A SINGLE-STAGE LINEAR AMPLIFIER FOR 50MHz

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INTRODUCTION

Since the allocation of 50MHz in the UK, activity and interest in this fascinating band has increased, and it is obvious from working on the band that many stations have realized that a four- or five-element Yagi is larger than anticipated, and have had to resort to a rotatable dipole. Although there is no substitute for large antenna arrays, the simple dipole can, and will, provide good performance while remaining unobtrusive. Due to its wide horizontal and vertical radiation pattern, it has also proven to be a very effective antenna for meteor scatter working.

So why all the preamble about antennas when this article is about amplifiers?

50MHz is our only allocation which is limited by erp. When using ssb the limitation is 20dBW p.e.p or 100W p.e.p erp. When using a dipole antenna the limit is 100W p.e.p at the antenna feedpoint.

The purpose of this article is to describe the design and construction of a single-stage transistor amplifier for the 50MHz band. The gain of the amplifier is approximately 12dB, and the quoted output of 100W p.e.p will be obtained with 6–7W of drive, which is within the capability of nearly all commercially-manufactured transceivers and transverters. Although the design is for use on 50MHz, it is hoped that the general design philosophy presented will encourage others to design and build simple, single-stage amplifiers for other frequencies using readily-available manufacturers' data.

There are many differing solutions which may be used to match transistors to their input/output sources and loads. The approach has been deliberately kept as simple as possible, which allows all matching component values to be realized using a simple pocket calculator.

Amplifier input and output are designed for 50Ω termination, and the unit contains built-in coaxial changeover relays. Transmit and receive signals are directly fed through when the amplifier is switched off. Power supply requirements are 48V at approximately 5A.

OBJECTIVES

Although I am a self-confessed valve man where high-power linear amplifiers are concerned, certain aspects favour the use of transistors in medium-power linear amplifiers. An equivalent valve amplifier would need to use the QQV-0750 family with the increased complexity of multiple voltage power supplies. For the constructor with a junk box brimming over with valves, bases and high voltage transformers, the question of cost does not arise, but for those who need to purchase these components, prices may be a little higher than expected!

The BLW96 used in this design is capable of excellent linearity at the quoted output power, and is ideally suited for operation at 50MHz.

POWER SUPPLY CONSIDERATIONS

Why 48V? Historically, 13·8V devices have been used for mobile installations. There are few power transistors capable of delivering 100W output from a 13·8V supply with good linearity. Some 28V devices exist which will give the required power, but if it is decided to move away from the 13·8V convention, then it is far better to use 48V. It is far easier to construct a power unit supplying 48V at 5A than one for 13·8V at 18A.

Details of a suitable 48V regulated psu which provides foldback current limiting, short-circuit shutdown and over-voltage protection are given later.

AMPLIFIER DESIGN

Input matching

The input matching consists of two T-networks which match the required drive impedance of 50Ω to the complex input impedance of the BLW96. T-networks are used in preference to the more simple L arrangement, as there is no control of the working Q when using the latter. As the Q of an L network increases with the ratio of transformed impedances, high circuit Qs can exist. This makes adjustment very critical and temperature sensitive and also causes unwanted narrow bandwidth and high circulating rf currents. Inspection of the manufacturer's data for the BLW96 shows an input impedance of 0·37 + j0·15Ω (equivalent series components) at 50MHz. The reactive part of the input is very small (0·47nH) and in this instance may be ignored. However, other devices may have a much higher reactive component in the input impedance, and for this reason it will be taken into account as part of the input matching network in order to demonstrate the method.

To start the design, it is first necessary to decide the intermediate resistance between the two T-match sections (see Fig 2). Assuming both sections have the same Q, the intermediate resistance is the geometric mean value of source and load resistance R1 and R2'. Therefore, intermediate resistance (RI) = $\sqrt{0{\cdot}37 \times 50}$ = 4·3Ω. Working from the base, the first T-match section transforms from 0·37Ω to 4·3Ω, and the second section from 4·3Ω to the design source impedance of 50Ω.

The next requirement is to define the working Q of the sections. If too low a value is selected, component values become impractical; too high a value leads to the unwanted problems described earlier. The minimum working Q for given values of R1 and R2', is when $R1(Q^2+1)/R2$ is greater than 1. If variables are selected which give a value less than 1, the value of A cannot be solved. In order to satisfy this requirement, and avoid high Q values for the reasons described earlier, a Q of 4 in both sections is used for this design.

Referring to Fig 1, the following formulas are applied to obtain the values for XL, XC1 and XC2.

$XL = QR1$ ... (1)
$XC2 = AR2$ ... (2)
$XC1 = B/Q - A$ ... (3)
Where $A = (B/R2 - 1)^{0{\cdot}5}$ ... (4)
$B = R1(Q^2+1)$ ... (5)

Inserting values into the formulas gives the following values:
From (5) .... $B = 0{\cdot}37\,(4^2+1)$. Therefore $B = 6{\cdot}29$.
From (4) .... $A = (6{\cdot}29/4{\cdot}3 - 1)^{0{\cdot}5}$. Therefore $A = 0{\cdot}68$.
From (3) .... $XC1 = 6{\cdot}29/(4 - 0{\cdot}68)$ Therefore XC1' = 1·89Ω.
From (2) .... $XC2 = 0{\cdot}68 \times 4{\cdot}3$. Therefore XC' = 2·92Ω.
From (1) .... $XL1 = 4 \times 0{\cdot}37$. Therefore XL1' = 1·48Ω.

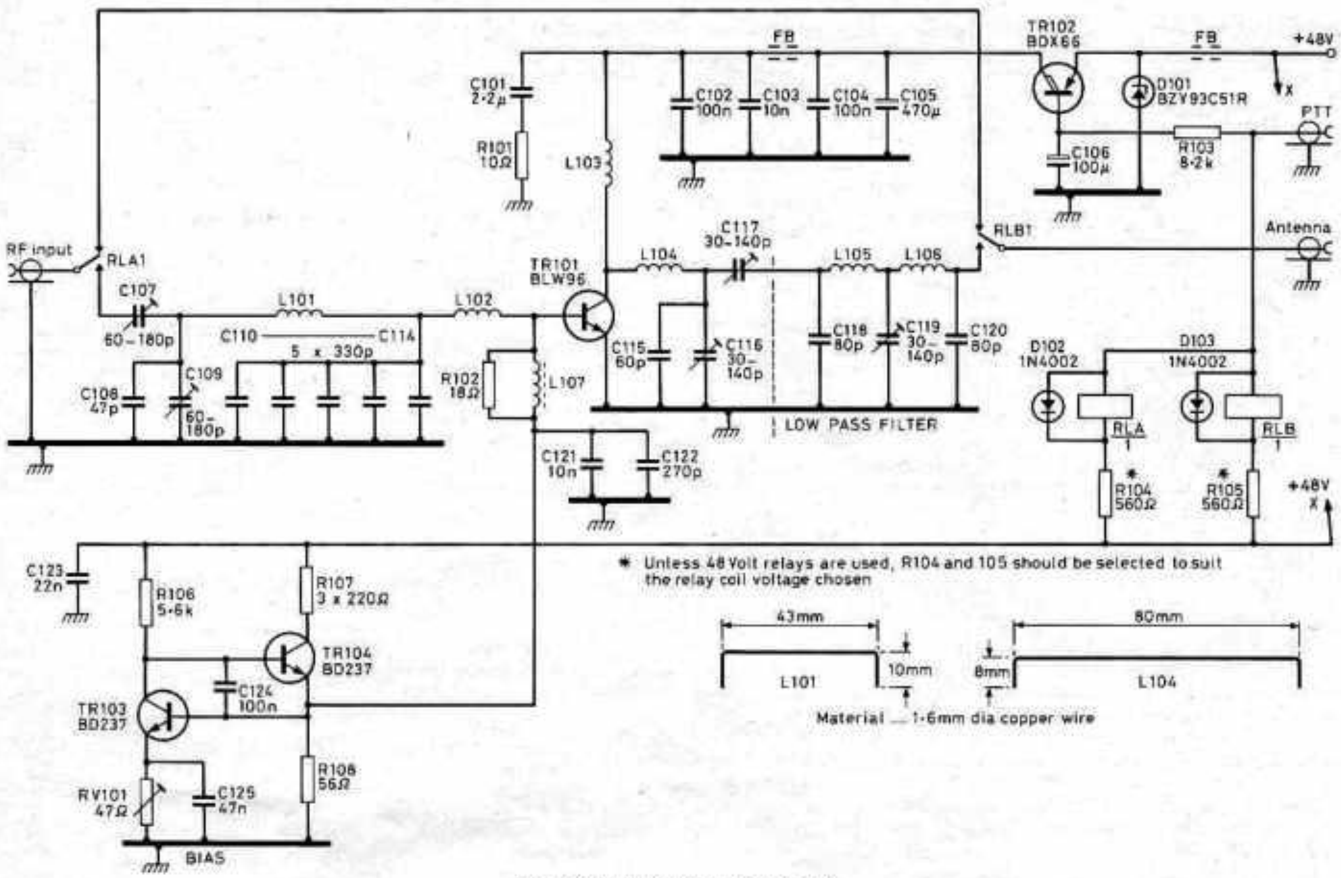


Fig 1. Circuit diagram of the amplifier

To calculate the component values:

$C = 1/2\pi f XC$ ... (6)

$L = XL/2\pi f$ ... (7)

Therefore $C1 = 1/2\pi \times 50 \times 10^6 \times 1{\cdot}89 = 1{,}684\text{pF}$

and $C2 = 1/2\pi \times 50 \times 10^6 \times 2{\cdot}92 = 1{,}090\text{pF}$

$L1 = 1{\cdot}48/2\pi \times 50 \times 10^6 = 4{\cdot}71\text{nH}$

The final configuration and values for the first matching section are shown in Fig 3(a). Although in this case the inductive reactance of the transistor input is insignificant from a practical point of view, it will be subtracted from the value of L1 in order to keep the example correct. The value of L1 then becomes $4{\cdot}7 - 0{\cdot}47 = 4{\cdot}23\text{nH}$ ($XL1 = 1{\cdot}48 - 0{\cdot}15\Omega$). The second matching section, transforming the intermediate value of $4{\cdot}3\Omega$ up to the required driving impedance of 50Ω, is obtained in the same way.

A working Q of 4 is used again. Refer to Fig 2.

From (5) .... $B = 4{\cdot}3(4^2+1)$. Therefore $B = 73{\cdot}1$.

From (4) .... $A = (73{\cdot}1/50 - 1)^{0{\cdot}5}$. Therefore $A = 0{\cdot}68$.

From (3) .... $XC1' = 73{\cdot}1/4 - 0{\cdot}68$. Therefore $XC1' = 22\Omega$.

From (2) .... $XC2' = 0{\cdot}68 \times 50$. Therefore $XC2' = 34\Omega$.

From (1) .... $XL1' = 4 \times 4{\cdot}3$. Therefore $XL1' = 17{\cdot}2\Omega$.

From (6) .... $C1' = 1/2\pi \times 50 \times 10^6 \times 22$. Therefore $C1' = 145\text{pF}$.

$C2' = 1/2\pi \times 50 \times 10^6 \times 34$. Therefore $C2' = 94\text{pF}$.

From (7) .... $L1' = 172/2\pi \times 50 \times 10^6$. Therefore $L1' = 55\text{nH}$.

The circuit configuration, with values, is shown in Fig 3b. In practice the series components C2 and L1′ are combined into one component. The capacitive reactance value ($2{\cdot}92\Omega$) is subtracted from the value of inductive reactance ($17{\cdot}2\Omega$). This leaves an effective inductive reactance of $14{\cdot}28\Omega$,

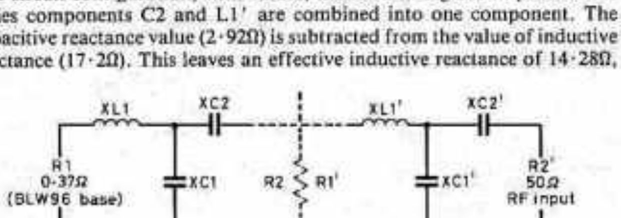


Fig 2. Input matching arrangement

which represent a value of 45·5nH. The values shown in the circuit diagram represent this effective inductance. The calculated values tend to be rather impractical of course, and far from standard value components. Where practical, mica compression trimmers are used based on the calculated component values.

Output matching circuit

Before work on the output matching is started, it will be necessary to determine the output impedance of the BLW96. This may be done in two ways. The first, by taking the values directly from the manufacturer's data sheets (only valid for a specific power level); or second, if these are not available, by means of a simple calculation.

Most data sheets include a simple graph of resistance and reactance plotted against frequency for a given output power. For the BLW96 at 50MHz the equivalent series load impedance is $4 + j3\Omega$. Unfortunately this value is quoted at the wrong power level for this design. It should be made clear that these values are the complex conjugate of the transistor load impedance and represent the *load* required to match the device correctly. In this particular case, the transistor is represented by a 4Ω resistor in a series with a 1,060pF capacitor. If a full data sheet is not available or, as in this

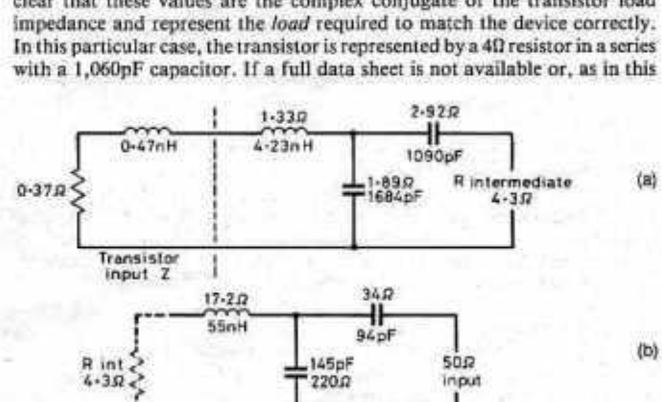
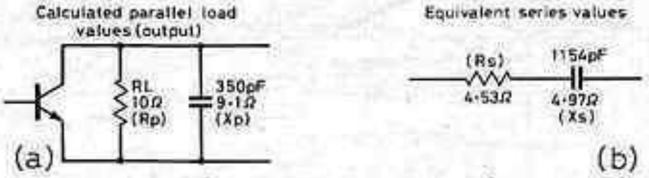


Fig 3. (a) First T-match section (input). (b) Second T-match section (input)



$$Rs = Rp\frac{Xp^2}{Xp^2 + Rp^2} \qquad Xs = Xp\frac{Rp^2}{Rp^2 + Xp^2}$$

Fig 4. Output load conversions. (a) Calculated parallel load values. (b) Equivalent series values

example, values are quoted at the wrong power level, a close approximation may be made by using the following formula in conjunction with the output capacitance.

$$RL = (Vcc - Vsat)^2/2 \times Pout$$

Based on a saturation voltage of 2V, and a power output of 100W p.e.p, the value of RL is:

$$(48 - 2)^2/2 \times 100. \text{ Therefore } RL = 10\cdot58\Omega.$$

The collector capacitance against voltage will normally be shown in the form of a graph, or published in tabular form.

As large changes in capacitance occur over the range of collector voltages, a general rule of thumb is to take the value shown at 50 per cent of the supply. In this case, Cc amounts to 350pF (Xc = 9Ω) at a Vcb of 25V. This value is in parallel with the load resistance of 10Ω previously calculated.

For ease of matching, and to enable a comparison to be made with the published figures, the parallel circuit must be converted into an equivalent series circuit. The conversion formulas with the calculated values are shown in Fig 4. It will be seen that these figures differ slightly from the values in the data sheet as the calculation was carried out at a different power level. A good degree of accuracy is obtained if no other data is available at a specific power level.

Now that the required collector load impedance is defined, the output matching circuit can be designed to match from 4·5–j5Ω to the required output of 50Ω. Unlike the input matching, only one T-match section will be required as the impedance step up ratio is lower.

Referring back to equations (1)–(5), and using a Q of 4:

$B = 4\cdot5(4^2 + 1)$. Therefore $B = 76\cdot5$.
From (5) .... $A = (76\cdot5/50 - 1)^{0\cdot5}$. Therefore $A = 0\cdot728$.
From (1) .... $XL = 4 \times 4\cdot5$. Therefore $XL = 18\Omega$.
From (2) .... $XC2 = 0\cdot728 \times 50$. Therefore $XC2 = 36\cdot4\Omega$.
From (3) .... $XC1 = 76\cdot5/4 - 0\cdot728$. Therefore $XC1 = 23\cdot38\Omega$.
From (6) .... $C1 = 1/2\pi \times 50 \times 10^6 \times 23\cdot38$. Therefore $C1 = 136pF$.
From (6) .... $C2 = 1/2\pi \times 50 \times 10^6 \times 36\cdot4$. Therefore $C2 = 87\cdot4pF$.
From (7) .... $L = 18/2\pi \times 50 \times 10^6$. Therefore $L = 57\cdot3nH$.

The final matching circuit values are shown in Fig 5. An additional 5Ω must be included in the value of XL, making a total of 23Ω. This additional reactance, being opposite sign, cancels the capacitive reactance part of the transistor output impedance (–j5Ω). Both the 16nH and 57nH inductors are combined into a single component.

Bias

As this amplifier is required for linear service, a temperature compensated adjustable bias supply is required. It is important to ensure that the output impedance of the biasing circuitry is low enough to supply adequate base current at any power level within the design ratings of the amplifier. The zero signal quiescent current is set at 100mA.

Two BD237 transistors are used in the bias supply; these are TO126 plastic types which are mounted onto the heatsink. Although the temperature compensation sensing transistor (TR103) is dissipating only a small amount of power, a power transistor must be used as a low Vbe is required to enable adequate range of bias adjustment.

The bias arrangement used is based on circuits published in the Mullard technical handbook on rf devices. It has proved successful in past amplifier projects, and provides good compensation over a wide range of temperatures. Adjustable bias is fed via L107 to the base of the BLW96. RV101 is adjusted for a quiescent collector current of 100mA.

DC supply and switching

Due to the very high available gain of most hf/vhf transistors at low frequencies (the BLW96 has 30dB gain at 1MHz), it is important to provide adequate supply decoupling, from hf to vhf. C201 to C205 provide this in conjunction with a ferrite suppression bead. Supply for the power amplifier is fed via the rfc L103. The value of L103 is not critical but should present a high impedance relative to the output impedance of the power amplifier

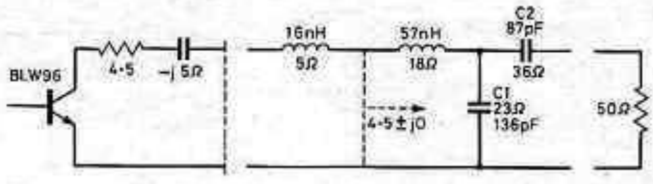


Fig 5. Output matching with values

transistor. If made too high, unwanted resonances can occur, and if too low, gain reduction will be evident. A good approximation is to use an inductive reactance of 15 to 25 times the collector impedance.

In this example, the BLW96 represents 4Ω at the collector. A reactance of 100Ω at 50MHz gives a value of 0·3μH for L103. During receive, supply voltage for the amplifier is removed by the series Darlington switch TR202. When the ptt line is grounded, TR202 switches on, C206 and R204 providing a short time delay (about 40ms) for the amplifier collector supply, which allows time for the coaxial relays to operate before rf power is applied to them. Any pnp Darlington device capable of switching the required voltage and current may be used for TR202, or one may be made up from two discrete devices if preferred.

Lowpass filter

Output from the amplifier is fed via a five-pole Tchebyscheff lowpass filter with a 3dB cut-off frequency of 60MHz. Input and output impedances are 50Ω. The inclusion of this filter reduces the second harmonic content at the output to better than –50dB, and the third, better than –55dB. It is not the intention of this article to venture into the design of filters, as many books are written on this subject alone! Values for the lpf included in this design were taken from normalized tables which may be found in references [1] and [2]. The normalized values with the appropriate formulas are shown in Fig 5.

Inductance formulas

Having calculated various inductance values in the design of matching sections and filters, the problem then arises of converting the theoretical values into physical inductors. For low values such as those used in the input and output matching, straight wire inductors were chosen. In order to calculate the inductance of a straight wire the following formula is used:

$$L = 0\cdot002\, l\, (\log_e 4l/d - 1)$$

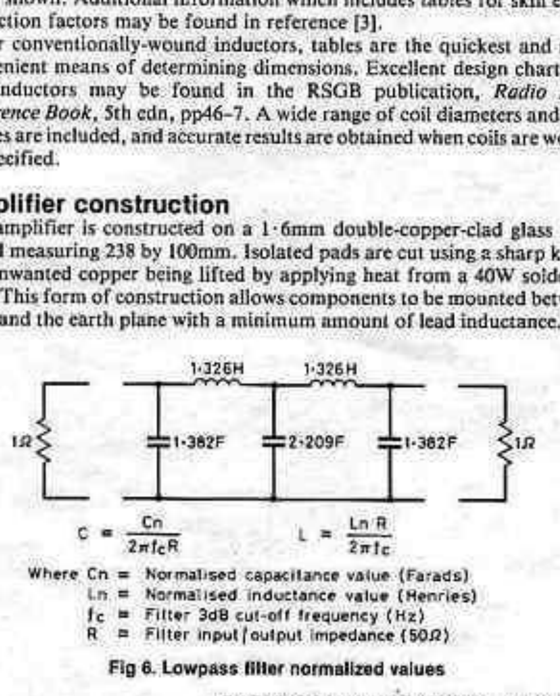
Where: L = Inductance in microhenries
l = Length of wire in centimetres
d = Diameter of wire in centimetres

It must be emphasised that the value of l used is not exact, and does vary according to the frequency. To be precise, a skin effect correction factor should be included. However, at vhf this effect is catered for by using the value shown. Additional information which includes tables for skin effect correction factors may be found in reference [3].

For conventionally-wound inductors, tables are the quickest and most convenient means of determining dimensions. Excellent design charts for vhf inductors may be found in the RSGB publication, *Radio Data Reference Book*, 5th edn, pp46–7. A wide range of coil diameters and wire gauges are included, and accurate results are obtained when coils are wound as specified.

Amplifier construction

The amplifier is constructed on a 1·6mm double-copper-clad glass fibre board measuring 238 by 100mm. Isolated pads are cut using a sharp knife, the unwanted copper being lifted by applying heat from a 40W soldering iron. This form of construction allows components to be mounted between pads and the earth plane with a minimum amount of lead inductance. The

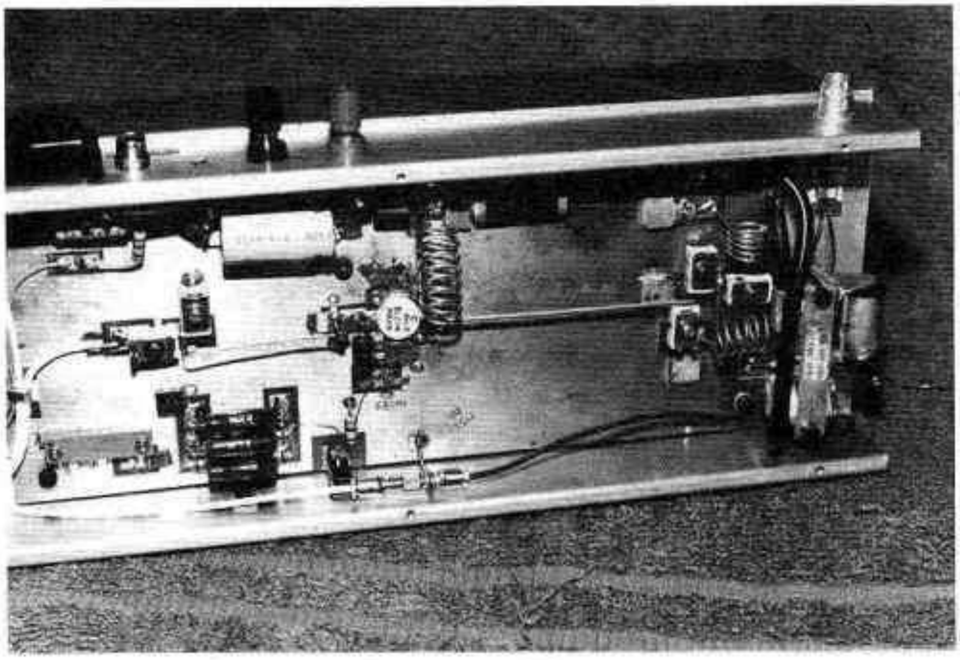


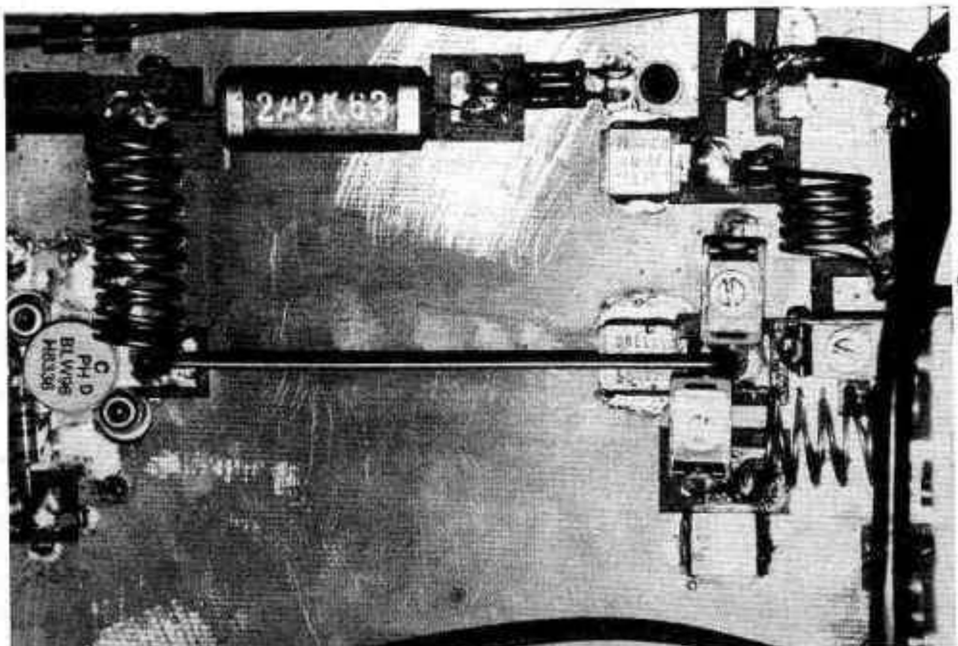
$$C = \frac{Cn}{2\pi fcR} \qquad L = \frac{Ln\, R}{2\pi fc}$$

Where Cn = Normalised capacitance value (Farads)
Ln = Normalised inductance value (Henries)
fc = Filter 3dB cut-off frequency (Hz)
R = Filter input/output impedance (50Ω)

Fig 6. Lowpass filter normalized values

The underside view of the amplifier. The output components can be seen on the right of the amplifier with the lowpass filter components adjacent to the coaxial relay. Bias components are located in the left-hand corner. Input matching components can be seen on the left-hand side





Close up view of the output matching components and the lowpass filter

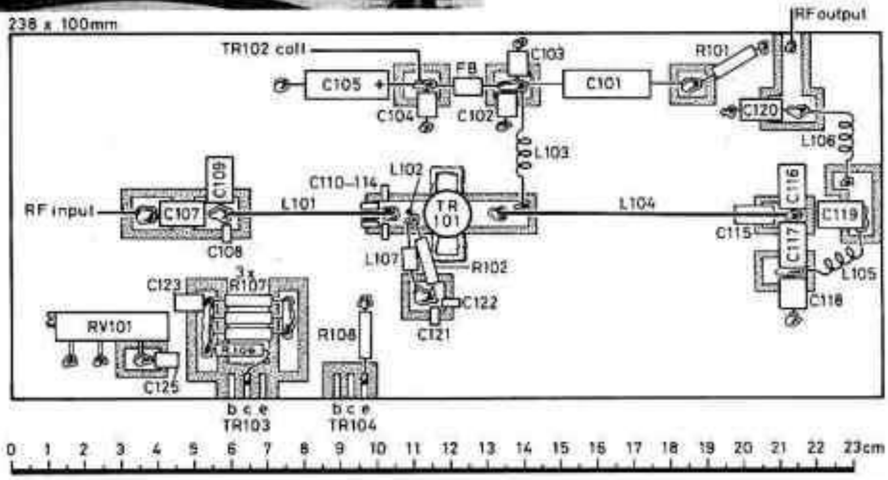


Fig 7. Main board layout

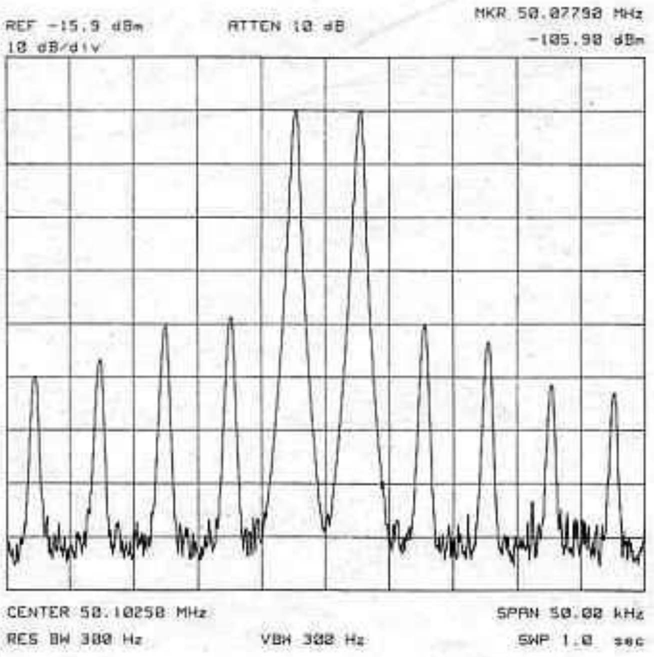


Fig 8. Intermodulation products at 100W p.e.p

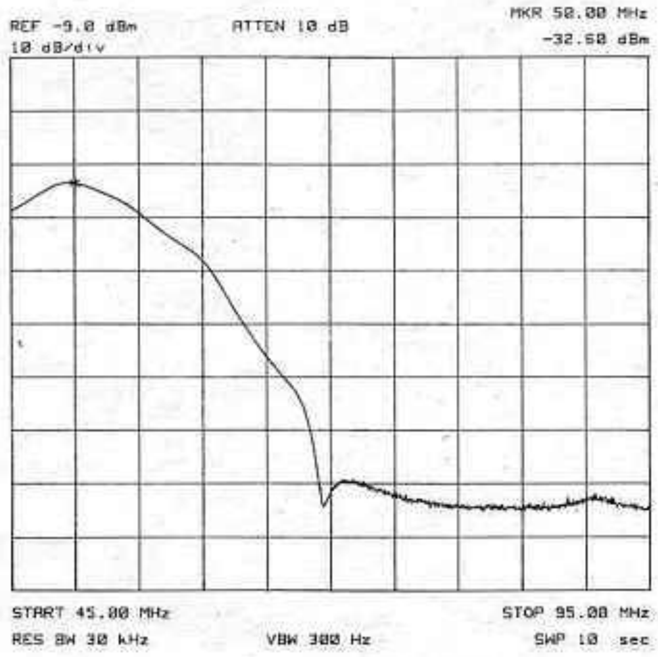


Fig 9. Amplifier frequency response

underside of the board is left as a complete earth plane except for one small area. A square of copper must be removed beneath the pad at the junction of L104, C115 and C117. Connections between earth-planes on each side of the board are made by drilling and pinning four holes around each emitter lead of the BLW96. Component locations and dimensions for the board are shown in Fig 6.

The completed amplifier was mounted on a large finned heatsink with a thermal resistance of approximately 0·3°C/W. This provides more than adequate cooling for normal ssb and cw operation. Total thermal resistance between the transistor junction and air, when using the above-mentioned heatsink rating, is 1·13°C/W. In real terms this means that 150W can be dissipated at an ambient temperature of 30°C before the maximum junction temperature is exceeded.

In common with all power devices, great care should be taken to ensure that the mounting surface is perfectly flat. After drilling and tapping the mounting holes, do not de-burr the holes by counter sinking, as this action degrades the thermal resistance. Any burrs should be removed with a safety-razor blade or with a flat block and fine-grade wet and dry.

Bias adjustment

Remove the earthy end of L103 and turn RV101 to minimum resistance. Apply 48V to the dc input socket and earth the ptt line. With 48V input, TR102 collector should be at 46·8V. Release the ptt and TR102 collector should return to 0V. Connect an ammeter in series with L103, terminate rf input and output with 50Ω loads, and earth the ptt line. Adjust RV101 for a standing current of 100mA, remove the dc supply and replace L103.

Alignment

The input matching capacitors C107 and C109 should be adjusted for optimum vswr using a low-level drive source (1–2W). After the initial adjustment, C116 and C117 may be adjusted for maximum power output. Increase the drive level to 4W and repeat the input adjustments for best

Components List

AMPLIFIER

R101	10Ω 0·5W cf		
R102	18Ω 0·5W cf		
R103	8·2kΩ 0·25W cf		
R104, 105	Select to suit relays used		
R106	5·6kΩ 0·5W cf		
R107	3 × 220Ω 6W w/w		
R108	56Ω 0·5W cf		
RV101	47Ω ct		
C101	2·2µf 63V pc	**C115**	60pF 250V Um
C102, 104	100nF 100V mc	**C116, 117, 119**	30–140pF mct
C103, 121	10nF 100V mc	**C118, 120**	80pF 250V Um
C105	470µF 63V te	**C122**	270pF 100V mc
C106	100µF 63V te	**C123**	22nF 100V mc
C107, 109	60–180pF mct	**C124**	100nF 100V mc
C108	47pF 50V cc	**C125**	47nF 100V mc
C110-114	330pF 50V cc		
D101	BZY93 C51R zener		
D102, 103	1N4002		
TR101	BLW96		
TR102	BDX66 pnp Darlington		
TR103, TR104	BD237		
FB	Suppression bead. Material, 3S2 (blue)		
L101	See separate drawing		
L102	15 × 7mm pad on pcb		
L103	12t 1·2mm copper wire, 9mm id, 28mm long		
L104	See separate drawing		
L105, 106	4·5t 1·2mm copper wire, 10mm id		
L107	2·5t 0·5mm enam copper wire wound through six-hole ferrite bead		
RLA, RLB	50Ω coaxial Type CX120P		

POWER SUPPLY UNIT

R201, 202	0·5Ω 10W w/w	**R209**	330Ω 25W a/c (see text)
R203	390Ω 2·5W w/w	**R210**	1·5Ω 10W w/w
R204, 213	3·0kΩ 0·25W cf	**R211**	120Ω 0·25W cf
R205	2·2kΩ 0·25W cf	**R212**	180Ω 0·25W cf
R206	39kΩ 0·25W cf	**R214**	220Ω 0·5W cf
R207	27kΩ 0·25W cf	**RV201**	1·0kΩ ct
R208	560Ω 5W w/w	**RV202**	10kΩ ct
C201	10,000µF × 3 100V		
C202	10nF 100V		
C203	470pF 100V		
C204	15µF a/e 63V		
C205	10µF a/e 63V		
C206	100nF 100V		
C207, 208	1nF c 50V		
C209	0·1µf pc 250V		
TR201, 202	BDX67 npn Darlington		
IC201	uA723		
ZD201	36V 5W zener		
ZD202	See text		
TH201	BTY79-400R		
SW201	DPDT 250V ac at 5A		
SW202	Momentary push 250V ac at 5A		
RLC	48V dc coil. 5A contact rating		
Mains tfmr	55V ac 300VA		
Rectifier	25A 200V bridge module		
Filter	5A mains rfi filter.		

Abbreviations
w/w = wirewound. cf = carbon film.
a/c = aluminium clad. ct = cermet trimpot.
a/e aluminium electrolyte. c = ceramic. cc = ceramic chip.
e = electrolytic. mc = monolythic ceramic.
mct = mica compression trimmer. pc = polycarbonate.
te = tubular electrolytic. Um = Unelco mica or atc.

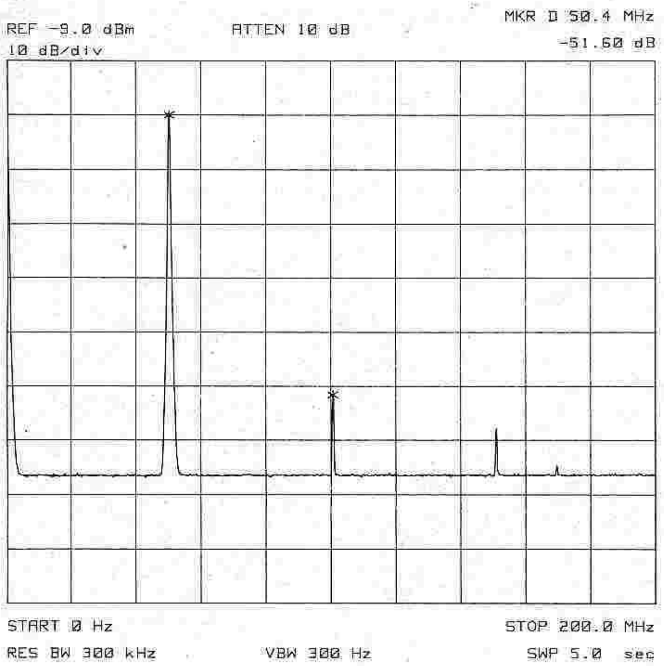


Fig 10. Amplifier harmonics

vswr, and the output tuning for maximum power. At this stage the turns on L105 and L106 can be compressed or opened out and C119 adjusted to obtain maximum output power. If all is well, 60–70W should be available with 4W of drive.

Final adjustments can be made at 100W p.e.p output, but beware, power transistors can also act as very high speed fuses! Severe detuning at high power levels can lead to instant destruction of any solidstate pa. Having said this, it must be stated that the BLW96 is a very rugged device and will withstand a vswr of 50:1 through all phases up to 150W p.e.p at 28MHz.

Linearity of this amplifier is very good and plots of intermodulation product are shown in Fig 8. Frequency response characteristics are shown in Fig 9. Harmonic levels at the rated output are shown in Fig 10.

POWER SUPPLY UNIT

It is not intended to give a detailed description of the power supply used, as most constructors seem to have their own ideas, particularly where protection circuitry is concerned. However, it is hoped that the circuit shown in Fig 11 may provide ideas for those not wishing to duplicate exactly what is shown here.

This psu is identical in design to one powering the amplifier just described. It is also used to power a similar power amplifier for the 70MHz band. Voltage regulation with current foldback is controlled by a 723 precision regulator.

Two npn Darlington power transistors with current sharing resistors (R201 and R202) are used as series-pass elements. Owing to the very high gain of the BDX67 it is possible to drive the bases directly from the Vz output of the 723. Foldback current limiting is set by the potential divider RV201 and R207. Output voltage is controlled by RV202.

As the 723 has a maximum input rating of 40V, it is not possible to use it directly across the supply. For this reason the 723 is connected as a floating regulator across the series-pass Darlington transistors. Zener diode ZD201 and resistor R203 limit the voltage across the 723 to 36V. Over-voltage protection is provided by a simple thyristor crowbar, zener diode combination, TH201 and ZD202. If the value of ZD202 is selected for the required over-voltage protection value, reliable operation will be obtained, but do not rely on the marked zener voltage for the trip value. In the prototype, two series zeners (33 + 18V) were used to provide shutdown at 50V.

Specialized ICs are available to drive a thyristor which do a similar job, ie the MC3423, but have a maximum protection threshold of 45V. If thought necessary, it should be possible to use one of these ICs. However, the negative supply and associated components will need to be returned to a point which is about 10V with respect to chassis. This will allow the protection threshold to be set at 50V.

When an over-voltage condition occurs, the crowbar thyristor fires and RLC de-energizes, switching the psu off. If excessive current is demanded, the current foldback circuit operates and reduces the output voltage to a low level. As before, RLC is de-energized, and the mains supply removed.

R209 reduces the switch-on surge current, and the value shown in the

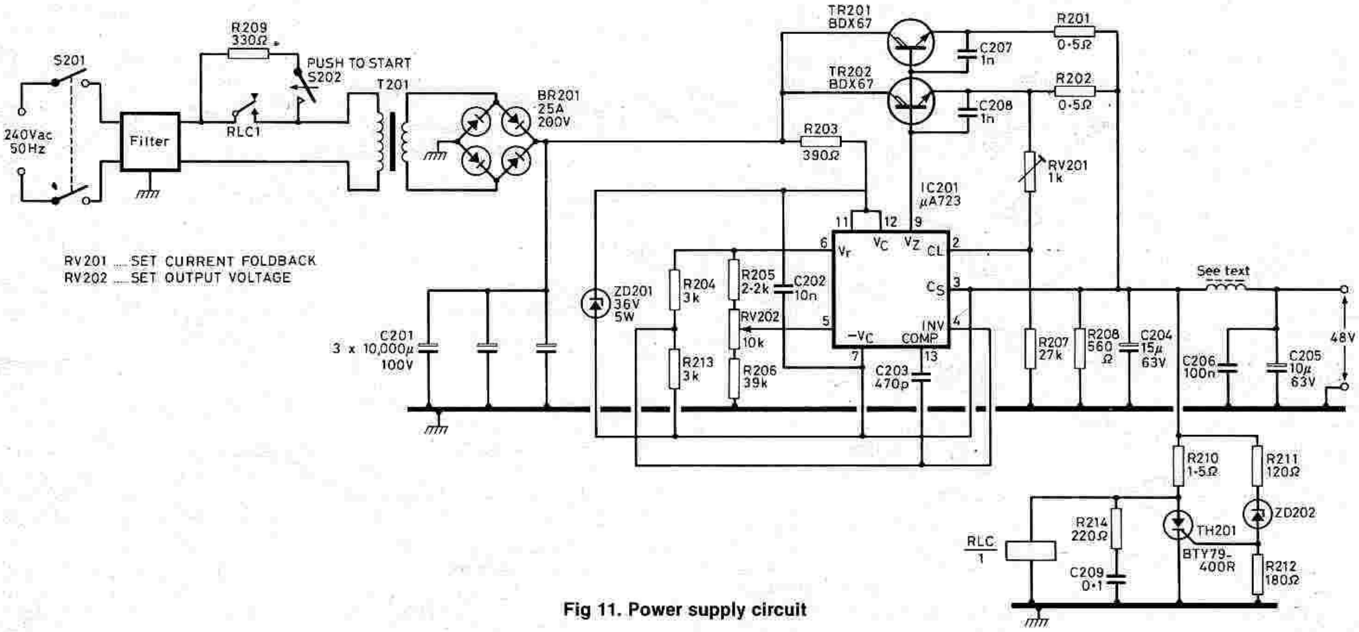


Fig 11. Power supply circuit

circuit diagram (330Ω) is only suitable when starting with low load current. If high current is demanded when starting, R201 will need to be reduced in value to allow the output voltage to rise.

The transformer used was "ex-equipment", and gave 55Vrms when loaded at 8A. Power rating will need to be 300VA. Mains input to the unit is fed via a 5A rfi filter. DC output leads are filtered by winding five turns of the lead-out wire around a toroid core placed as close as possible to the output terminals.

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