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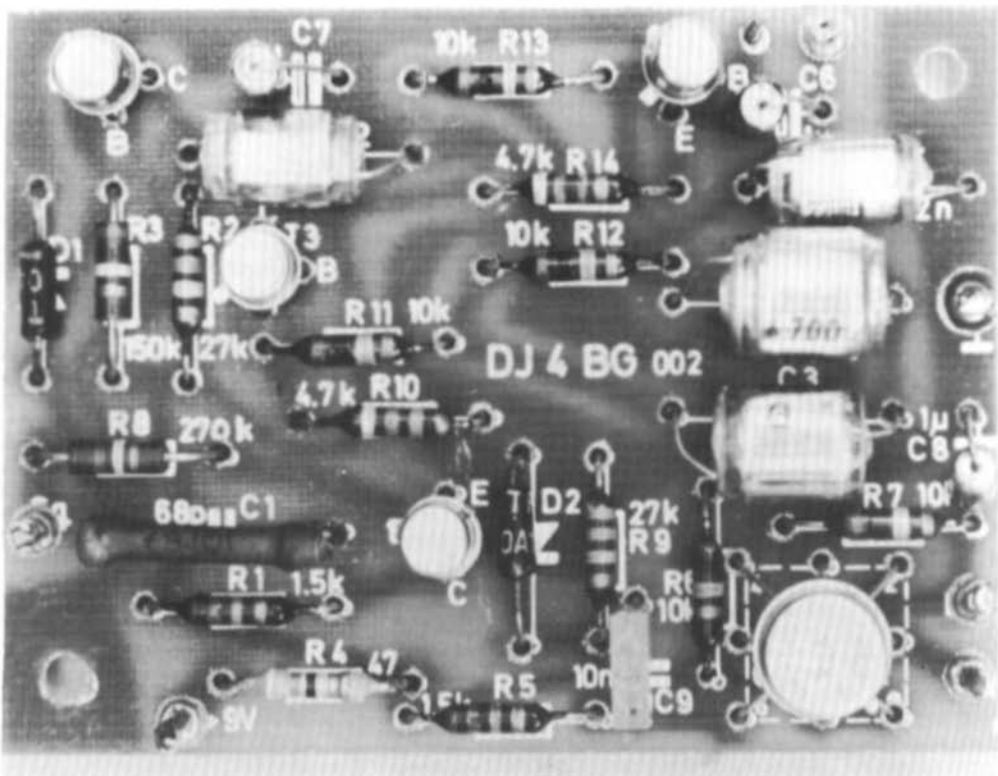

Digital FM Discriminator

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A SSB-TRANSCEIVER WITH SILICON TRANSISTOR COMPLEMENT

PART 2 : THE 9 MHz TRANSCEIVER

by G. Laufs, DL 6 HA

1. INTRODUCTION

Part 2 of this SSB transceiver article describes the circuit, PC-board and complement of the 9 MHz transceiver as well as the alignment of this module.

This transceiver is built up of individual modules which can be used for other applications besides being used in the VHF transceiver; the possibility of constructing a short wave transceiver was mentioned in Part 1 (1).

Figure 1 shows the block diagram of the complete transceiver system. The MOSFET converter module DL 6 HA 001 was already described in Part 1.

Part 3 and 4 will describe the 9 MHz-14 MHz transmit-receive converter; the 14 MHz-144 MHz transmit converter; VFO and low-pass filter as well as the power supply and AF amplifier modules.

2. CIRCUIT DETAILS OF THE 9 MHz TRANSCEIVER

The circuit diagram of the 9 MHz transceiver is shown in Figure 2. The frequency processing of the transmit signal, the receive IF amplifier with automatic gain control and the demodulators are also shown in this diagram. The carrier oscillator (BFO) is accommodated on a separate printed circuit board (Fig. 3), which has the following advantages: The unwanted coupling of the carrier signal to the filter output is reduced on the transmit side. This means that a more favourable carrier suppression is obtained without requiring screening plates. On the receive side, it is necessary for the BFO to be well decoupled from the IF amplifier input, if the automatic gain control is to be driven from the intermediate frequency - as it is in our case. One of the main reasons why the control voltage is often obtained from the audio voltage are the difficulties involved in keeping the relatively high voltage of the carrier oscillator from the sensitive IF amplifier.

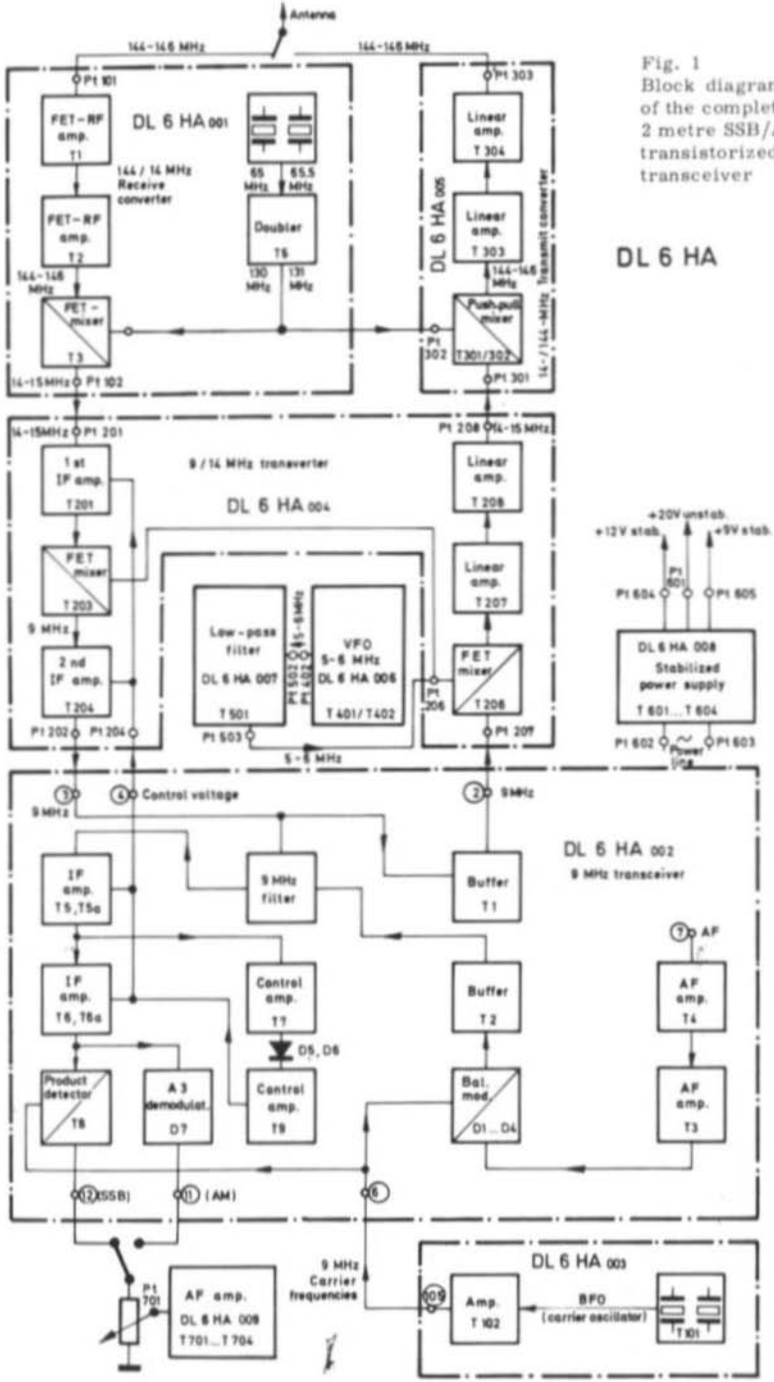


Fig. 1
Block diagram
of the complete
2 metre SSB/AM
transistorized
transceiver

DL 6 HA

In the described circuit, a common carrier oscillator and crystal filter are used for both transmit and receive. The filter is not switched; it is continuously connected to the transmit and receive portion. A 9 MHz crystal filter type XF-9A (of KVG) was used in the prototype. In spite of the good results obtained with this filter, the author recommends the use of a XF-9B type crystal filter. The crystal filter XF-9A is, of course, sufficient for the demands of the transmit portion, but it is possible that certain receive conditions would require the use of the higher performance XF-9B type.

A two-stage amplifier comprising transistors T 5 and T 6 (BF 173) is used in the receive mode. The transistor BF 173 exhibits a very low feedback capacitance and a high transit frequency. The automatic gain control is made using pass transistors in the emitter branch. Cheap AF transistors (Silicon-NPN) can be used here, the type is uncritical.

The control voltage amplifier (T 7) is connected to the base of the second IF amplifier. This means that the second amplifier stage is controlled by forward-action. The decoupling of the BFO signal at the base of the second IF amplifier is sufficiently great that no control voltage is generated by the BFO.

After amplification, the control voltage is rectified in a voltage doubler circuit (D 5, D 6) and fed to transistor T 9. This transistor is biased so that it just does not take current. If this transistor is driven, collector current will flow and will simultaneously alter the base bias voltage of transistor T 10. Under quiescent conditions, this transistor will conduct. The base bias voltage of the transistor can be varied by transistor T 9 till it is switched off in extreme cases. A voltage of +12 V will result on the control voltage line if transistor T 10 is conducting; if this transistor is switched off, zero volts will result. The pass transistors in the IF amplifier (T 5a, T 6a) will be overloaded under full gain conditions, whereas under fully controlled conditions, the base connections will be grounded via resistors R 35, R 38 and R 44.

The demodulation in the A 1 (CW) and A 3 J (SSB) mode is made in a product detector equipped with the double gate MOSFET TA 7150 (T 8). The output impedance of the product detector is formed using the transistor AF driver transformer Ch 2. An envelope demodulator using the diode D 7 is provided for the demodulation of A 3 (AM) signals.

In the transmit portion, a two stage AF amplifier equipped with transistors T 3 and T 4 is used to increase the output signal of a dynamic microphone to the level required to drive the balanced modulator (D 1, D 2, D 3, D 4). The carrier signal of approx. 0.8 V which is continuously connected to the product detector and balanced modulator, is suppressed in the balanced mixer. The resulting double sideband signal with suppressed carrier is fed to transistor T 2 which is used as an impedance converter allowing it to be matched to the crystal filter. A buffer stage (T 1) is provided at the filter output so that the subsequent mixer cannot cause a mismatch condition for the filter.

A carrier injection device is provided for the operating modes A 1 and A 3 (single sideband with carrier injection). In order to inject a carrier, a variable DC voltage can be placed across the balanced modulator via connection point Pt 5.

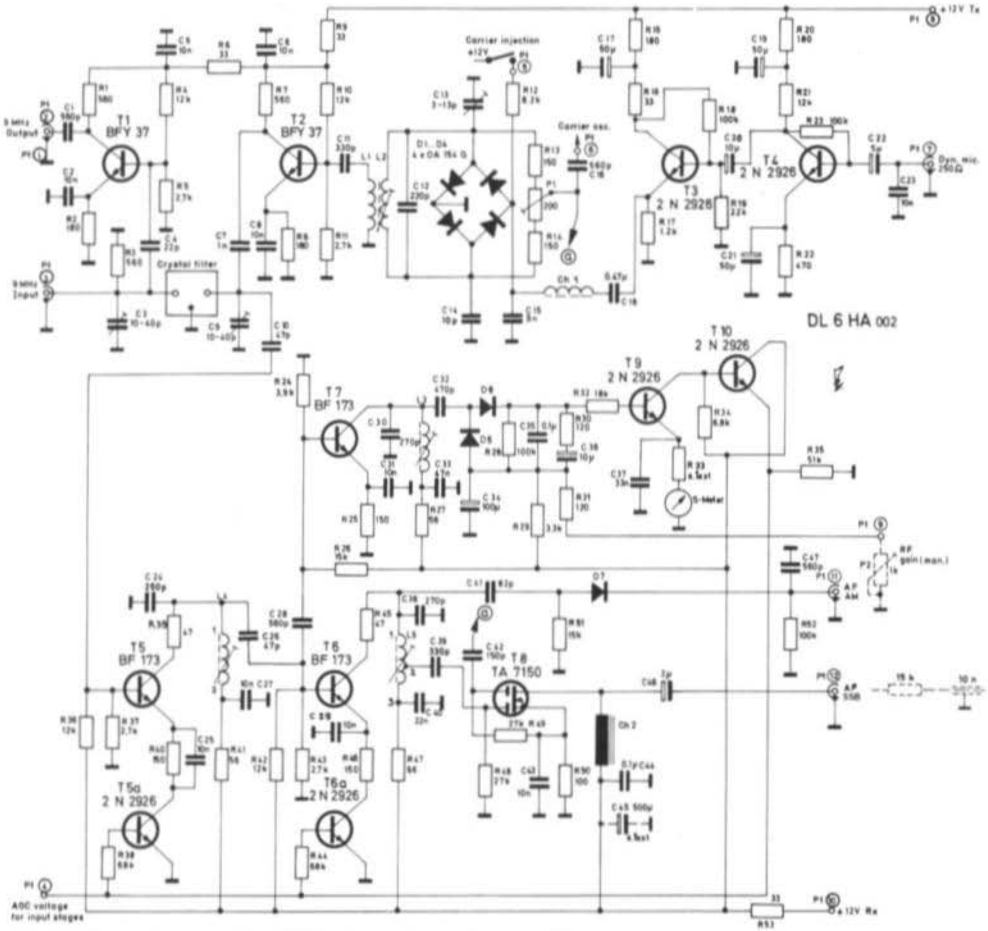
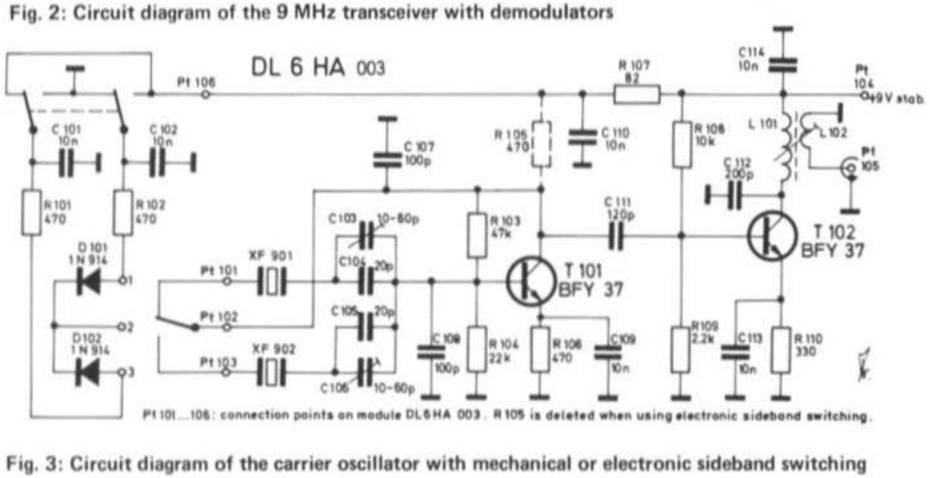


Fig. 2: Circuit diagram of the 9 MHz transceiver with demodulators



Pt 101 ... 106: connection points on module DL6HA 003. R 105 is deleted when using electronic sideband switching.

Fig. 3: Circuit diagram of the carrier oscillator with mechanical or electronic sideband switching

The circuit diagram of the carrier oscillator is shown in Figure 3. The two crystals for the upper and lower sideband oscillate in series-resonance between the collector and the base of transistor T 101. The sideband switching can either be carried out mechanically using a relay or switch, or electronically with the aid of two diodes. The required circuitry is given in the circuit diagram and on the component location plan. The carrier signal is fed via a buffer stage (T 102) to connection point Pt 105 where it is available at a voltage of approx. 0.8 V. The operating voltage of the carrier oscillator should be stabilized.

3. MECHANICAL ASSEMBLY

As was already mentioned in the previous section, the 9 MHz transceiver is built up on two printed circuit boards. Figure 4 shows a photograph of the authors prototype.

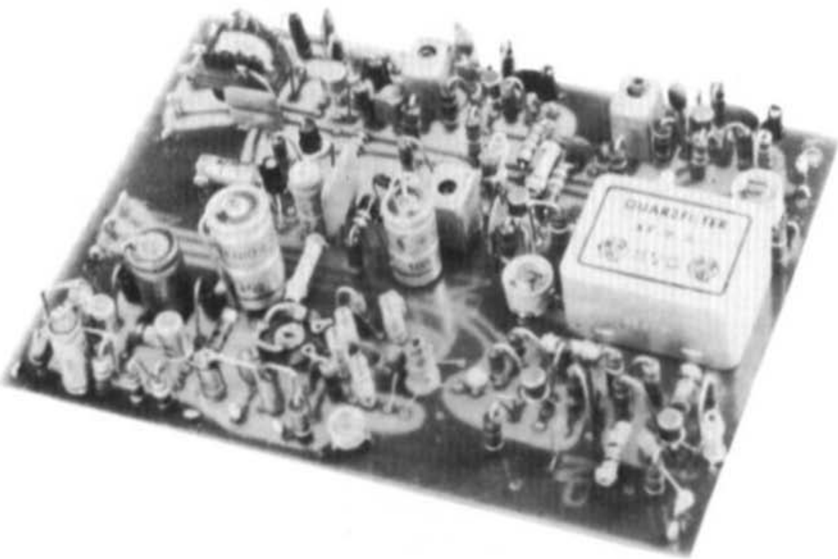


Fig. 4 a: Photograph of DL 6 HA 002

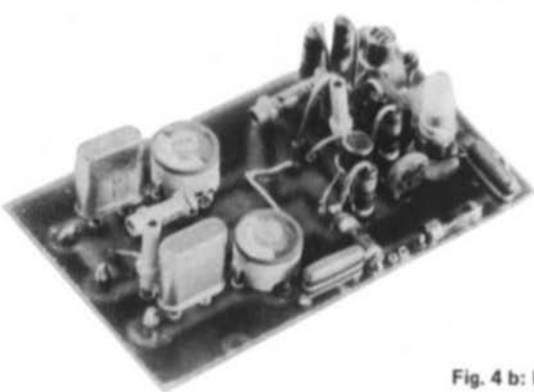


Fig. 4 b: Photograph of DL 6 HA 003

The larger of the two printed circuit boards has the dimensions 147 mm by 103 mm. This printed circuit board which accommodates the components shown in Fig. 2, has been designated DL 6 HA 002 and is illustrated in Fig. 5. The PC-board has been designed so that both the old and the new version of the crystal filter XF-9A or XF-9B can be used. The smaller printed circuit board accommodates the components of the carrier oscillator. The printed circuit board, which has been designated DL 6 HA 003, is illustrated in Figure 6. The dimensions are 70 mm by 40 mm.

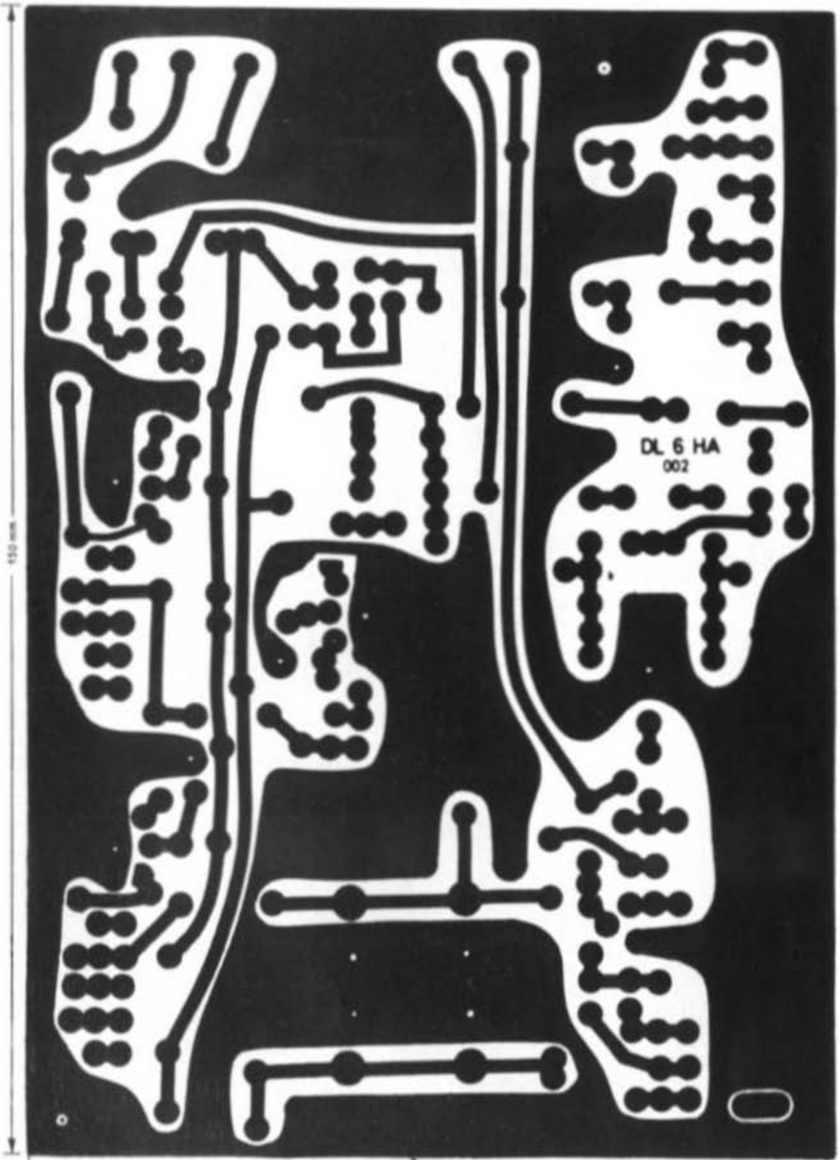


Fig. 5: Printed circuit board DL 6 HA 002

The component location plans of the two printed circuit boards are given in Figures 7 and 8.

Transistor T 8 should be the last component to be mounted. The same precautionary measures should be taken for this MOSFET transistor as were given in Part 1 (1).

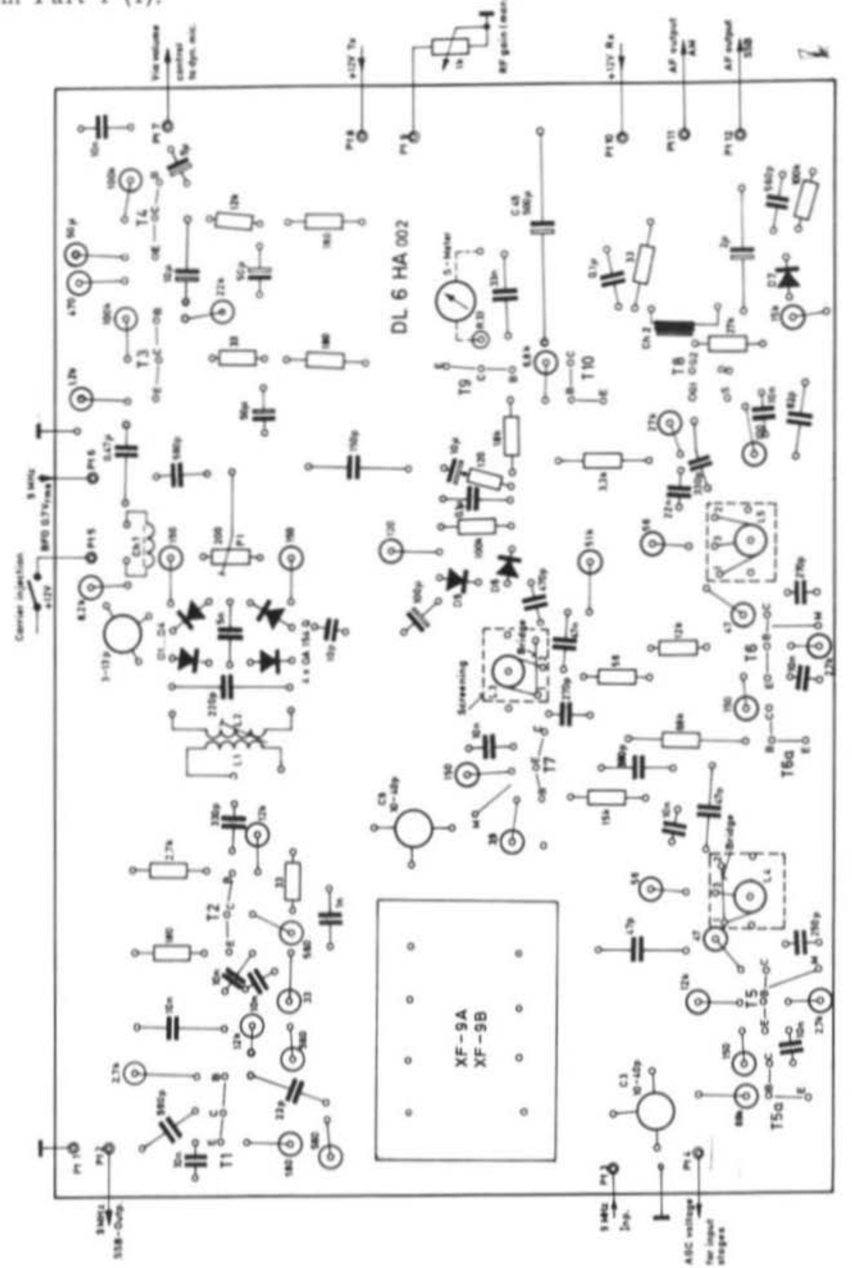


Fig. 7: Component location plan to PC-board DL 6 HA 002

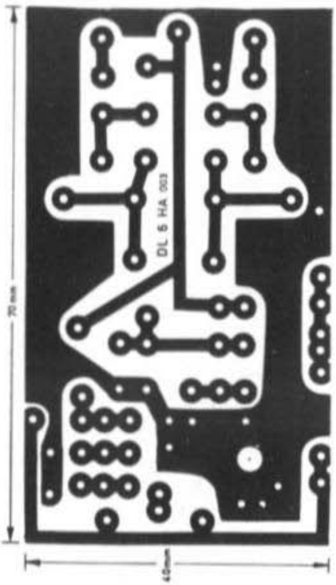


Fig. 6: Prindet circuit board DL 6 HA 003

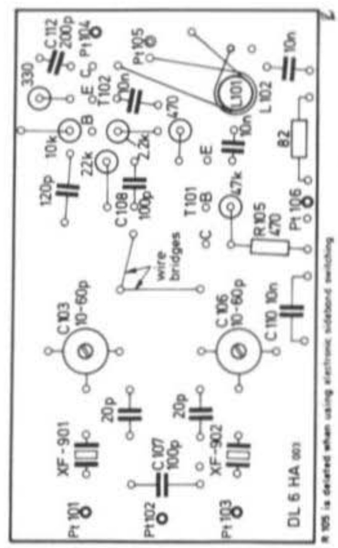


Fig. 8: Component location plan to PC-board DL 6 HA 003

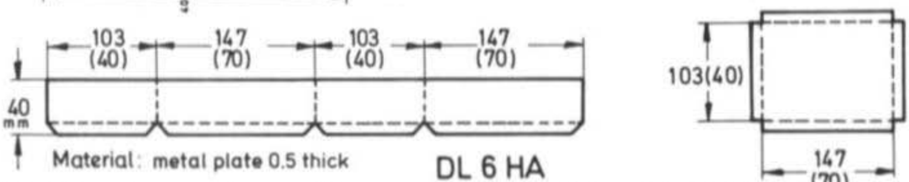


Fig. 9: Screening case with cover. Dimensions in brackets for PC-board DL 6 HA 003

The following points should be observed regarding the external connections of the printed circuit boards: The input connection Pt 7 and the output connection Pt 11 should be connected via feedthrough-capacitors of approximately 150 pF. The output connection Pt 12 is fed via a low-pass filter of 15 kΩ and a 10 nF feedthrough capacitor. This low-pass filter is provided to suppress the residual carrier frequency voltage and to stop it reaching the AF amplifier. In addition to this, the DC voltage connection points Pt 4, Pt 5, Pt 8 and Pt 10 as well as Pt 104 and Pt 106 should be fed via the screening using feedthrough capacitors of 10 nF each.

The required screening is made from thin tin-plate. It is possible to either form a complete casing whereby a spacing of 8 mm must be maintained between the conductor lanes and the bottom of the casing, or for a 4 cm tin-plate strip to be bent around the PC-board so that it can be soldered to the ground of the printed circuit board. This can then be provided with a cover also made out of tin-plate. The circuit lanes then remain accessible. Fig. 9 shows such a screening arrangement in the form of a sketch.

The effective screening of each individual printed circuit board, in conjunction with the use of feedthrough capacitors and coaxial cables for the RF connections, results in a very high spurious signal rejection in the transmit and receive mode. Fundamentally speaking, two oscillators should never be enclosed in the same screening casing in order to avoid generation of combination oscillations due to interaction.

4.1. COIL DATA

Coil formers: 4.3 mm outer diameter, 14 mm long with SW core
Wire: 0.1 mm diameter (38 AWG) enamelled, silk-covered copper wire

L 1 : 4 turns onto the centre of L 2; L 2 = 16 turns
L 3 : 15 turns, in screening can 10 mm by 10 mm, 16 mm high
L 4 : 15 turns, in screening can as for L 3
L 5 : 15 turns, coil tap at 3 turns. Coil length 5 mm, screening can as for L 3
L 101 : 20 turns; L 102 = 6 turns wound onto L 101

Ch 1 : RF choke, approx. 50 μH
Ch 2 : high impedance side of a driver transformer for a transistor AF amplifier

4.2. SEMICONDUCTORS

T 1, T 2 : BFY 37, 2 N 706 (BF 224, 2 N 918)
T 3, T 4 : 2 N 2926, BC 108, BC 183
T 5, T 6, T 7 : BF 173, BF 167, 2 N 4934 (RCA)
T 5a, T 6a : 2 N 2926, BC 108, BC 183
T 8 : TA 7150, TA 7151, 40604 (Dual-gate-MOSFET, RCA)
T 9, T 10 : 2 N 2926 yellow, BC 108 B, BC 183 B
T 101, T 102 : BFY 37, 2 N 706 (BF 224, 2 N 918)

D 1 to D 4 : 4 x OA 154 Q (AEG-Telefunken)
D 5, D 6, D 7 : AA 112 or similar germanium demod. diode
D 101, D 102 : 1 N 914, BAY 38

4.3. CAPACITORS

C 3, C 9 : 10 - 40 pF, ceramic disc trimmer, 10 mm dia.
C 13 : 3 - 13 pF ceramic disc trimmer, 7 mm dia.
C 103, C 106 : 10 - 60 pF ceramic disc trimmer, 10 mm dia.

All capacitors of less than 1 nF are ceramic tubular or disc types.
All bypass capacitors over 1 nF are ceramic disc types.
Capacitors below 1 μF are plastic foil types.

5. ALIGNMENT

Before switching on any of the modules, attention should be paid that the components have been mounted correctly.

5.1. ALIGNMENT OF THE CARRIER OSCILLATOR

A voltage of 9 V is connected to point Pt 104 of printed circuit board DL 6 HA 003. The RF-probe of a valve voltmeter (VTVM) is connected to the RF output Pt 105. One of the two crystals is now switched on using a wire bridge. This is followed by aligning inductance L 101 for maximum output voltage. If a frequency counter or other frequency measuring instrument is available, the crystal-controlled frequency should be aligned to $\pm$ 1.6 kHz from the centre frequency of the crystal filter. If only one sideband is to be used, the other crystal can be adjusted to a frequency spaced 1.3 kHz from the centre frequency of the filter. This allows a higher output voltage to be obtained after the filter in the A 3 and A 1 modes. If no frequency measuring instrument is available, the alignment should be made as follows:

Terminate the input and output of the crystal filter with a 560 Ω resistor. The carrier oscillator is now coupled to the filter via a capacitor of approximately 10 pF. The RF voltage is measured at the output. By varying trimmer capacitor C 103 or C 106, one is able to tune within the passband of the crystal filter. The trimmer for the upper sideband crystal should now be increased until the output voltage is reduced by approximately 10 to 30% of the maximum value. The trimmer for the lower sideband crystal should be reduced until the same output voltage is obtained.

An adjustment of the carrier frequencies to compensate for the voice and the microphone in question should not be made until the transmitter is complete.

5.2. ALIGNMENT OF THE TRANSCEIVER

The transmit portion of this module is firstly brought into operation. The operating voltage is connected to point Pt 8. The carrier is injected by connecting a voltage of +12 V to the carrier injection connection Pt 5. One should now ascertain whether the carrier oscillator provides a voltage of approximately 0.8 V at the balanced modulator. The VTVM is now connected to the output Pt 2. The resonant circuit of the balanced modulator is now aligned with the aid of inductance L 2 for maximum output voltage. Due to the damping by the balanced modulator, the resonance curve is relatively broad. The voltage at the carrier injection connection is now removed and trimmer potentiometer P 1 as well as trimmer capacitor C 13 of the balanced modulator aligned for minimum output voltage. This alignment of the carrier suppression can only be made coarsely; a receiver must be connected for the final adjustment.

The microphone is now connected. The level should be adjustable with the aid of a potentiometer. It is now possible to observe the output voltage whilst talking into the microphone. If an audio signal generator is available, it is possible to align the passband range of the filter for minimum ripple. Trimmer capacitors C 3 and C 9 should be adjusted so that a virtually constant output voltage is obtained within the passband range of 300 to 2700 Hz.

This is followed by alignment of the receive portion. The operating voltage is connected to point Pt 10, and point Pt 9 of the RF gain connected to ground. A meter having 2 mA FSD should now be connected via a trimmer potentiometer of 1 kΩ to the S-meter connection. A voltage of +12 volt is fed via a 10 kΩ resistor to the control voltage connection Pt 4. At first, the BFO is not provided with voltage. A 9 MHz signal generator should now be connected to the input Pt 3. If no signal generator is available, the BFO must be brought into operation and connected to the input via a capacitor having a few picofarads. The VTVM can then be connected to the collector of each stage and the individual resonant circuits aligned for maximum. The S-meter will show a deflection although the control circuit is still not operative. The operation of the control circuit can only be checked when the voltage on the control voltage line has been removed. The control voltage can be measured using a high-impedance DC voltmeter. The control voltage amounts to approximately 10 V under no-signal conditions. With strong signals, the voltage will fall to approximately +1 V. The control characteristic can be influenced with the variable resistor R 33 connected in series with the S-meter. If the resistance is too small, the receiver will be over-controlled. This means that strong signals will produce less AF power than weak ones. If on the other hand, the resistor is too great, a limiting effect will occur even on weak stations and cause the audio frequency to be distorted. The final value of resistor R 33 must be found by experiment since it is dependent on the transistor used and on the impedance of the S-meter. Finally, the BFO is coupled in and an AF amplifier connected to the product detector. If a hum or other interference is audible, it will be necessary for the power supply filtering to be improved by connecting a 500 μF/15 V capacitor (C 45) between the filter resistor R 53 and ground. After the module has been aligned, it is possible to connect it to the other modules and the operating controls.

6. MEASURED VALUES

The following values were measured by the author:

6.1. TRANSMIT PORTION

Output voltage in the A 3 J mode (SSB) : 0.5 V rms
RF passband : 300 to 3000 Hz
Passband ripple : less than 3 dB
Carrier suppression : approx. 50 dB
Linear crosstalk (sideband suppression) at 1000 Hz using the XF-9A filter : approx. -40 dB
Non-linear crosstalk (intermodulation) : approx. -46 dB

6.2. RECEIVE PORTION

Sensitivity : 10 μV for 10 dB S/N
Control range : 60 dB
Control fall time : approx. 1 s
Band width : according to the filter

The 9 MHz receiver input is not decoupled by a buffer stage from the crystal filter. When connecting a receive converter, it is therefore necessary for the crystal filter matching to be maintained. If the crystal filter is incorrectly matched, this will cause an increase of the passband ripple.

7. NOTES

Further details regarding the control elements and components not accommodated on the described printed circuit boards will be given after describing the further modules.

8. AVAILABLE PARTS

The printed circuit boards DL 6 HA 001 to DL 6 HA 009, the coil formers and the trimmer capacitors as well as complete kits of parts and the crystal filter are available from the publishers or their national representatives. Please see advertising page.

9. REFERENCES

(1) G. Laufs : A SSB Transceiver with Silicon Transistor Complement
Part 1: The 144 MHz Converter
VHF COMMUNICATIONS 2 (1970), Edition 1, Pages 1-11

STABLE REFERENCE VOLTAGES

by H.-J. Franke, DK 1 PN

INTRODUCTION

A great number of variable frequency oscillators are now in use on the two metre band. A large number of these oscillators use varactor diodes for tuning; the tuning voltage for these varactor diodes must be kept extremely constant. If one assumes that all other points that affect the stability of a VFO have been taken into consideration, the temperature response of the reference voltage element and thus of the tuning voltage remains. Since little has been published in this respect for the radio amateur, the author would like to describe the design and construction of reference voltage elements that are stable in both the voltage and temperature sense.

1. REQUIRED VOLTAGE STABILITY FOR VARACTOR-TUNED VFOs

In order to obtain an idea of how constant the voltage must be for varactor-tuned VFOs for use on the two metre band, this is to be described in conjunction with a 8 MHz VFO:

The frequency range of this oscillator is from 8.0 MHz to 8.12 MHz; the output frequency of this oscillator is multiplied by 18 to obtain a frequency in the two metre band. Assuming, for instance, that a voltage variation of 4.5 V is needed to obtain the frequency variation of 120 kHz, a voltage variation of only 1 mV will cause a frequency shift of 27 Hz. After considering the required frequency multiplication, this will result in a frequency variation of 480 Hz/mV on the two metre band.

As can be calculated, a 72 MHz VFO is no better in this respect. This value may not seem considerable, however, the following considerations will show how difficult it is to keep the reference voltage source as constant as this.

2. ZENER DIODES

Before it is possible to discuss the means of reducing the voltage fluctuation of reference voltage sources, it is necessary that the characteristics of zener diodes themselves are examined. Figure 1 shows a typical characteristic curve of a zener diode. A zener diode operates in the reverse-bias range and the operating point is to be found somewhere on the steep portion of the characteristic curve. Two important characteristics can be observed from this illustration: Firstly that the characteristic curve is not vertical and secondly that it is varied as a function of temperature. The first point means that the zener voltage U_Z also varies as a function of the zener current I_Z. This is characterized by the so called differential resistance r_Z. The more diagonal the characteristic curve, the greater will be the voltage variations ΔU_Z as a result of current fluctuations ΔI_Z.

The second point means simply that the zener voltage U_Z also varies as a function of the temperature T. The magnitude of this dependence is characterized by the temperature coefficient TC. Table 1 gives examples based on a popular zener diode series that are to be explained in the following sections.

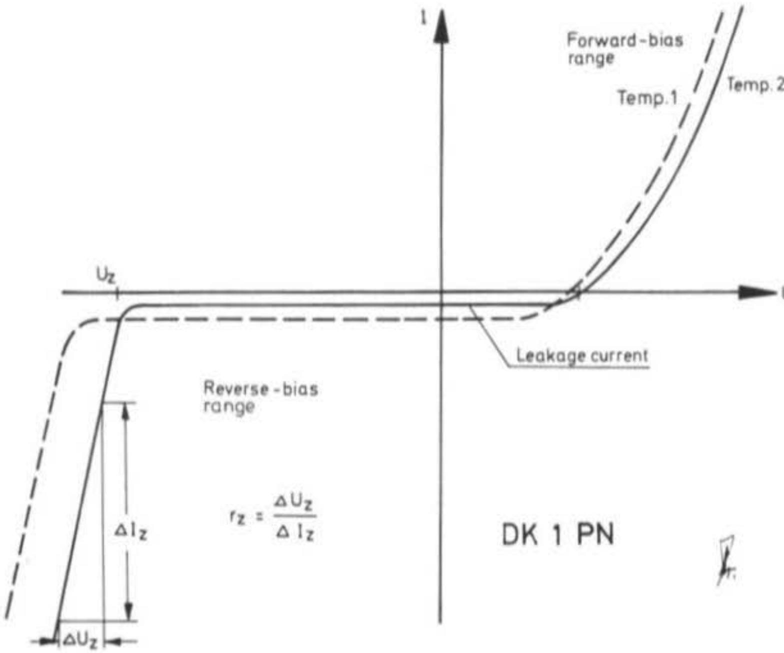


Fig. 1: Principle characteristic curve of zener diodes at two different temperatures

3. THE EFFECT OF OPERATING VOLTAGE FLUCTUATIONS

From the two previously mentioned, unwanted characteristics of zener diodes, it is the variation of U_z as a function of fluctuations of the operating voltage U_B that is to be examined firstly. Various circuits are to be described which allow this to be avoided as far as possible.

3.1. SIMPLE STABILIZATION CIRCUITS

Figure 2 shows the basic circuit for voltage stabilization using zener diodes. For our examination, only the residual voltage fluctuation as a function of input voltage variations is to be calculated:

$$\Delta U_z = \frac{\Delta U_B \times r_z}{R_d}$$

Practical examples are now to be given to show the efficiency of the following circuits using this calculation:

<u>Example 1:</u> The circuit according to Fig. 2 has the following values:

$U_B = 30\ V \pm 10\%$, $R_d = 3\ k\Omega$, $U_z = 15\ V$, $r_z = 25\ \Omega$
(from Table 1 for BZY 85/C15)

This allows the voltage fluctuation ΔU_z to be calculated at an operating voltage variation of 10% (= 3 V) :

$$\Delta U_z = \frac{3 \times 25}{3000} = 25 \times 10^{-3}\ V = \underline{25\ mV}$$

Type	U_z Volt	r_z Ohm	TC %/°C	TC mV/°C
BZY 85/C3V6	3.4 to 3.8	70 < 80	-0.6	-20.4 to -22.8
C3V9	3.7 to 4.1	60 < 80	-0.45	-16.6 to -18.4
C4V3	4.0 to 4.6	55 < 75	-0.25	-10.0 to -11.5
C4V7	4.4 to 5.0	50 < 70	-0.04	- 1.76 to - 2.0
C5V1	4.8 to 5.4	43 < 65	-0.025	- 1.20 to - 1.35
C5V6	5.2 to 6.0	32 < 55	-0.0003	- 0.15 to - 0.18
C6V2	5.8 to 6.6	16 < 35	+0.015	+ 0.87 to + 0.99
C6V8	6.4 to 7.2	4.5 < 8	+0.030	+ 1.92 to + 2.16
C7V5	7.0 to 7.9	2.0 < 7	+0.040	+ 2.80 to + 3.16
C8V2	7.7 to 8.7	2.8 < 7	+0.047	+ 3.62 to + 4.10
C9V1	8.5 to 9.6	4.7 < 10	+0.054	+ 4.60 to + 5.20
C 10	9.4 to 10.6	7 < 15	+0.059	+ 5.55 to + 6.25
C 11	10.4 to 11.6	10.5 < 20	+0.063	+ 6.55 to + 7.30
C 12	11.4 to 12.7	15 < 25	+0.066	+ 7.52 to + 8.38
C 13	12.5 to 14.0	20 < 30	+0.068	+ 8.50 to + 9.52
C 15	13.8 to 15.5	25 < 35	+0.070	+ 9.66 to +10.85
C 16	15.3 to 17.0	30 < 40	+0.071	+10.85 to +12.10
C 18	16.8 to 19.0	35 < 45	+0.072	+12.1 to +13.7
C 20	18.8 to 21.0	40 < 50	+0.073	+13.7 to +15.3
C 22	20.8 to 23.0	45 < 55	+0.074	+15.4 to +17.0

Table 1: Data of the zener diode series BZY 85 (AEG–Tfk); all values are based on I_Z = 5 mA

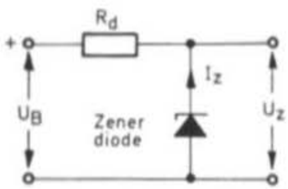


Fig. 2: Basic circuit for voltage stabilization using zener diodes

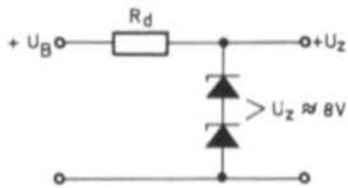


Fig. 3: Series-connection of zener diodes

This shows that it is not advisable to use such a simple circuit for stabilization of the operating voltage for a tuning diode, since the possible operating voltage variation of 10% is probably too optimistic. It is therefore to be examined if an improvement can be made by increasing the operating voltage and thus the value of the dropper resistor R_d.

Example 2: The circuit according to Fig. 2 has the following values:

U_B = 100 V ± 10%, R_d = 17 kΩ, all other values as in Example 1.

The voltage fluctuations ΔU_z at an operating voltage variation of 10% (= 10 V) now amount to:

$$\Delta U_z = \frac{10 \times 25}{17000} = 14.7 \times 10^{-3} \text{ V} = \underline{14.7 \text{ mV}}$$

It will be seen that a noticeable improvement has been achieved, however, it is still not satisfactory. Furthermore, an operating voltage of 100 V is not practical for battery driven equipment.

3.2. SERIES CONNECTION OF ZENER DIODES

After studying Table 1, zener diodes will have been found that possess an especially low differential resistance r_z. This is due to physical-technological considerations and is therefore independent of the manufacturer. It may seem advisable, to obtain higher zener voltage values by series connection of several zener diodes with voltages in the order of 8 V. Figure 3 shows such a circuit; Example 3 shows the effects of this.

Example 3: The circuit shown in Fig. 3 has the following values:

U_B = 30 V ± 10%, R_d = 3 kΩ, U_z = 15 V, r_z = 2 Ω + 2 Ω = 4 Ω
(Taken from Table 1 for two BZY 85/C7V5 diodes in series)

At an operating voltage variation of 10% (= 3 V) the following variation of the zener voltage is obtained:

$$\Delta U_z = \frac{3 \times 4}{3000} = 4 \times 10^{-3} \text{ V} = \underline{4 \text{ mV}}$$

This improvement is very considerable, however, it is still not sufficient to satisfy the demands made upon it. Furthermore the example is based on the optimistic assumption that two 7.5 zener diodes provide the required voltage, which is not always the case in practice.

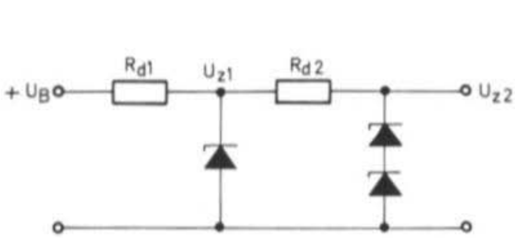


Fig. 4: Double stabilization together with series-connection ot two zener diodes

3.3. DOUBLE STABILIZATION

Figure 4 shows what is meant with the term double stabilization. The operating voltage of the second stabilizing stage (U_{z1}) is already stabilized by the first stabilization stage. Example 4 will explain the degree of stabilization obtained with such a circuit:

Example 4: $U_B = 30\ V \pm 10\%$, $R_{d1} = 800\ \Omega$, $U_{z1} = 22\ V$, $r_{z1} = 45\ \Omega$
(Taken from Table 1 for BZY 85/C22);
$R_{d2} = 1.4\ k\Omega$, $U_{z2} = 15\ V$, $r_{z2} = 4\ \Omega$
(Taken from Table 1 for two BZY 85/C7V5 in series).

$$\Delta U_{z1} = \frac{3 \times 45}{800} = 0.17\ V$$

The variation of U_{z2} is now to be calculated:

$$\Delta U_{z2} = \frac{0.17 \times 4}{1400} = 0.482 \times 10^{-3}\ V = \underline{0.48\ mV}$$

The required independence of operating voltage fluctuations is now obtained with this rather elaborate circuit comprising three zener diodes and double the current consumption (2 x 5 mA).

The result of these examples can also be obtained by analyzing the voltage fluctuation formula: If the value of the dropper resistor (R_d) were to obtain the value infinity, the voltage fluctuation ΔU_z would be zero. This, however, only means that R_d must be replaced by a constant current source. This most favourable solution is to be given in Section 5.

4. THE EFFECT OF TEMPERATURE FLUCTUATIONS

In Section 3, only the effect of operating voltage fluctuations was examined. However, the temperature stability of the voltage is equally important. It should be pointed out that already temperature compensated diodes are available on the market. Some examples are given in Table 2. Unfortunately, the prices of such diode combinations are somewhat expensive (Approximately DM 20,-- to DM 35,-- according to the TC value). The following sections, therefore, describe compensation possibilities that can be achieved by amateur means.

Type	U_z Volt	P_{tot} mW at 45 °C	I_z mA	r_z Ohm	TC mV/°C
BZX 51	8.6 ± 0.4	250	10	9	± 0.086
BZX 52	"	250	10	9	± 0.043
BZX 53	"	250	10	9	± 0.017
BZX 54	"	250	10	9	± 0.008
BZY 22	8.4 ± 0.4	200	5 ± 0.5	15 25	± 0.084
BZY 23	"	200	"	"	± 0.042
BZY 24	"	200	"	"	± 0.017
BZY 25	"	200	"	"	± 0.008

Table 2: Two series of temperatur compensated zener diodes. BZX 51 to 54 : AEG-Telefunken, BZY 22-25: ITT-Intermetall

4.1. ZENER DIODES WITH OPPOSITE TC VALUES

Table 1 shows that the TC value of zener diodes is sometimes positive and sometimes negative. This is also dependent on physical considerations. The TC value itself is dependent on the operating point. The reversal point between positive and negative TC values (TC $\approx$ 0) is approximately 5.6 V. However, this value is not the same for each technology, which means that it varies somewhat from manufacturer to manufacturer.

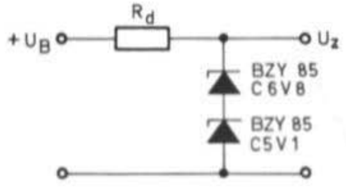


Fig. 5: Simple circuit for TC compensation of zener diodes

Two zener diodes having opposite TC values can be connected in series to obtain temperature compensation. Fig. 5 shows such a circuit. Examples 5 and 6 are to show how great the voltage fluctuations caused by a temperature variation of only 1 °C are.

> Example 5: The 15 V zener diode used in the circuit shown in Fig. 2 possesses, according to Table 1, a TC value of = +0.07 %/°C, or when expressed in mV: TC = +10.5 mV/°C. This means that the zener voltage will increase by 10.5 mV for each degree (°C) of temperature increase.

With the given voltage dependence of the VFO frequency given at the commencement of this article of 480 Hz/mV, it is easily possible to calculate that a temperature variation of only 1 °C would result in a frequency variation of 5 kHz.

> Example 6: The total TC value of the circuit shown in Fig. 5 is:
>
> BZY 85/C6V8: TC = +0.03 %/°C = +2.04 mV/°C
> BZY 85/C5V1: TC = -0.025 %/°C = -1.28 mV/°C
>
> U_z = 11.9 V: TC_{tot} : +0.76 mV/°C

This TC value is extremely good for amateur means even when the industrial TC compensated zener diodes possess values one or two orders of magnitude better (Table 2). Unfortunately, this circuit limits the user to a zener voltage of approximately 11.9 V.

For later discussion, the value of the differential resistance r_Z is to be calculated for the circuit given in Fig. 5 :

According to Table 1, the following will result: r_{Ztot} = 4.5 Ω + 43 Ω = 47.5 Ω

4.2. TEMPERATURE COMPENSATION USING DIODES

The fact that all germanium and silicon diodes exhibit a negative TC value in the forward voltage range, means that they can be used for temperature compensation of zener diodes having a positive TC value ($U_Z \gtrless 6.2$ V).

The TC value of diodes is dependent on the forward current I_F. The individual values can be determined from Figure 6.

When only considering the temperature compensation, it will be seen that all diodes can be used. Since, however, the voltage dependence examined in Section 3 must also be considered, the differential resistance values r_Z are given for some diode types used as examples. The table shows that germanium demodulation diodes are, due to their high differential resistance r_Z, unsuitable. On the other hand, germanium gold-bonded diodes, such as OA 180, OA 182 and AA 135 are more favourable. Practically all silicon junction and planar diodes are suitable.

Diode type		r_Z at I_F = 5 mA
BAY 67	(Silicon)	9.5 Ω
BZ 102	(Silicon)	13 Ω
BZY 87	(Silicon)	8 Ω
OA 182	(Germanium)	13 Ω
AA 111	(Germanium)	100 Ω

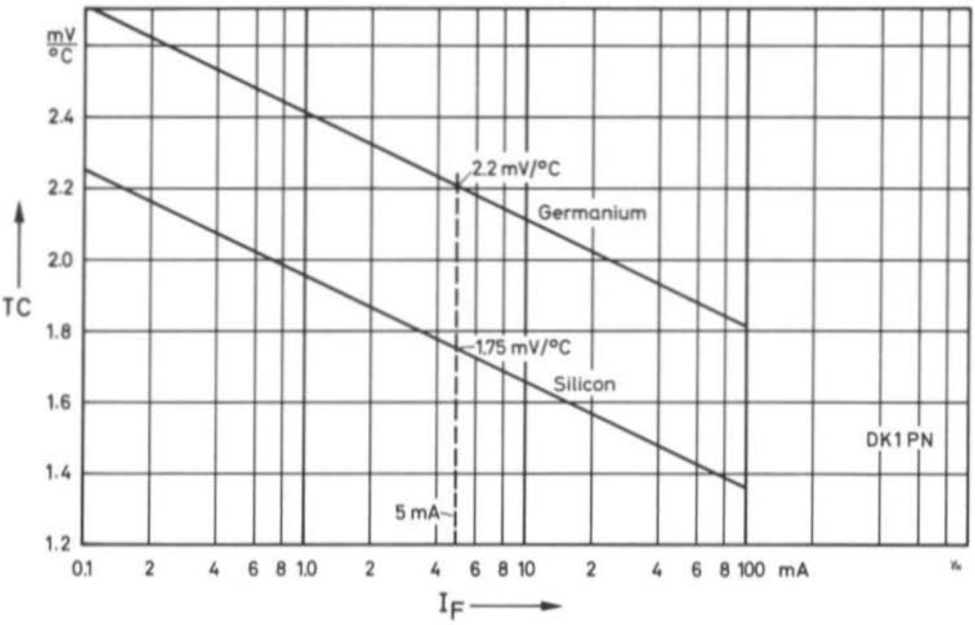


Fig. 6: Temperature coefficient of diodes as a function of the forward current I_F

From the great number of diode combinations that can be used for TC compensation, three possibilities have been chosen for the following examples:

Example 7: The circuit shown in Fig. 7 operates in conjunction with the zener diode BAY 85/C6V8 and the silicon diode BAY 67. The dropper resistor R_d allows a current of $I_z = 5$ mA to flow. The TC value of the compensating diode is, according to Fig. 6, therefore TC = -1.75 mV/°C. The TC value of the zener diode is taken from Table 1 which results in a total value of

$$\begin{array}{rr} & +\ 2.04\ \mathrm{mV/^oC} \\ & \underline{-\ 1.75\ \mathrm{mV/^oC}} \\ TC_{tot} = & \underline{+\ 0.29\ \mathrm{mV/^oC}} \end{array}$$

The total differential resistance amounts to:

$$r_{z\ tot} = 4.5\ \Omega + 9.5\ \Omega = \underline{14\ \Omega}$$

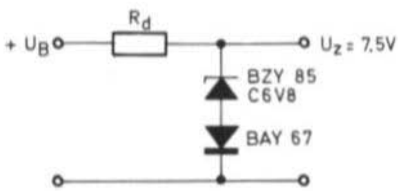


Fig. 7: TC compensation using a silicon diode

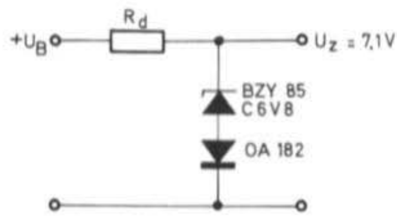


Fig. 8: TC compensation using a germanium diode

Example 8: The circuit shown in Fig. 8 operates according to the same considerations as that given for Fig. 7. However, in this case, a germanium diode OA 182 is used for temperature compensation. According to Fig. 6, the TC value is TC = -2.2 mV/°C. The reference voltage source therefore has a TC value of :

$$\begin{array}{rr} & +\ 2.04\ \mathrm{mV/^oC} \\ & \underline{-\ 2.20\ \mathrm{mV/^oC}} \\ TC_{tot} = & \underline{-\ 0.16\ \mathrm{mV/^oC}} \end{array}$$

The differential resistance of this circuit combination amounts to:

$$r_{z\ tot} = 4.5\ \Omega + 13\ \Omega = \underline{17.5\ \Omega}$$

Example 9: The third compensation circuit, shown in Fig. 9, uses one germanium and one silicon diode together with a 8.2 V zener diode. As in all other examples, the dropper resistor R_d is also dimensioned so that the zener current I_z is 5 mA. The TC values given in the previous examples are therefore still valid.

This combination achieved:

$$TC_{tot} = \underline{-\ 0.1\ \mathrm{mV/^oC}}$$

$$r_{z\ tot} = 2.8\ \Omega + 9.5\ \Omega + 13\ \Omega = \underline{25.3\ \Omega}$$

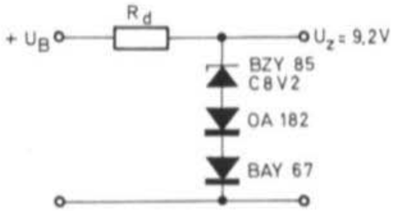


Fig. 9: TC compensation using a silicon and germanium diode

The differential resistance values given in all examples show that the described compensation circuits result in a high differential resistance. They therefore have an unfavourable behaviour with respect to fluctuations of the operating voltage. In order to achieve both temperature compensation and a high independence of operating voltage fluctuations, the dropper resistor R_d must be replaced by a constant current source as was already mentioned in Section 3.

5. CONSTANT CURRENT SOURCES

Constant current sources can be easily realized using pentodes, transistors or field effect transistors. The principle is based on the high impedance of these components. The effect of this is that a constant anode (collector or drain) current will flow if the grid (base or gate) voltage is kept constant. The current is therefore virtually independent of the anode (collector or drain) voltage. Prerequisite, however, is that the voltage is greater than the residual voltage.

These relationships are graphically shown by the family of curves given in Fig. 10. The impedance value is characterized by the slope of the characteristic curve. The flatter the curve, the higher will be the impedance. Parameter for this is the voltage to be stabilized at the control electrode (grid, base, gate). The values are valid for junction field effect transistors where the residual voltage is termed "pinch-off voltage" and amounts to approximately 2 V.

A constant current source is to be described that is equipped with a field effect transistor. In principle, any of the characteristic curves given in Fig. 10 could be used. However, the transistor is operated at a gate-source voltage U_{GS} of 0 V in order to avoid having to generate an additional stabilized voltage for the gate (Fig. 11). However, the drain current fluctuates greatly between individual transistors. In order to achieve a certain current suitable for the reference voltage source, a resistor will be required.

In addition to this, the drain current of the field effect transistor is temperature dependent (as with all other semiconductors). This, however, can be easily compensated using a diode. Such a circuit is given in Fig. 12. The compensation diode alters the gate voltage, which in turn reduces the drain current. This is easily compensated using the trimmer resistor.

DK1PN

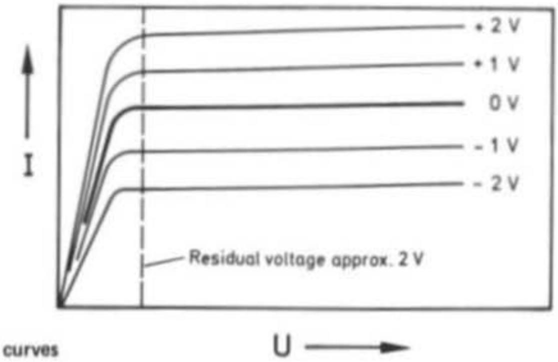


Fig. 10: Principle family of characteristic curves

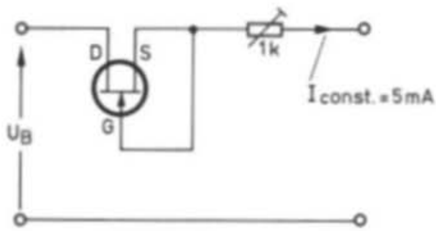


Fig. 11: Principle of a constant current source using a junction FET

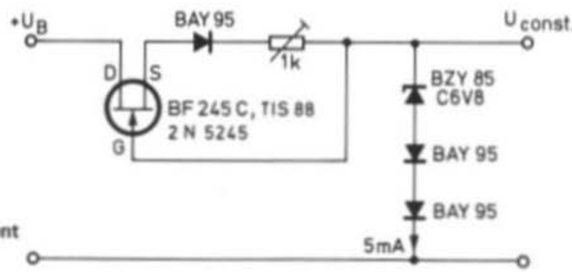


Fig. 12: A reference voltage source that is TC compensated and independent of operating voltage fluctuations

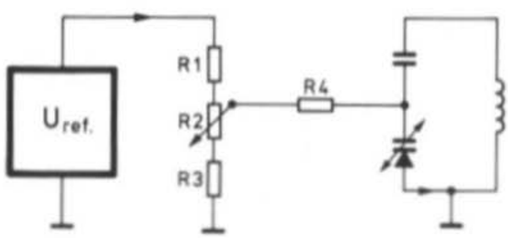


Fig. 13: Principle of a varactor tuning circuit

5.1. MEASUREMENTS MADE ON A COMPLETED REFERENCE VOLTAGE SOURCE

The current of the reference voltage source is adjusted to 5 mA. The temperature coefficient of the zener diode was compensated using two silicon diodes BAY 95 manufactured by AEG-Telefunken (can be replaced with 1 N 914). Although on paper, only one diode should be sufficient to obtain compensation, different TC values were observed in practice for the zener diode than were given in Table 1. The following measured values finally resulted:

> Fluctuation of the reference voltage on varying the operating voltage between 12 and 25 V: None could be determined on a four position digital voltmeter. Temperature coefficient of the reference voltage in the temperature range of -40 °C to +50 °C: TC = -0.5 mV/°C.

6. APPLICATION OF THE REFERENCE VOLTAGE SOURCE

The high-impedance voltage source according to Figure 12 can unfortunately not be loaded to any great extent. This is, of course, not a disadvantage when using the source in conjunction with varactor diodes because a high-resistance potentiometer can be used. The potentiometer should be a high-quality type having a constant TC value over the whole resistive surface as well as allowing a very fine adjustment; helical type potentiometers are very favourable. It is not possible to compensate the temperature response of the tuning diode (varactor) with amateur means. It is therefore advisable for it to be located together with the other frequency-determining components of the VFO in some form of constant-temperature device.

Finally, a few points are to be mentioned in conjunction with Figure 13 that affect the dimensioning:

The current flowing via the voltage divider should, firstly not load the reference voltage noticeably, but should, secondly, be approximately ten times greater than the highest reverse current of the varactor diode. A value of 1 μA has therefore been chosen as the voltage divider current. This is obtained with: R 1 + R 2 + R 3 = 8.2 MΩ. Resistor R 4 is only used as a RF choke; a suitable value is 100 kΩ.

A UNIVERSAL VHF-UHF TRANSMITTER FOR AM AND FM

by R. Lentz, DL 3 WR

1. INTRODUCTION

The following article is to describe a transistorized transmitter which can be switched for VFO or crystal-controlled operation. The transmitter can also be switched to amplitude or frequency modulation and possesses, in addition, varactor triplers for 432 MHz and 1296 MHz. The required operating voltage is in the range of 10 V to 13.5 V for AM or between 10 V and 18 V for FM and telegraphy. The 145 MHz power amplifier provides a RF carrier output power of approximately 1.7 W at an operating voltage of 12 V. At 18 V, the output will increase to approximately 3.5 W. The varactor triplers for UHF exhibit efficiencies of somewhat over 50% which means that an output power of about 1 W (2 W at 18 V) is available at 432 MHz and 0.5 W (1 W at 18 V) at 1296 MHz.

For portable operation, the transmitter can be operated from nine series-connected 1.5 V batteries (60 mm long, 33 mm dia.) which will still not be exhausted after eight hours of intensive contest operation.

For home station operation, the author uses a stabilized power supply. At an operating voltage of 18 V, the described transmitter provides sufficient output to drive a 4 X 150 in a resonant line PA stage for 145 MHz (1) fully into class C operation. If 70 cm operation is required, a tripler, with or without amplifier stage equipped with the tube EC 8020 (2) can be used. Such a circuit will provide peak-power levels of 10 W in the FM and CW modes. This power is sufficient to drive a 4 X 150 resonant line PA stage for 432 MHz (3).

The authors VHF-UHF transmitter (Fig. 1) weighs 2 kg (with cover but without batteries, microphone and coaxial relay). The minus pole of the operating voltage is grounded.

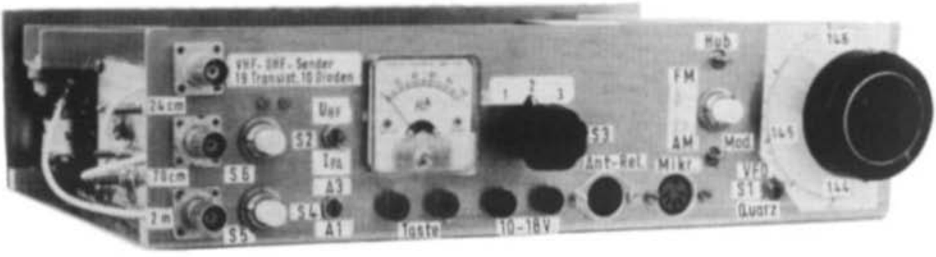


Fig. 1: Front view of the VHF-UHF transmitter

2. CIRCUIT DESCRIPTION

The VHF-UHF transmitter consists of six sub-assemblies, which are joined together according to Fig. 2 to form the complete transmitter. The heart of this transmitter is the printed circuit board DL 3 WR 003 which accommodates the five-stage transmitter and the three-stage AM modulator complete with modulating transformer.

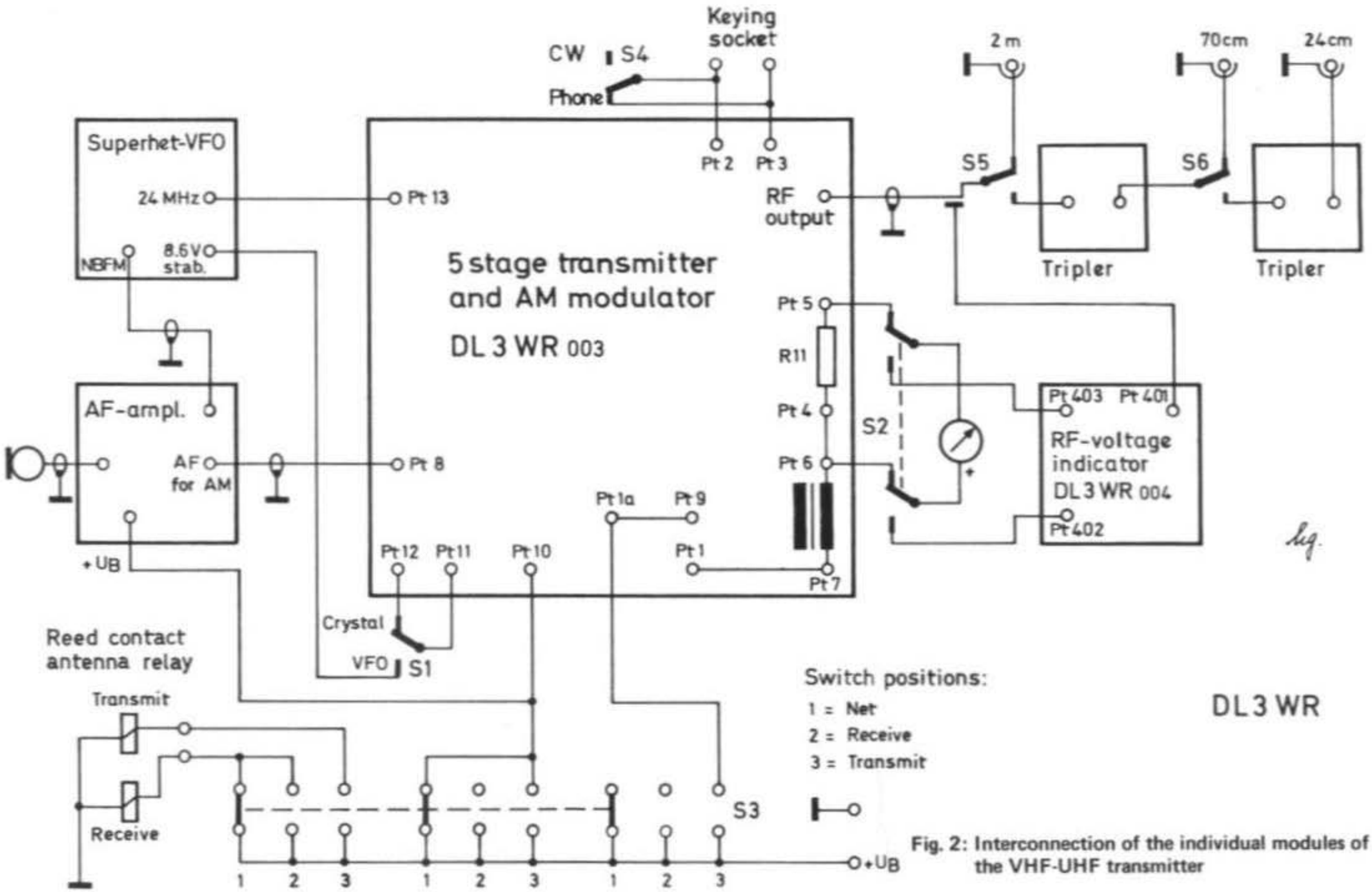


Fig. 2: Interconnection of the individual modules of the VHF-UHF transmitter

The variable frequency oscillator (VFO) provides an output frequency of 24.0 MHz to 24.34 MHz. This VFO is also provided with an FM accessory.

The AF preamplifier is equipped with an active audio filter as described in (4) which provides an attenuation of 6 dB at 320 Hz and 4 kHz. A change-over switch for "AM-FM" switches the output of this filter to a level potentiometer for each of the two modulating modes.

A home-made wire-wound resistor (R 11) is provided so that the collector current of the final transistor can be measured as the voltage drop across it. The meter can also be switched to allow indication of the RF voltage. This is made using the printed circuit board DL 3 WR 004, which possesses a resonant circuit comprising a printed inductance, a demodulator diode and a RF filter.

Two frequency triplers equipped with the varactor diode type BAX 11 (AEG-Telefunken) are connected to the output of the transmitter. Two push-button switches suitable for VHF are used so that the output power is available in two metre, 70 cm, or in the 23 cm band.

A reed-contact relay as described in (5) is used for the antenna change-over switching. It is provided with voltage from the transmitter using an additional plug and socket, so that it can be used in the appropriate antenna cable.

The following sections are to give detailed descriptions of the individual sub-assemblies.

3. THE INDIVIDUAL SUB-ASSEMBLIES

3.1. TRANSMITTER AND MODULATOR BOARD DL 3 WR 003

Figure 3 shows the circuit diagram of the five-stage, crystal-controlled two metre transmitter (transistors T 1 - T 5) and the frequency tripler stage with transistor T 6 for VFO operation as well as a voltage stabilizer with transistor T 7 and a three-stage collector-voltage modulator comprising transistors T 8 to T 11. These stages are accommodated on the printed circuit board DL 3 WR 003. The RF voltage indicator, whose circuit is also given in Figure 3, is built up on a separate small printed circuit board (DL 3 WR 004).

3.1.1. CIRCUIT DETAILS

The crystal-controlled transmitter represents a further development of the two metre transmitter UTS 5 (6) which provided an output power of 2 W at an operating voltage of 12 V. The crystal-controlled oscillator (T 1) and the 72 MHz buffer stage (T 2) are equipped with high gain transistors of the 2 N 918 type instead of the originally used 2 N 708 types. This means that the feedback and the coupling need not be so tight and that the RF power required for the doubler stage (T 3) can be achieved at a far lower input power. In addition to this, the crystal holder capacitance is neutralized by inductance L 1.

A further electrical modification to the UTS 5 transmitter is the use of the transistor 40290 in the frequency doubler stage (T 3) instead of the more expensive 2 N 3553.

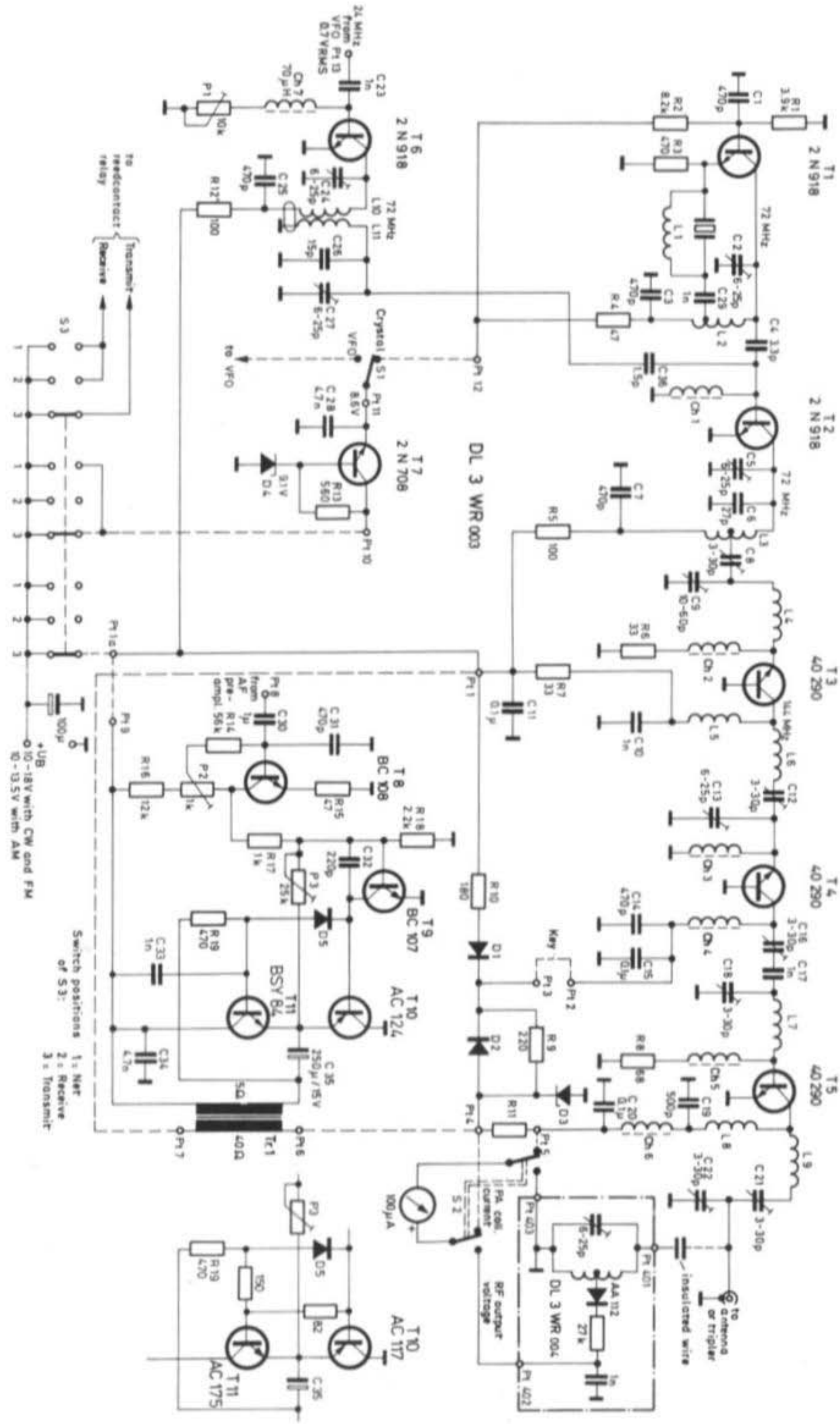


Fig. 3: Circuit diagram of the 2 metre transmitter DL 3 WR 003 for VFO and crystal-controlled operation; with modulator and RF-indicator (DL 3 WR 004)

Furthermore, the modulation of the driver stage is made using a diode-resistor network (D 1, D 2, R 9, R 10) in order to avoid the tap on the modulating transformer. This circuit is based on information given in the data sheet for transistors 40290 - 40292 (RCA).

Further circuit details were given in more detail for the transmitter UTS 5 in (6). It should be noted, however, that some of the component numbers may have changed.

Since the transmitter was also to be used from a 24 MHz VFO, an additional frequency tripler (T 6) was required. The high-gain transistor type 2 N 918 is also used in this stage so that only low RF and DC power levels are required. An input resonant circuit is not provided because the author assumes that all variable frequency oscillators possess output resonant circuits. This stage is provided with a bandpass filter for the tripled frequency (inductances L 10, L 11). An inductive coupling has been chosen because a coupling capacitor would provide an increasing low impedance path for higher order harmonics. Since the RF power levels from the crystal controlled stage and the tripler are sufficiently great, it is possible for both stages to be connected together at high impedance (capacitors C 4 and C 36). The low capacitance values ensure that interaction onto the tuning is small. This means that the switching between crystal-controlled and VFO operation can be made easily by switching the operating voltage. This is achieved by switching the stabilized voltage of 8.6 V provided from the voltage stabilizer circuit, equipped with the transistor T 7. The stabilized voltage then feeds either the crystal-controlled oscillator or the VFO.

The principle of the three-stage collector voltage modulator is known as a ferrous-free audio amplifier equipped with the complementary transistor pair AC 117/AC 175. After several failures caused by overvoltages, the transistor AC 117 has been replaced by the higher breakdown-voltage type AC 124, and the type AC 175 by the silicon NPN transistor BSY 84. The circuit for the base bias voltage for the complementary output pair is simplified in this manner; however, it is necessary for a certain high frequency blocking to be made (capacitors C 33, C 34). The audio output power is less than when using the transistor AC 175 due to the higher residual voltage of the silicon transistor. However, the power is just sufficient to fully modulate the transmitter. This means that an overload of the modulator will not cause overmodulation, but only a higher AF harmonic distortion factor.

The silicon transistor T 11 must have a permissible collector current of 1 A and a current amplification of at least 40 at this current value. Transistors having too low a current amplification at peak collector current will not be driven sufficiently from the driver stage, which means that the maximum output power will not be reached. It is, of course, possible to use the original modulator circuit, which is the reason why it is also given in Figure 3.

In order to save space and weight, as well as to have all components on the main printed circuit board, a small type C core is used for the modulation transformer (Tr 1). Of course, a normal transformer using dynamo sheet can also be used and mounted beside the printed circuit board. The specifications for these two transformer types are given in Section 3.1.3.

3.1.2. MEASURED VALUES FOR THE SUB-ASSEMBLY DL 3 WR 003

U_B	8.6 V stab.	10 V	12 V	13.5 V	15 V	18 V
T 1	3.5					
T 2		12	13	14	15	16
T 3		30	35	40	43	47
T 4		36	60	80	100	135
T 5		150	190	220	240	310
T 6		5.7	6	6.2	6.5	7.3
D 4		1	5	8	12	15
I_{rmod}		14	16	17	18	20
I_{tot}		250	330	390	450	560

Table 1: Orientation values of the individual stages when adjusting for optimum amplitude modulation with 50 Ω termination.

The current values of driver and PA stage transistors T 4 and T 5 can greatly differ from the values given in Table 1 if tuning and drive power are different or if the load deviates from 50 Ω. $I_{r\ mod}$ is the quiescent current of the whole three-stage modulator; I_{tot} is the total current of the transmitter, where the current values for VFO and audio preamplifier are included. U_B is the operating voltage.

Transistors	Transformer Tr 1	U_B V	$I_{r\ mod}$ mA	I_{max} mA	$P_{AF\ max}$ into 40 Ω W	efficiency %	AF input voltage mV
AC 117, AC 175	without	12	17	350	2.7	64	
" "					(into 4.7 Ω)		
" "	normal	12	17	330	2.0	50	
" "	type C	12	17	250	1.7	56	30
AC 124, BSY 84	normal	12	16	260	1.0	32	
" "	type C	12	16	240	1.25	43	

Table 2: Values measured on the AM speech amplifier

The values of the output power are obtained at a magnetic bias current of 300 mA. The audio power of 1.25 W amounts to exactly half the input power of transistor T 5 at 12 V. The half-power points of the three-stage modulator when using a type C core transformer are at 150 Hz and 15 kHz. The audio bandwidth of the whole transmitter is determined by the active audio filter in the preamplifier.

3.1.3. SPECIAL COMPONENTS FOR SUB-ASSEMBLY DL 3 WR 003

T 1, T 2, T 6: 2 N 918 (RCA), BFX 73 (SGS)
T 3, T 4, T 5: 40290 (RCA)
T 7: 2 N 708 or similar silicon NPN small power transistor
T 8, T 9: BC 107/BC 108, BC 147/BC 148, BC 182/BC 183, 2 N 3903, 2 N 3904
T 10: AC 117, AC 124 (AEG-Telefunken), AC 188 K (Siemens or Philips)
T 11: AC 175 (AEG-Telefunken), AC 187 K (Siemens or Philips) or
BSY 84, BSY 86 (ITT-Intermetall), BFX 97 (SGS), 2 N 2219A
D 1, D 2: BAY 86 (AEG-Telefunken), 1 N 2858 (RCA), (any Silicon rectifier diode for 50 V/200 mA)
D 3: BZY 92/C33 (AEG-Telefunken), ZD 33, ZM 33 (ITT-Intermetall)
1 N 3032 A, 1 N 4174 A (33 V/ 1 W zener diode)
D 4: BZY 85/C9V1, BZY 85/D10, OA 126/10 (AEG-Telefunken), 1 N 1932
(9.1 V or 10 V zener diode)
D 5: BZY 87, 1 N 914 or similar silicon diode

Ch 1, Ch 2, Ch 6: 6 mm dia. ferrite core, 10 mm long with 6 axial holes: 2.5 turns through the ferrite (Z = 850 Ω)
Ch 3, Ch 5: as Ch 1 but only 1.5 turns (Z = 450 Ω)
Ch 4: approx. 1.5 μH; wound on the SW core of a 6 mm dia. coil former; the thread is fully filled with 0.5 mm dia. (24 AWG) enamelled copper wire (approx. 16 turns)
Ch 7: Wideband ferrite choke of 30 to 70 μH

L 1: 10 turns of 0.6 mm dia. (23 AWG) wound on a 6 mm former, self-supporting, coil length 10 mm

All further coils are from 1 mm dia. (18 AWG) silver-plated copper wire, self-supporting.

L 2: 6 turns on a 8 mm former, tap two turns from ground end
L 3: 4 turns on a 8 mm former, coil length 13 mm with centre tap
L 4: 5 turns on a 8 mm former, coil length 13 mm
L 5: 3 turns on a 5 mm former, coil length 15 mm
L 6, L 7: 2 turns on a 8 mm former, coil length 6 mm
L 8: 3 turns on a 5 mm former, coil length 15 mm
L 9: 3.75 turns on a 8 mm former, coil length 13 mm
L 10, L 11: 7 turns on a 8 mm former, coil length 20 mm, coupling link: enlongated O from insulated copper wire between the 1st and 2nd turn from the cold end. Inserted approx. half way.

C 2, C 5, C 13, C 24, C 27: 6 - 30 pF ceramic disc trimmers of 10 mm dia.
C 8, C 12, C 16, C 18, C 21, C 22: 3 - 30 pF air-spaced (Philips) trimmers
C 9: 10 - 60 pF ceramic disc trimmer of 10 mm dia.

Modulation transformer Tr 1 :

Type C core: SM 30b; sheet material: Permalloy N 2. With air gap. (1 x scotch-film). Suitable for PC-board mounting (available from the publishers)

Primary: 75 turns of 0.4 mm dia. (26 AWG) enamelled copper wire
Secondary: 225 turns of 0.4 mm dia. (26 AWG) enamelled copper wire
Conventional type transformer using transformer laminated plates (EI 48) with air gap.
Primary: 65 turns of 0.6 mm dia. (23 AWG) enamelled copper wire
Secondary: 195 turns of 0.5 mm dia. (24 AWG) enamelled copper wire

3.1.4. MECHANICAL ASSEMBLY OF DL 3 WR 003

The two metre transmitter shown in Figure 3 is, with exception of the PA stage comprising transistor T 5 and the RF voltage indicator (DL 3 WR 004), accommodated on a printed circuit board having the dimensions 210 mm by 80 mm. This printed circuit board is 5 mm shorter but 35 mm wider than the printed circuit board of the two metre transmitter UTS 5 (DJ 1 NB 004). Printed circuit board DL 3 WR 003 is illustrated in Figure 4; the corresponding component location plan is given in Figure 5.

The cases of transistors T 3 and T 4 must be directly adjacent to the printed circuit board; all other transistors can be soldered into place leaving approximately 5 mm long connections. Transistor T 4 is provided with cooling fins. Ceramic tubular capacitors are used for bypassing; the connections of these capacitors should be as short as possible. The capacitance values should not be less than 390 pF and not more than 1 nF.

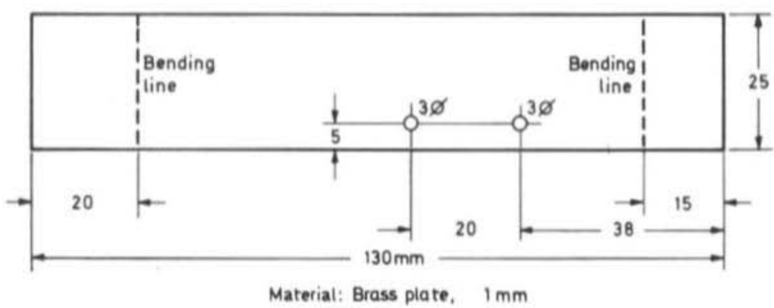


Fig. 6: Cooling and screening plate for the AM modulator

The two circuit possibilities for the modulator output stage are given in the component location plan. The germanium transistors are enclosed in a rectangular heat sink which is screwed to the screening and cooling plate shown in Fig. 6. If a silicon output transistor is used, it should be provided with somewhat flattened cooling fins;an insulated mounting bracket for screw mounting will be more favourable. The heat sink plate itself should be soldered to two solder tags (ground as shown in Figure 5).

If a normal transformer is used instead of the type C core transformer, the non-required corner of the printed circuit board can be removed. This provides enough room to mount the modulation transformer beside the PA stage.

The PA stage equipped with the transistor T 5 is mounted on a brass bracket that has been prepared according to Figure 7. A second bracket, which is also shown in Fig. 7, is used to complete the screening and to mount the transmitter into the cabinet. These metal brackets are smaller than those used in the transmitter UTS 5 (DJ 1 NB 004) and also somewhat simpler. In addition to this, the PA stage has been rotated by 180° and mounted on the printed circuit board in such a manner that the mechanical build-up is on the component side. This means that the conductor side remains free, allowing the printed circuit board to be close-spaced to the cabinet plate. In spite of these modifications, it is possible to use the final amplifier of the transmitter UTS 5 if this should be available.

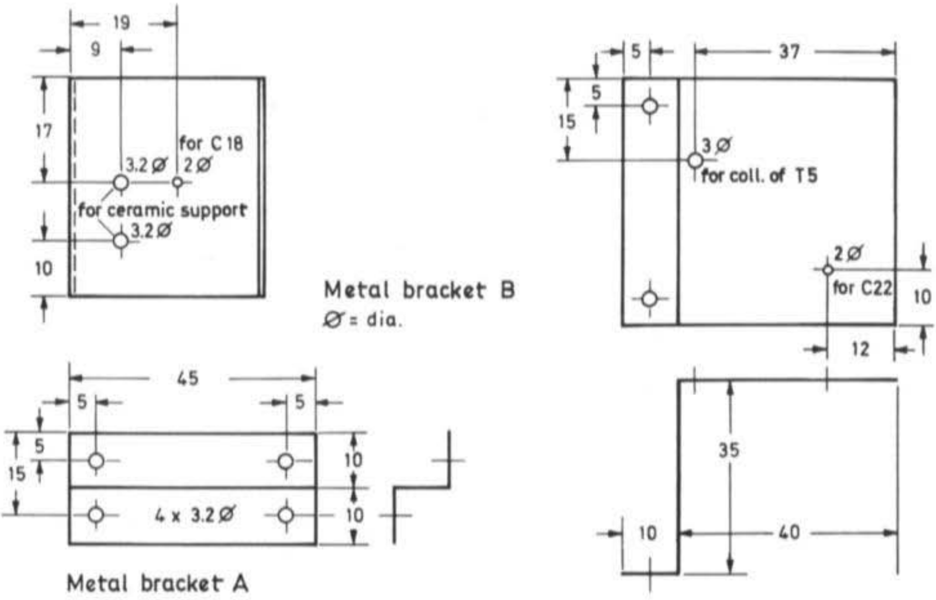


Fig. 7: Metal brackets for the transmitter output stage

After having mounted the PA stage, the metal brackets are screwed to the printed circuit board so that a good contact to the ground area is made. Finally, they are soldered using a large soldering iron (80 W). A thin solder joint that has been soldered quickly will be sufficient.

Figures 8 to 11 illustrate the preparations and mounting of the PA stage and should provide assistance during construction. They show the PA stage graphically or as a photograph in three different perspectives. It will be seen that the collector connection of the final transistor is very short and is soldered to a small teflon (PTFE) feedthrough. This is more favourable than to feed the collector connection lead through the hole. The emitter connection is also very short and is connected to the metal plate. The large heat sink of the final transistor - which is very necessary for FM operation at 18 V - is mounted onto the metal bracket using a nylon screw and ceramic washer. A good thermal connection between the transistor and the heat sink can be made by depressing the aluminium carefully using a centre punch. This rather elaborate cooling arrangement is used in order to avoid the more expensive transistor type 40291, which could also be used.

The connection points Pt 4 (at one end of the printed circuit board) and Pt 5 (on the metal břacket) can bee seen in the photographs. Screw clamps are soldered to these connections onto which the home-made resistor R 11 is connected (This resistor cannot be soldered). The resistor itself is covered with insulating tube.

The individual operating voltage connections are connected in the given order during the alignment process.

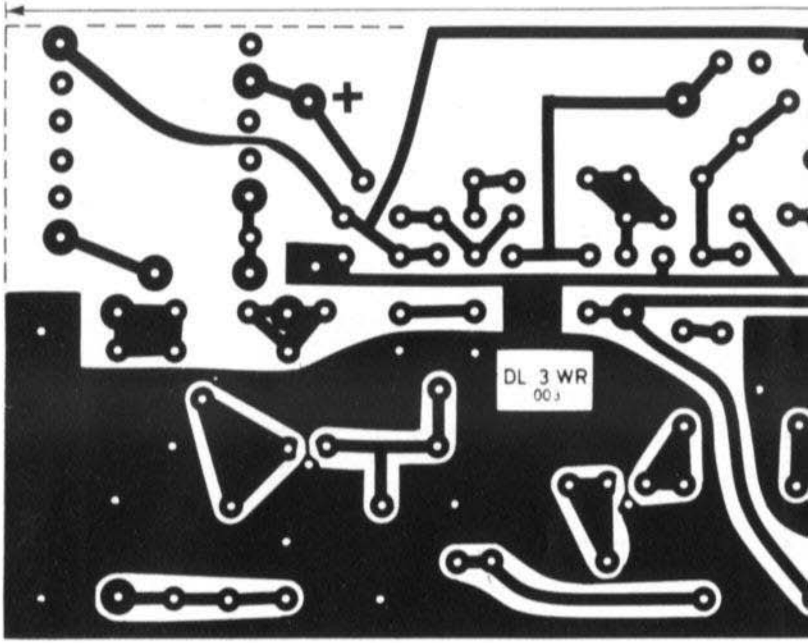


Fig. 4: Printed circuit board DL 3 WR 003 for the 2 metre transmitter including modulato

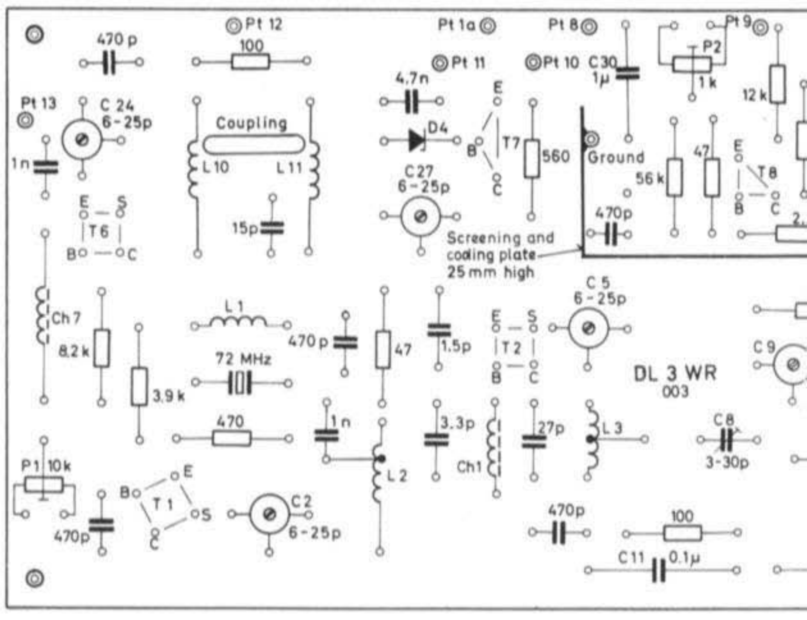
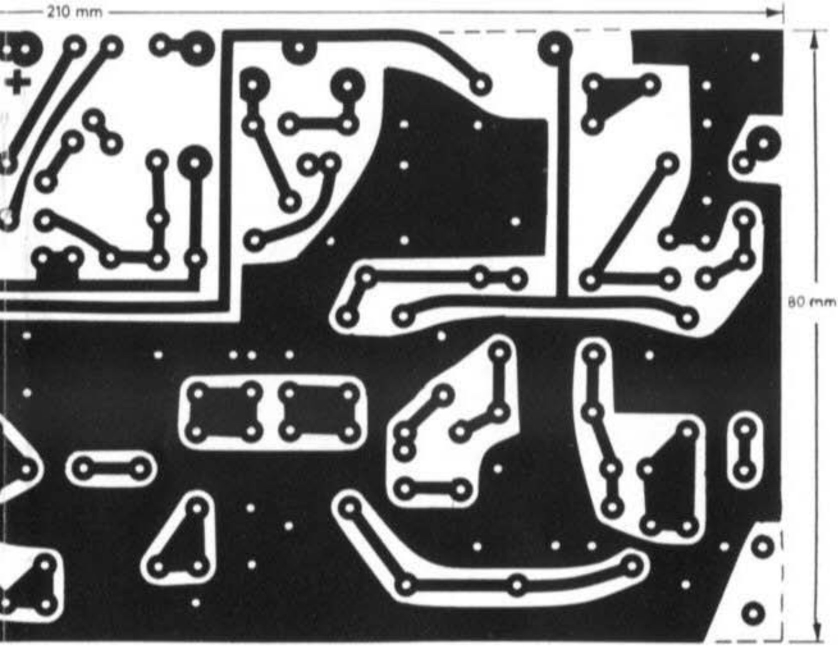
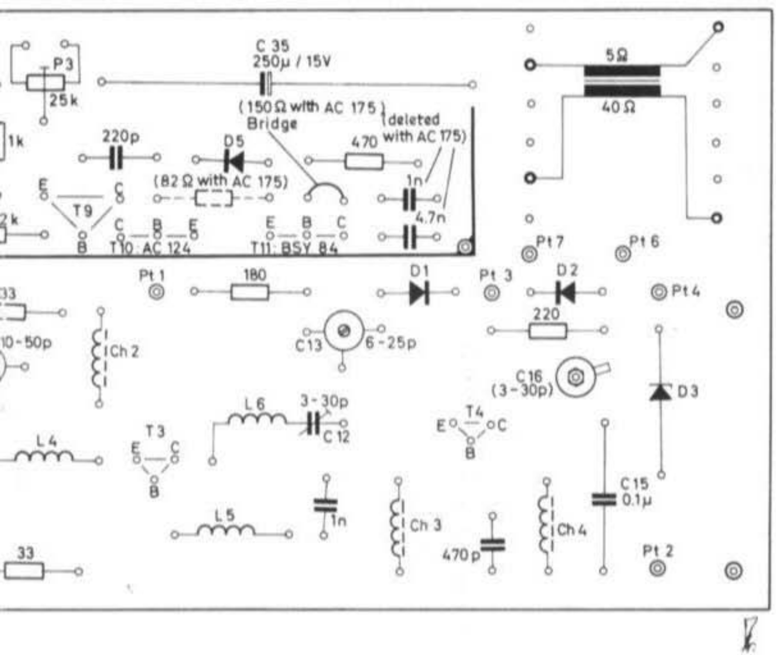
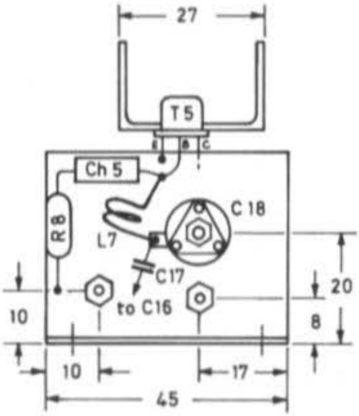


Fig. 5: Component location plan to PC-board DL 3 WR 003

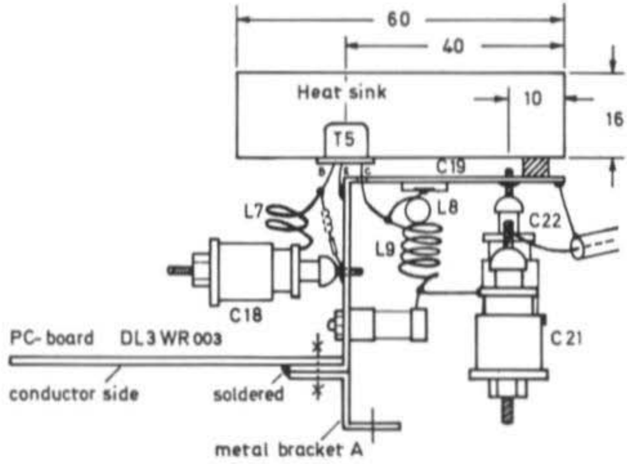


r and VFO matching stage (210 mm by 80 mm)





Output stage (metal bracket B)
as seen from the PC-board



Output stage, side view

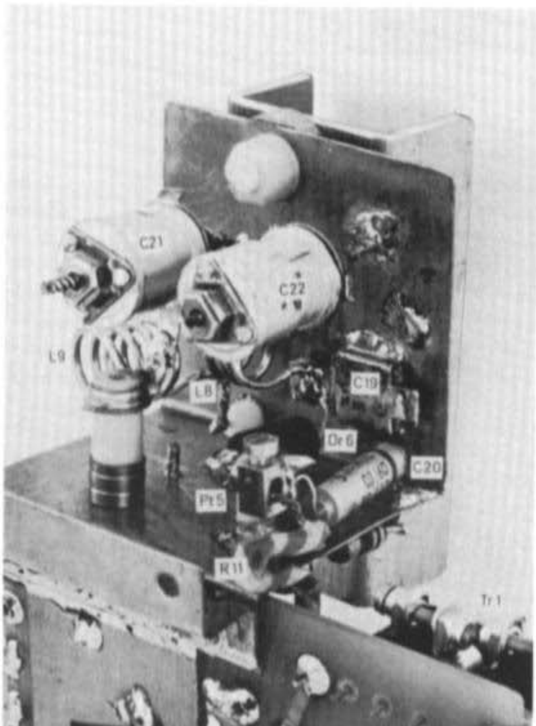
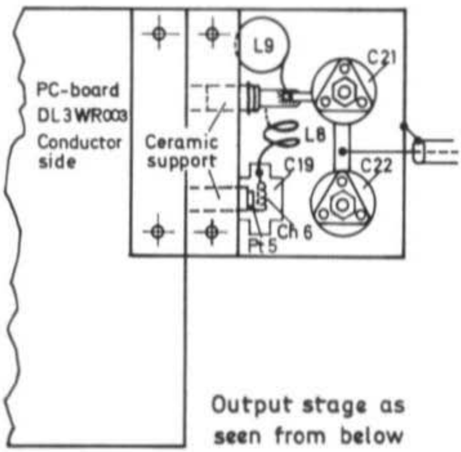


Fig. 9



Output stage as seen from below

Fig. 8: Assembly and mounting of the transmitter

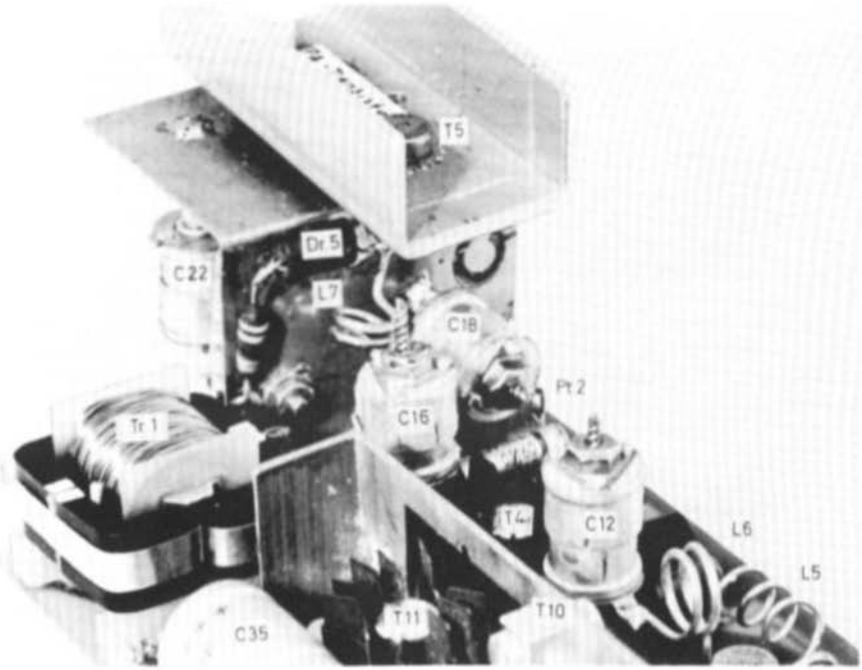

T5
Dr 5
C22
L7
C18
Pt 2
Tr 1
C16
T4
C12
L6
L5
T11
T10
C35

Fig. 10

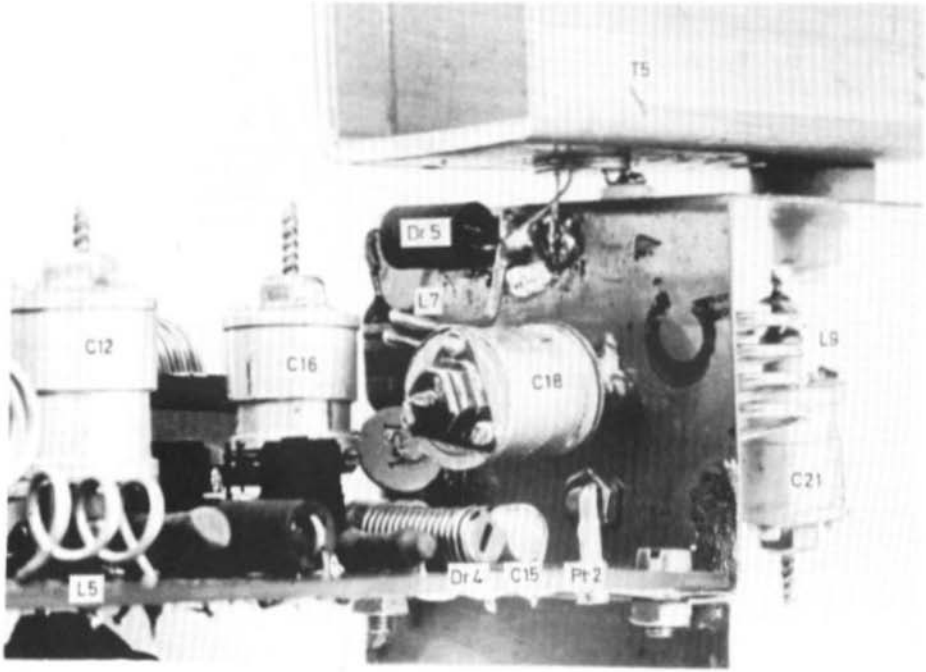

T5
Dr 5
L7
C12
C16
C18
L9
C21
L5
Dr 4
C15
Pt 2

Fig. 11

3.1.5. ALIGNMENT OF TRANSMITTER AND MODULATOR

The alignment is to be explained more extensively because it does not represent a simple "tune for maximum" process. Some general notes are also to be given.

A two metre receiver should be used to monitor the crystal-controlled frequency of the transmitter during the alignment process, with which the band is observed for spurious signals. The alignment can either be made at 12 V, or 13.5 V (so that the RF indication lamp lights quicker when aligning the first stages). The total operating voltage range should be tested after completing the alignment to find the most favourable compromise between the adjustments for the highest and lowest voltage. The alignment of trimmer capacitors C 8 and C 12 is somewhat critical. The bandwidth of the transmitter is more than sufficient; however, the final alignment should be made at approximately band centre. The current values of the first three stages and the tripler stage with transistor T 6 can be measured as the voltage drop across the filter resistors R 4, R 5, R 7 and R 12.

The operation of the stabilizer is firstly tested by checking its output voltage which should not vary noticeably when loading with approximately 500 Ω.

The crystal-controlled oscillator chould cease oscillation if capacitor C 2 is detuned. The inductance L 1 will require more inductivity if the oscillator still operates on detuning towards lower frequencies, and vice-versa.

The first two stages with transistors T 1 and T 2 are aligned whilst observing a 6 V/0.6 W lamp which is closely-coupled to inductance L 3; trimmer capacitor C 8 should be at its point of minimum capacitance.

To align the doubler stage comprising transistor T 3, the lamp is now closely-coupled to inductance L 6 and the collector current measured. This is followed by increasing the capacitance of trimmer capacitor C 8 slightly and aligning trimmer capacitors C 5 and C 9 for maximum collector current of transistor T 3. These two trimmers then remain in this position. Trimmer capacitor C 8 can now be rotated to approximately half its capacitance, and trimmer capacitors C 12 and C 13 aligned for maximum lamp brightness.

The driver and final amplifier stage are aligned together. The collector current values should be measured using separate meters and the common operating voltage should be fed via an electronic fuse such as described in (7). A 6 V/2.8 W lamp is now soldered to the output of the transmitter. The driver and doubler stages are now aligned for maximum current through the PA transistor. The PA stage itself is aligned for maximum lamp brightness. The alignment of trimmer capacitors C 12, C 16 and C 18 is somewhat critical. The drive to the three power transistors can be regulated using trimmer capacitor C 8. Too high a drive will cause an unnecessary high current (especially in the driver stage); on the other hand, if the drive is too low, the transmitter will not operate at operating voltages of 12 V or 11 V, or will not allow itself to be positively modulated.

After this preliminary peak-power alignment, the tripler stage (Transistor T 6) should be considered. The collector current of this transistor and transistor T 2 should be indicated on separate meters. The trimmer potentiometer P 1 is firstly adjusted to its maximum value. The VFO is tuned so that a weak signal is audible at 145 MHz. The value of potentiometer P 1 is now reduced until the current given in Table 1 flows via transistor T 6. If the drive voltage is too great, a resistor should be soldered between the VFO output and ground. The 72 MHz bandpass filter is now coarsely aligned using trimmer capacitors C 24 and C 27 for maximum collector current via transistor T 2. After this, inductance L 11 is carefully damped with a finger so that the collector current of transistor T 2 is still just indicated on the meter. This is followed by aligning trimmer capacitor C 24 for maximum meter reading. Inductance L 10 is now damped, and trimmer capacitor C 27 aligned for the highest current reading. The VFO is then tuned from one band limit to the other whilst observing the collector current of transistor T 2. If it does not remain constant, it will be necessary for the coupling link to be varied and the alignment procedure to be repeated. The correct position is where the loosest coupling resulting in a constant current via transistor T 2 is obtained.

The collector current of transistor T 2 should be as high when driven from the VFO as it is when driven by the crystal-controlled oscillator (or even higher). This is achieved by finding the most favourable adjustment of the trimmer potentiometer P 1. It may be necessary for the crystal-controlled oscillator to be finally trimmed due to the altered impedance of the resonant circuit comprising inductance L 11 and capacitors C 26 and C 27.

The peak-power alignment can now be carefully corrected at 145 MHz, using the VFO. The trimmer capacitors C 5 and C 9 may only be adjusted when trimmer capacitor C 8 is at virtually minimum capacitance.

For alignment of the modulator, a voltmeter should be connected between the minus pole of the electrolytic capacitor C 35 and ground. With potentiometer P 3 in its centre position, the voltage reading at this point is adjusted, using potentiometer P 2, to half the operating voltage: to 6 V for 12 V operation. This is valid for the germanium complementary pair; if a silicon transistor is used for transistor T 11, the voltage reading should be adjusted to 5 V. At the same time, the quiescent current of the whole modulator is aligned to the values given in Table 1 by adjusting potentiometer P 3. A more accurate adjustment for maximum output power without "flat-topping" can only be made with the aid of an oscilloscope. In order to do this, a resistor of approximately 40 Ω/2 W is soldered between connection points Pt 6 and Pt 7, and the connection to the transmitter disconnected.

The transmitter is now to be aligned for the most favourable amplitude modulation: The modulation depth is kept low at first. Trimmer capacitors C 12, C 16, C 18, C 21 and C 22 are adjusted for maximum light brightness. The modulation depth is increased and the alignment repeated for each step. The modulation should be removed at regular intervals to check whether the brightness really increases during modulation. No splatter should be audible on a receiver tuned to a frequency approximately 10 to 25 kHz from the transmit frequency when fully modulating.

Since the lamp does not represent a true 50 Ω load, it is necessary for the alignment to be corrected with modulation on a well-matched antenna or dummy load. It should only be necessary to realign trimmer capacitors C 21 and C 22. It is now possible to check the operating voltage range of the transmitter; to achieve the large range of 10 V to 13.5 V, some compromises in the tuning will be required.

3.2. RF VOLTAGE INDICATOR DL 3 WR 004

The circuit of the RF voltage indicator is given in Fig. 3. The resonant circuit of this indicator, which is tuned to the transmit frequency range, has the advantage in comparison to the normal wideband demodulation circuit that harmonic and spurious power is not indicated. If the transmitter is defective i.e. if it generates power at a number of different frequencies, the indicated voltage level will fall. Any defective operation will be indicated in this manner on the RF voltage indicator.

In addition to this, a DC oscilloscope can be connected to the output of this circuit (point Pt 402 and 403) and be used to check the modulation depth and distortion. This is done by disconnecting the meter and replacing it with a 15 kΩ resistor. If this resistor was not provided, the time constant would not be correct and the modulation would not be visible.

The resonant circuit should be aligned to the band centre frequency. However, due to its relatively high Q, the indication will not be completely constant when tuning the VFO from one band limit to the other.

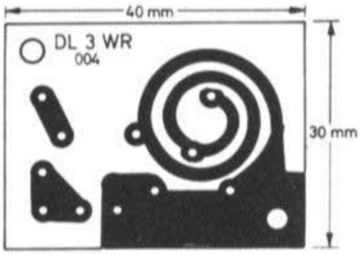


Fig. 12: RF-indicator board DL 3 WR 004

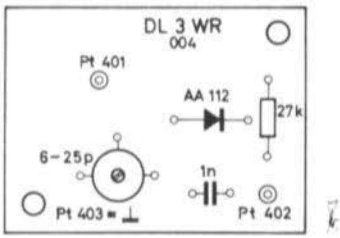


Fig. 13: Component location plan to DL 3 WR 004

3.2.1. ASSEMBLY OF THE RF VOLTAGE INDICATOR

The RF voltage indicator is built up on the printed circuit board DL 3 WR 004 whose dimensions are 40 mm by 30 mm. Figures 12 and 13 show the printed circuit board and the corresponding component location plan. The trimmer capacitor is of the same type as used in the transmitter; the diode is a conventional germanium type. The printed circuit board is mounted in the vicinity of the output socket for the two metre antenna. A short piece of wire is fed from connection point Pt 401 to near the position where the coaxial cable from the transmitter to the change-over switch "Ant. socket - 70 cm Tripler" is fed.

to be continued

REFERENCES

(1) D. Grossmann: Simple, Compact PA Stages for Two Metres
VHF COMMUNICATIONS 2 (1970), Edition 1, pages 45-55

(2) H. J. Franke: A Ten Watt Transmitter for 70 cm
VHF COMMUNICATIONS 1 (1969), Edition 4, pages 243-248

FIELD EFFECT TRANSISTORS IN THE 28/144 MHz TRANSVERTER DJ 6 ZZ 001

by F. Weingärtner, DJ 6 ZZ

A transistorized transverter was described in Edition 4/1969 of VHF COMMUNICATIONS (1) which allowed a short wave SSB transceiver to be extended for VHF operation. In its original state, the transverter only provided an output power of approx. 200 mW if normal silicon transistors (e.g. BF 224) were used in the transmit mixer. Mr. P. Griebel, DJ 9 PC, suggested the possibility of using two junction field effect transistors in the transmit mixer in such a manner that it was still possible to achieve the output power of 200 mW, and to obtain a considerably higher intermodulation ratio. The circuit was tested and the required modifications are now to be described. Since only one conductor lane has to be broken, it is possible to use the original printed circuit board (DJ 6 ZZ 001). If the printed circuit board has been completed, the modification will not take more than one hour.

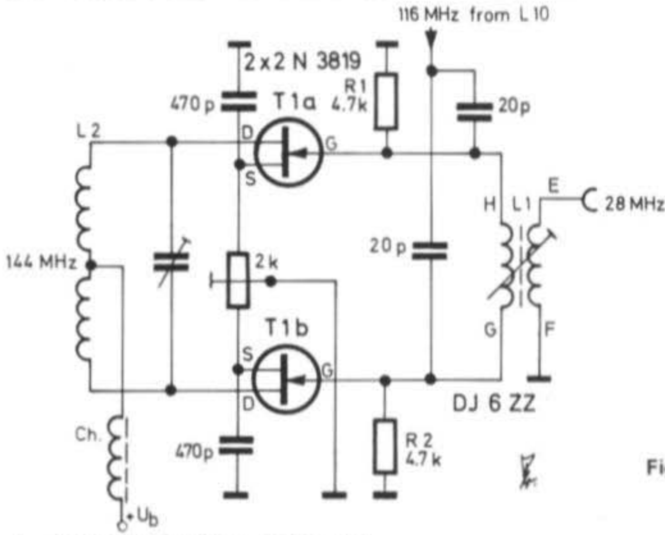


Fig. 1: Circuit extract showing the new FET mixer

1. THE MODIFIED CIRCUIT

A circuit diagram of the new FET-transmit mixer is given in Fig. 1. The oscillator voltage is fed to the two gate connections in push-push via two 20 pF capacitors. The series-connection of these two capacitors also formes the resonant circuit capacity for the 10 metre input circuit.

Two 4.7 kΩ resistors (R 1 and R 2) represent the DC connection between the gate connections and ground. The required bias voltage is generated as the voltage drop across a resistor in the source circuit. In order to balance the operating point, the resistor is in the form of a potentiometer. Each source connection is individually bypassed with a 470 pF capacitor. The drain connections are connected to the push-pull resonant circuit, which is not altered.

2. MODIFICATIONS OF THE ASSEMBLY

As already mentioned, only a few, simple modifications are necessary on the original printed circuit board DJ 6 ZZ 001 and its complement. The required modifications are now to be explained briefly; we recommend that the original description (1) be also studied for comparison.

2.1. The resistors R 1 and R 2 (originally 220 Ω) must be removed, and replaced by a wire bridge on the conductor side. Two new holes should now be drilled in the adjacent ground area. These holes are required for the new resistors R 1 and R 2 (4.7 kΩ) which are mounted vertically.

2.2. The output coupling links A - B and C - D of inductance L 1 are no longer required; they should be unsoldered and carefully removed. The inductance connections G and H are also unsoldered and resoldered into the holes where the connections A and D were soldered. No modification of the coupling link E - F is required.

The 470 pF capacitor between points B and C and the two trimmer potentiometers P 1 and P 2 are no longer required and can be removed.

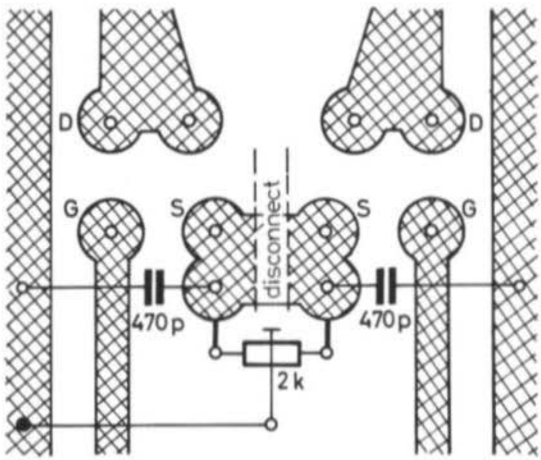


Fig. 2: Modifications to be made in the lower righthand corner of PC-board DJ 6 ZZ 001. Components as seen from the conductor side.

2.3. The resistor R 3 (560 Ω) and the parallel-connected 4.7 nF bypass capacitor are now unsoldered and replaced by two 470 pF capacitors. The conductor lane between these two capacitors is now disconnected (broken) using a knife. The location on the printed circuit board is shown in Fig. 2. This illustration also shows where the additional holes for the balancing potentiometer should be drilled. The wiper of the potentiometer is connected to ground using a wire bridge; the other two connections can be bent and soldered to the corresponding conductor lanes.

2.4. The author found that the field effect transistors required a higher oscillator voltage than conventional silicon transistors. The load exhibited by the mixer stage is relatively low; the oscillator voltage can be increased easily by placing the coil tap of inductance L 10 two turns towards the "hot" end (3 turns instead of 1 turn) and by increasing the value of resistor R 12 from 120 Ω to 270 Ω.

3. REFERENCES

(1) F. Weingärtner: A 28 MHz/144 MHz Transistorized Transverter
VHF COMMUNICATIONS 1(1969), Edition 4, Pages 189-195

A DIGITAL DISCRIMINATOR ACCESSORY FOR FM DEMODULATION

by D. E. Schmitzer, DJ 4 BG

INTRODUCTION

A special demodulator is required for the reception of frequency modulated signals. Such a demodulator converts the frequency variations of the RF or IF signal into voltage-amplitude variations. The most popular form of FM demodulators is the ratio detector. The disadvantage of this type of detector is that the ratio filter requires a certain amount of measuring instruments for the alignment process. A circuit that is less known but is gaining popularity in modern FM systems is a circuit operating in a digital manner. The advantage of this circuit is that no alignment is required and that it possesses a very linear relationship between input frequency and output voltage amplitude.

In order to indicate the industrial interest in this new circuit technology, it should be pointed out that a special integrated circuit (TAA 710 manufactured by ITT-Intermetall) has been developed. This integrated circuit converts a television sound IF signal of 5.5 MHz to approximately 200 kHz where it is demodulated using a digital discriminator.

1. THEORY OF OPERATION

A basic circuit diagram of the digital discriminator is shown in Figure 1. The most decisive point for its operation are the cut-off frequencies of the RC-links R 1/C 1 and R 2/C 2. The cut-off frequency f_{c1} is:

$$f_{c1} = \frac{1}{2\pi \times R\,1 \times C\,1}$$

It must be somewhat higher than the highest input frequency, whereas the cut-off frequency f_{c2} :

$$f_{c2} = \frac{1}{2\pi \times R\,2 \times C\,2}$$

should be chosen so that it is far greater than the highest audio frequency to be demodulated. For an intermediate frequency of 500 kHz and a highest modulation frequency of 3 kHz, the dimensioning should be as follows:
$f_{c1} \approx 1.6$ MHz (R 1 = 1 kΩ, C 1 = 100 pF) and $f_{c2} \approx 6.5$ kHz (R 2 = 47 kΩ, C 2 = 2.2 nF).

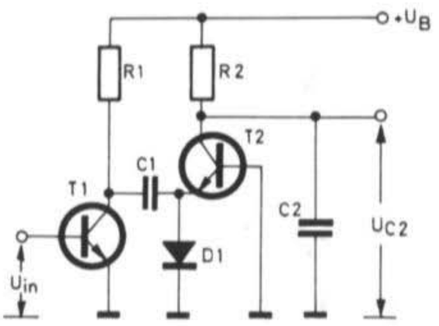


Fig. 1: Principle of a digital discriminator

Under non-drive conditions, both transistors of the digital discriminator will be blocked. The capacitor C 1 is charged via the resistor R 1 and diode D 1 to the value of the operating voltage U_B. The same is valid for capacitor C 2 which is charged via the resistor R 2. If a positive voltage U_{in} of more than 0.7 V appears at the input of the circuit, this will cause transistor T 1 to conduct. Due to the collector current, the voltage drop across resistor R 1 will be so great that the voltage across transistor T 1 will be reduced to the value of the residual voltage. This voltage surge is passed via capacitor C 1 to the emitter of the second transistor. The emitter of transistor T 2 becomes thus negative with respect to the base which effects a current flow through transistor T 2. This current continues to flow via T 2 until capacitor C 1 is discharged. Since the collector current exhibits practically the same value as the emitter current, the short period of the collector current flow reduces the charge of capacitor C 2 by the value that was previously stored in capacitor C 1. This causes the voltage across C 2 (output voltage U_{C2}) to fall. If the voltage U_{in} ceases after this charge transfer process, the current flow via transistor T 1 will also cease and capacitors C 1 and C 2 will recharge themselves to the initial value (U_B).

If an alternating voltage having a suitable frequency and amplitude is connected to the input instead of the DC pulse, the charge cycle of the charge transfer and the recharge will be repeated in rhythm with the input frequency. The cut-off frequencies or the time constants are now dimensioned according to the previously mentioned rules so that the time between the charge transfer periods is not sufficient to fully charge capacitor C 2. The charge process via resistor R 2 and the discharge due to the current pulses via transistor T 2, which are in turn caused by the input voltage, therefore have opposite effects. Since the current pulse rate is proportional to the input frequency, the higher the input frequency, the lower will be the charge across capacitor C 2. The voltage across capacitor C 2, which represents the output voltage of the digital discriminator, thus falls on increasing the frequency of the input AC voltage. This represents the required conversion of frequency variations into corresponding voltage amplitude variations.

2. PRACTICAL CIRCUIT OF A DIGITAL DISCRIMINATOR

The complete circuit diagram of a FM demodulator accessory for an existing receiver is given in Figure 2. The FM accessory is equipped with an integrated circuit as a wideband amplifier, a digital discriminator comprising transistors T 1 and T 2, as well as a 3 kHz active low-pass filter (transistor T 4) and a buffer stage with transistor T 3 at the input of the RC low-pass filter. The circuit shown in Figure 2 is suitable for any intermediate frequency between 100 kHz and approximately 550 kHz. The selectivity must be carried out previous to this accessory; the bandwidth should not be narrower than 12 kHz. Such a bandwidth corresponds to the values given in reference (1) for a modulation index of M = 1.

The intermediate frequency is fed to the FM accessory via a coupling link to point E and E'. The number of turns and the degree of coupling should be chosen so that the weakest audible signals result in an IF voltage of approximately 10 mV at points E and E'. Stronger signals are limited in the integrated circuit and by diode D 2 in conjunction with the base-emitter diode path of transistor T 1. This limiting process suppresses amplitude modulated signals, such as ignition interference.

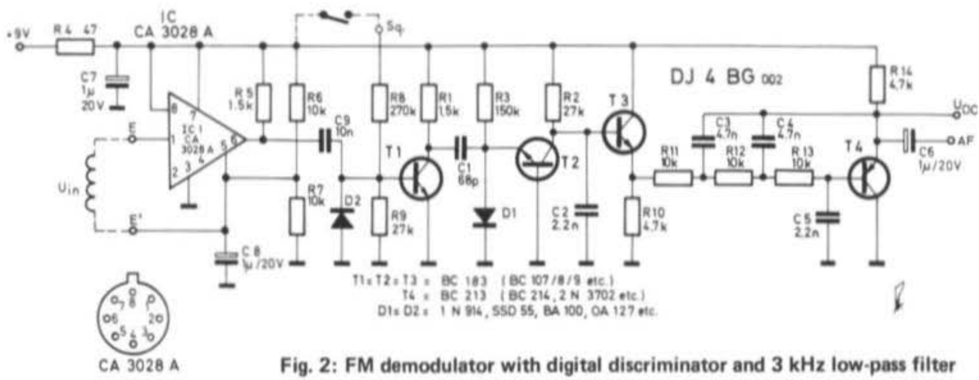


Fig. 2: FM demodulator with digital discriminator and 3 kHz low-pass filter

In order to ensure that the decisive components of the actual digital discriminator circuit can be easily found, they are designated in the same manner in Figure 2 as in the principle circuit diagram given in Fig. 1. The previously mentioned cut-off frequencies have been dimensioned so that intermediate frequencies of approximately 100 to 550 kHz can be processed. The conversion characteristic curve of the circuit is shown in Figure 3. The voltage U_{out} is thus the voltage at the DC output (U_{DC}). The subsequent AF amplifier only receives the AC component via capacitor C 6. The amplitude of this AC voltage is dependent on the frequency deviation and on the transconductance of the conversion slope. The slope of the characteristic curve is in turn dependent on the dimensioning of the RC-link R 1/C 1 and thus on the highest input frequency for which it is dimensioned. The higher the upper frequency limit, the lower will be the values for resistor R 1 and capacitor C 1, and therefore the flatter the characteristic curve will be. Since, however, a relatively flat characteristic curve only produces a low AF voltage at a certain frequency deviation, it is advisable to select a low intermediate frequency in order to obtain the highest possible AF voltage and to suitably dimension the digital discriminator.

The characteristic curve shown in Figure 3 shows that the described circuit provides an AF voltage of approximately 1.3 V peak-to-peak at a frequency deviation of ± 50 kHz. Division of these values by 20 shows that amateur signals with a frequency deviation of ± 2.5 kHz will only produce an AF signal of approximately 65 mV. This corresponds to a mean value (for sinusoidal voltages) of approximately 23 mV. An AF amplifier should follow the accessory, whose sensitivity is approximately that required for a dynamic microphone.

The resistor R 3, not shown in the principle circuit diagram, linearizes the characteristic curve to a certain degree. However, the linearity of the circuit is sufficiently good so that no differences can be observed between the characteristic curves of the principle circuit and the complete circuit diagram shown in Figure 3.

As has been briefly mentioned previously, the discriminator is followed by an impedance converter equipped with the transistor T 3 and an active AF filter. The low-pass filter is provided firstly for suppression of the residual IF voltage and secondly to limit the AF range to approximately 3 kHz. This means that frequency modulated noise components that are passed through the IF pass-band (which must be more than 3 kHz) and converted by all types of discriminator circuits into AF noise voltage, are suppressed at frequencies higher than 3 kHz. The dimensioning of the low-pass filter for other applications of the digital discriminator can be made as described in reference (2). The RC-link comprising R 2/C 2 should also be considered.

Figure 3 also indicates excellent temperature stability of the circuit. This is especially due to the use of complementary transistors for T 3 and T 4. The operating point drift of the individual transistors is therefore compensated.

Figure 4 indicates the output DC voltage U_{out} of the circuit (Fig. 2) as a function of the input signal voltage U_{in} at three different intermediate frequencies. It will be seen that an excellent AM suppression is provided at IF levels of more than 20 mV (measured between points E and E').

At this point, a special circuit feature should be described: As shown in Fig. 4, the circuit exhibits a relatively steep threshold at input voltages between 10 and 15 mV (continuous lines). This threshold can be used as a squelch circuit by dimensioning the coupling link to the IF circuit so that the IF noise voltage is below the threshold in the order of 5 mV.

The threshold shows itself when transistor T 1 operates in class C without bias voltage. This means that the threshold can be suppressed (dashed lines) by providing transistor T 1 with a positive bias voltage via resistor R 8 so that transistor T 1 just takes collector current (class B). In order to make the squelch circuit switchable, a connection marked Sq. is provided.

3. CONSTRUCTION OF THE DIGITAL DISCRIMINATOR

The FM accessory shown in Figure 2 can be built up on the printed circuit board DJ 4 BG 002 given in Fig. 5. The dimensions of this PC-board are 68 mm by 50 mm. The corresponding component location plan for the PC-board DJ 4 BG 002 is given in Figure 6. The construction is uncritical; it is only necessary for attention to be paid that no spurious signals are introduced into the input of the sensitive wideband amplifier IC 1, and that the operating voltage U_B should be well filtered due to the high sensitivity of the subsequent AF amplifier.

3.1. SPECIAL COMPONENTS

IC 1 : Linear integrated circuit CA 3028 A from RCA
T 1 to T 3 : BC 183 (TI) or BC 108, 2 N 2926, 2 N 3903
T 4 : BC 213 (T I) or 2 N 3905, 2 N 3702, 2 N 2907
D 1 and D 2 : BA 100, OA 127-131, BAY 86-91, 1 N 914 or similar types
C 1 : 68 pF ceramic or styroflex
C 2 : 2.2 nF styroflex etc.
C 3 and C 4 : 4.7 nF ± 5% styroflex etc.
C 5 : 2.2 nF ± 5% styroflex etc.
C 6 to C 8 : 1 μF/20 V tantal electrolytic
C 9 : 10 nF ceramic disc
R 1 to R 10 : as given in circuit diagram; tolerance ± 10%
R 11 to R 13 : 10 kΩ ± 5%
R 14 : 4.7 kΩ ± 10%

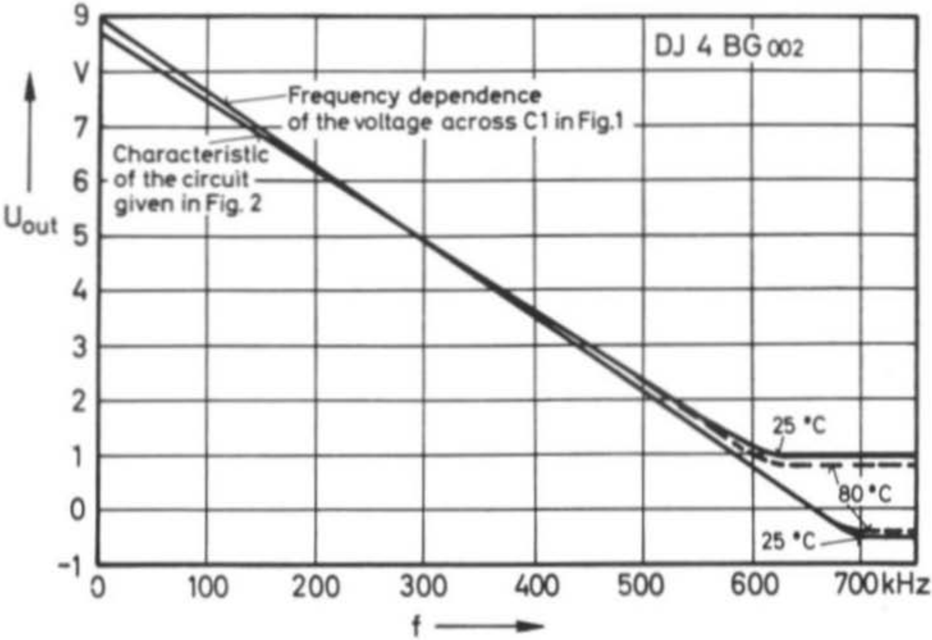


Fig. 3: Output voltage-frequency curve of the principle circuit shown in Fig. 1 and the practical circuit in Fig. 2 at 25° C and 80° C.

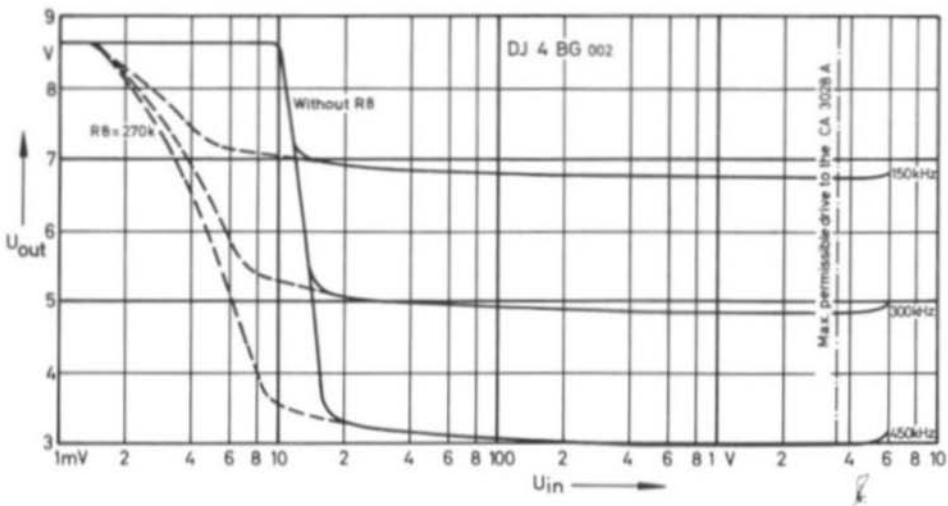


Fig. 4: Dependence of the output voltage U_{out} on the input voltage U_{in} at various frequencies

3.2. AVAILABLE PARTS

The printed circuit board DJ 4 BG 002 as well as all semiconductors are available from the publishers. Please see advertising page.

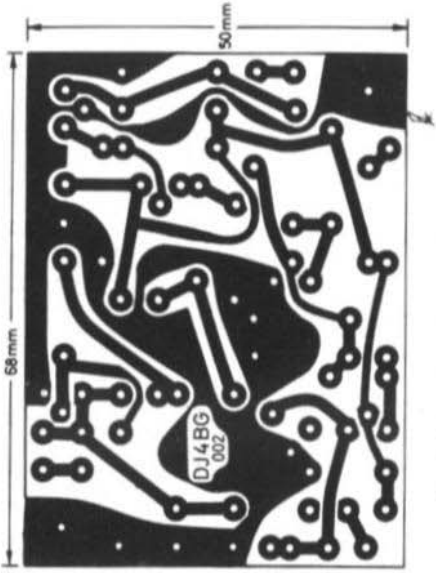


Fig. 5: Printed circuit board DJ 4 BG 002

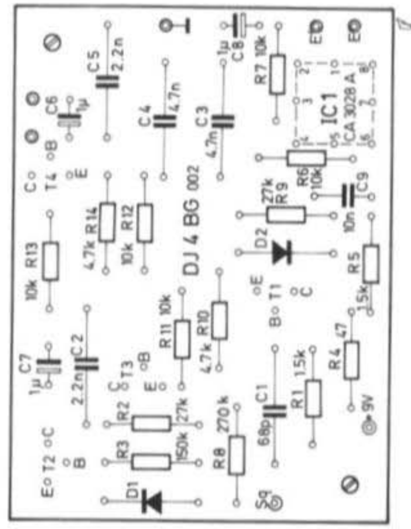


Fig. 6: Component location plan to DJ 4 BG 002

4. EDITORIAL NOTES

Where it may be impossible to use an additional coupling link, the digital discriminator DJ 4 BG 002 can also be connected to the IF circuit using a coupling capacitor. Since the input impedance of the discriminator is relatively low (several kΩ), a very low capacitance value must be used if only the non-loadable "hot-end" of the IF circuit is available (tubed equipment). The resulting voltage division, means that a relatively high IF voltage level is required if enough voltage is to be available at input E via the low coupling capacitor. It is therefore more favourable to make the connection via a larger capacitance connected to a capacitive voltage divider across the IF circuit.

In addition to this, a DC path must be available between connections E and E' when using a capacitive coupling. This can be made by connecting a resistor having a maximum of 4.7 kΩ between these two points.

If the IF amplifier of the receiver does not exhibit the required bandwidth of 12 kHz, very unpleasant distortions will be audible. Frequency modulation has a different behaviour in this respect than AM where too low a bandwidth will only cause the response to be rather bassy due to the missing treble tones.

The described digital discriminator converts a frequency to a DC voltage. This process can, of course, also be used for other applications. It is, for instance, possible to use the discriminator in a revolution counter, frequency meter as well as in an AF discriminator such as required for RTTY. However, this is not to be dealt with at this point.

5. REFERENCES

(1) D. E. Schmitzer: Is Frequency Modulation Advantageous on the VHF bands ?
VHF COMMUNICATIONS 2 (1970), Edition 1, pages 21-24

(2) D. E. Schmitzer: Active Audio Filters
VHF COMMUNICATIONS 1 (1969), Edition 4, pages 218-235

SIMPLE, COMPACT PA STAGES FOR TWO METRES

by D. Grossmann, DJ 4 RX

PART II

Continuation from VHF COMMUNICATIONS 2 (1970), Edition 1

5.2. A PA STAGE WITH HELICAL INNER CONDUCTOR

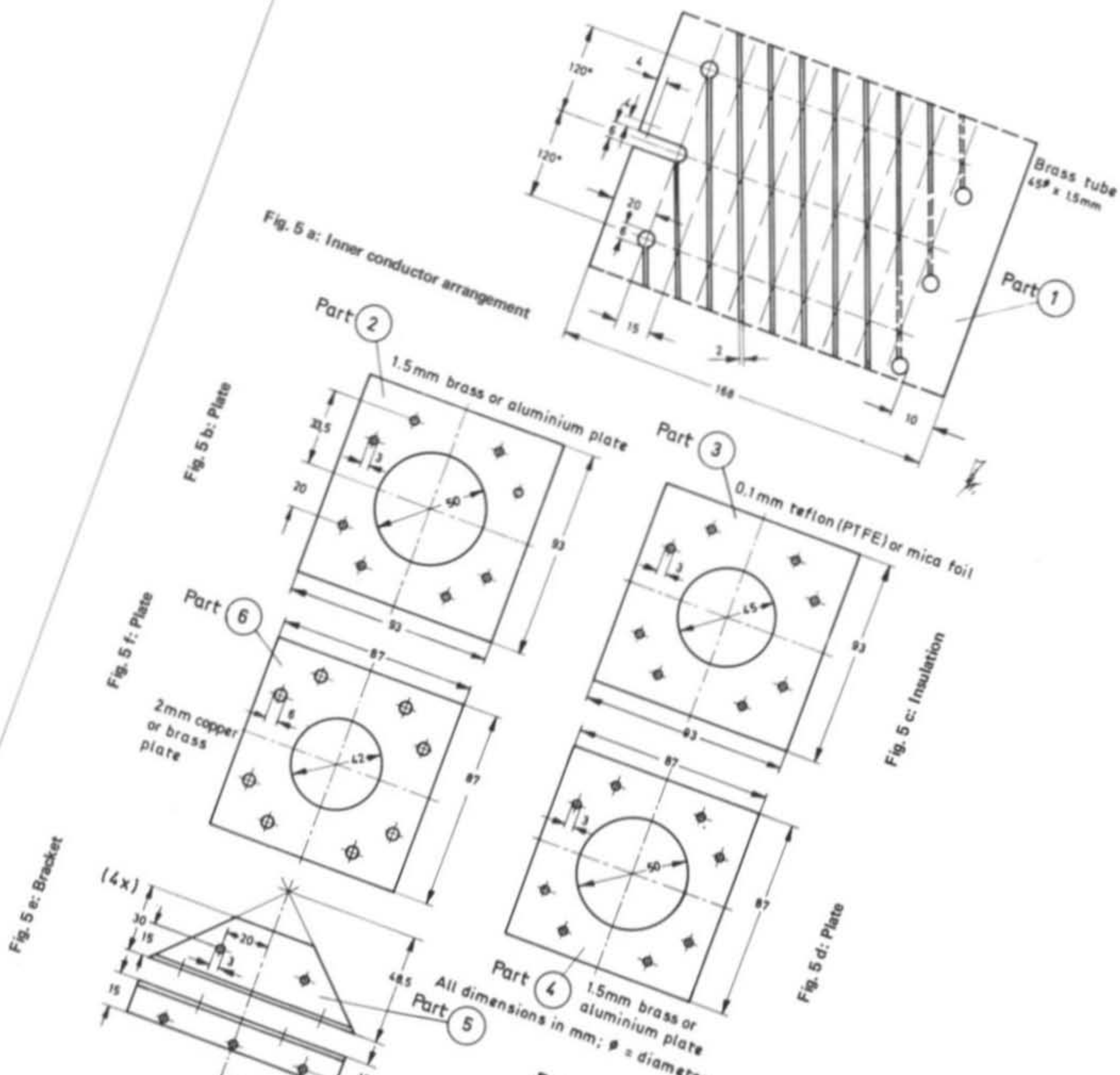
The dimensions of PA stages for the two metre band can be decreased still further if somewhat more elaborate metalwork is made. The impedance of the helical line used is approximately three times that of a normal inner conductor, which is also true of the electrical length (Z approx. 150 Ω). In order to obtain a more constant current distribution than that obtainable with a simple helix, the inner conductor has been built up in the form of a triple helix.

5.2.1. CONSTRUCTION OF THE INNER CONDUCTOR

Figure 5 illustrates the construction of the inner conductor and helix. Fig. 5a shows a flat projection of the inner conductor, which is in fact constructed from a 45 mm diameter brass tube. The inner conductor is firstly cut to the correct length of 168 mm and the end filed so that it is exactly vertical to the surface of the tube. The points indicated by the thin crosses in Figure 5a are now marked on the outer surface of the tube. Since it is very difficult to cut the helix in the round surface of the tube, this can be made more easily by joining each centre point of the individual crosses using a self-adhesive tape which is passed around the tube in the helical direction. The tube is now cut at one edge of this adhesive tape. The three cuts cease at both ends of the inner conductor in 6 mm diameter holes; one of the helical cuts stops in an elongated 6 mm cut-out which reaches to the end of the tube.

Unfortunately, conventional tubes possess considerable mechanical stress which will show itself by distorting the helix. It is therefore necessary to bend the tube back into its original form after the sawing process.

The 42 mm inner diameter of the brass tube used (see Fig. 5a) is somewhat greater than the outer diameter of the 4 X 150 tube's anode radiator (approx. 41.2 mm diameter). Furthermore, the inner conductor will expand somewhat after cutting the helix. Thus, if the PA tube and the inner conductor are mounted exactly concentric to another, no galvanic contact will exist at first between these two parts. In order to clamp the helical tube over the anode radiator of the 4 X 150 tube, two small brackets are mounted at each side of the previously mentioned elongated 6 mm cut-out (see Fig. 5a, left) which are joined together by a tensioning screw in the manner shown in Figure 6. One of these brackets is screwed from the inner surface of the helical tube using a countersunk screw. The second bracket is soldered to the other side of the cut-out using a soft solder having a low melting point. If the PA tube is overloaded, the anode radiator and thus also the corresponding end of the inner conductor will be heated to such an extent that the solder holding the bracket will melt and the helical tube will expand. Since the anode voltage of the tube is fed via the helical tube, the whole arrangement operates as a thermal fuse which will interrupt the anode voltage supply under severe overload conditions.



DJ 4 RX

Fig. 5: Individual pieces for the 144 MHz power amplifier with helical inner conductor (see parts list)

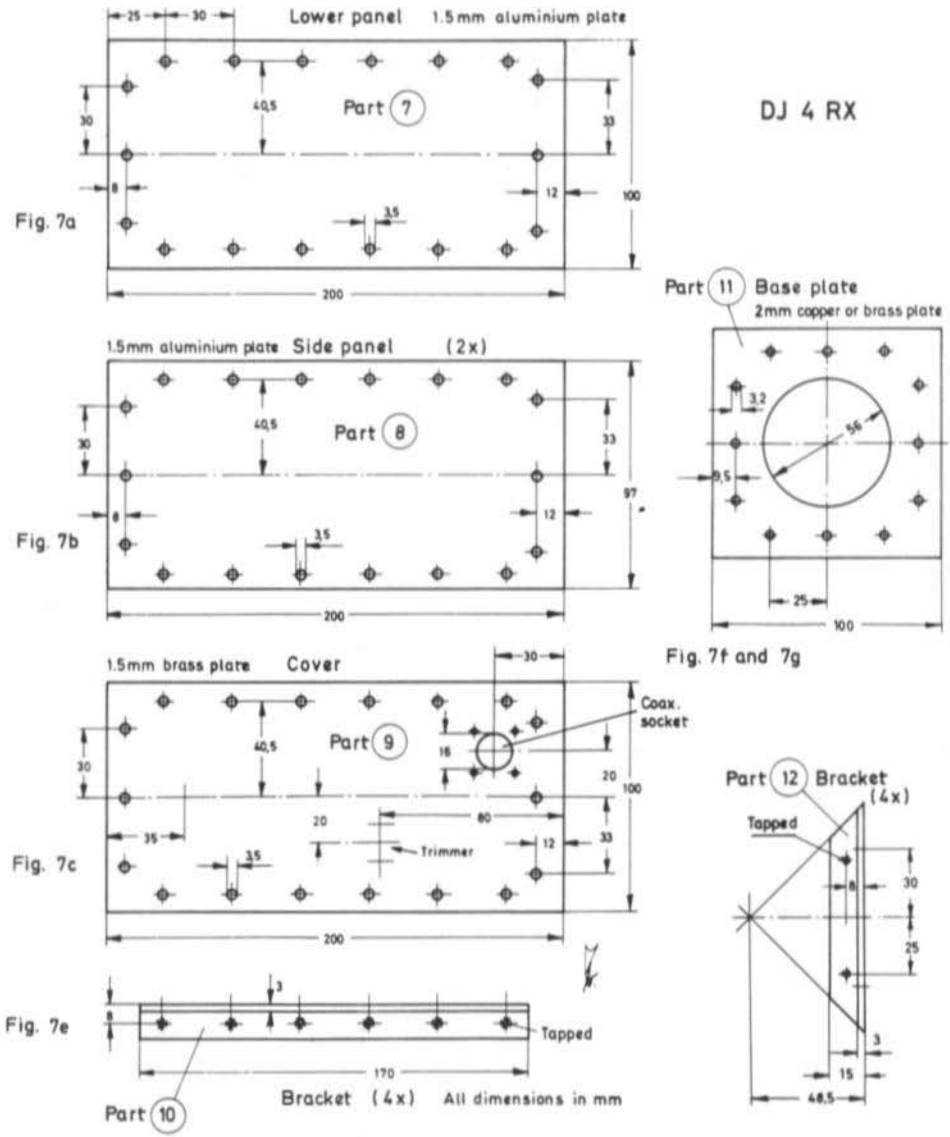


Fig. 7: Individual pieces for the 144 MHz power amplifier with helical inner conductor

5.2.2. CONSTRUCTION OF THE COMPACT PA STAGE WITH HELICAL INNER CONDUCTOR

The inner conductor (Part 1 in Fig. 5a) is placed vertically with its non-slotted end to the centre of the plate (Part 6) shown in Fig. 5f and soldered. After all parts have been manufactured according to the following parts list or Figures 5 and 7, it is possible to commence mounting the $\lambda/4$ anode circuit. The completed construction is shown in Figure 8.

The inner conductor flange (Part 6) is insulated on both sides using the insulation (Part 3 shown in Fig. 5c). After this, Part 6 is screwed to the plates (Parts 4 and 2 shown in Fig. 5d and 5b) as well as to the four brackets (Part 5). This should be made as shown in Fig. 8. This represents the RF bypass arrangement at the cold end of the anode line. In order to obtain the required insulation, the holes in the flange (Part 6) are greater than those of Parts 2, 3 and 4. These cut-outs are centred by placing insulation discs having an outer diameter of 6 mm and an inner diameter of 3 mm (2 mm thick) into the 6 mm holes of Part 6.

The base, side and covering plates (Parts 7, 8 and 9 in Fig. 7) can now be mounted onto the brackets(Part 5). The four brackets (Part 10 in Fig. 7e) provide stability for the outer conductor casing. At the other end of the casing, four brackets (Part 12 in Fig. 7f) and the base plate (Part 11) are screwed to the tube socket of the 4 X 150. A ceramic chimney is used for the cooling ventilation. The output coupling is mounted on the covering plate (Part 9), which consists of the coaxial socket, the coupling loop and a trimmer.

Photographs of the completed PA stage are given in Figures 9 and 10. Since the prototype PA stage was, with the exception of plates 3 and 4, constructed from aluminium plate, a great number of screws and brackets were necessary. A considerable amount of metalwork could be saved if copper or brass plate were used for the outer conductor casing. With the exception of Parts 3 and 4 as well as one cover (for instance Part 8), it is then possible for the whole casing to be constructed from one piece and soldered. The tube socket can be soldered or clamped into place.

The operative Q of 19 exhibited by this PA circuit results in such a large bandwidth that no capacitive tuning is required in the two metre range. If the resonant frequency of 145 MHz is not obtained, this can be adjusted by slightly bending the helical inner conductor or by filling one of the helical slots at the cold end with solder (a few mm are usually sufficient). If other tube types (for instance 4 X 250) are to be used, it is recommended for the resonant frequency to be adjusted to 148 MHz using a 4 X 150 tube and to load the anode side of the circuit capacitively using a disc in the normal manner (see Fig. 4).

If the mechanical parts are made from copper or brass, it is possible for them to be silver plated. The parts for the grid circuit are to be discussed in Section 7.

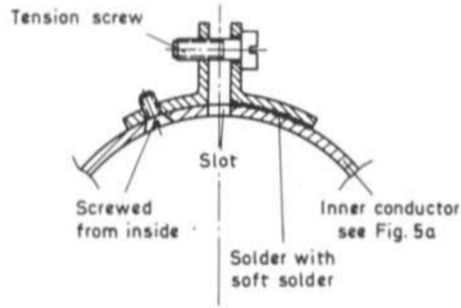


Fig. 6: Tensioning device for the inner conductor tube with thermal fuse

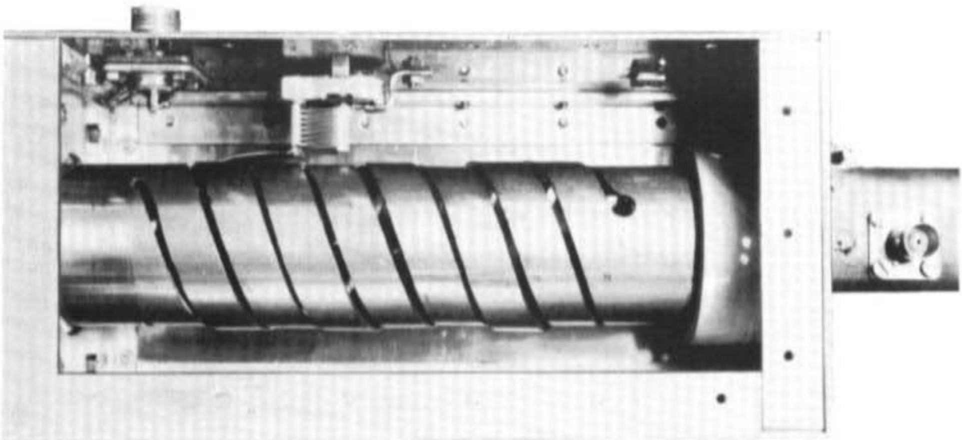


Fig. 9: λ /4 anode line circuit

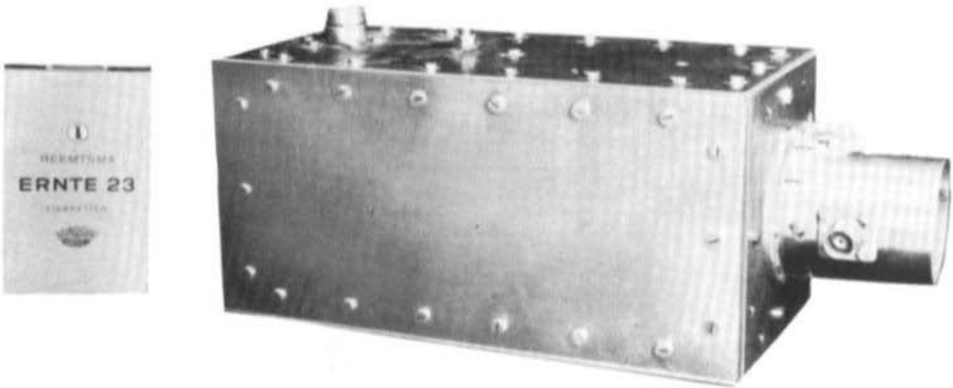


Fig. 10: The complete power amplifier

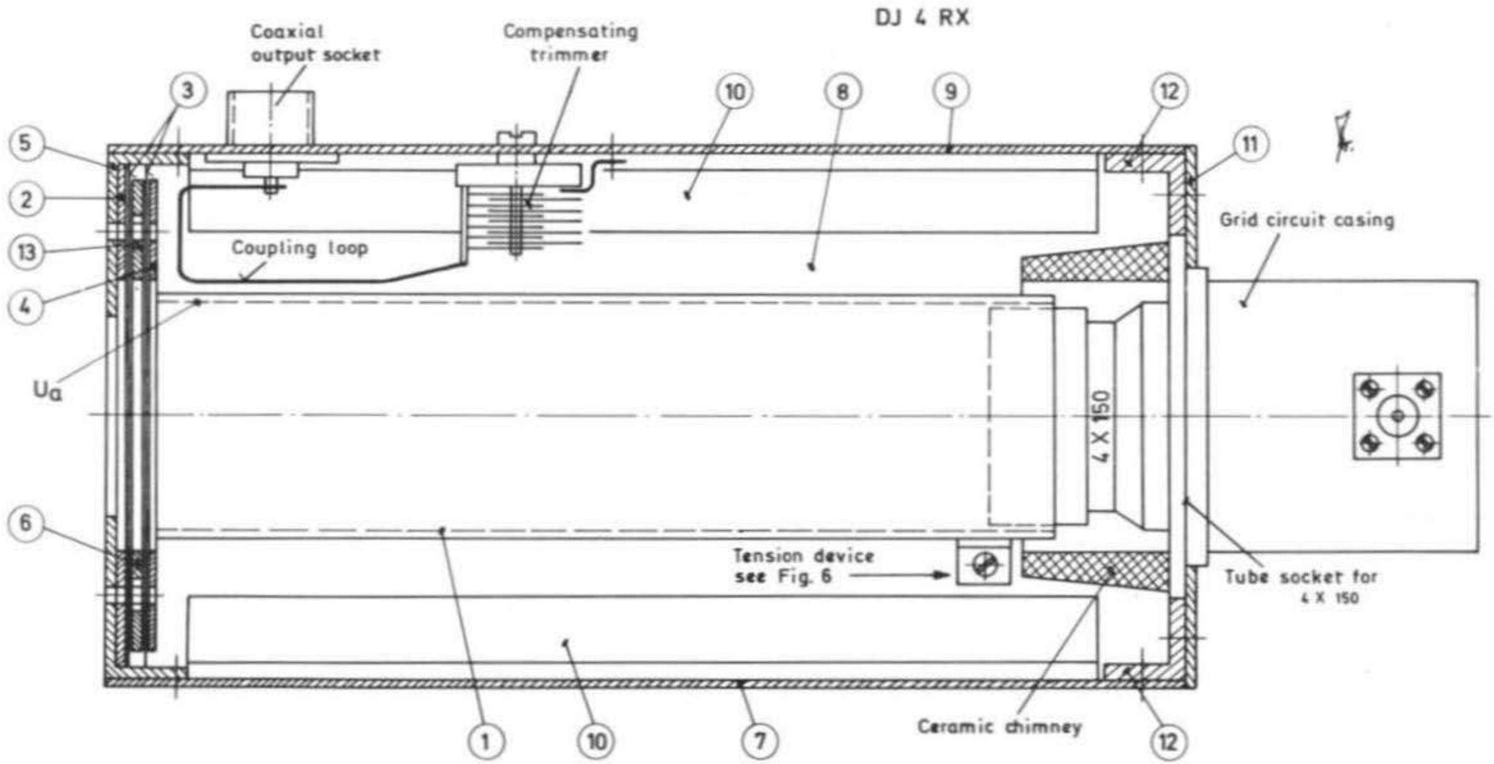


Fig. 8: The $\lambda/4$ anode circuit of the compact 144 MHz power amplifier

5.3. PARTS LIST FOR THE λ/4 ANODE CIRCUIT

Part	Figure	Number	Designation and material
1	5a	1	Helix, Brass tube 45 mm outer diameter, 1.5 mm thick
2	5b	1	Plate, (RF-bypassing), aluminium or brass plate 1.5 mm thick
3	5c	2	Insulating foil, teflon or mica 0.1 mm thick (Transparent plastic)
4	5d	1	Plate (RF-bypassing), aluminium or brass plate 1.5 mm thick
5	5e	4	Brackets, 30 x 15 x 2 mm aluminium
6	5f	1	Plate, copper or brass plate 2 mm thick; is soldered to Part 1
7	7a	1	Base plate, aluminium or brass plate 1.5 mm thick
8	7b	2	Side plates, aluminium or brass plate 1.5 mm thick
9	7c	1	Covering plate, aluminium or brass plate 1.5 mm thick
10	7e	4	Brackets, 15 x 15 x 3 mm aluminium
11	7f	1	Base plate, copper or brass plate 2 mm thick
12	7g	4	Brackets, 15 x 15 x 3 mm
13	8	8	Insulating discs, 6 mm outer and 3 mm inner diameter, 2 mm thick

Further items that are required:

1 4 X 150 tube socket with bypass capacitor for the screen grid, type SK-600 or similar
1 4 X 150 tube or similar (4 CX 250 B)
1 ceramic chimney for the tube socket (usually supplied with the socket)
1 coaxial socket, as required
A number of screws, tapped rivits and possibly washers

If a tuning disc is to be provided for the circuit, a 40 mm diameter, 1.5 mm thick copper or brass disc will be sufficient. Furthermore, a 40 mm long screw as well as a nut and locking nut will be required. In this case, a hole must be provided in Part 9, where the nut is soldered into place. The screw is soldered to the centre of the disc. The disc itself serves as a capacitive electrode (see Fig. 4).

6. THE SCREEN GRID SUPPLY

Tetrodes and pentodes require a well stabilized screen grid supply in class AB_1. If this is not the case, a poor linearity will result. With its maximum dissipation power of 12 W, the screen grid of the 4 X 150 is very sensitive to overload. In addition to this, the screen grid current can become negative at low anode current values (low residual currents). The screen grid supply should therefore offer:

a) a constant voltage during normal operation

b) be able to cope with negative screen grid currents

c) be able to avoid an overload of the screen grid and thus avoid possible damage to the PA tube.

It is the screen grid that is most endangered when an anode voltage failure should occur. This is because it attempts to take over the anode current.

These considerations do not allow an electronically stabilized power supply. Due to the negative current, a certain degree of load must be provided. An automatic overload protection would be, of course, too elaborate. Figure 11 recommends a circuit which is very simple. The maximum power that can be taken from the circuit is limited to approximately eleven watts. This ensures that an overload cannot occur. The voltage of 600 V is easily obtainable from another source.

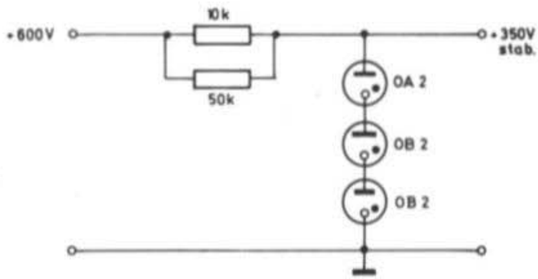


Fig. 11: Screen grid supply of the 144 MHz power amplifier

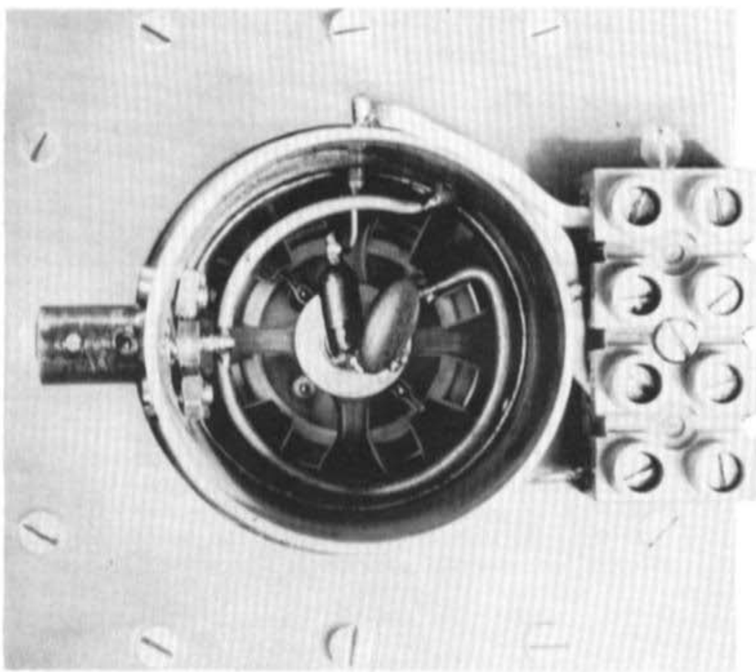


Fig. 13:

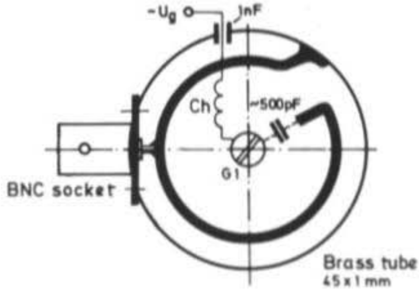
Photograph of the grid circuit

7. THE GRID CIRCUIT

Fundamentally speaking, all characteristics of the output circuit are also valid for the grid circuit. Especially the bandwidth should be great enough so that the circuit need not be tuned when varying the frequency within the two metre band.

The grid bias voltage is most favourably fed via a shunt fed choke. In this case, the circuit possesses ground potential at all points. Figure 12 and 13 show the practical grid circuit as a sketch and photograph. As can also be seen in Figure 8, a brass tube of 45 mm diameter, 1 mm thick and 50 mm in length is fixed to the tube socket on the opposite side to the anode circuit. An inner conductor made out of 2 mm diameter (12 AWG) silver-plated copper wire is soldered into this tube according to Figure 12. This represents a $\lambda/4$ line circuit with an impedance Z of approximately 100 Ω, which is greatly shortened by the input capacitance of the 4 X 150 tube. By bending the inner conductor of the grid circuit (altering the line impedance), it is possible to tune the circuit. The injection of the RF drive power from the exciter is made galvanically via a coaxial socket. The free-end of the line circuit is connected via a capacitor of 500 pF to the grid connection of the tube socket. The grid leak is taken from this point via a $\lambda/4$ choke and a feed-through capacitor of 1 nF from the tube. The open end of the grid circuit tube is connected to a ventilation blower.

Fig. 12: The grid circuit
The inner conductor is made from 2 mm dia. (12 AWG) silver-plated copper wire according to drawing



A second possibility is for the grid to be matched using a piece of RG 59 A/U coaxial cable. Due to its higher impedance of 75 Ω, RG 59 A/U is more suitable than RG 58 A/U. The determination of the length is shown graphically in the form of a Smith Chart diagram in Figure 14; the dimensions are given in Figure 15. These considerations assume a drive power of approximately 2 W at the grid of the PA tube.

The dimensions given in Figure 15 differ slightly from those determined in the Smith Chart. This is because the lead from the actual grid circuit to the grid must also be considered.

7.1. IMPROVING THE GRID CIRCUIT MATCHING

Generally speaking, one should not place too strict requirements when checking the matching or standing wave ratio at the exciter side of the PA circuit. Naturally a SWR of 1:1 would be very favourable. However, if the PA stage is to be used in various modes, for instance in class AB_1 and class C, it is only possible to align the grid circuit for optimum in one mode. This is because

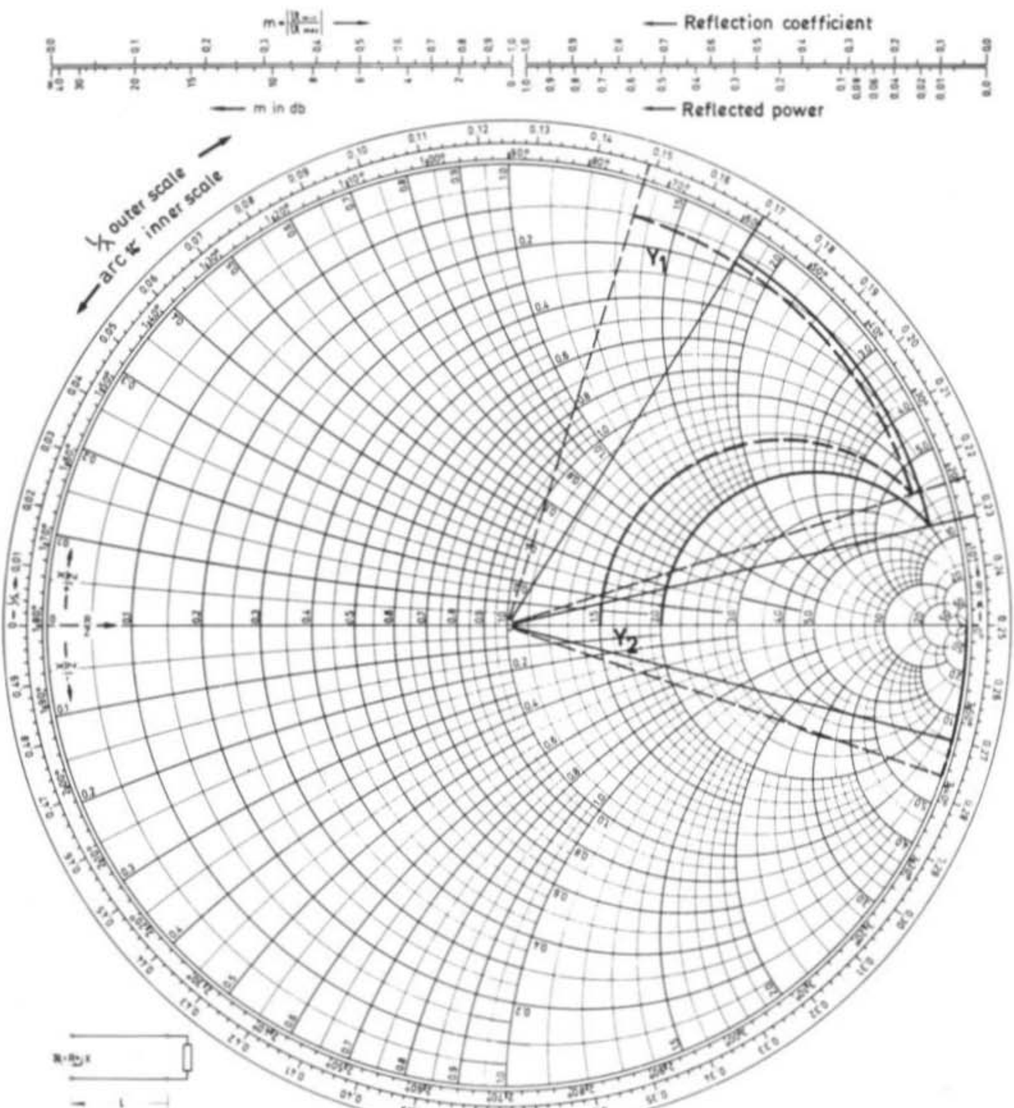
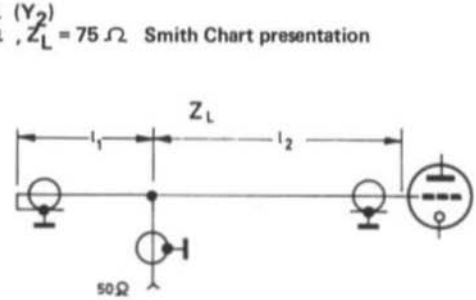


Fig. 14: Matching the grid conductance Y_1 to 50 Ω (Y_2)
Impedance of the grid circuit: Z_L = 100 Ω , Z_L' = 75 Ω Smith Chart presentation

Fig. 15: Grid circuit of a 144 MHz power amplifier with a 4 x 150 A tube Z_L = 75 Ω , l_1 = 36 mm, l_2 = 75 mm



the grid impedance is greatly dependent on the operating point and the grid current. This can only be avoided using an additional matching link. Such a matching link consists basically of three trimmers of approximately 20 pF each and two λ/4 coaxial cables. Figure 16 shows such an arrangement with which a maximum standing wave ratio of 3:1 can be compensated.

A matching link as shown in Figure 16 also allows a mismatch condition to be avoided which can be caused by a coaxial relay possessing impedance irregularities. The circuit given in Figure 17 simplifies matters when the resistance to be matched is greater than the impedance of the cable. Such a matching link is of great assistance in improving the standing wave ratio, especially when the PA stage is to be used for several modes. It is only necessary for the preliminary alignment to be made in the range having the maximum mismatch.

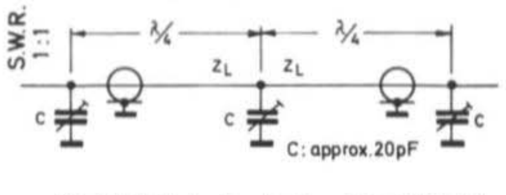


Fig. 16: Grid circuit as load **Max. SWR: 3:1**

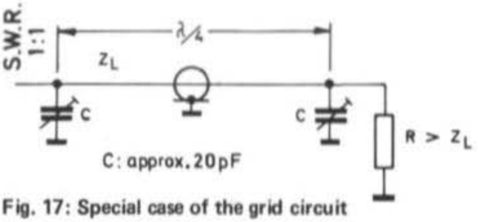


Fig. 17: Special case of the grid circuit

8. ALIGNMENT OF THE COUPLING LOOP

Additional instruments are required to find the most favourable adjustment of the coupling loop:

a) A variable signal generator for 145 MHz; maximum RF output power approximately 3 to 4 W.
b) A relative RF indicator, e.g. a reflectometer
c) A dummy load (if it possesses a calibrated wattmeter, "b" will not be required).
d) A variable grid bias source to adjust the residual anode current.

The alignment is best made at half the anode voltage since the power dissipation of the PA tube can be very high on commencing the alignment process. In the first approximation, the PA stage behaves as an ohmic resistance, which is the reason why this procedure is permissible. The operating data are:

U_a = 1250 V I_{ao} = 60 mA

I_a = 200 mA (single tone under full-drive conditions)

The alignment is carried out at:

U_a = 600 V I_{ao} = 30 mA I_a = 100 mA

Figure 2 in Part 1 of this article shows the operating slope in a simplified family of characteristic curves for a tetrode, even for half the anode voltage. The operating voltages of the PA stage are switched on in the following order:

1. Ventilation and heater voltage
2. Grid bias voltage
3. Anode voltage
4. Screen grid voltage (at the 600 V side; the stabilizing tubes remain connected to the PA tube at all times).

The residual anode current of the PA tube is now adjusted with the aid of the grid bias voltage. If the dummy load has been connected, slowly increase the drive until the anode current is slightly increased. This is followed by tuning the anode and the coupling capacitor alternately for maximum output power. As soon as a maximum has been found, it is possible for the most important measurement to be made: The drive is increased still further whereby both anode current and RF output voltage are observed. At first, both instruments will increase in the same manner until a point is reached where the output voltage is not, or only slowly, increased. If the output voltage is traced as a function of the anode current, the resulting curve will be similar to that shown in Figure 18. The most decisive point for the matching is the value of $I_{a\ max}$, which is the point where the output voltage slope is still proportional to the drive. In our example, $I_{a\ max}$ must correspond to 100 mA in order to achieve the most favourable impedance for operation. If the measured current is higher, this will mean that the output coupling is too great and that the coupling loop must be shortened.

The measurement is repeated with the shortened coupling loop. If $I_{a\ max}$ is too small, it will be necessary to use the opposite process until the characteristic curve of the output voltage U_{RF} as a function of I_a bends at 100 mA.

Finally, the full anode voltage of 1250 V can be connected. The residual current should be increased to approximately 60 mA. Using the same procedure, the bend of the characteristic curve as shown in Figure 18 is adjusted to correspond to approximately 200 mA. In this position, the most favourable output coupling will have been found.

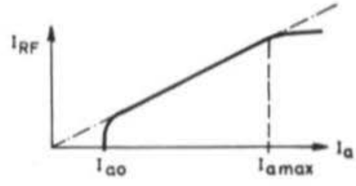


Fig. 18: Relationship between RF output voltage and anode current on aligning the power amplifier

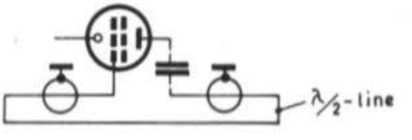


Fig. 19: Neutralizing circuit of the power amplifier

9. NEUTRALIZATION

A tendency to self-oscillation was noticed very seldom with previously constructed PA stages. When they appeared, they were very easy to neutralize. The impedance of a tuned-plate tuned-grid (TPTG) oscillator always has an inductive reactive component at the resonant frequency whereas the grid impedance is slightly capacitive. If the resonant frequency of the grid circuit is adjusted to be slightly above that of the output circuit (if the operative Q are comparable), conditions favourable to oscillation will not occur at the grid-anode capacitance of the 4 X 150.

The circuit given in Figure 19, is advantageous when neutralization is necessary. A λ/2 line is more suitable than a push-pull grid circuit, since the latter cannot be classed as ideal phase reversal link due to the unbalanced load.

The λ/2 line represents practically no load to the grid circuit.

CHEAP VARACTOR DIODES FOR THE 70 cm TRANSMITTER USING AN EC 8020 TUBE

by H.-J. Franke, DK 1 PN

A 70 cm transmitter was described in Edition 4/1969 (1) of VHF COMMUNICATIONS whose PA tube was driven by a varactor tripler. A professional, high power varactor diode (BAX 11 or MA 4061 B) was used in the frequency tripler stage from 144 to 432 MHz. Since these diodes are rather expensive, they are not readily available to the amateur. This was the reason why the author examined whether cheap tuning diodes, as used in the radio and television technology, could be used. The results were as follows:

Fundamentally speaking, the UHF tuning diode types BA 110 (ITT-Intermetall), BA 121 (AEG-Telefunken) or MA-45034 (Microwave Associates) also provide a sufficient power of approximately 0.5 W for driving the EC 8020 tube. However, it is necessary for a certain amount of experiment to be made with several examples of the chosen type to find one that is suitable for this application. Suitable diodes will provide enough drive to allow a plate current of 60 to 70 mA with optimum tuning and matching. Another indication is that the diode with short wire connections does not get hot. As was given in (1), the anode voltage at this point is 200 V and the drive power at 144 MHz is approx. 1.5 W.

Unsuitable diodes will be indicated firstly by a lower plate current and secondly by a further reduction of the output power that will be seen a few seconds after switching on the drive. This effect is caused by the diode heating up, which can also be felt.

Interesting was the fact that good results were also achieved using VHF and VHF/FM radio diode types. Types BA 124 and BA 150 (AEG-Telefunken) were found to be suitable. The equivalent type MA-45043 of Microwave Associates should also be suitable but was not tried by the author. However, the percentage of suitable diodes seems to be lower for these types.

A series connection of two identical diodes was tried in order to reduce the power dissipation. However, this was not found to be favourable; in fact, since this arrangement could not be tuned correctly, virtually the whole drive power was converted into heat. The authors experiments have shown that it may be more favourable to purchase several cheap tuning diodes instead of one of the expensive professional types and to experiment to find the most favourable tuning and matching. However, no reserve of power dissipation is available to allow the tripler to be "overdriven" by a 2 m transmitter having more than 2 W output.

REFERENCES

(1) H. J. Franke: A Ten Watt Transmitter for 70 cm
VHF COMMUNICATIONS 1 (1969), Edition 4, pages 243-248

CORRECTIONS AND IMPROVEMENTS TO THE 9 MHz SSB CONVERTER WITH INTEGRATED CIRCUITS DJ 9 ZR 005

by G. Stroessner, DJ 2 VN

Unfortunately, some errors were included in the description of the DJ 9 ZR 005 IF-AF portion with ICs in Edition 3 of VHF COMMUNICATIONS 1 (1969). The following differences exist between the circuit diagram (Fig. 3) and the component location plan (Fig. 4c) which we would like to correct:

1. CORRECTIONS TO THE CIRCUIT DIAGRAM (Fig. 3)

1.1. The designations of transistors T 1 and T 2 were exchanged. The transistor in the upper right-hand corner below the S-meter is T 1, the lower transistor is T 2.

1.2. The capacitor connected between connection 4 of integrated circuit IC 2 and ground should have a value of 330 pF (and not 68 pF).

1.3. A 100 kΩ resistor must be drawn between PC-board connection 6 (+9 V bar) and the connection point of diode D 8, the 100 Ω resistor and the 4.7 nF capacitor (Fig. 3).

1.4. The capacitor between PC-board connection 3 and ground (parallel to the S-meter) has a value of 10 nF (and not 10 μF).

1.5. A 5.6 kΩ resistor is connected parallel to the two AF-limiter diodes (AA 143 or 1 N 277). This resistor was incorrectly marked in the circuit diagram as being 56 kΩ.

1.6. The 820 Ω resistor in the collector circuit of T 1 is not connected to PC-board connection 6 (+9 V bar) but to connection 2 (+9 V Rx).

2. CORRECTIONS TO THE COMPONENT LOCATION PLAN (Fig. 4c)

2.1. The ground-end capacitor of the capacitive divider across L 1 should have a value of 330 pF (not 68 pF).

2.2. The ground-end resistor of the base voltage divider should be 1.2 kΩ (and not 5.6 kΩ as given in the component location plan). The correct value was given in the circuit diagram.

2.3. The capacitor of the series RC-link connected parallel to choke Ch 2 should be 0.47 μF and not 10 μF.

2.4. A wire bridge must be made between the capacitor given in 2.3. and the 560 Ω resistor so that the "+9 V bar" is connected to the 1.8 kΩ resistor.

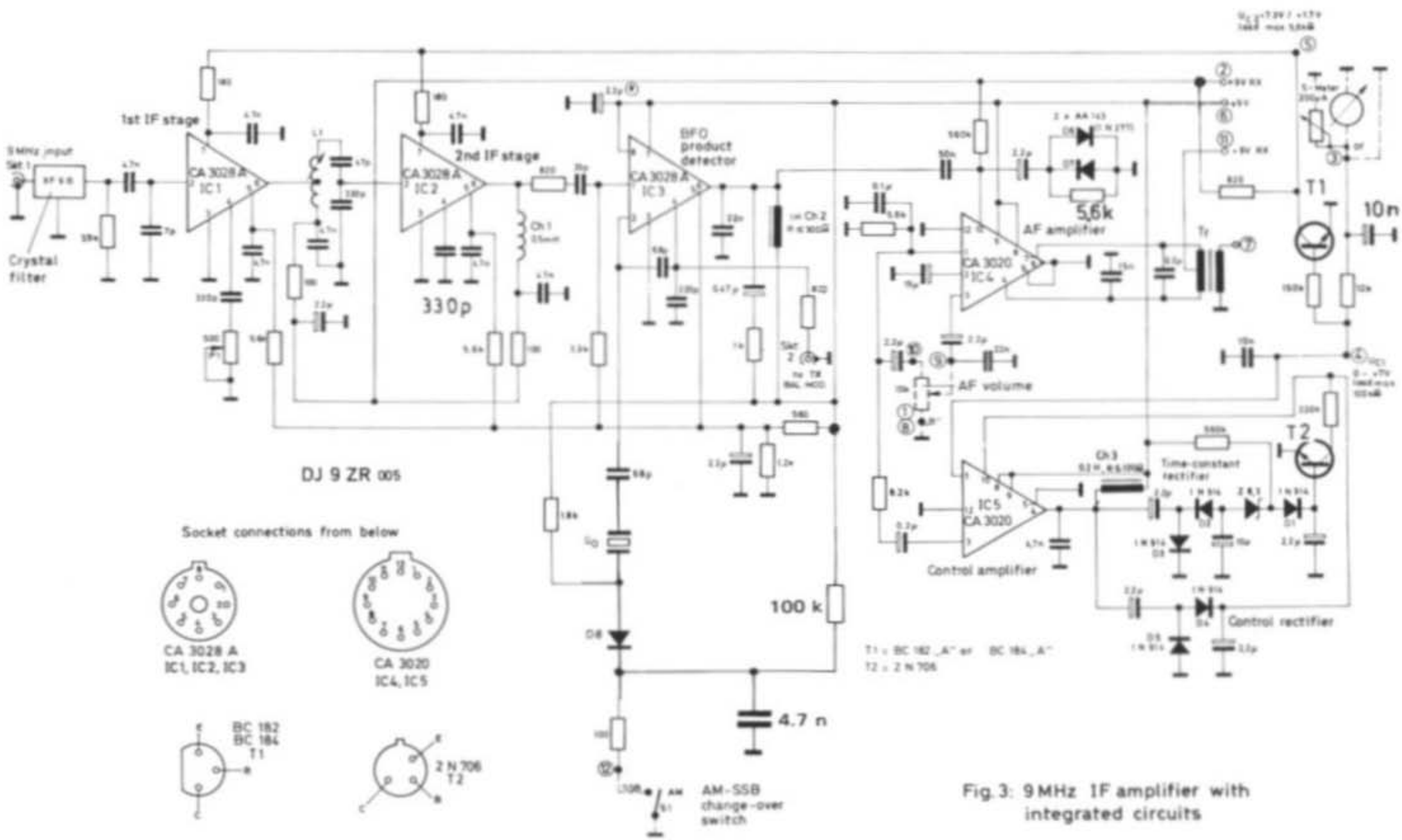
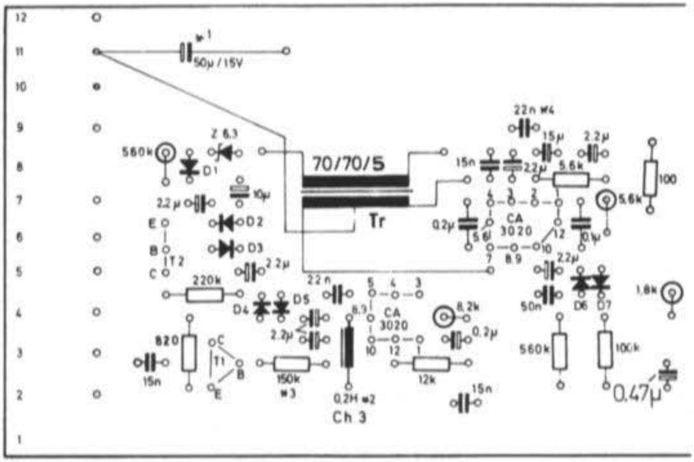


Fig. 3: 9 MHz IF amplifier with integrated circuits



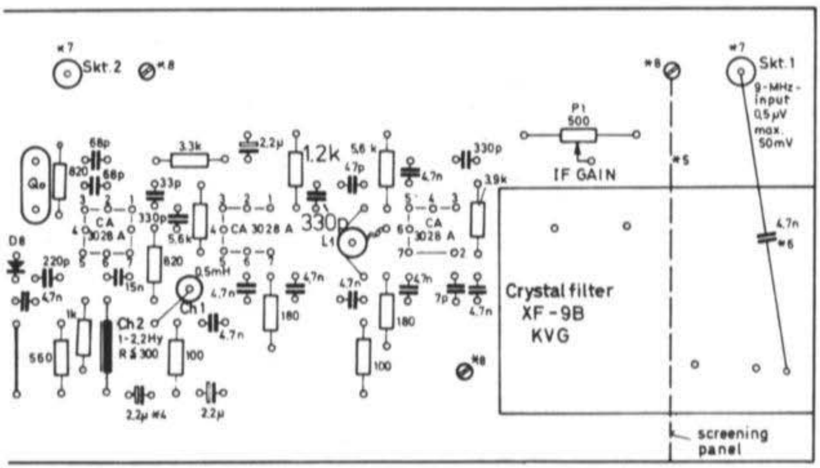


Fig. 4c: Component location plan

of the IF-AF amplifier DJ 9 ZR 005

VHF CONGRESS WEINHEIM(W. GERMANY) 1970

We would like to point out that the annual VHF Congress is once again taking place in Weinheim, near Heidelberg/W. Germany. This year the conference is to be held on the 19th and 20th of September. It offers continuous lectures by outstanding European VHF/UHF/SHF amateurs as well as facilities for discussion groups on diverse topics appertaining to amateur radio at the higher frequencies. We extend a cordial welcome to all VHF/UHF amateurs.

SUMMER HOLIDAY

The Publishers and the Material Sales Department will be taking their summer holiday during the month of August 1970. Since we are not able to dispatch orders or answer queries in this period, some delays could be encountered in receiving your orders or answers. If you require items within this period, please order them well before hand.

PRICE LIST OF NEW EQUIPMENT, DESCRIBED IN VHF COMMUNICATIONS 2/1970
For earlier equipment, see price list in edition 1/1970
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DL 6 HA 001/28 — 2 metre MOSFET converter (IF 28-30 MHz)	
Parts and kit as in VHF COMMUNICATIONS 1/1970	
Completed converter, ready to operate	DM 134.80
DL 6 HA 002 — 9 MHz SSB Transceiver	
PC-board only	DM 16.--
Coil former set	DM 3.50
Trimmer set	DM 2.30
Semiconductors	DM 51.10
Crystal filter XF-9A with both sideband crystals	DM 106.--
Crystal filter XF-9B with both sideband crystals	DM 137.--
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DL 6 HA 003 — 9 MHz Carrier oscillator	
PC-board only	DM 3.--
Coil former	DM -.40
Trimmer set	DM 1.70
Transistors	DM 6.20
Kit: PC-board, coil former, trimmers, semiconductors	DM 11.30
DL 3 WR 003 — 2 metre Transistor AM Transmitter	
PC-board only	DM 17.50
Semiconductors (with germanium complementary pair)	DM 91.90
Modulation transformer, core & PC-board mounting	DM 7.50
72 MHz quartz crystal, please state exact required frequency, delivery time 4 - 6 weeks	DM 21.50
Kit: PC-board, set of chokes, trimmer set, semiconductors, special drill	DM 123.80
DL 3 WR 004 — 2 metre RF voltage indicator	
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Kit: PC-board, trimmer, diode	DM 4.20
DJ 4 BG 002 — Digital FM Discriminator	
PC-board only	DM 5.--
Semiconductors	DM 22.--
Kit: PC-board, semiconductors	DM 27.--
DJ 4 BG 003 — Calibration-Spectrum Generator	
Parts and kit as in VHF COMMUNICATIONS 1/1970	
Trimmer only not DM -.40 but	DM -.80
DK 1 PN — Ten Watt Transmitter for 70 cm	
Tube EC 8020	DM 27.--
UHF Tube Socket, made from silicon-glassfiber	DM 2.90
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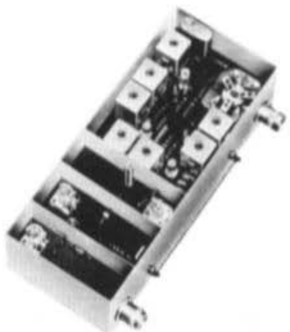

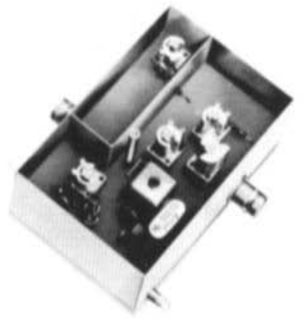

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This unit represents — in conjunction with a two metre station — the quickest and simplest means of becoming active on 70 cm. It has been especially developed for portable operation.
No antenna switching is required! The transverter is simply connected between the two metre station and the 70 cm antenna. Completely screened, silver-plated brass casing. The 432 MHz circuit is a true stripline circuit with 10 μ silver plating.
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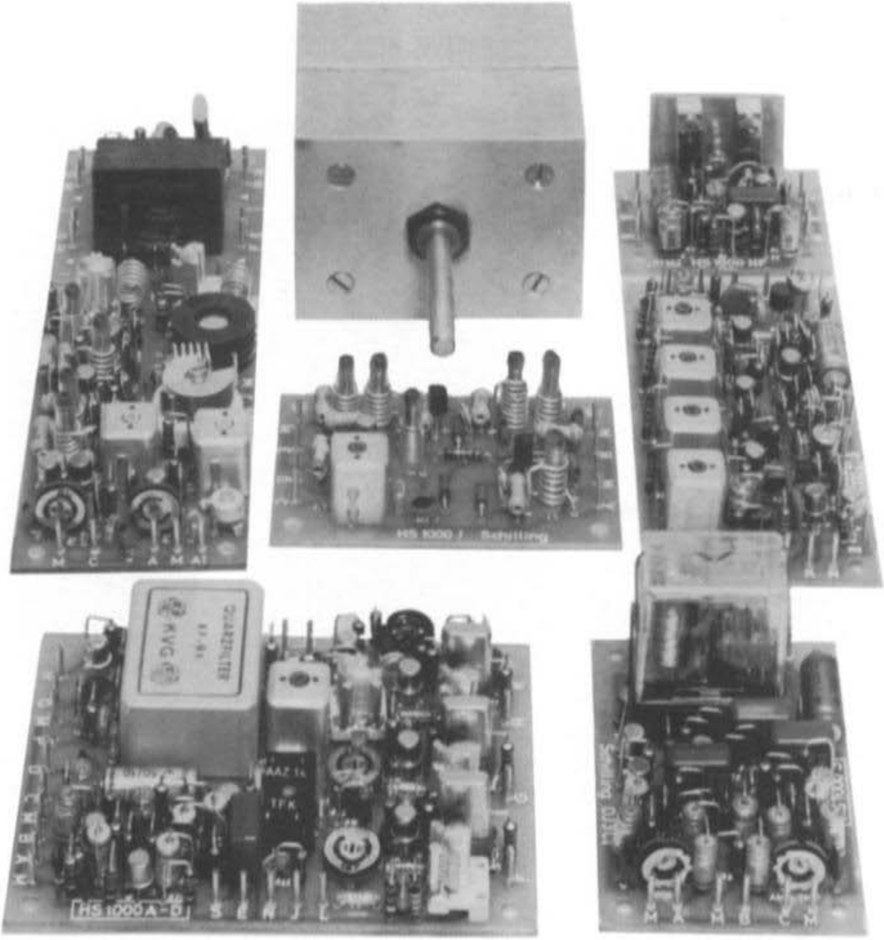
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