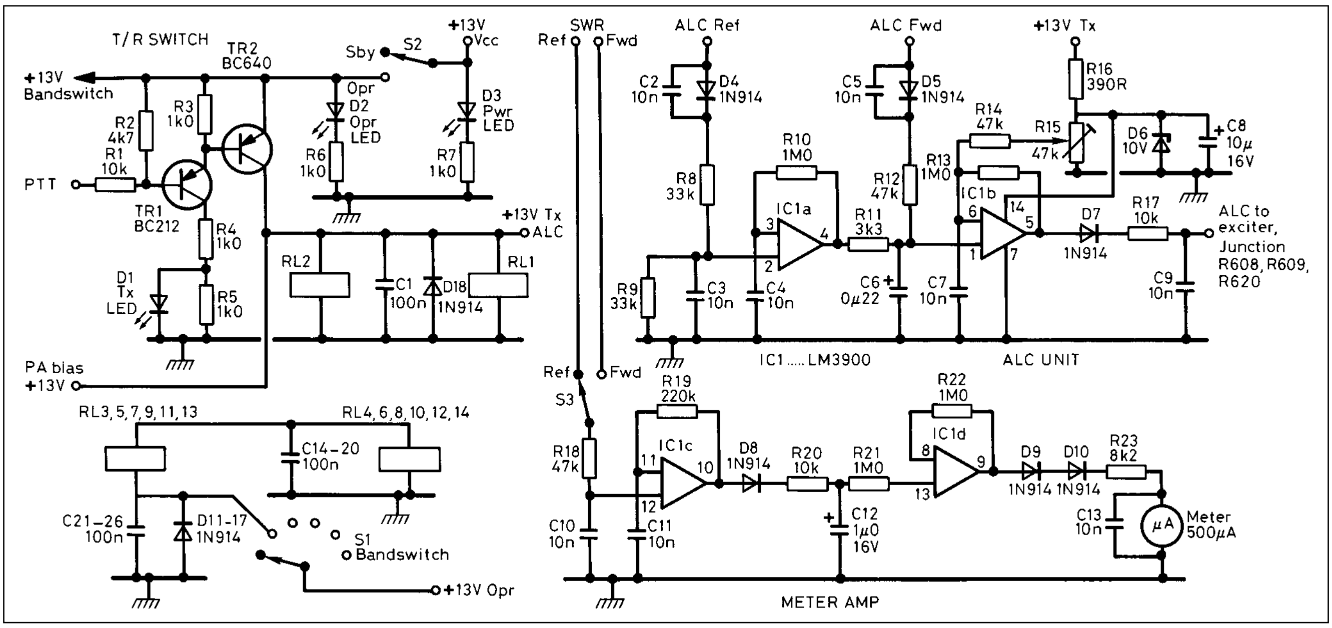


Fig 7.65: Transmit/receive switching, ALC and SWR circuits

Interfacing Amplifier to Exciter

The G3TSO modular transceiver was used as the drive source for the AN762 solid-state amplifier, 5W of drive producing 140W output from the linear. A little more drive and 200W came out. This was rapidly reduced by setting the ALC threshold. Initially the spectrum analyser showed the primary signal, harmonics reduced by the action of the elliptic low-pass filter, but alas a response at 26MHz and not many decibels down on the fundamental. A check with the general-coverage transceiver revealed that there really was something there, while a finger on the 80m low-pass filter showed quite a lot of heat being generated.

Investigations revealed quite clearly that this type of broad-band amplifier cannot be operated with a capacitive-input, low-pass filter at either end without it going into oscillation at some frequency, usually well above the cut-off frequency of the filter. The filter input impedance decreases with frequency, and with two such filters located at either end of the amplifier there comes a point where the input circuit and output circuit resonate at the same frequency and a spurious oscillation occurs. Removal of either filter solves the problem.

A direct connection between the exciter and linear amplifier is the preferred solution, but is not always practicable in the case of add-on amplifiers where there is already a low-pass filter installed in the exciter. Another solution is to provide a resistive termination at the input of the amplifier; this is far simpler to effect and is used in several commercial designs. The network used comprises a 50Ω carbon resistor across the input of the amplifier which reduces the input impedance to 25Ω, so a 30Ω resistor is placed in series with the drive source to present a near-50Ω impedance to the exciter. Power from the exciter will be absorbed in these resistors which must be built from a number of lower-wattage resistors in parallel, ie five 150Ω resistors make a 30Ω resistor with five times the power rating. In addition it was found necessary to add a ferrite bead to the input and output leads to the amplifier to effect a complete cure to the parasitic problem which was at its worst on 21MHz, the parasitics occurring above 40MHz. An alternative solution would be to use an inductive-input low-pass filter (Fig 7.27(c) & Table 7.9).

Conclusion

The construction of a solid-state high-power amplifier is very simple, especially as the parts are obtainable in kit form via international mail order from the USA.

The use of a spectrum analyser, no matter how simple, greatly eases the setting up of such an amplifier and ensures peace of mind when operating it.

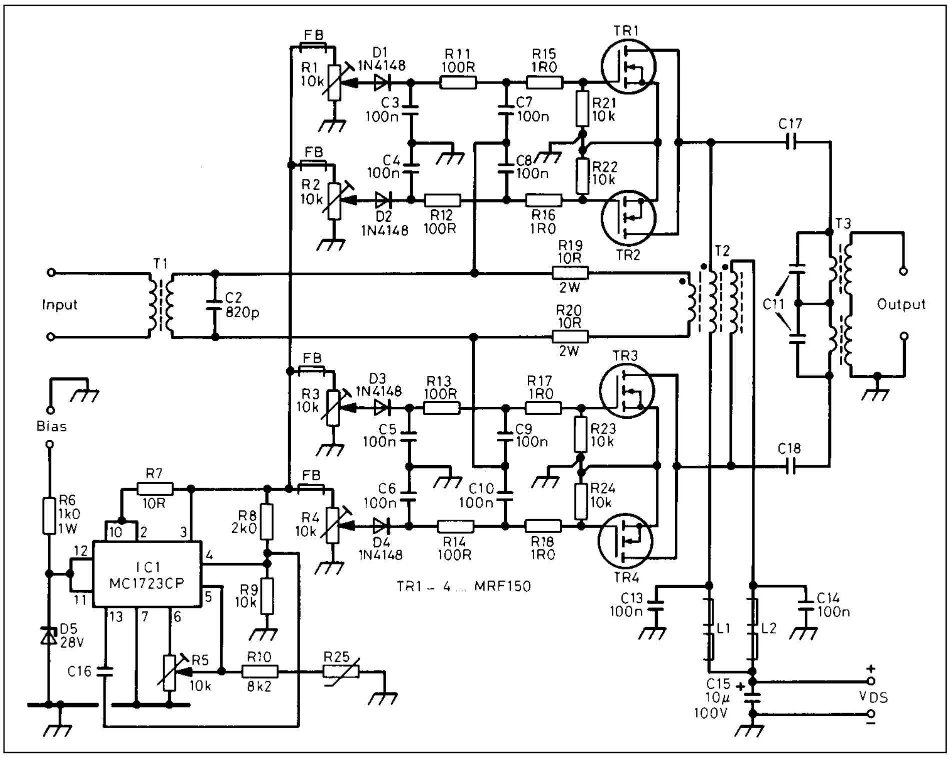


Fig 7.66: 600W output RF amplifier

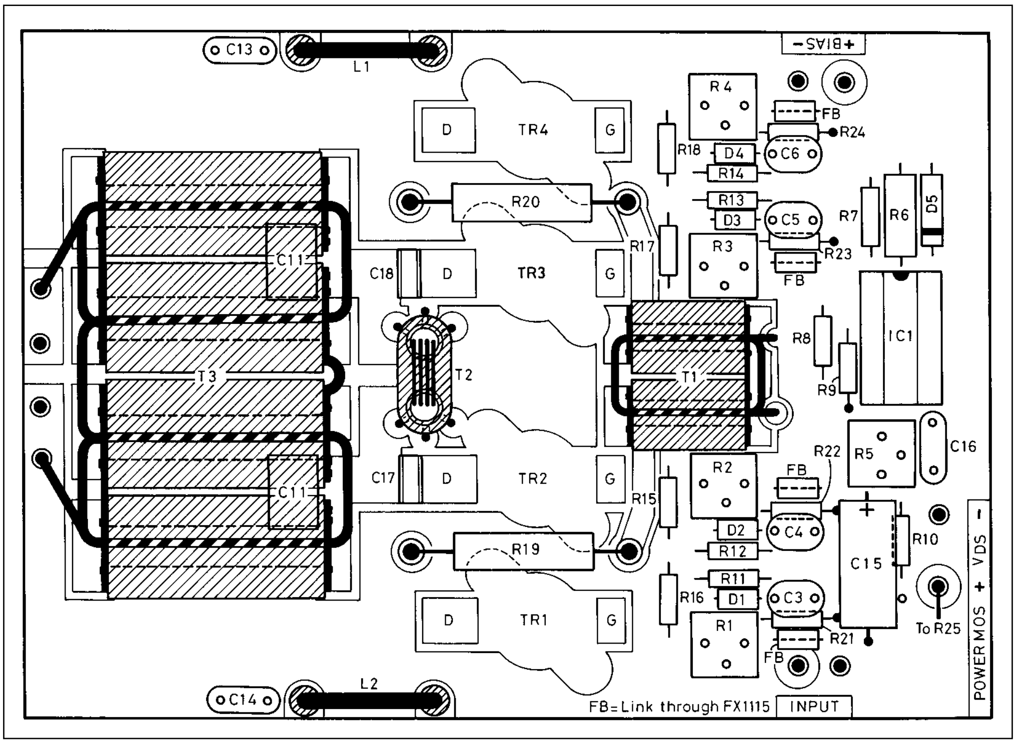


Fig 7.68: 600W amplifier component layout

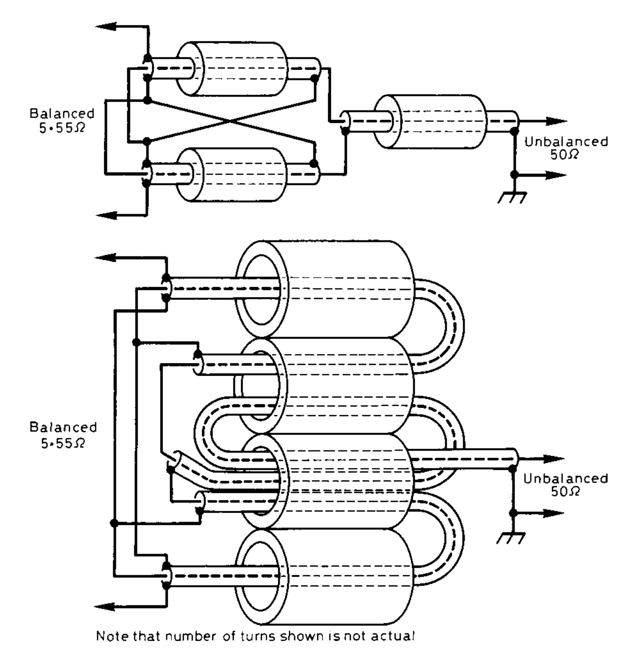


Fig 7.69: Winding the transformers

Amplifiers are available for a number of different power levels, and can be combined to provide higher power levels. The use of 13V supplies practically limits powers to about 180W and below, while 28 or 50V facilitates higher-power operation without the need for stringent PSU regulation.

The full-power solid-state linear is now possible at a price showing a considerable saving on the cost of a commercially made unit.

There is a great similarity between all the Motorola designs: the 300W amplifier described in AN758 is virtually the same as the AN762 amplifier, the fundamental difference being the 28V DC supply voltage.

The 600W MOSFET amplifier (**Figs 7.66, 7.67** in Appendix B, **7.68 and 7.69**, plus **Table 7.25**) described in EB104 represents a real alternative to the valve linear amplifier and operates from a 50V supply, but considerably more attention must be paid to the dissipation of heat.

R1-R5	10k trimpot
R6	1k, 1W
R7	10R
R8	2k
R9, 21-24	10k
R10	8k2
R11-14	100R
R15-18	1R
R19, 20	10R, 2W carbon
R25	Thermistor, 10k (25°C), 2k5 (75°C)
All resistors ½W carbon or metal film unless otherwise noted.	
C1	Not used
C2	820p ceramic chip
C3-6, 13, 14	100n ceramic
C7-10	100n ceramic chip
C11	1200p each, 680p mica in parallel with an Arco 469 variable or three or more small mica capacitors in parallel
C12	Not used
C15	10µ, 100V elec
C16	1000p ceramic
C17, 18	Two 100n, 100V ceramic each (ATC 200/823 or equivalent)
D1-4	1N4148
D5	28V zener, 1N5362 or equivalent
L1, 2	Two Fair-Rite 2673021801 ferrite beads each or equivalent, 4 H
T1	9:1 ratio (3t:1t)
T2	2µH on balun core (1t line)
T3	See Fig 7.64
TR1-4	MRF150
IC1	MC1723CP

Table 7.25: Components list for 600W output RF amplifier

CHOOSING A COMMERCIAL TRANSCEIVER

Apart from low power radios, the majority of transceivers used in the shacks of UK amateurs are commercial, usually manufactured in the far east. Many of these are reviewed in the RSGB members' magazine RadCom*, by Peter Hart, G3SJX. The following is his guide to finding your way round the bewildering range of choices available. Further information and detailed reviews are included in [17].*

Buying an HF transceiver can be one of the bigger purchases which the amateur is likely to make. The large number of radios on the market and the various features and functions which they provide can be rather daunting, especially to the newcomer and the following guidelines may help in the decision making process.

Evolution of the Modern Transceiver

The SSB/CW transceiver, as distinct from a separate receiver and transmitter first started to make an appearance during the 1960s. At the beginning of the 1980s, the first models to include a microcontroller were introduced. This had a profound effect on the architecture of the radio, possibly unsurpassed since Edwin Armstrong invented the superhet a century ago. The microcontroller made possible the single tuning knob digital frequency synthesiser and with this the ability to make stable oscillators at VHF. This in turn made viable the up-conversion broadband superhet architecture which forms the basis of virtually all successive designs up until the present day.

This was a very timely introduction as in 1979 we had gained three more HF bands, the so-called WARC bands, adding three more band positions and complexity to the traditional bandswitched design.

These new generation radios featured memories, twin VFOs and variable IF bandwidth. Valved power amplifiers were replaced by broadband semiconductor designs as power transistor technology was becoming more mature and cheaper.

As we progressed into the 1990s, the Direct Digital Synthesiser made its appearance, significantly reducing the complexity of the multi-loop PLL and improving performance. Digital Signal Processing started to be used in the audio stages giving degrees of filtering and noise reduction facilities previously unattainable by analogue technologies. With a greater realisation of the importance of dynamic range we started to see radios with substantially improved performance.

As the decade progressed the level of features provided snowballed. DSP started to be used in the IF stages giving a far greater range of bandwidth options and removing the need to fit expensive extra IF filters. LCD panels of increasing size and complexity began to dominate over LED and fluorescent display technology. HF radios also started to appear with extensions into the VHF and UHF bands and some even catered for full duplex satellite operation and coverage up as far as the 23cm band.

Entering the 21st century, radios were just as well-featured for the data modes as for the traditional modes with some top end models including RTTY and packet decoders. A fresh look

Table 7.26: Comparison of the categories of commercially manufactured amateur radio transceivers

Type of radio	Characterised by	Principle use	Typical models
Base stations mains and 12V	Largest size, most features, highest performance, versatile, highest cost, may need external mains PSU	Home use, DXing, contesting, general use, all modes	FTDX5000 IC-7800 TS-950D Ten-Tec Orion
Mid size 12V	Mid size, needs external mains PSU, many features, mid/high performance, mid cost	General all round use principally at home.	FT-950 IC-7600 TS-2000 Elecraft K3
Small size 12V	Small size, multifunction knobs, mid performance, mid/low cost, external PSU for mains use	General all round use at home, portable, transportable, lightweight DXpeditions	FT-897 IC-7200 TS-480
Mobile	Dash mounting or detachable panel, 12V operated, multifunction knobs, mid cost, mid performance	Mobile, transportable	FT-857 IC-7000 TS-50S
Battery operated portables	Small size, lightweight, low power, fewer features, internal or external batteries	Hand portable, take anywhere	FT-817 IC-703
Software Defined Radios SDR	Unlimited features, downloadable software, needs PC, keyboard and mouse control, spectrum display with point and click tuning	Home use, an experimenters radio	SDR-1000 FLEX-5000A

at performance and features has resulted in a new range of really high-end designs and currently the state of the art in dynamic range.

At the other end of the scale, a range of lightweight models appeared aimed at the mobile user or traveller and currently the only true HF (plus VHF and UHF) hand portable. Computer interfacing had for some time featured strongly and now black boxes started to appear using the PC as the sole user interface.

The close affinity with PCs continues and we are now seeing radios sporting VGA connectors for external displays, LAN connectors and USB ports for control and interface lines, and upgradeable firmware downloadable from the Internet.

The current generation of radios use DSP for all signal processing functions with comprehensive channel filtering, tailoring of the passband shape and noise reduction modes. To achieve the very highest close-in dynamic ranges for the toughest of crowded conditions there is a move away from up-conversion architectures back to lower frequency IFs with narrow roofing filters as low as 300Hz in some radios.

Software Defined Radio (SDR) is the latest major milestone in the evolution of the radio transceiver with all processing functions performed in software and with minimal hardware. First implementations used down-conversion and the PC soundcard as the radio 'engine'.

ALL RADIOS	HIGHER END RADIOS	TOP CLASS RADIOS
Multimode	Selectable IF filters	Excellent RF performance
Twin VFOs	Variable bandwidth	Excellent audio performance
Memories	Notch filters	Top class channel filtering and filtering armoury
Clarifier	AF Filters	Ability to operate in multi station environment
Switchable front end	Dual watch capability	Uncompromised dual receiver capability
Variable Power	CW keyer	Interfacing to QSK linears
Interfacing	Auto ATU	Extensive interfacing to accessories
Switchable AGC	50MHz and higher coverage	Ergonomic use in contest and expedition environments
Many software features	More comprehensive displays	Upgradeable software
Noise blankers	More user customisation	Use with receive only antennas
Computer interface	DSP filters and noise reduction	
	Improved data capability	
	More features	
	Spectrum display	

Table 7.27: Features provided on HF transceivers

More recently direct digital sampling using A to D conversion at signal frequency is used removing the need for a high quality sound card. Feature implementation is limited only by imagination, and with open source code and enthusiastic developers this continues to be a fast moving area.

See the earlier chapter on Software Defined Radios for more information.

Turning our attention now to choosing a radio, **Table 7.26** summarises the different types of radio available for HF use. There are several main factors to consider when choosing a transceiver, and these are described below.

Intended Use

Will the radio be used at home, portable, in the car or on a DXpedition, or a multiple of these uses? This determines the overall size and weight. Is the main use for general operating or for competitive activities such as contesting and DXing? A higher performance radio is desirable for contesting and DXing, whereas a lower performance and cheaper radio is entirely satisfactory for more casual operating.

Is the main use on a specific mode such as SSB, CW or Data modes? Some radios have more features suited to different

RECEIVE PERFORMANCE	
ON-CHANNEL	Sensitivity
	Distortion / AF quality
	Bandwidth
	AGC
OFF-CHANNEL	Stopband selectivity
	Spurious responses
	Non-linearity / intermod
	Phase noise

Table 7.28: Key receive performance parameters

TRANSMIT PERFORMANCE	
ON-CHANNEL	Power output
	Distortion / AF quality
	Bandwidth
	Keying characteristics
OFF-CHANNEL	Spurious outputs
	Sideband splatter/clicks
	Non-linearity
	Phase noise
	Wideband noise

Table 7.29: Key transmit performance parameters

modes, such as full break-in and built-in keyers for CW and extensive PC/audio interfacing and dedicated data mode selection and filters for RTTY, PSK and other specialist modes.

Features and Functions

Most modern radios are very well equipped with all the features you are likely to need for most general purpose operation. See **Table 7.27**. Filtering functions and noise reduction capabilities are better implemented on the higher and top end models. DXing frequently uses split frequency operation and the ability to tune and receive on both channels is important. Dual receiver models are ideal for this purpose.

The larger base station radios provide more extensive interfacing capabilities to multiple and receive-only antennas, linears, PC and audio lines etc. Radios at home are more likely to be used in combination with a linear, PC control and sound card for data, CW and voice keyers, transverters etc. If using a small radio at home make sure that it has the interfaces that are needed.

The new breed of software defined radios (SDR) offers limitless possibilities but using the PC as the user interface with keyboard control and mouse tuning is not to everyone's liking. SDR inherently provides high resolution spectrum displays and panadaptors with 'point and click tuning' and in conjunction with further processing software, such as *CW Skimmer*, intelligent analysis and multiple decoding of CW and data signals across a wide band is possible. Expect rapid progress in this area.

Ease of Use

This is most important if you are to obtain maximum enjoyment from the use of your radio. Large well-spaced control knobs and buttons, dedicated controls rather than menu or context switched controls, and clear displays all make for easy operation.

Compromises are, however, inevitable on the smaller sized radios. Ease of use is particularly important in minimising fatigue in extended operating periods such as in contests. Try to check out the radio on air before buying.

TRANSCEIVER		Phase Noise @10kHz	ILDR @5kHz	ILDR @50kHz
Yaesu	FTDX5000	-142dBC/Hz	105dB	105dB
Kenwood	TS-950S	-145dBC/Hz	103dB	104dB
Elecraft	K3	-134dBC/Hz	101dB	101dB
Ten-Tec	Eagle	-131dBC/Hz	89dB	98dB
FlexRadio	FLEX-5000A	-124dBC/Hz	97dB	97dB
FlexRadio	SDR-1000	-135dBC/Hz	94dB	94dB
Ten-Tec	ORION 1&2	-129dBC/Hz	93dB	93dB
Yaesu	FTDX9000D	-137dBC/Hz	93dB	98dB
Elecraft	K2/100	-128dBC/Hz	91dB	95dB
Ten-Tec	CORSAIR	-132dBC/Hz	90dB	90dB
Yaesu	FT-2000D	-125dBC/Hz	89dB	98dB
Ten-Tec	OMNI-VI	-130dBC/Hz	88dB	88dB
Icom	IC-7800	-134dBC/Hz	88dB	111dB
Icom	IC-7600	-124dBC/Hz	87dB	105dB
Ten-Tec	OMNI-VII	-127dBC/Hz	84dB	89dB
Kenwood	TS-950	-136dBC/Hz	83dB	102dB
Yaesu	FT-1000MP	-128dBC/Hz	82dB	97dB
Yaesu	FT-950	-128dBC/Hz	82dB	97dB
JRC	JST-245	-115dBC/Hz	80dB	92dB
Icom	IC-737	-138dBC/Hz	80dB	102dB
Yaesu	FT-747	-119dBC/Hz	80dB	97dB
Icom	IC-725	-117dBC/Hz	79dB	95dB
Yaesu	FT-990	-127dBC/Hz	78dB	97dB
Kenwood	TS-930	-123dBC/Hz	77dB	95dB
Icom	IC-746	-124dBC/Hz	77dB	98dB
Icom	IC-707	-127dBC/Hz	77dB	91dB
Kenwood	TS-940	-132dBC/Hz	76dB	94dB
Yaesu	FT-450D	-123dBC/Hz	76dB	94dB
Drake	R8E	-123dBC/Hz	76dB	92dB
Icom	IC-736/8	-128dBC/Hz	76dB	100dB
Kenwood	TS-850	-134dBC/Hz	75dB	98dB
Kenwood	TS-480	-123dBC/Hz	75dB	98dB
Ten-Tec	JUPITER	-112dBC/Hz	75dB	89dB
Icom	IC-751A	-139dBC/Hz	75dB	104dB
Alinco	DX-70TH	-112dBC/Hz	75dB	92dB
Icom	IC-756PROIII	-127dBC/Hz	74dB	105dB
Icom	IC-756	-125dBC/Hz	74dB	89dB
Icom	IC-7000	-125dBC/Hz	74dB	88dB
Icom	IC-756PROII	-126dBC/Hz	73dB	98dB
Icom	IC-756PRO	-126dBC/Hz	73dB	92dB
Icom	IC-7400	-127dBC/Hz	73dB	97dB
Icom	IC-703	-118dBC/Hz	73dB	91dB
Yaesu	FT-1000MP mk5	-129dBC/Hz	73dB	95dB
Icom	IC-775DSP	-134dBC/Hz	72dB	93dB
Yaesu	FT-920	-131dBC/Hz	72dB	96dB
Kenwood	TS-50	-129dBC/Hz	69dB	96dB
Yaesu	FT-900	-125dBC/Hz	69dB	97dB
Yaesu	FT-890	-125dBC/Hz	69dB	95dB
Kenwood	TS-2000	-124dBC/Hz	68dB	95dB
Yaesu	FT-817	-119dBC/Hz	68dB	90dB
Yaesu	FT-100	-121dBC/Hz	68dB	90dB
Yaesu	FT-847	-124dBC/Hz	67dB	94dB
Yaesu	FT-897	-116dBC/Hz	65dB	92dB
Yaesu	FT-857	-118dBC/Hz	65dB	93dB
Yaesu	FT-1000	-125dBC/Hz	65dB	96dB
Kenwood	TS-430	-116dBC/Hz	62dB	96dB
Icom	IC-729	-134dBC/Hz	62dB	101dB
Kenwood	TS-870	-131dBC/Hz	61dB	94dB
Yaesu	FT-767	-107dBC/Hz	59dB	95dB
Icom	IC-781	-137dBC/Hz	57dB	103dB
Yaesu	FT-757	-119dBC/Hz	55dB	88dB

Table 7.30: Dynamic range receive performance of HF transceivers

Maker	Model	Type	Bands	Introduced	Supply	HF Power	*RadCom* Review	Other key features
Alinco	DX-70	mobile	HF 50	1995	13V	100W	August 1995	10W on 6m
Alinco	DX-70TH	mobile	HF 50	1999	13V	100W	August 1999	100W on 6m
Elecraft	K2	base small	HF	1999	13V	15/100W		Kit construction. Optional PA
Elecraft	K3	base small	HF 50	2008	13V	100W	July 2008	Twin receiver option
FlexRadio	FLEX-1500	SDR	HF 50	2010	13V	5W	April 2011	USB interface to external PC
FlexRadio	FLEX-3000	SDR	HF 50	2009	13V	100W	August 2009	External PC needs fireware interface
FlexRadio	FLEX-5000A	SDR	HF 50	2007	13V	100W	January 2008	External PC needs fireware interface
FlexRadio	SDR-1000	SDR	HF 50	2005	13V	1/100W	June 2006	External PC needs soundcard
Icom	IC-7000	mobile	HF 50 144 430	2005	13V	100W	April 2006	Spectrum scan. 2m 50W, 70cm 35W
Icom	IC-703	mobile	HF 50	2003	9-15V	5/10W	October 2003	Transportable, detachable front panel
Icom	IC-706	mobile	HF 50 144 430	1995	13V	100W	June 1997	70cm on G version 20W, 2m 50W
Icom	IC-707	base small	HF	1993	13V	100W	April 1994	Front speaker
Icom	IC-7200	base small	HF 50	2008	13V	100W	January 2009	Water resistant. USB interface
Icom	IC-725	base small	HF	1988	13V	100W	September 1989	FM/AM transmit module option
Icom	IC-726	base small	HF 50	1989	13V	100W	February 1990	As IC-725 plus 10w on 6m
Icom	IC-728	base small	HF	1992	13V	100W	April 1993	FM/AM transmit module option
Icom	IC-729	base small	HF 50	1992	13V	100W	April 1993	As IC-728 plus 10W on 6m
Icom	IC-736	base mid	HF 50	1994	mains	100W	May 1995	100W on 6m
Icom	IC-737	base mid	HF	1993	13V	100W	September 1993	Auto ATU, keyer"
Icom	IC-738	base mid	HF	1994	13V	100W	May 1995	As IC-736 without mains PSU or 6m
Icom	IC-7400	base mid	HF 50 144	2002	13V	100W	October 2002	100W 6m,2m. spectrum scan, IF DSP
Icom	IC-7410	base mid	HF 50	2011	13V	100W		Feature packed. USB interface
Icom	IC-746	base mid	HF 50 144	1997	13V	100W	March 1998	100W 6m,2m. spectrum scan, AF DSP
Icom	IC-756	base large	HF 50	1997	13V	100W	May 1997	AF DSP, dual watch, Spectrum display
Icom	IC-756PRO	base large	HF 50	1999	13V	100W	March 2000	IF DSP, dual watch, feature packed
Icom	IC-756PROII	base large	HF 50	2002	13V	100W	June 2002	IF DSP, dual watch, feature packed
Icom	IC-756PROIII	base large	HF 50	2004	13V	100W	February 2005	IF DSP, dual watch, feature packed
Icom	IC-7600	base large	HF 50	2009	13V	100W	June 2009	Successor to IC-756PRO series
Icom	IC-775DSP	base large	HF	1995	mains	200W	January 1996	AF DSP, dual watch, auto-ATU, keyer
Icom	IC-7700	base large	HF 50	2008	mains	200W	June 2008	IF DSP, single RX, feature packed
Icom	IC-7800	base large	HF 50	2004	mains	200W	August 2004	IF DSP, dual RX, feature packed
Icom	IC-781	base large	HF	1987	mains	150W	July 1990	Dual watch, CRT display
Icom	IC-9100	base mid	HF 50 144 432 1.3G	2011	13V	100W	April 2011	Duplex satellite, DSTAR, 23cm options
JRC	JST-245	base large	HF 50	1994	mains	150W	October 1997	150W on 6m
Kenwood	TS-2000	base mid	HF 50 144 432 1.3G	2000	13V	100W	April 2001	Data TNC, duplex satellite, 23cm option
Kenwood	TS-450S	base mid	HF	1992	13V	100W	November 1992	
Kenwood	TS-480	mobile	HF 50	2003	13V	100/200W	March 2004	Remote panel. 100W / 200W versions
Kenwood	TS-50S	mobile	HF	1993	13V	100W	May 1993	
Kenwood	TS-570D	base mid	HF	1996	13V	100W	December 1996	Auto ATU, keyer, AF DSP
Kenwood	TS-590S	base mid	HF 50	2010	13V	100W	January 2011	Twin architecture
Kenwood	TS-680S	base mid	HF 50	1988	13V	100W	March 1989	10W on 6m
Kenwood	TS-690S	base mid	HF 50	1992	13V	100W	November 1992	50W on 6m
Kenwood	TS-850S	base large	HF	1991	13V	100W	October 1991	Options for ATU, message stores, DSP
Kenwood	TS-870S	base large	HF	1995	13V	100W	April 1996	IF DSP, ATU
Kenwood	TS-950S	base large	HF	1989	mains/13V	150W	April 1990	SDX version has AF DSP
Ten-Tec	EAGLE	base small	HF 50	2010	13V	100W	July 2011	RF performance, simple controls
Ten-Tec	JUPITER	base mid	HF	2000	13V	100W	January 2004	IF DSP. Spectrum scope.
Ten-Tec	OMNI-VI	base large	HF	1992	13V	100W	January 1994	Ham bands only
Ten-Tec	OMNI-VII	base mid	HF 50	2007	13V	100W	September 2007	Remote LAN operation. Spectrum scan
Ten-Tec	ORION 1	base large	HF	2003	13V	100W	June 2004	Dual RX, IF DSP, narrow roof, top end
Ten-Tec	ORION 2	base large	HF	2006	13V	100W	August 2006	Dual RX, IF DSP, narrow roof, top end
Yaesu	FT-100	mobile	HF 50 144 430	1999	13V	100W	June 1999	Spectrum scan. 2m 50W, 70cm 20W
Yaesu	FT-1000	base large	HF	1990	mains	200W	June 1991	Dual RX, auto-ATU, keyer
Yaesu	FT-1000MP	base large	HF	1995	mains/13V	100W	January 1996	Dual RX, AF DSP, auto-ATU, keyer"
Yaesu	FT-1000MP mk5	base large	HF	2000	mains	200W	October 2000	Separate PSU, Class AB/A PA, Dual RX
Yaesu	FT-2000	base large	HF 50	2006	mains/13V	100W		Dual RX, IF DSP, DMU display features
Yaesu	FT-2000D	base large	HF 50	2007	mains	200W	March 2008	Separate PSU, Class AB/A PA, Dual RX
Yaesu	FT-450	base small	HF 50	2007	13V	100W	October 2007	IF DSP, optional ATU
Yaesu	FT-450D	base small	HF 50	2011	13V	100W		Style upgrade. ATU included
Yaesu	FTDX5000	base large	HF 50	2010	mains	200W	June 2010	3 variants, top end, feature packed
Yaesu	FT-747GX	base small	HF	1988	13V	100W	May 1989	FM optional, low cost"
Yaesu	FT-817	portable	HF 50 144 430	2001	9-13V	5W	June 2001	Int batteries / ext supply
Yaesu	FT-840	base small	HF	1993	13V	100W	February 1994	Optional FM
Yaesu	FT-847	base mid	HF 50 144 430	1998	13V	100W	August 1998	50W on 2m/70cm, duplex satellite, 4m
Yaesu	FT-857	mobile	HF 50 144 430	2003	13V	100W	June 2003	Detachable panel. 50W/2m 20W/70cm
Yaesu	FT-890	base small	HF	1992	13V	100W	September 1992	Optional ATU
Yaesu	FT-897	base small	HF 50 144 430	2003	mains/13V	100W	April 2003	20W with int batteries.
Yaesu	FT-900	base small	HF	1994	13V	100W	November 1994	Optional ATU
Yaesu	FT-920	base large	HF 50	1997	13V	100W	August 1997	ATU, keyer, voice store, AF DSP
Yaesu	FT-950	base mid	HF 50	2007	13V	100W	December 2007	IF DSP, ATU, DMU display features
Yaesu	FT-990	base large	HF	1991	mains	100W	April 1992	Auto ATU, keyer
Yaesu	FTDX9000D	base large	HF 50	2005	mains	100W	December 2006	Dual RX, microtune filters, top end

Table 7.31: Some of the most popular models of HF transceiver over the last 25 years

Performance

The most important performance parameters for both the receiver and the transmitter are shown in **Tables 7.28 and 7.29**. A high performance receiver is most beneficial in contesting and DXing where wanted signals can be weak, bands crowded and unwanted signals strong. A 100dB dynamic range in SSB bandwidths represents a target for the very best receivers, 95dB is very good, 85-95dB is typical for mid price radios and 80-85dB for budget priced radios.

Third order intermodulation and reciprocal mixing are the principal performance measuring parameters of significance for dynamic range. Achieving a respectable dynamic range at frequency offsets greater than 20kHz from the receive frequency is achieved by most commercial amateur receivers these days, but closer in, particularly inside the roofing filter bandwidth, the performance of most receivers degrades sharply. There are very few radios which achieve even 80dB dynamic range at 5kHz offset, and cost is not necessarily an indicator of best performance at these close spacings. **Table 7.30** summarises the performance of receivers measured for *RadCom* equipment reviews over the last 20 years in respect of their dynamic range range in 2.4kHz bandwidth due to third order intermodulation at 5kHz and 50kHz offset (ILDR - intermodulation limited dynamic range). Reciprocal mixing figures are quoted in terms of phase noise in dBC/Hz at 10kHz offset. Subtract 34dB to give the dynamic range in 2.4kHz bandwidth or 27dB to give the dynamic range in 500Hz bandwidth.

Cost

Cost is generally related to performance and features. Second hand purchases can be a good buy as radios do not normally wear out and significant savings can be made. Some useful guidance on buying second hand is contained in the RSGB Publication *The Rig Guide*, edited by Steve White., G3ZVW

Frequency Coverage

All HF transceivers cover the bands 1.8 to 30MHz but many also provide coverage of the VHF and UHF bands. A growing number provide coverage of 50MHz and a lesser number cover 144, 432 and even 1296MHz. If your interests cover VHF and UHF as well as the HF bands then a multiband radio may be of particular interest. Many such radios allow simultaneous reception and transmission (full duplex) between main band groupings eg between HF, 144 and 432MHz and some with frequency tracking making them particularly suitable for satellite working.

RF Power Output

Most HF transceivers provide a transmit output power of 100W unless they are intended for use on internal batteries in which case the power is significantly less. A few provide a higher power of 200W and even 400W and this extra power can be an advantage. However where a linear amplifier is also used the extra power is unnecessary, and indeed unwanted where there is a danger of overdrive of the linear amplifier.

Available Radios

The "big three" Yaesu, Icom and Kenwood have over the years produced the greatest range of available models of HF transceivers. However the US suppliers Ten-Tec and more recently Elecraft have also produced popular models, in many cases specialising in good RF performance. **Table 7.31** lists the key features of many of the most popular models produced over the last 25 years.

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About the Author

Bill Mantovani, G4ZVB, *first became interested in electronics and radio as a short wave listener whilst still at school. He turned his interest into a career within the telecommunications field where, after his studies, he joined the team that designed the original digital telephone exchange. During this time he started writing articles for a variety of publications, initially on his other interests of motor sport and classic car restoration but later on radio and electronics.*

A change of career and a move back North in the early 1980s saw Bill become a college lecturer. This offered him more time for radio and writing and the opportunity to finally switch from being a SWL to a licensed amateur. Keen on homebrew, more articles and a number of books on various subjects followed as well as talks to clubs around the Yorkshire area. Despite keeping up to date with modern technology, Bill is a fan of older amateur radio equipment and relishes the challenge of updating various circuitry whilst still retaining the good old valve PA!