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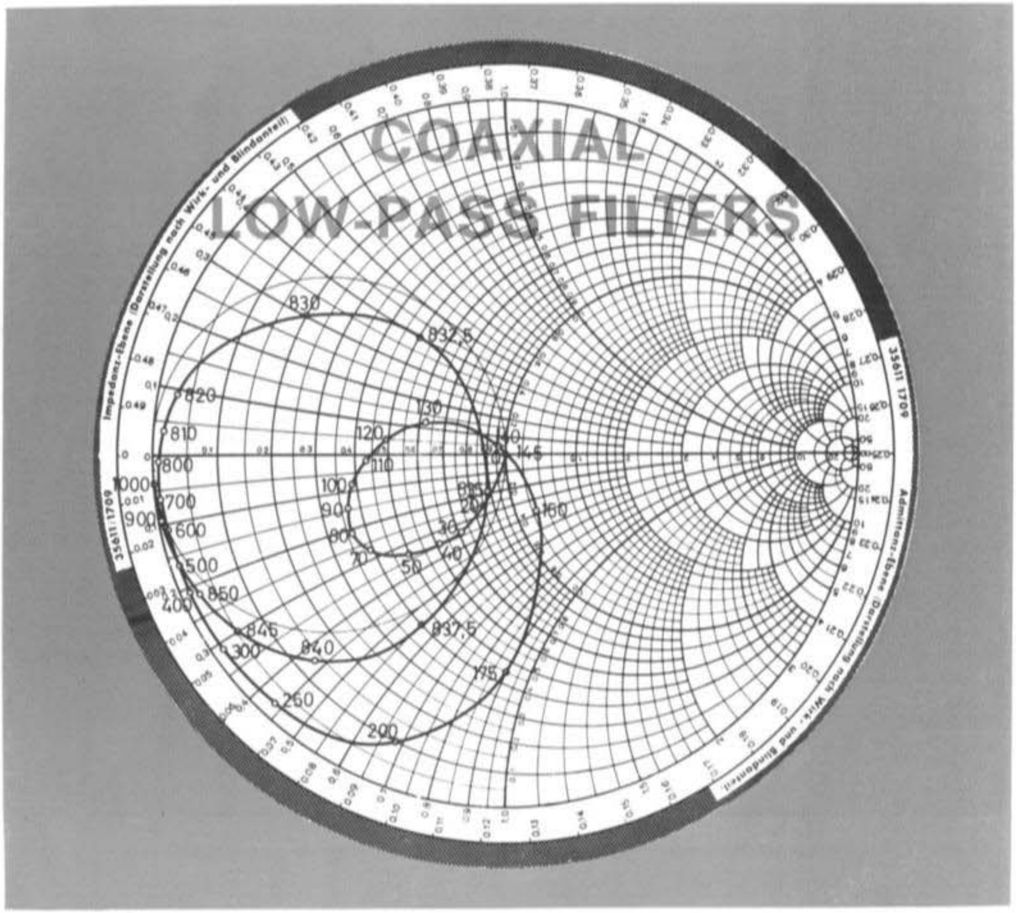

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A SSB TRANSCEIVER WITH SILICON TRANSISTOR COMPLEMENT

PART 3: THE 9 MHz-14 MHz TRANSMIT-RECEIVE CONVERTER 14 MHz-144 MHz TRANSMIT CONVERTER; VFO AND LOW-PASS FILTER

by G. Laufs, DL 6 HA

1. INTRODUCTION

Part 3 of this SSB transceiver article describes the 9 MHz-14 MHz transmit-receive converter, the 14 MHz-144 MHz transmit converter module with linear amplifier, as well as the 5 MHz VFO and matching low-pass filter.

These items are, of course, individual modules which can be used for other applications besides being used in the VFO transceiver system; the possibility of constructing a shortwave 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 has already been described in Part 1.

Part 2 described the 9 MHz transceiver module DL 6 HA 002 (2) and Part 4 will describe the power supply and AF amplifier modules as well as a special PC board for the two crystal oscillators of the modified MOSFET converter with an IF of 14 MHz.

2. THE 9 MHz-14 MHz TRANSMIT-RECEIVE CONVERTER

2.1. CIRCUIT DETAILS

The circuit diagram of this module is given in Figure 2. The transmit-receive converter possesses a complete field effect transistor complement; junction field effect transistors in the amplifier stages and dual-gate MOSFET types in the mixer stages. A further important feature having an effect on the behaviour of the total transceiver are the narrow band characteristics. This is obtained by a synchronous tuning of all 14 MHz circuits with the aid of varactor diodes. The tuning is made by means of a (external) potentiometer. This is more favourable with respect to a high spurious suppression (for reception: high ultimate selectivity) than a wideband amplification. The high input and output impedance of field effect transistors allow the high Q of this narrow band design.

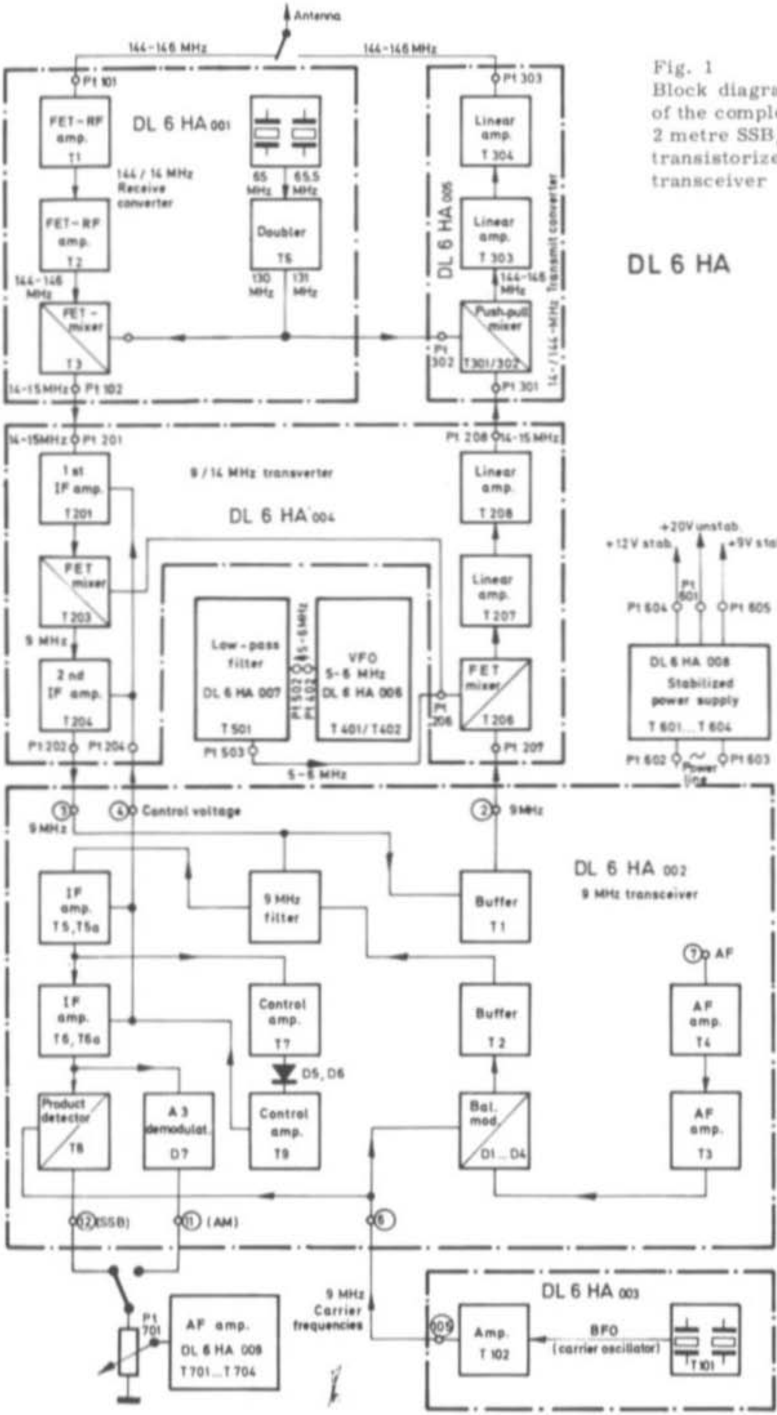


Fig. 1
Block diagram
of the complete
2 metre SSB/AM
transistorized
transceiver

DL 6 HA

On the receive side, the 14 MHz signal is fed via a 9 MHz trap to the input of the preamplifier stage equipped with transistor T 201. This trap is, however, only necessary when connecting the receive converter directly to an antenna for 14 MHz reception. For VHF operation, the two metre converter suppresses the 9 MHz signal to such an extent that the trap will not be required.

The 14 MHz preamplifier operates in a very stable common-gate circuit. The gain of this stage is controlled in the source branch by the pass transistor T 202. The 9 MHz amplifier stage (equipped with transistor T 204) subsequent to the mixer stage, also operates in this manner. The matching to the following crystal filter (and thus the pass band ripple) is considerably affected by the capacitor designated with ∗ in the drain circuit of this stage.

The oscillator voltage (via Pt 206) is attenuated for the transmit mixer (transistor T 206) by means of a capacitive voltage divider. The reduced drive of the mixer characteristic obtained in this way ensures that unwanted conversion products are not generated. The transmit mixer is followed by a two-stage selective amplifier in a common-gate circuit. This amplifier tracks the tuning of the selective amplifier in the receive converter.

2.2. MECHANICAL ASSEMBLY

The transmit-receive converter is accommodated on a printed circuit board having the dimensions 150 mm by 105 mm. This PC board, which has been designated DL 6 HA 004, is illustrated in Figure 3; the corresponding component location plan is given in Fig. 4. The transistors are the very last components to be mounted. The same protective measures should be made for the dual-gate MOSFETs as were given in (1). The connections of the transistors are shown from below. Attention should be paid to the correct polarity of the varactor diodes and that all DC operating voltage lines are bypassed using feed-through capacitors. A photograph of the DL 6 HA 004 module is given in Fig. 5.

2.2.1. SPECIAL COMPONENTS FOR THE DL 6 HA 004 MODULE

T 201, T 204, T 207, T 208: BF 245 C, BF 244 (Texas Instruments), 2 N 5284
T 202, T 205: 2 N 2926, BC 148 A or B, BC 108 A or B
T 203, T 206: 40603 (TA 7150), 40604 (TA 7151), (RCA)
D 201 to D 205: BA 111 (ITT-Intermetall), BA 150, BA 124 (AEG-Telefunken) or MA-45043 (Microwave Associates)
Trimmer capacitors (5 required): 10-40 pF or 10-60 pF ceramic disc trimmers

All coils are close-wound onto the eight 5 mm diameter coil formers with SW core. When not otherwise mentioned, 0.1 mm dia. (38 AWG) silk-covered enamelled copper wire is used.

L 201, L 204, L 209: 20 turns, coil length: 5 mm
L 202: 3 turns of 0.3 mm dia. (29 AWG) silk-covered enamelled copper wire wound onto L 203
L 203, L 207, L 208: 20 turns; coil tap 5 turns from cold end; coil length 5 mm
L 205: 27 turns; coil tap 6 turns from the cold end; coil length 7 mm
L 206: as L 205 but without tap
L 210: 3 turns of 0.3 mm dia (29 AWG) of silk-covered enamelled copper wire wound onto L 209

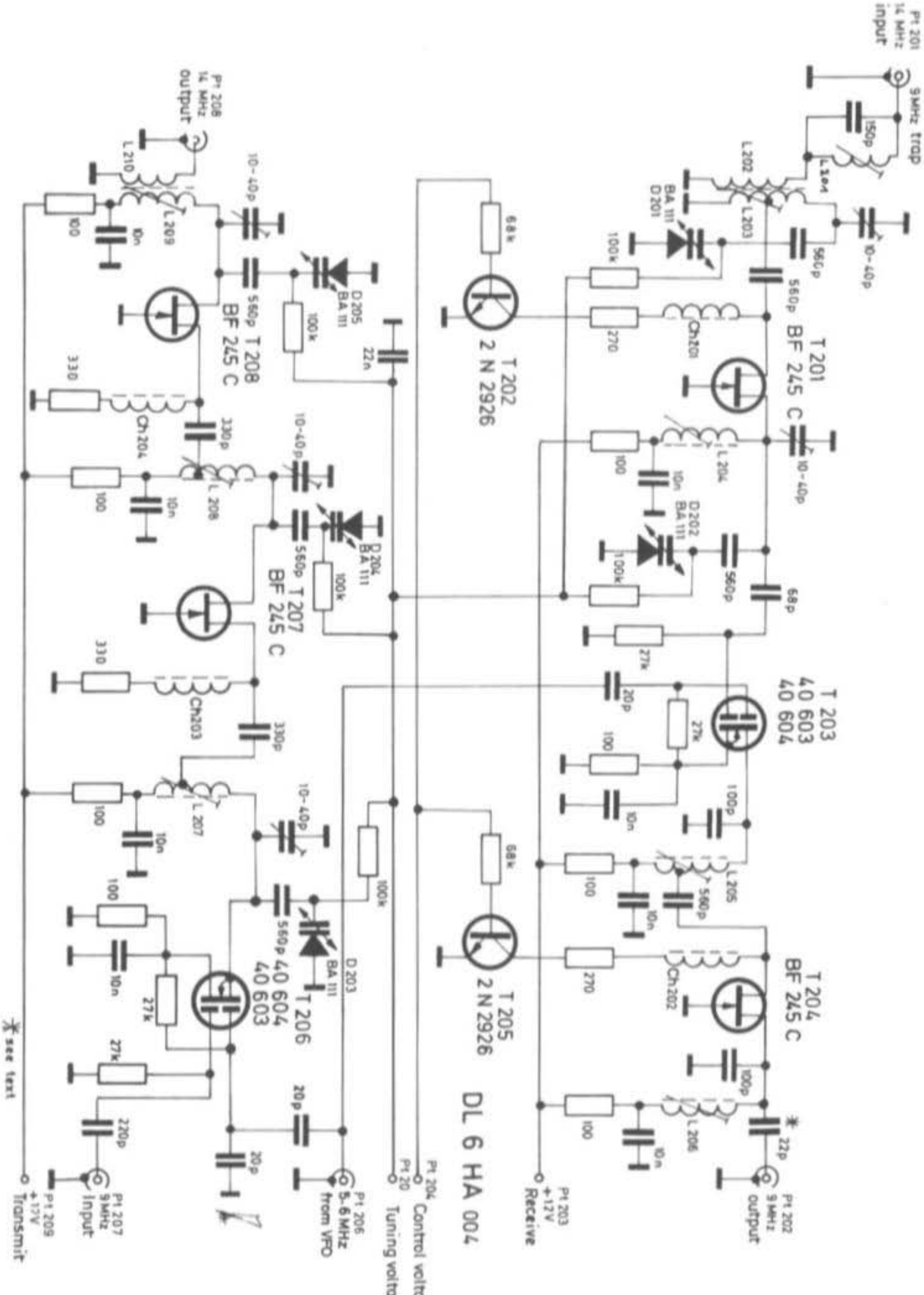


Fig. 2: Circuit diagram of the 9-MHz — 14-MHz-transmit-receive converter

Ch 201, Ch 202: Wideband ferrite choke having approx. 50 μH (50 turns of 0.1 mm (38 AWG) silk-covered enamelled copper wire wound on a 5 mm dia., 10 mm long ferrite core)

Ch 203, Ch 204: Wideband Ferroxcube chokes (2.5 turns through a 6 hole bead, Z = 850 Ω, e.g. 4312 020 36700 from Philips)

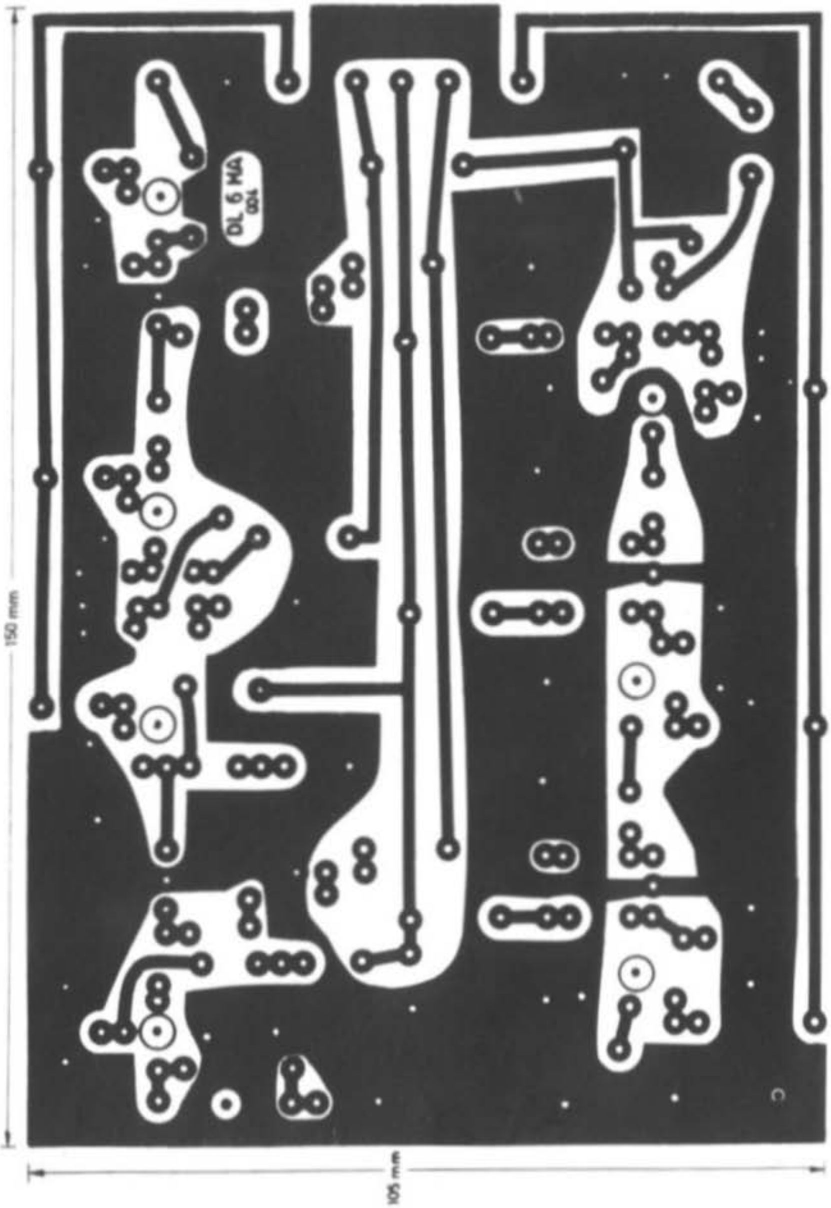


Fig. 3: Printed circuit board DL 6 HA 004

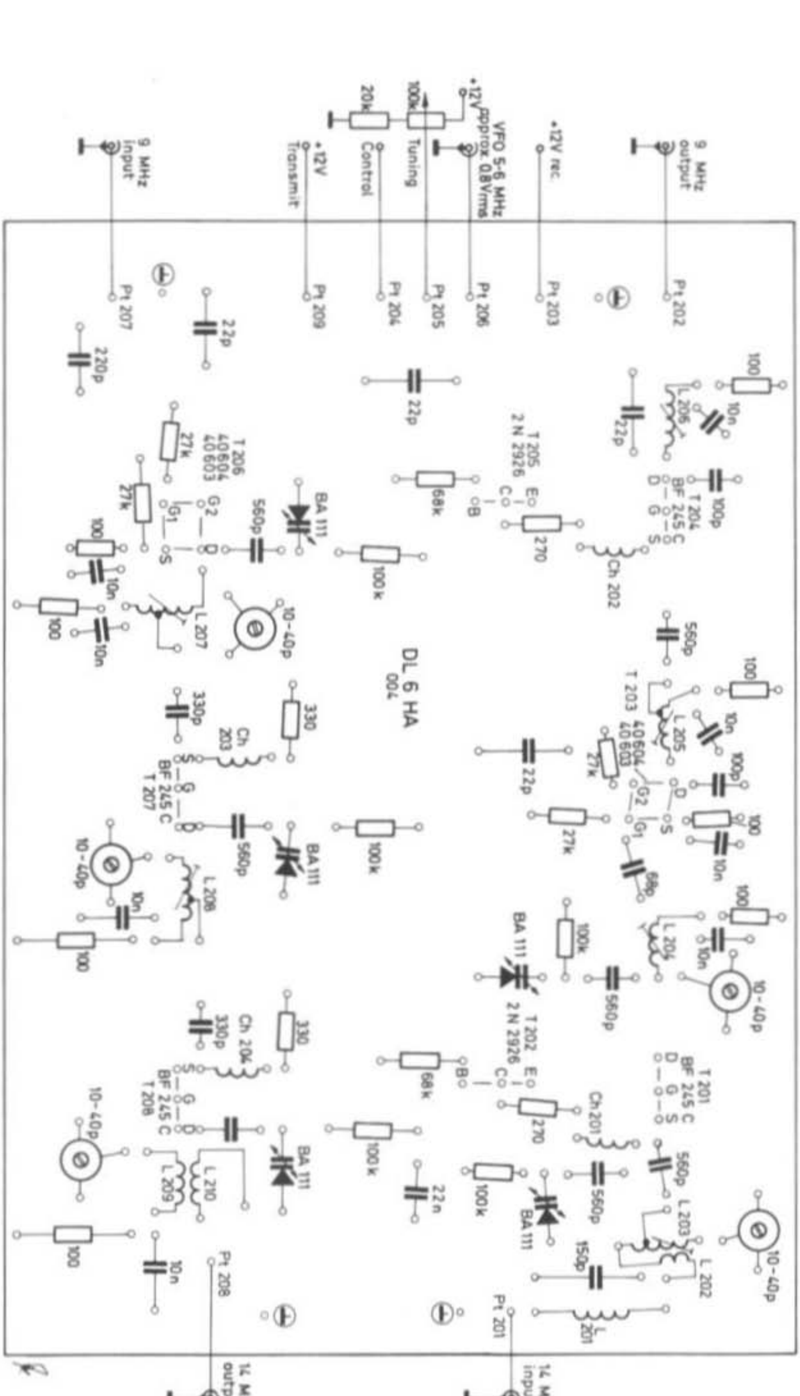


Fig. 4: Component location plan to DL 6 HA 004

2.3. ALIGNMENT

The following order is recommended for the alignment: Connect a DC voltage of 12 V to point Pt 203 of the receive converter and the minus pole to ground. A DC voltage of +12 V is also connected to the control voltage connection Pt 204 and the tuning voltage connection Pt 205. Using a high-impedance multimeter (20.000 Ω per Volt), the following voltages should be measured for orientation: Approximately 0.3 V - 0.5 V at the source connection of the receive mixer (T 203), 0 V at gate 1 and somewhat less than the source voltage at gate 2. If higher voltages are measured at gate 1 or gate 2 than at the source electrode, this will indicate that the transistor is defective (probably destroyed when mounting onto the printed circuit board). A voltage of between 1.5 V and 1.8 V should be measured at the source electrodes of the junction field effect transistors T 201 and T 204. Small variations can occur at this point due to series fluctuations of these transistors. Modification of the source resistors, however, allows this to be compensated. If the control voltage connection Pt 204 is connected to ground instead of to +12 V, the voltage at the source electrode of the above mentioned transistors should rise to between +4 V and +5 V.

Connection point Pt 204 is now reconnected to +12 V and the two input circuits comprising inductances L 203 and L 204 are aligned with the aid of a dip meter to 15 MHz; the IF circuit (inductance L 206) is aligned to 9 MHz.

This is followed by operating the transmit converter. A voltage of +12 V is connected to point Pt 209; point Pt 205 remains connected to +12 V. Using the DC multimeter, determine whether 0.3 V - 0.5 V can be measured at the source electrode of the transmit mixer transistor T 206. The same voltages are valid at gate 1 and gate 2 of this transistor as for the receive mixer. At the source electrodes of the two common-gate circuits, it should be determined whether a voltage of 1.5 V - 2 V can be measured. After this, the three circuits comprising inductances L 207, L 208 and L 209 should be aligned with the aid of a dip meter to 15 MHz. This should not be achieved with the trimmers in their minimum capacitance position, If the IF trap is used, it should be aligned to 9 MHz in conjunction with a dip meter before mounting.

The following alignment procedure requires that the 9 MHz transceiver module DL 6 HA 002 has been completed and aligned. This module was described in the last edition of VHF COMMUNICATIONS.

The alignment in conjunction with the 9 MHz transceiver module is made as follows: Connect +12 V to point Pt 10 of printed circuit board DL 6 HA 002 and Pt 203 of DL 6 HA 004. The 9 MHz output (Pt 202) is now connected to the 9 MHz input Pt 3 of DL 6 HA 002. A short screened RF cable should be used for this interconnection. Connect the control voltage connection Pt 204 with point Pt 4 of module DL 6 HA 002. The tuning voltage connection Pt 205 is connected to the wiper of a 100 kΩ potentiometer that is connected as shown in Figure 4. The variable frequency oscillator (module DL 6 HA 006) is connected to point Pt 206. After commencing operation, a RMS value of approx. 0.7 V should be measured at gate 2 of the mixer transistor T 203. A modulated signal of approximately 15 MHz is now connected to the 14 MHz input connection Pt 201 and the VFO tuned to approximately 6 MHz.

The signal will be audible on a temporarily connected audio amplifier connected to point Pt 11, or on a headset. If the signal is unmodulated, the carrier oscillator on module DL 6 HA 003 can be switched into operation, point Pt 105 connected to Pt 6 and the audio amplifier connected to Pt 12.

The voltage of the 14 MHz signal is now adjusted so that the S meter just shows a reading. After this, the resonant circuits comprising inductances L 203, L 204, L 205 and L 206 should be aligned for maximum meter reading. For this alignment process, the tuning potentiometer should be at maximum (12 V).

The input frequency is now tuned to approximately 14 MHz and the VFO tuned correspondingly to approximately 5 MHz. The tuning potentiometer is now adjusted for maximum S meter reading and the potentiometer position marked for the transmit alignment. This is followed by realigning the resonant circuits comprising inductances L 203 and L 204 for maximum. This alignment is repeated until the required tracking is obtained.

A stable, unmodulated input signal is required for checking the crystal filter matching. The VFO is slowly tuned over the receive frequency, whilst observing the S meter - which should indicate approximately 1/3 of FSD. The fluctuations of the S meter reading indicate the ripple in the pass band range. If the ripple is greater than 3 dB (1/2 S-point), it will be necessary for the capacitor designated with a + in Figure 2 to be replaced by a 10 - 40 pF trimmer capacitor which is then adjusted for minimum ripple.

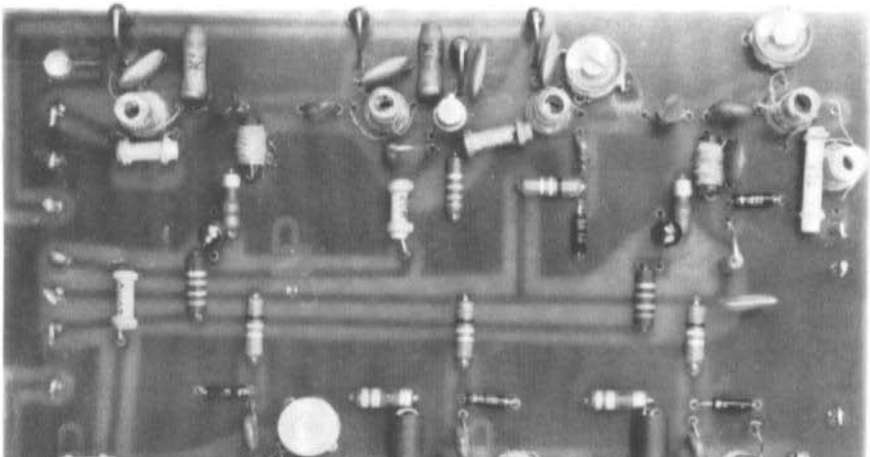


Fig. 5: Photograph of the transmit-receive converter

In order to align the transmit converter, a voltage of +12 V is connected to point Pt 209, as well as to Pt 8 of the DL 6 HA 002 module. The 9 MHz input (Pt 207) is connected using a short coaxial cable to Pt 2 of DL 6 HA 002. The RF probe of a VTVM is connected to the 14 MHz output (Pt 208) of module DL 6 HA 004. The carrier injection device in the 9 MHz transceiver module (Pt 5) allows a carrier signal to be generated. The tuning potentiometer should be at its 12 V position. The VFO is now tuned to a frequency of approximately 6 MHz and the resonant circuits comprising inductances L 207, L 208 and L 209 are aligned with the aid of the corresponding trimmers to 15 MHz, i.e. for maximum reading on the VTVM. If the voltage is still not sufficient for a reading, it is possible for a receiver tuned to 15 MHz to be used for the indication.

The VFO is now tuned to approximately 5 MHz, the tuning potentiometer tuned to the position marked during the receive converter alignment and the previously mentioned circuits aligned once again for maximum reading - this time with the aid of the coil cores. As was the case with the receive converter alignment, this procedure must be repeated several times until the most favourable tracking is achieved. The carrier injection can now be switched off and the quality of the SSB signal may be checked on a receiver. It is, of course, necessary for the carrier oscillator (on module DL 6 HA 003) to be connected. In most cases, the SSB signal obtained from module DL 6 HA 002 is too great. With the aid of a small trimmer potentiometer of 1 kΩ, whose wiper is connected to the 9 MHz input, it is possible for the level to be reduced to the required value of approximately 100 - 150 mV.

To conclude the alignment process, the tracking between the receive and the transmit converter should be checked. If the tuning potentiometer is aligned for maximum S meter reading of a receive signal, the transmit signal should also be at maximum on switching to transmit.

Finally a very effective manual RF control is to be mentioned: If the control voltage (Pt 204) is connected to the wiper of a 250 kΩ potentiometer whose one end is connected to ground and the other to the control voltage output of module DL 6 HA 002 (Pt 4), an extremely effective manual gain control will be obtained.

2.4. MEASURED VALUES

The author measured the following values in conjunction with the 9 MHz transceiver module DL 6 HA 002 and module DL 6 HA 003.

2.4.1. TRANSMIT CONVERTER

Linear cross-talk : -40 dB at 1000 Hz when using the crystal filter XF-9A
(This was the same as measured for the 9 MHz transceiver)
Non-linear cross-talk : approx. -40 dB
Spurious signal suppression: approx. -75 dB

2.4.2. RECEIVE CONVERTER

Sensitivity: 0.5 µV into 60 Ω for 15 dB signal-to-noise ratio

IF breakthrough	without 9 MHz trap: approx.	-60 dB
	with 9 MHz trap: approx.	-80 dB
Image rejection:	approx.	-70 dB
Control range:	approx.	100 dB

Cross modulation 1) : required signal: 2 μV
An interfering signal of 18 mV having a modulation depth of 30% generates a cross modulation of 1% when spaced 50 kHz from the required signal.

Desensitation 1) : required signal: 2 μV
An unmodulated interfering signal of 30 mV causes a 3 dB reduction of the required signal when spaced 50 kHz from the required frequency.

1) These values should be better when using a crystal filter XF-9B because the XF-9A filter is quite wide at approximately 45 dB. This means that interfering signals can be passed to the subsequent transistors.

3. THE 14 MHz - 144 MHz TRANSMIT CONVERTER

The DL 6 HA 005 module is the transmit counterpart to the two metre receive converter described in Part 1. It is provided to transpose the 14 MHz SSB signal to 144 MHz. The subsequent two-stage amplifier provides an output power of approximately 300 mW. The required crystal-controlled signals of 130 MHz and 131 MHz are obtained from the receive converter described in (1) after having carried out the modification to the original converter to obtain an intermediate frequency of 14 MHz. Of course, the transmit converter can also be used together with other receive converters that provide a sufficient oscillator voltage. In addition to this, it is also possible for the transmit converter to be used in other equipment or with a different frequency synthesis, for instance in conjunction with a 10 metre signal. In this case, it is only necessary to redimension inductance L 302.

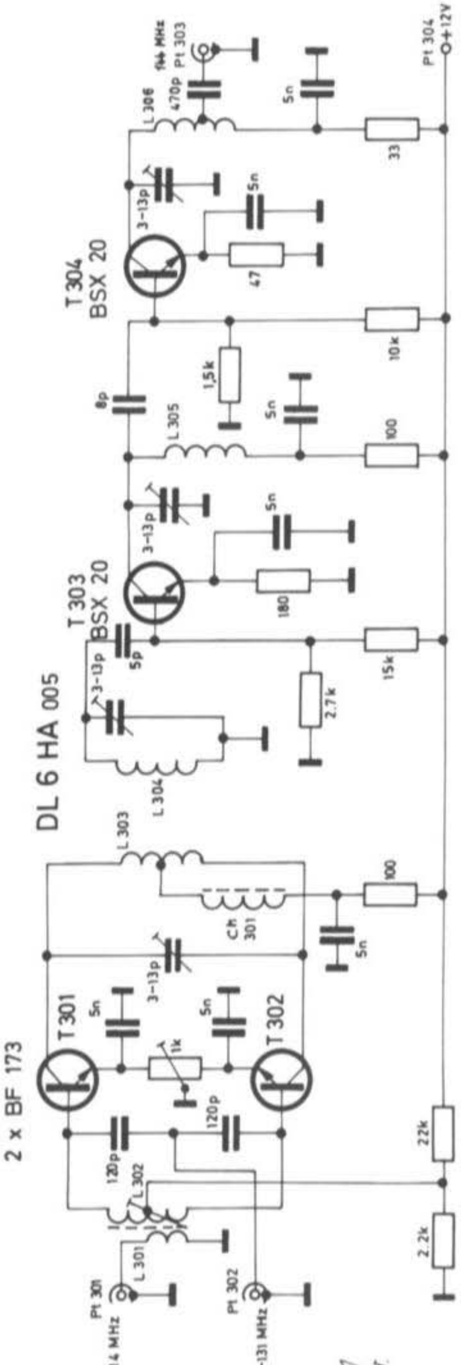
3.1. CIRCUIT DETAILS

The circuit diagram of this module is given in Figure 6. The 14 MHz signal is fed in push-pull to the push-pull mixer comprising the low-reactive transistors T 301 and T 302; the oscillator signal is fed in push-push. The DC balance can be adjusted with the aid of the potentiometer in the emitter circuit of the mixer transistors. The balance of the resonant circuits comprising inductances L 302 and L 303 also affects the value of the oscillator frequency suppression.

The subsequent two-stage linear amplifier improves the selectivity and amplifies the two metre signal to approximately 300 mW. The two transistors of the linear amplifier (T 303 and T 304) are one of the lowest power-transistor types. They are enclosed in the small TO 18 casing, but are able to be driven up to a collector current of 500 mA. The transit frequency f_T is at least 500 MHz, whereas the collector-emitter breakdown voltage is only 15 V when the base is floating.

3.2. MECHANICAL ASSEMBLY

The 14 MHz - 144 MHz transmit converter is accommodated on a printed circuit board having the designation DL 6 HA 005. The dimensions of this board, which is illustrated in Figure 7, are 62 mm by 114 mm; the corresponding component location plan is given in Fig. 8. The three stages are screened from another by a U-shaped screening plate (approx. 20 mm high). A photograph of the DL 6 HA 005 module is given in Figure 9.



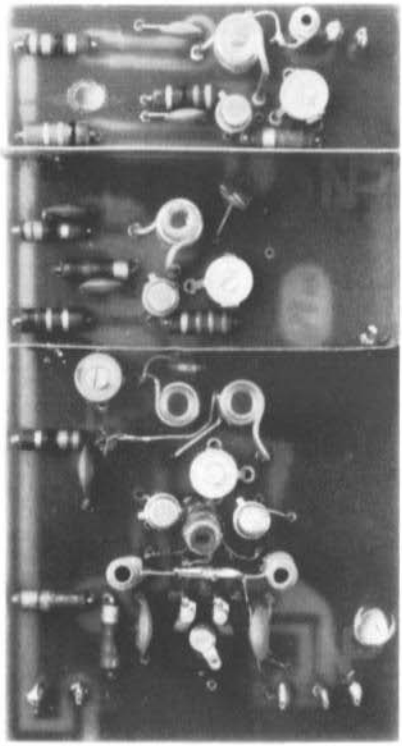


Fig. 9: Photograph of the 14 MHz –

144 MHz transmit converter

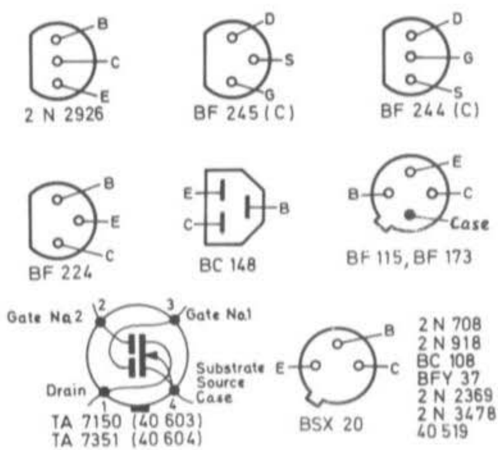


Fig. 6: Circuit diagram of the 14 MHz – 144 MHz transmit converter

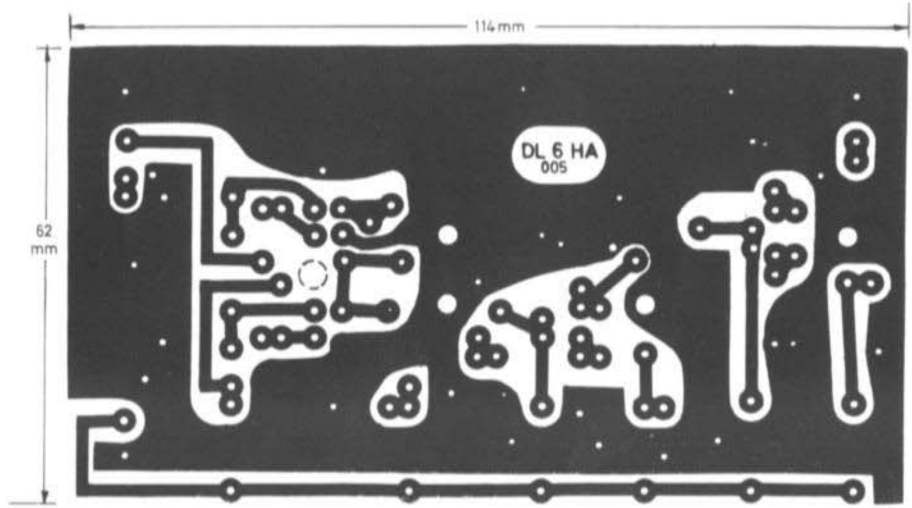


Fig. 7: Printed circuit board DL 6 HA 005

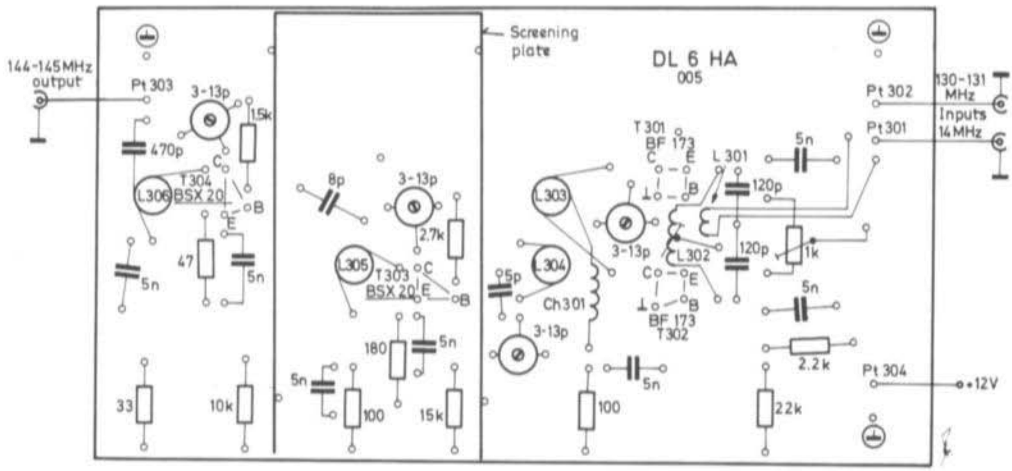


Fig. 8: Component location plan to DL 6 HA 005

3.2.1. SPECIAL COMPONENTS FOR THE DL 6 HA 005 MODULE

T 301, T 302 : BF 173 (Philips, Siemens, AEG-Telefunken),
BF 224 (Texas Instruments)

T 303, T 304: BSX 20 (Philips), 2 N 2369, 40 519 (RCA)

Trimmer capacitors (4 required) 3 - 13 pF, ceramic micro disc trimmers (7 mm dia. as in DL 6 SW converter).

L 301 : 3 turns of 0.3 mm dia. (29 AWG) enamelled copper wire wound onto L 302. Approx. 3 cm connections should be left and twisted together.

L 302 : 14 turns of 0.1 mm dia. (38 AWG) silk-covered enamelled copper wire on a 5 mm diameter coil former with SW core. Coil length: 4 mm.

All other inductances are wound with 1 mm dia. (18 AWG) silver-plated copper wire wound on 5 mm diameter coil formers without core. The supplied coil formers should be shortened.

L 303: 6 1/2 turns, 11 mm long
L 304 and L 305: 4 1/2 turns, 8 mm long
L 306: 6 1/2 turns, 12 mm long, tap at 2 1/2 turns

Ch 301: Wideband Ferroxcube choke (2.5 turns through a 6 hole bead, Z = 850 Ω) e.g. 4312 020 36700 from Philips. Mounted vertically.

3.3. ALIGNMENT

Terminate the 144 MHz output connection Pt 303 with a 50 Ω resistor. The 12 V operating voltage is connected to point Pt 304, and the minus pole to ground. The current drain should be approximately 35 mA.

A 145 MHz signal is now fed at approximately 0.5 V to the oscillator frequency connection Pt 302. This signal can be obtained from a dip meter or another VHF transmitter. The 145 MHz signal allows the amplifier stages to be aligned if the trimmer potentiometer is removed somewhat from its centre position. The RF probe of a VTVM is connected across the terminating resistor, after which the resonant circuits comprising inductances L 303, L 304, L 305 and L 306 are aligned for maximum reading.

This is followed by feeding a 130 MHz signal from the receive converter module DL 6 HA 001 instead of the 145 MHz signal. The voltage of the 130 MHz signal should be in the order of 0.5 V to 0.7 V. All voltage values are RMS values. The potentiometer is now adjusted for minimum reading on the VTVM.

A 14 MHz signal is now fed to the 14 MHz input (Pt 301); this signal can be obtained from the previously described module DL 6 HA 004, providing that the VFO module (DL 6 HA 006) is already in operation. Using the 14 MHz signal, the resonant circuit comprising L 302 is aligned for maximum reading. The resonance of this circuit is quite wide, which means that a detuning of 1 MHz will not cause any significant power loss.

The transmit converter should now be monitored on a SSB receiver, after which it will be ready for operation. If the signal level provided by module DL 6 HA 004 is too great, a 1 kΩ potentiometer can be used in a similar manner to that described in Section 2.3.

4. THE VFO MODULE

The frequency of the variable frequency oscillator is between 5 MHz and 6 MHz; it is used as the local oscillator for the frequency conversion from 9 MHz to 14 MHz - 15 MHz, or vice-versa for the receive conversion. The VFO module is designated DL 6 HA 006.

4.1. CIRCUIT DETAILS

The variable frequency oscillator shown in Figure 10 operates using a Clapp circuit. In order to reduce the dependence on load impedance variations, a buffer stage has been provided. The VFO module is operated from a stabilized 9 V supply which has been especially provided in the stabilized power supply.

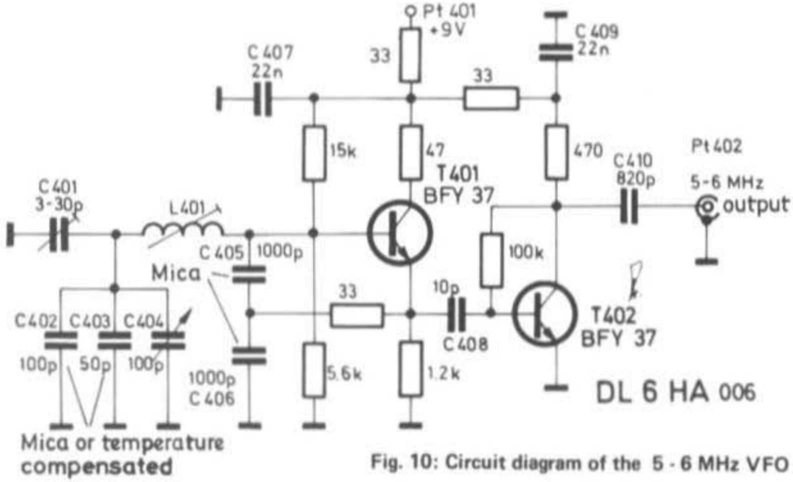


Fig. 10: Circuit diagram of the 5 - 6 MHz VFO

4.2. MECHANICAL ASSEMBLY

The VFO is accommodated on a printed circuit board having the dimensions 62 mm by 114 mm and is designated DL 6 HA 006. This printed circuit board is shown in Figure 11; the corresponding component location plan is given in Fig. 12. A photograph of the completed module allows further details to be gained (Fig. 13). Enough room has been provided for mounting a suitable variable capacitor. The mounting of the variable capacitor on the printed circuit board has the advantage that a better ground connection to the other resonant circuit elements is guaranteed. In addition to this, a more constant temperature is provided for all components, especially when the module is to be enclosed in a temperature insulated compartment. However, if it is more favourable for the variable capacitor to be mounted in another position, the non-required portion of the printed circuit board can be removed. it is with this module that the author has the least influence on the satisfactory operation of the equipment. This is because the selection and condition of the frequency-dependent components have a greater effect on the frequency stability of an oscillator than the dimensioning of the circuit and printed circuit board. The author therefore recommends that special care should be paid with respect to these components.

4.2.1. SPECIAL COMPONENTS FOR THE DL 6 HA 006 MODULE

T 401, T 402: BFY 37 (ITT-Intermetall), 2 N 918 (RCA),
BF 173 (Philips, Siemens, AEG-Telefunken),
BF 224 (Texas Instruments) or 2 N 708

C 401: 3 - 30 pF air-spaced trimmer e.g. spindle trimmer 2222803 20001(Philips)
C 402: 100 pF mica or ceramic capacitor for TC compensation
C 403: 50 pF as for C 402
C 404: 10 - 110 pF (Δ C = 100 pF) air spaced variable capacitor,
e.g. 2222 805 00005 (Philips)
C 405, C 406: 1000 pF mica capacitor

L 401: 32 turns of 0.3 mm dia. (29 AWG) silk-covered enamelled copper wire wound on a 10 mm dia. ceramic or similar coil former. Coil length approx. 10 mm close wound. SW core. The coil should be fixed with a little dual-component adhesive and aged by temperature cycling.

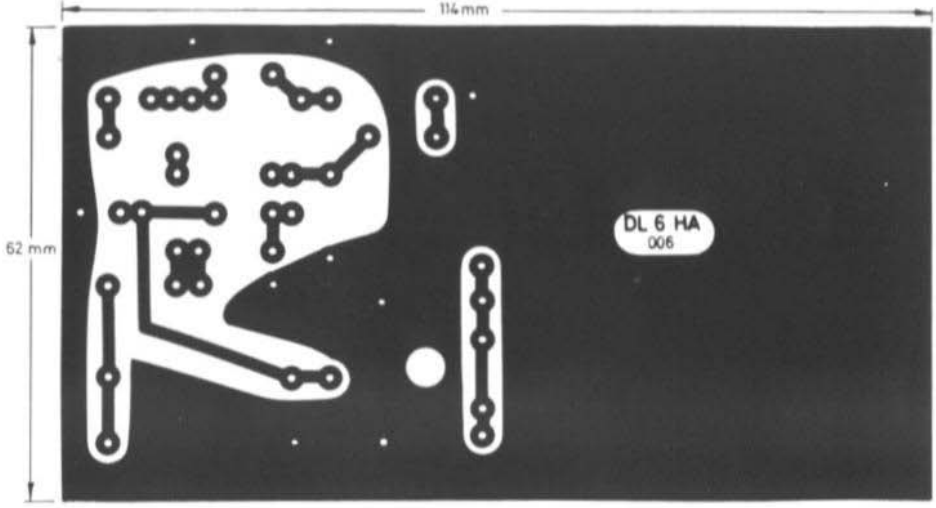


Fig. 11: Printed circuit board DL 6 HA 006

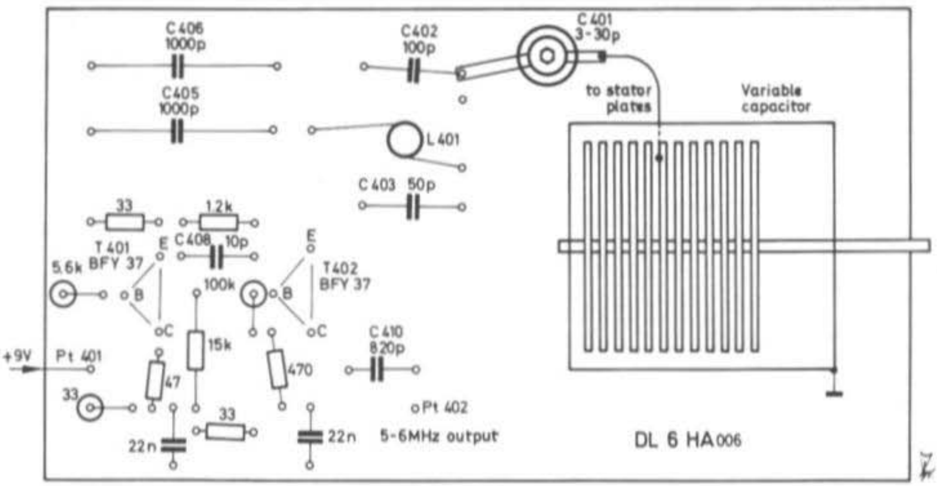


Fig. 12: Component location plan to DL 6 HA 006

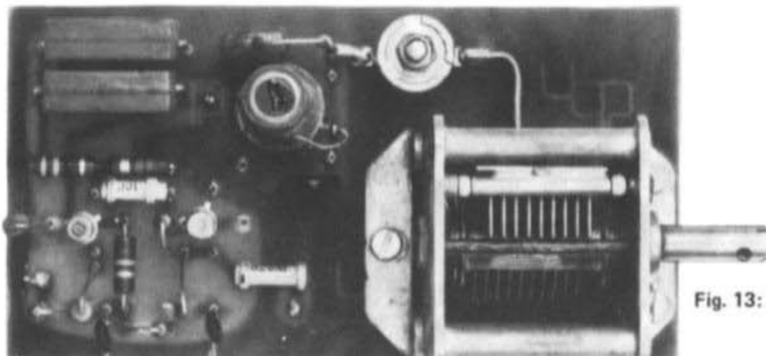


Fig. 13: Photograph of the 5 - 6 MHz VFO

4.3. ALIGNMENT

The alignment is very simple if, as was the case of the author, no temperature compensation is made.

A stabilized voltage of 9 V is connected to point Pt 401; the RF probe of the VTVM is connected to the output connection Pt 402. The oscillator should immediately commence oscillation if it has been correctly assembled and no components are defective. It is only necessary for the tuning range of the oscillator to be adjusted with the aid of the coil core and trimmer capacitor C 401. A shortwave receiver can be used during the frequency measurement. As has been previously mentioned, the author did not carry out any temperature compensation. Capacitors C 402 and C 403 are mica types. The VFO module, however, was contained in Styrene foam which ensured that no rapid temperature variations could occur.

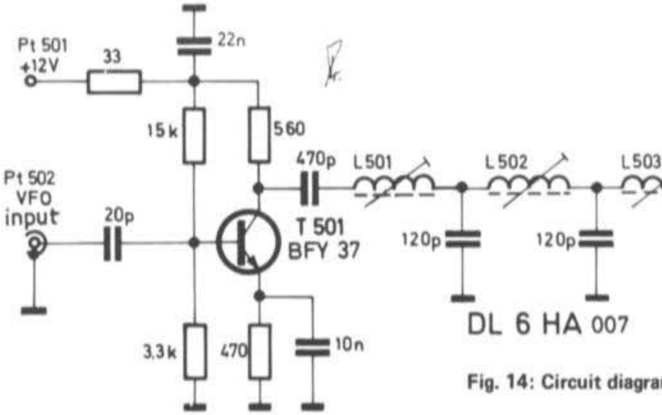


Fig. 14: Circuit diagram of the low-pass filter

5. THE LOW-PASS FILTER

The low-pass filter circuit shown in Figure 14 is provided to suppress the harmonic spectrum of the variable frequency oscillator. This is necessary in order to obtain a high spurious rejection of the transmitter and receiver.

The described low-pass filter represents a double T-link following a transistor buffer. No further amplifier stage should be provided between the output of the low-pass filter and the mixer stages, since otherwise it could be possible for new harmonics to be formed by the non-linear transconductance of the amplifier.

5.1. MECHANICAL ASSEMBLY

The low-pass filter is accommodated on a printed circuit board having the dimensions 70 mm by 40 mm. The printed circuit board, which has been designated DL 6 HA 007, is shown Figure 15; the corresponding component location plan is given in Fig. 16. The coupling capacitor between transistor T 501 and inductance L 501, as well as the terminating resistor of 560 Ω should not be mounted during assembly but after alignment.

A U-shaped screening plate (35 mm high) screens the three inductances from another. The photograph of the completed low-pass filter, given in Figure 17, shows further constructional details.

The output connection Pt 503 is connected after the alignment process by a short low-capacitive (high Z) cable to point Pt 206 of the DL 6 HA 004 module. A large capacitive or resistive load at the output of the filter will detune it to such a degree that it is no longer effective.

During the construction of the whole transceiver, attention must be paid to obtain a good screening between the variable frequency oscillator and the low-pass filter so that no unwanted coupling can occur.

5.1.1. SPECIAL COMPONENTS FOR THE DL 6 HA 007 MODULE

T 501: BFY 37 (ITT-Intermetall), 2 N 708, BF 115 (Philips, Siemens, AEG-Telefunken), BF 224 (Texas Instruments) or similar types

0.1 mm dia (38 AWG) silk-covered enamelled copper wire is used for all inductances.

L 501, L 503: 60 turns wound on to a 5 mm diameter coil former. Coil length 12 mm. SW core.

L 502: 90 turns wound onto a 5 mm diameter coil former. Coil length 22 mm. SW core.

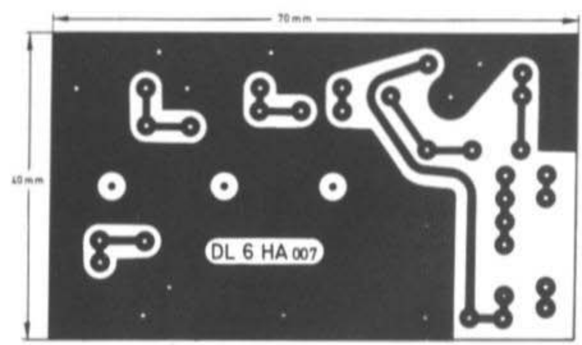


Fig. 15: Printed circuit board DL 6 HA 007

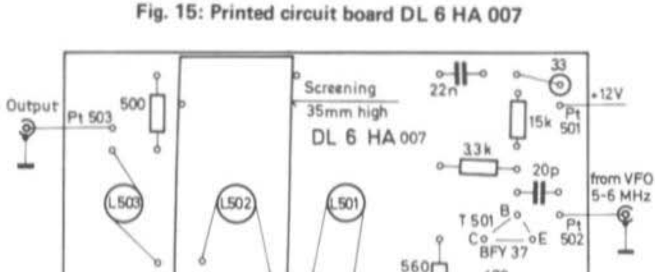
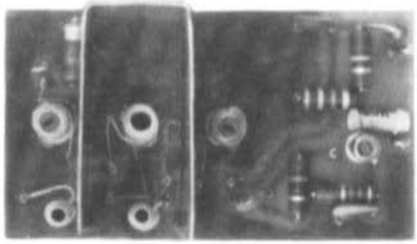


Fig. 16: Component location plan to DL 6 HA 007

Fig. 17: Photograph of the low-pass filter



5.2. ALIGNMENT

The free connection on inductance L 501, which is to be used later for the 470 pF coupling capacitor, is connected to ground and inductance L 501 is aligned to 6 MHz with the aid of a dip meter. After this, the free end of inductance L 503 (output connection Pt 503) is connected to ground and also aligned to 6 MHz The two wire bridges are now removed and the centre inductance L 502 is aligned to approximately 5 MHz.

The circuit is then completed with the coupling capacitor and the terminating resistor, and a signal is fed from the dip meter to the input connection Pt 502. This is done by winding a few turns of wire around the coil of the dip meter and soldering the twisted ends to the low-pass filter input and ground. If the operating voltage of +12 V is now connected to point Pt 501, the low-pass filter module will be ready for operation. A VTVM connected to the output will indicate the relative attenuation curve if the dip meter is tuned in the range of 5 MHz to approximately 18 MHz. Since a large number of dip meters vary their output voltage with frequency, this could cause the measurement to be falsified. However, the most important point is that a great increase in attenuation appears at approximately 6 MHz and that a high attenuation remains up to at least 18 MHz.

5.3. MEASURED VALUES

The following values were measured by the author on the described low-pass filter:
Voltage amplification in the pass band (f < 6 MHz) : approx. 3 dB
Pass band ripple : approx. 1 dB
Attenuation at 10 MHz : approx 43 dB ref. to the output voltage at 5 MHz
Attenuation at 15 MHz : approx. 57 dB ref. to the output voltage at 5 MHz.

6. AVAILABLE COMPONENTS

The printed circuit boards, special components, as well as kits of parts are available from the publishers or their national representatives (see advertising page).

7. REFERENCES

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

(2) G. Laufs: A SSB Transceiver with Silicon Transistor Complement
Part 2: The 9 MHz Transceiver
VHF COMMUNICATIONS 2 (1970), Edition 2, Pages 65-75

EXPERIMENTS WITH A CRYSTAL DISCRIMINATOR

by D. E. Schmitzer, DJ 4 BG

INTRODUCTION

Crystal discriminators, which are extensively used in commercial communications equipment, represent an alternative to the digital discriminator described in (1). These types can be used successfully for the demodulation of frequency modulated transmissions; they have the advantage of requiring no alignment which means that they are expecially useful for amateur radio applications. At present, crystal discriminators are available for low intermediate frequencies (around 500 kHz), for 10.7 MHz and 9 MHz. The latter type is especially suitable for amateur use because it allows modern SSB receivers to be modified simply and elegantly for FM reception. This article describes a number of experiments made with a 10.7 MHz crystal discriminator to determine the (small) effects of mis-match conditions.

1. ASSEMBLY AND USE OF CRYSTAL DISCRIMINATORS

Simply speaking, one could say that crystal discriminators operate in a similar manner to ratio discriminators; where the resonant circuits are replaced by the crystals. The critical alignment of the hump frequencies of the discriminator curve is no longer required because this is made by the manufacturer by the combination of accurately measured crystals. Since the crystal discriminator does not possess any limiting characteristics, it is necessary for it to be driven from a limited preamplifier. Values given in the following diagrams indicate the input and output levels.

An FM accessory equipped with a crystal discriminator will be built up as follows: The signal is tapped off previous to the SSB crystal filter, fed through a crystal filter with a bandwidth of approx. 12 kHz and a wideband amplifier with limiter characteristics (integrated circuit) after which it is passed to the crystal discriminator. It is favourable to connect an AF impedance converter, and possibly a 3 kHz low-pass filter (active audio filter) to the output of the crystal discriminator. This should provide optimum FM reception. It is true that this concept is somewhat more expensive, but also more elegant than a conversion of the higher intermediate frequency to approx. 500 kHz, passing it through a subsequent amplifier with band pass characteristic to a digital discriminator. Of course, the latter arrangement is suitable for receivers that already possess a low intermediate frequency.

Finally, it should be mentioned that a modulation index of 1 was agreed for frequency modulated amateur transmissions (F 3) at the IARU Region I Conference 1969. Corresponding to this, the bandwidth of suitable FM filters will be 12 kHz (max. frequency deviation: 3 kHz, highest modulation frequency: 3 kHz). The relationships between modulation index, frequency deviation, highest modulation frequency and bandwidth were described in (2).

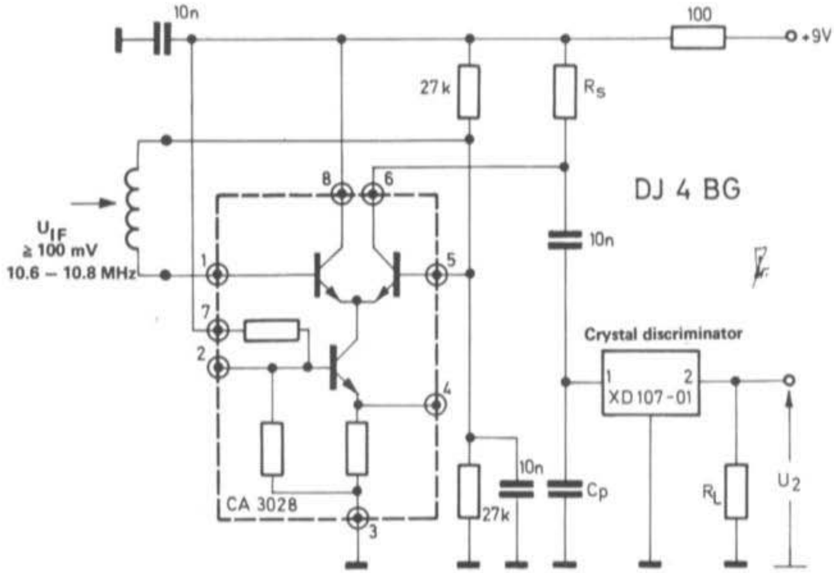


Fig. 1: Experimental circuit for measurements on a crystal discriminator

2. MEASUREMENTS ON A CRYSTAL DISCRIMINATOR

For his experiments, the author used a crystal discriminator type XD 107-01 manufactured by KVG. This type is dimensioned for a centre frequency of 10.7 MHz and possesses a hump spacing of approx. 40 kHz. This means that it is especially suitable for narrow band frequency modulation (NBFM) with a modulation index of 1 to 2. As can be seen in the associated specification sheet, a source impedance of 500 Ω parallel to 30 pF and a load impedance of at least 100 kΩ are prescribed for this crystal discriminator. The experimental circuit given in Fig. 1 was built up to determine how the frequency response alters when deviating from these values. The main characteristics of this circuit are the wideband preamplifier with limiting characteristics using the integrated circuit CA 3028, the collector resistor R_S, which represents the source impedance for the crystal discriminator, as well as capacitor C_p, which represents the parallel capacitance of the source resistor R_S. The circuit capacitances, of course, add themselves to C_p.

A voltage U_2 appears at the output of the crystal discriminator which is a DC voltage if a constant frequency is present at the input;with frequency modulation, an alternating voltage corresponding to the modulation frequency is available at the output. The resistor R_L represents the load impedance for the discriminator.

2.1. EFFECTS OF THE PARALLEL CAPACITANCE

The frequency response with 3 different values of the parallel capacitance C_p is given in Fig. 2. The other values were kept constant: Source resistor R_S = 510 Ω, the load resistor R_L = 270 kΩ.

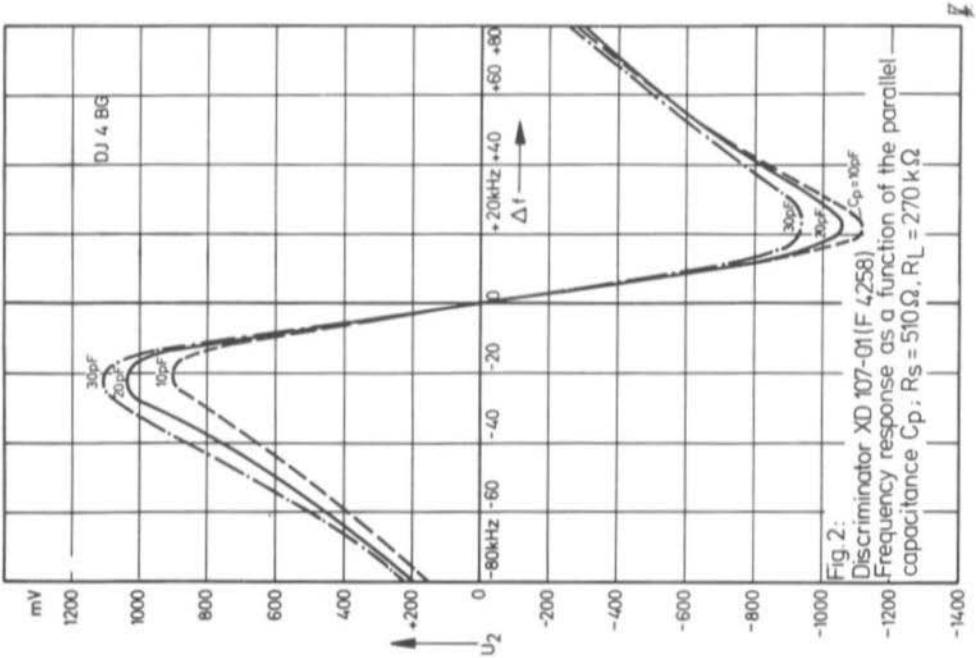


Fig. 2:
Discriminator XD 107-01 (F 42258)
Frequency response as a function of the parallel capacitance C_p: $R_S = 510\,\Omega$, $R_L = 270\,k\Omega$

As one can see, the straight centre portion of the curve is practically non-affected. The only effect is that the humps have different amplitudes and that the zero line is shifted slightly which, however, is completely unimportant for the planned application as a FM demodulator. Correct alignment is achieved when the voltages corresponding to the humps are equally great. This was the case at C_p = 20 pF which leads to the assumption that the circuit capacitance was approx. 10 pF (output capacitance of the CA 3028 : 3.5 pF). This means that an alignment is practically superfluous if a capacitor with approximately 20 - 25 pF is used to compensate for the circuit capacitance instead of the prescribed 30 pF.

2.2. THE EFFECTS OF THE SOURCE RESISTOR

The frequency response of the discriminator characteristic with three different values of the source resistor R_s is given in Fig. 3. The two other parameters are once again kept constant: C_p = 20 pF, R_L = 270 kΩ.

When altering the source impedance between 360 Ω and 680 Ω, no alternations of the curve shape will result. If the source impedance is increased, the only effect will be that the output voltage also increases since the source impedance of the crystal discriminator is simultaneously load resistor for the output transistor of the integrated circuit. This means, that by increasing the gain, (gain = slope x Z_{out}) a higher IF voltage will be provided for the discriminator.

The source impedance value of 500 Ω prescribed in the data sheet is thus very uncritical and can be increased to the next standard resistance value of 560 Ω, or even 680 Ω, in order to obtain a higher output voltage.

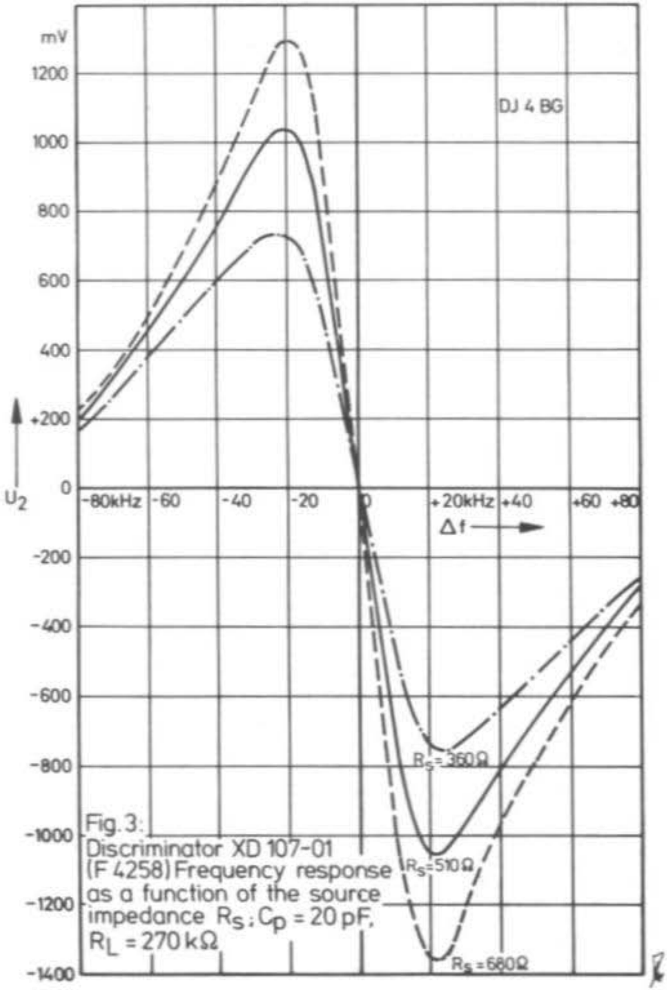


Fig. 3:
Discriminator XD 107-01
(F 4258) Frequency response as a function of the source impedance R_S; $C_P = 20\,pF$, $R_L = 270\,k\Omega$

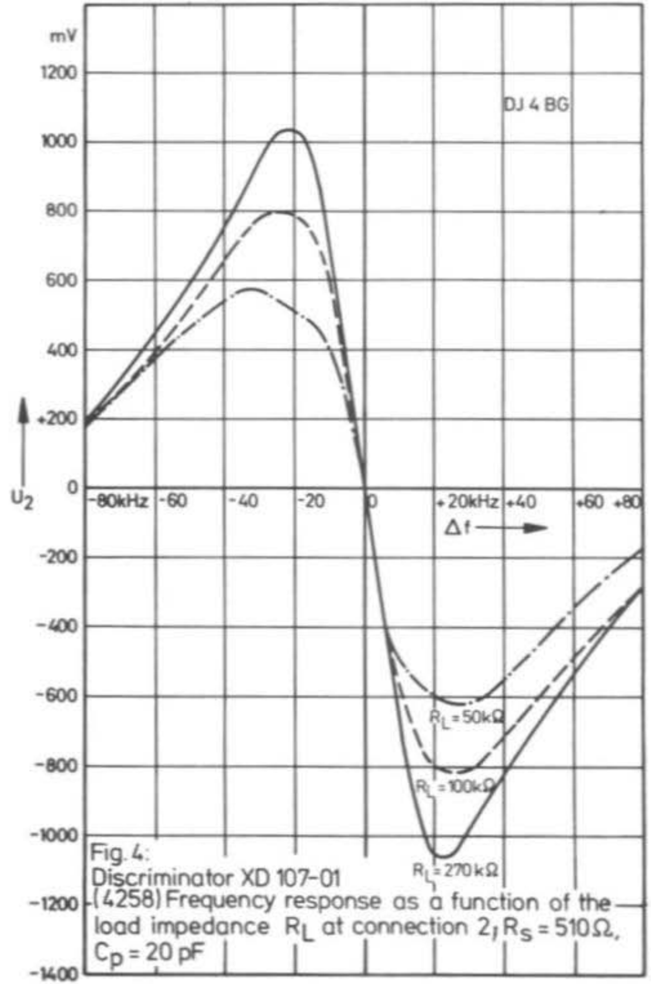


Fig. 4:
Discriminator XD 107-01
(4258) Frequency response as a function of the load impedance R_L at connection 2, $R_S = 510\,\Omega$, $C_P = 20\,pF$

2.3. EFFECTS OF THE LOAD RESISTOR

The prescribed load impedance for the audio frequency side of the crystal discriminator is at least 100 kΩ. The effects on the frequency responce curve for impedance values differing from this are given in Fig. 4: If the impedance value is too small, (R_L = 50 kΩ), a noticeable distortion of the characteristic (higher distortion factor) is caused and the available AF voltage will be less. A load impedance value of 100 kΩ should therefore always be provided which can be achieved elegantly using an impedance converter stage.

3. AUTOMATIC FREQUENCY CONTROL

It is, of course, possible to use the DC voltage component of the discriminator output voltage for automatic frequency control (AFC) of an oscillator. The output voltage of approx. ± 600 mV is possibly sufficient to directly control a varactor diode. In this case, the voltage is taken at high impedance from the discriminator and filtered. The time constants of the filtering links must be lower than the lowest required frequency so that it is not controlled.

A suitable circuit is given in Fig. 5. The previously mentioned impedance converter using a junction field effect transistor is included. The circuit does not require a great number of components since the DC path is already made within the crystal discriminator (approx. 60 kΩ).

If the centre frequency is to be indicated, this can be made using a centre zero meter having a full scale deflection of ± 5 μA or ± 10 μA, fed via a dropper resistor of approx. 270 kΩ or 120 kΩ respectively. For meters with higher current requirements, it will be necessary to provide a symmetrical DC voltage preamplifier.

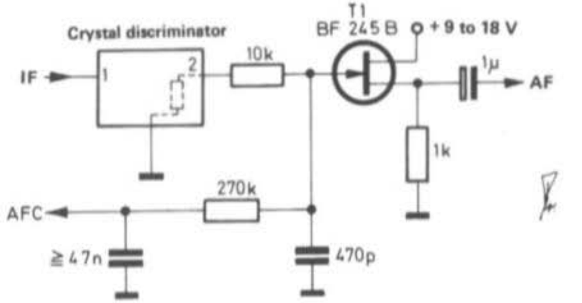


Fig. 5: Impedance converter and filtering of the AFC voltage at the output of the crystal discriminator

4. NOTES

The crystal discriminator type XD 107-01 (KVG) is very suitable for amateur applications because, if simple rules are followed, it will not require any alignment. A matching crystal filter with a bandwidth of 12 kHz (XF 107 A) is also available. These items are accommodated in the same hermetically sealed casings (27 mm x 36 mm x 18 mm) as the 9 MHz crystal filters. However, the connections of the crystal discriminator XD 107-01 are designed for wire connections; this means that extra care must be paid when mounting them on printed circuit boards.

4.1. CRYSTAL DISCRIMINATORS FOR 9 MHz

Two types of crystal discriminators are manufactured by KVG for the most common intermediate frequency for amateur applications of 9 MHz. (XD 09-01 and XD 09-02). These filters are somewhat too narrow for frequency demodulation, even at very low frequency deviation values, (the hump spacing is too small). However, these discriminators are very suitable for automatic frequency control. A 9 MHz crystal discriminator (XD 09-03) corresponding to the XD 107-01 type is now available as is the matching crystal filter having a bandwidth of 12 kHz (XF-9E).

5. AVAILABLE PARTS

The crystal discriminator XD 09-03 and the crystal filter XF-9E are available from the publishers or their national representatives. See advertising page.

6. REFERENCES

(1) D. E. Schmitzer: A Digital Discriminator Accessory for FM Demodulation
VHF COMMUNICATIONS 2 (1970), Edition 2, pages 105-110

(2) D. E. Schmitzer: Is FM Advantageous on the VHF-UHF Bands?
VHF COMMUNICATIONS 2 (1970), Edition 1, pages 21-24.

A UNIVERSAL VHF-UHF TRANSMITTER FOR AM AND FM

by R. Lentz, DL 3 WR

Continuation from VHF COMMUNICATIONS 2 (1970), Edition 2

3.3. THE VARIABLE FREQUENCY OSCILLATOR

A synthesis VFO is used, where a crystal-controlled frequency of 27.8 MHz is mixed with the frequency of the actual variable frequency oscillator in the range of 3.47 MHz to 3.8 MHz. These two signals are fed to a mixer where the difference frequency is selected. This results in an output frequency of between 24.0 and 24.33 MHz. A FM accessory is provided to allow frequency modulation.

3.3.1. CIRCUIT DETAILS

The circuit diagram of the VFO is given in Figure 14. The mixer stage (Transistor T 2) is equipped with a junction field effect transistor so that unwanted conversion products are low. It is favourable for the 24.0 to 24.33 MHz bandpass filter (Inductances L 2 and L 3) to be inductively or capacitively base-coupled, however, this filter is very difficult to align without a sweep measuring set.

Transistor T 3 should be a low-reactive type since it operates as a 24 MHz buffer. If difficulties are encountered in this respect, the collector circuit can be damped by connecting a low-resistive resistor to the output or to connect the base to a low impedance tap (increase the value of capacitor C 16).

The frequency modulation of the variable frequency oscillator is made with the varactor diode D 1. This diode is loosely coupled to the rest of the VFO using only 4.7 pF; this loose coupling ensures that the very great temperature coefficient of the varactor diode does not affect the resonant circuit of the VFO. On the other hand, a greater capacitance variation must be used in order to obtain the required frequency deviation. The circuit is dimensioned so that an AF voltage of 6.5 V (peak-to-peak) will result in a frequency deviation of approximately 3 kHz (on 144 MHz). To achieve this, the varactor diode is provided with a bias voltage of 4.15 V.

As has been already mentioned, the VFO is provided with a stabilized operating voltage of only 8.6 V. A built-in second stabilization circuit with transistor T 5 has an output of 7.5 V from which the actual variable frequency oscillator and the varactor diode are fed. This double voltage stabilization is very effective: A frequency shift can hardly be heard on a receiver tuned to 1296 MHz when varying the operating voltage in the range of 10 V to 18 V.

In contrast to the crystal-controlled oscillator of the synthesis VFO with its considerably stronger oscillation and thus thermal loading, the variable frequency oscillator can be keyed without chirp. This is the reason why the operating voltage of the actual VFO is fed via its own feedthrough capacitor.

The suppression of spurious signals was measured in an industrial laboratory: The crystal-controlled frequency of 27.8 MHz was found to be suppressed by 43 dB, the image frequency of 31.27 to 31.6 MHz by 64 dB.

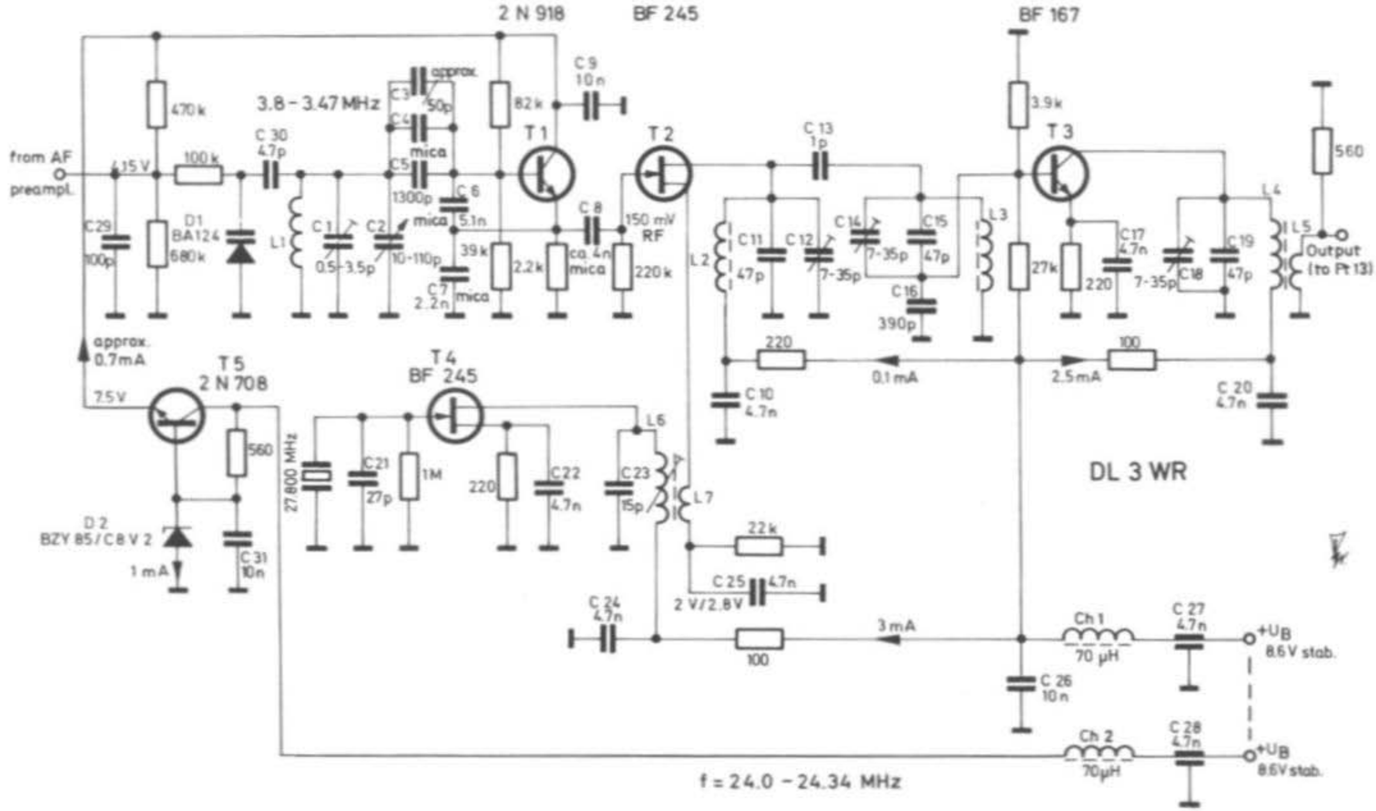


Fig. 14: Circuit diagram of the VFO, including FM accessory

The output voltage into an impedance of 560 Ω amounts to approximately 1 V (RMS). The total current drain is approx. 7 mA at an operating voltage of 8.6 V; an AF voltage of 2.3 V (RMS) is required for maximum frequency deviation. Some further measured values are given in the circuit diagram.

3.3.2. SPECIAL COMPONENTS FOR THE VFO

T 1 : 2 N 918, BFX 73
T 2, T 4 : BF 245 C, TIS 88, 2 N 5245
T 3 : BF 167, BF 173
T 5 : 2 N 708 or similar silicon NPN transistor

D 1 : BA 124, BA 150, BA 119, BA 111, MA-45043 (55 pF at 2 V)
D 2 : BZY 85/C8V2, ZF 8.2, 1 N 1520 A (8.2 V zener diode)

L 1 : 2.45 μH; 23.2 turns of 0.5 mm dia. (24 AWG) silk-covered, enamelled copper wire wound on a 9 mm diameter ceramic coil former. Coil length 15.5 mm. Glue with dual-component adhesive which should be hardened and artifically aged by a number of temperature variation cycles between -10 °C and +80 °C. Due to the temperature coefficient, no core is used. This means that the correct inductivity for the bandspreading must be found by experiment according to the variable capacitor and circuit capacitance.

All other coils are enclosed in pottet cores 11 by 7 mm; material:

K 12 (AL = 16) manufactured by Siemens. Wire as for L 1.

L 2, L 3, L 4: 680 μH; 6.5 turns
L 6 : 1.6 μH; 9.5 turns
L 5 : 3 turns on L 4
L 7 : 1 turn on L 6

C 1 : 0.5 to 3.5 pF glass tubular trimmer
C 2 : 10 to 110 pF air-spaced variable capacitor
C 3, C 4 : Together about 90 pF; ceramic tubular capacitors for alignment of the band limits and for TC compensation
C 5 : 1300 pF mica capacitor
C 6 : 5100 pF mica.capacitor
C 7 : 2200 pF mica capacitor
C 8 : Any value between 1 nF and 10 nF mica capacitor
C 12, C 14, C 18 and may be also C 23: 10-40 pF ceramic disc trimmer cap.

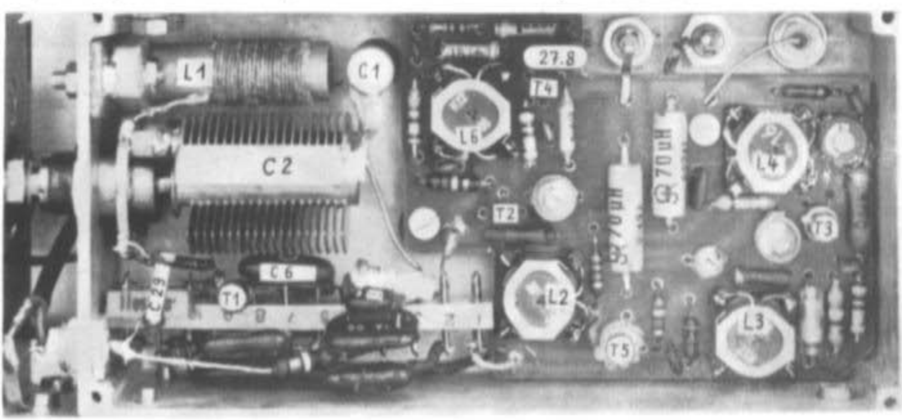


Fig. 15: Variable frequency oscillator from above

3.3.3. MECHANICAL ASSEMBLY OF THE VFO

The whole VFO is accommodated in an aluminium die-cast casing having the dimensions 125 mm by 65 mm by 35 mm; the casing is 2 mm thick. The actual variable frequency oscillator and the FM accessory are built up on a 12 pole ceramic tag board which is located beside the variable capacitor. The variable capacitor, the trimmer capacitor C 1 and the inductance L 1 are directly screwed to the casing. All other parts of the circuit are located on a small printed circuit board. Since the assembly is greatly dependent on the available casing and variable capacitor, no detailed description of the mechanical build-up is to be given. The photograph given in Figure 15 shows the authors prototype. The casing was provided with a plate metal cover and may also be thermally insulated using styrene foam.

3.4. AF PREAMPLIFIER

The AF preamplifier circuit given in Figure 16 consists of an active audio filter as was described in (4), an AM-FM change-over switch, a separate level potentiometer for each modulating mode and a further amplifier comprising transistor T 3 for generation of the voltage level required for the FM accessory. The buffer stage (T 1) previous to the active audio filter provides a certain amplification which means that the voltage from a dynamic microphone is sufficient to drive the modulator accommodated on printed circuit board DL 3 WR 003. The modulation mode switch shorts out the non-required output so that the non-used modulator can not be driven by any crosstalk or RF injection.

The amplifier stage equipped with transistor T 3 is dimensioned so that maximum frequency deviation is just reached.

All inputs and outputs of the AF preamplifier are bypassed with respect to RF voltage injection. In addition to this, the circuit, which is built up on a printed circuit board DJ 4 BG 001, is enclosed in a screened cabinet made from double-coated printed circuit board. The input socket, the potentiometers and the change-over switch are also accommodated in this cabinet. This ensures that the AF-preamplifier is RF-tight so that the VHF-UHF transmitter can be also used as an exciter for a high-power final amplifier stage.

3.4.1. VALUES MEASURED ON THE AUDIO PREAMPLIFIER

All measured values at U_B = 12 V and f = 2 kHz	Max. output voltage mV (r.m.s.)	Corresponding input voltage mV (r.m.s.)	Gain dB
AM-output	300	3	40
FM-output	1800 (sinusoidal)	0.9	66
	3200 (distorted)	1.6	66

Frequency (Hz)	20	100	320	400	1 k	2 k	3.4 k	4 k	7.6 k	20 k
Output level (dB)	-50	-20	-6	-3	-1	0	-3	-6	-20	-43

Table 3 : Output voltages, gain and frequency response of the audio preamplifier (see Fig. 16)

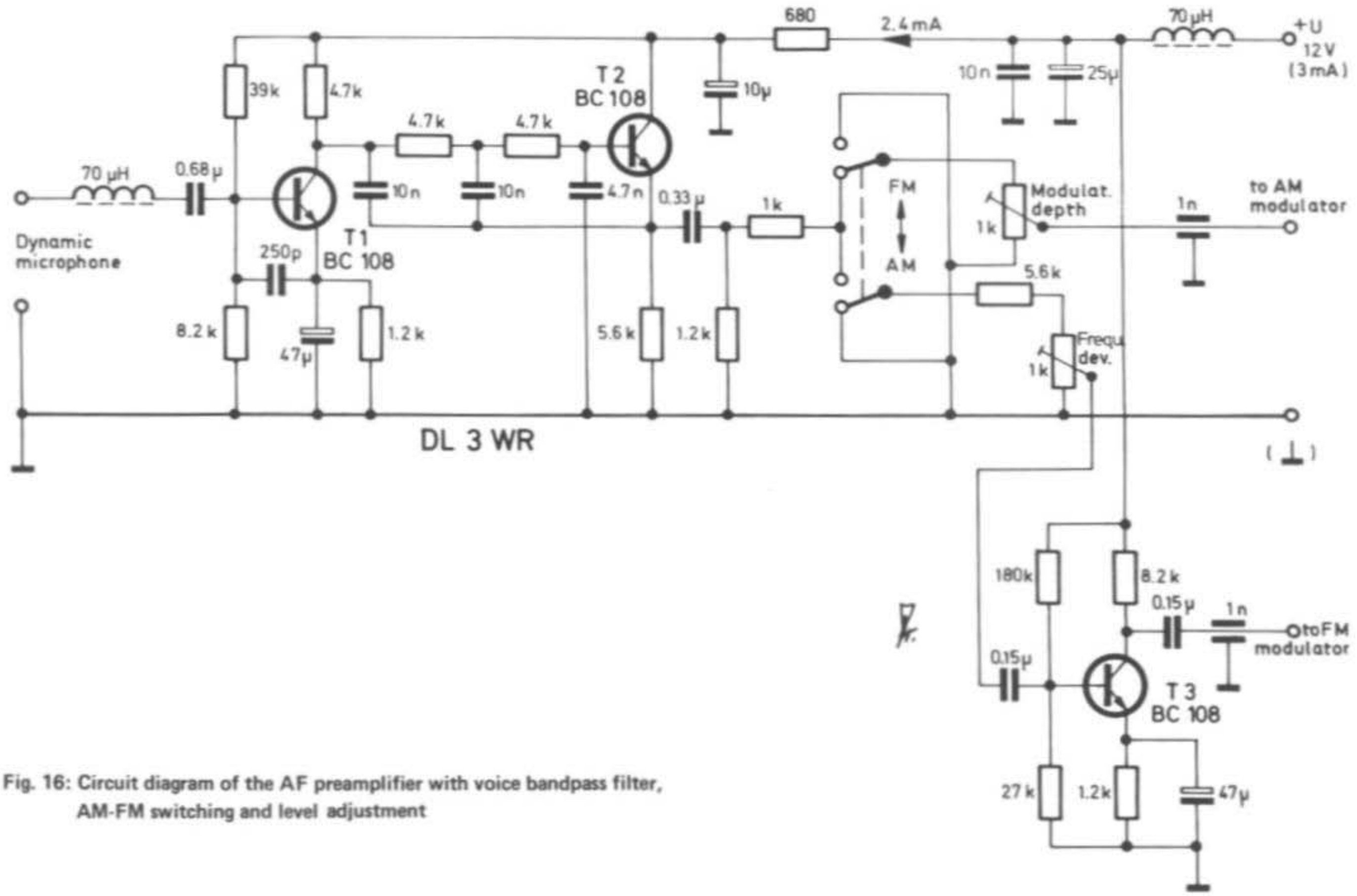


Fig. 16: Circuit diagram of the AF preamplifier with voice bandpass filter, AM-FM switching and level adjustment

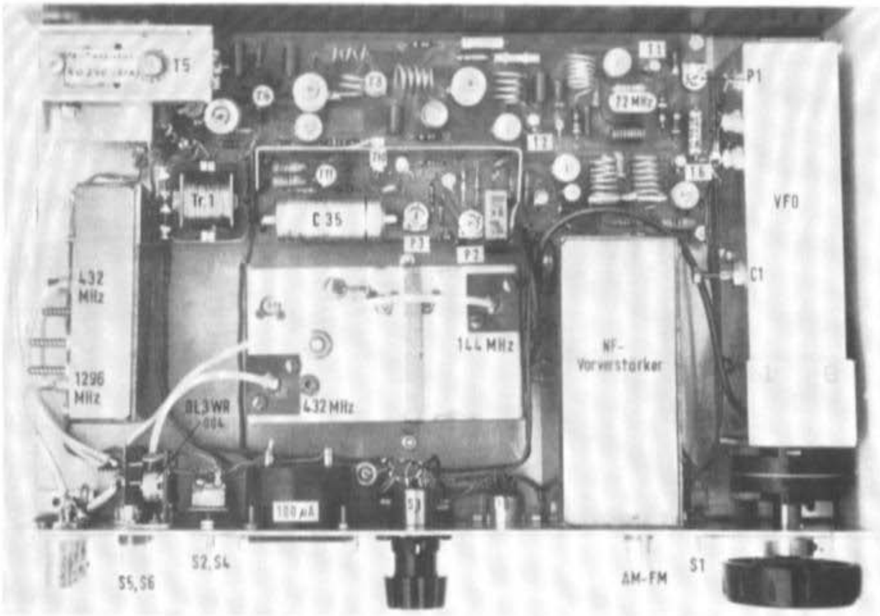


Fig. 17: View of the whole VHF-UHF transmitter from above

4. CONSTRUCTION OF THE COMPLETE VHF-UHF TRANSMITTER

The sub-assemblies described in the previous sections, as well as varactor triplers for 432 MHz and 1296 MHz are now to be combined as shown in Fig. 2. In the author's case, they were all enclosed in a U-shaped chassis made from aluminium plate having the dimensions 300 mm long, 180 mm deep and 75 mm high. A matching, also U-shaped cover is used to completely enclose the transmitter. Figure 17 shows a photograph of the completed transmitter. Since the VFO and its scale as well as the varactor triplers can differ considerably from those used by the author, a detailed construction description is not given.

5. FURTHER NOTES

The two modulator inputs (on the printed circuit board DL 3 WR 003 and on the VFO are connected using screened cable to the corresponding output of the AF preamplifier. The VFO output on the other hand, is connected in the shortest possible manner using normal wire to point Pt 13 of the transmitter. In addition to this, a low-impedance ground connection should be made between the cabinet of the VFO and the ground of printed circuit board DL 3 WR 003. This is made favourably with a short but wide piece of copper foil. In order to provide a certain amount of thermal insulation, the VFO is only connected to the chassis via two strong pertinax strips. It is then possible to also glue styrene foam to the lower portion of the VFO cabinet.

The actual transmitter board is mounted by the metal bracket holding the PA stage and - on the VFO side - by two long screws each having a metalic spacer bushing.

6. AVAILABLE PARTS

All printed circuit boards, the most important components, as well as a kit of parts are available from the publishers or their national representatives. Please see advertising page.

7. 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

(3) H. Dohlus: Einfache Topfkreis-Stufen für 435 MHz
UKW-BERICHTE, SONDERHEFT II (1969), pages 54-60

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

(5) E. Berberich: A Coaxial Relay with a High Coupling Attenuation and Good SWR
VHF COMMUNICATIONS 1 (1969), Edition 2, pages 124-125

(6) E. Fluegel: The 2 Metre Transmitter UTS 5 with 2 W Mean Output Power at an Operating Voltage of 12 V
VHF COMMUNICATIONS 1 (1969), Edition 3, pages 179-187

(7) R. Lentz: A Simple Electronic Fuse
VHF COMMUNICATIONS 1 (1969), Edition 3, pages 174-178

NOTES FOR OUR READERS

We would like to mention the following points:
We are receiving a large number of address alterations which should actually go to the national representative where the subscription is registered. This is because the address file is held by the representative and not by us.

Only those subscribers living countries where we do not have a representative should inform us directly.

We are receiving a number of money transfers without full address or mentioning the required items. This means that we are not able to contact the sender and we must wait until the customer informs us of his requirements or complains of non-delivery. Please ensure that your full address and list of requirements are included.

Now to the material position: We are pleased to say that we are now able to supply the Philips trimmers for the DJ 1 NB 004 and DL 3 WR 003 transmitters. These trimmers are available for DM 1.00 each, e.g. DM 6.00 for the set of six. The price of the complete trimmer set is now DM 10.00 for DL 3 WR 003 or DM 9.10 for DJ 1 NB 004.

Secondly, a printed circuit board has now been developed to accommodate the two crystal oscillators of DL 6 HA 001/14 MOSFET converter. This board is available under the designation DL 3 YK 001 at DM 4.00.

Verlag UKW-BERICHTE

70 cm - 23 cm STRIPLINE VARACTOR TRIPLER

by H.-J. Franke, DK 1 PN

After studying numerous application notes (1), (2), (3), (4) and after constructing several poor or non-operative prototypes, the author built up a small and simple tripler (Fig. 1) which is to be described. The diode used is a step-recovery diode type BAX 11/II (AEG-Telefunken). Since this designation is valid for a family of diodes with various storage and switching times, it is necessary for the required frequencies to be given on ordering.

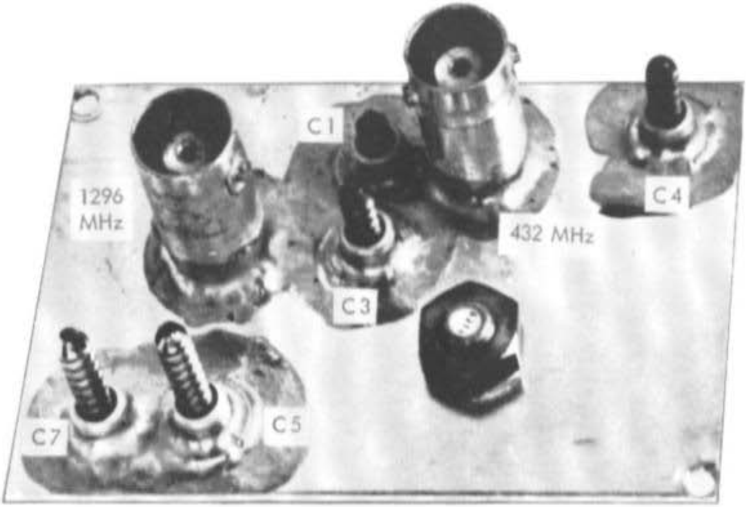


Fig. 1: Photograph of the varactor tripler from above

The maximum drive power for the diode used in this circuit is 10 W. At an efficiency of 50% - which the described tripler exhibited without difficulties - this results in a maximum output power of 5 W. This represents a considerable power for the 23 cm band. The transmitter described in (5) equipped with the EC 8020 tube in the PA stage is a very suitable exciter.

The given values are peak power values for CW and frequency modulated transmissions. For amplitude modulation, the drive power should be reduced to a maximum of 2.5 W. The peak envelope power (PEP) then obtains a value of 10 W at a modulation depth of 100%.

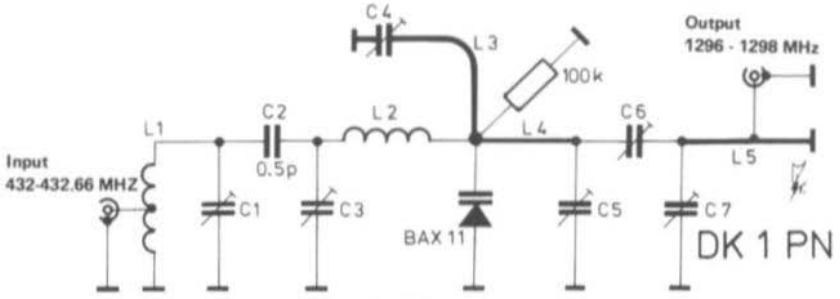


Fig. 2: Varactor tripler 432/1296 MHz

1. CIRCUIT DESCRIPTION

The circuit of this tripler (Fig. 2) does not differ greatly from those given in other publications. A parallel resonant circuit (L 1, C 1) is arranged at the 70 cm end, which is followed by a matching network consisting of the two capacitors C 2 and C 3 as well as inductance L 2. These elements on the 432 MHz side are built out of concentrated elements, e.g. inductances and capacitors.

The idler circuit L 3, C 4 which is tuned to the first harmonic of the input frequency and the output circuits (L 4, C 5; L 5, C 7) are capacitively shortened stripline circuits. The designations L 3, L 4 and L 5 are therefore not exactly correct; however, they are suitable for describing this circuit. In order to achieve rational line length and optimum Q values in spite of the inavoidable large capacitive loading of the circuits, the stripline circuits were designed for an impedance of approx. 60 Ω. This low value requires a very low spacing of the lines from the ground area, namely 2 mm with the dimensions chosen here. The advantage of this is that the field of the striplines is concentrated to a very small area which means that the vicinity of the circuits is hardly affected. It is therefore possible for the varactor tripler to be used without casing (Fig. 3).

2. SPECIAL COMPONENTS

Step-Recovery diode BAX 11/II (AEG-Telefunken)
Storage time: approx. 46 ns, switching time: approx. 0.6 ns;
Capacitance at 6 V: approx. 3 pF
Probably usable but not tested by the author: MV 1808 J (MOTOROLA)
or MA-4963 B (Microwave Ass.)

L 1: 3 turns of 1.2 mm dia. (17 AWG) silver-plated copper wire wound onto a 5.5 mm former (self-supporting). Coil tap: 1 turn from the cold end.
L 2: as L 1, but without coil tap
L 3, L 4, L 5: see Figure 4

C 1, C 3, C 4, C 5, C 7 : 0.8 - 6.8 pF ceramic tubular trimmers type 2222 801 20006 or 2222 801 96004 (Philips). Please note details given in Section 3. for replacement types.
C 2 : 0.5 pF ceramic disc capacitor
C 6 : See Figure 4.

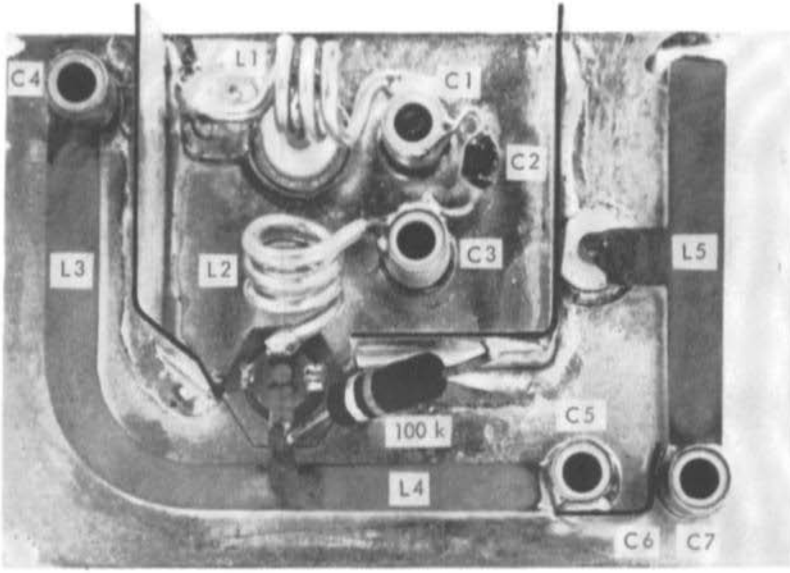


Fig. 3: Photograph of the varactor tripler from below

3. MECHANICAL ASSEMBLY

The tripler is built up on a metal plate with the dimensions 55 mm x 75 mm which serves as ground area for the stripline circuits and as cooling plate for the diode. Two vertical angle pieces are soldered onto the plate and screen the 70 cm end of the striplines. Figure 4 gives all dimensions necessary for assembly. They differ slightly from those of the prototype shown in the photograph. BNC connectors that can be directly soldered to the base plate are suitable. However, it is possible to directly connect the cable in order to avoid plugged connections. This mode is especially advantageous when installing the tripler within other equipment (6).

Now a few remarks regarding the tubular trimmers: Since the striplines are directly soldered onto the trimmers without being bent and are only spaced 2 mm from the base plate, trimmers should be used whose stators begin a maximum of 2 mm above the soldered ground connection. This is the case with the type given in Section 2. In order to have this spacing after having soldered the trimmers into place, the trimmers are, in contrast to the normal manner, passed through the base plate from the cold side and the stators soldered to the striplines. Of course, for trimmers C 1 and C 3 this is not necessary and other types could be used.

The striplines L 3 and L 4 are made out of one piece, which is electrically divided by the diode connection. The shape has been selected in order to obtain the smallest possible dimensions. It can also be shaped in a different manner; the only important facts are the length and the position of the diode connection.

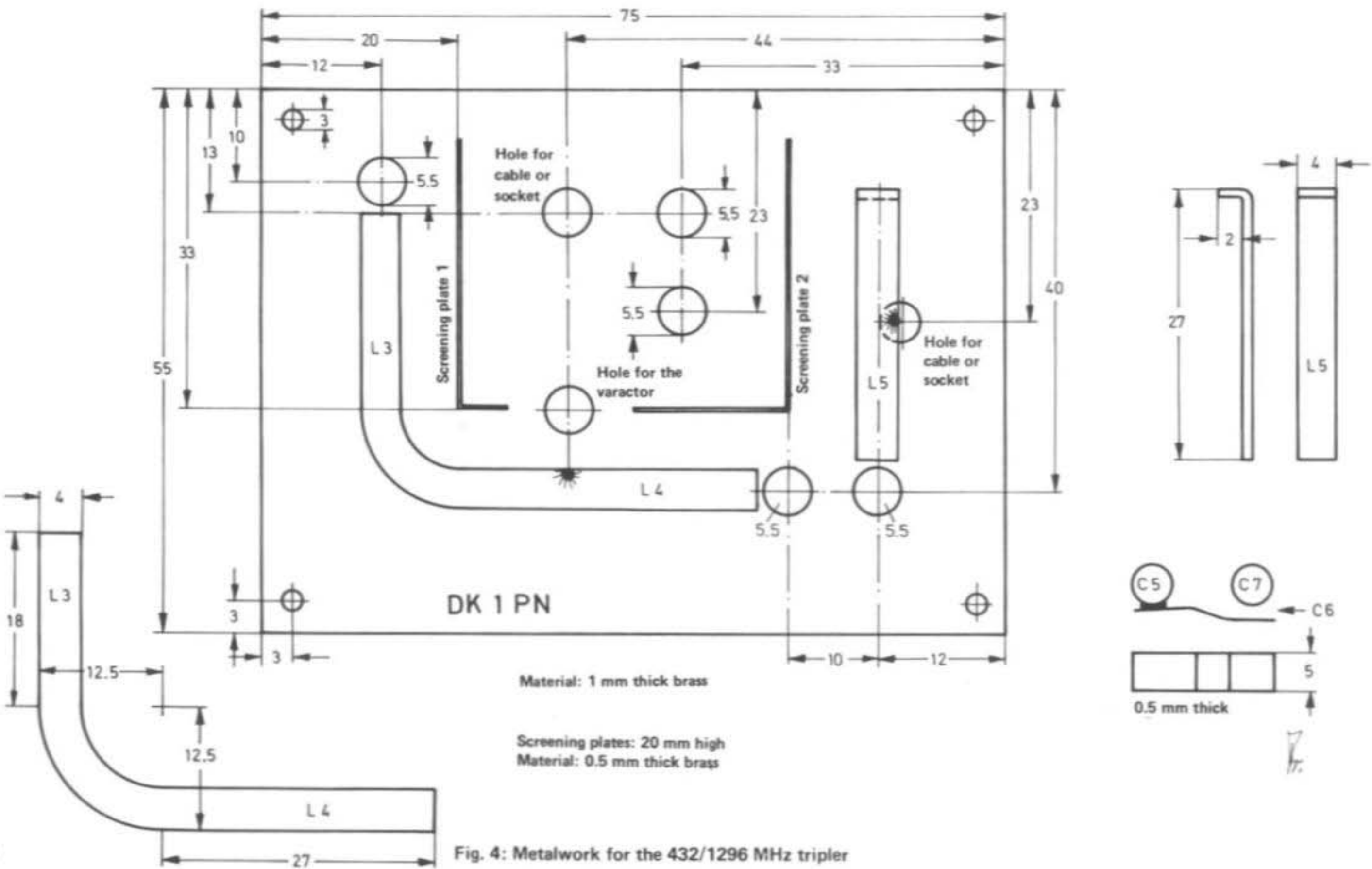


Fig. 4: Metalwork for the 432/1296 MHz tripler

After cutting, drilling and bending the metal parts according to Figure 4, the sockets, or cables and the two screening panels are soldered into place. The next components to be mounted are the trimmers which should be soldered into place as already described. The striplines are filed where they meet the trimmers so that the ends fit on the stator surface of the trimmers. Place a 2 mm thick piece of wood or pertinax underneath the striplines and, fixing them in the correct position, solder them into place using a small amount of solder. After inductance L 1 and the metal piece representing C 6 have been mounted, it will be possible for the diode to be installed. Inductance L 2 and the 100 kΩ resistor meet each other at the diode. The connection to the designated connection point on the stripline is made with the aid of a thin, 1 - 2 mm wide copper foil strip. Finally, the fixed capacitor C 2 and the input and output connections are soldered into place.

The tripler can be silver-plated before mounting the diode if teflon-insulated BNC sockets are used. If this is not the case, the metal pieces can be silver-plated before assembly. The effect of not silver-plating the tripler was not examined. Due to the large current areas and the low impedance, it is possible that the efficiency reduction would not be too great.

4. ALIGNMENT

Besides an exciter providing between 1 and 10 W in the 70 cm band, a bandpass filter for the output frequency, a terminating resistor and an output power meter (relative) are required for the alignment of the tripler.

A suitable exciter is described in (5). A coaxial filter or an interdigital filter (7) can be used. Such a filter is also suitable for use when operating the tripler. A well-matched antenna can be used as the terminating resistor. If an efficient reflectometer is not available for examining the matching, it is adviseable to assemble a folded dipole with balun. The dipole is not critical as long as the wire is not too thin. A VHF reflectometer can be used as power meter even when it does not work as a reflectometer at 1296 MHz. It is only a stable relative indication that is required.

The coupling capacitor C 6 should firstly be adjusted to its lowest value, e.g. the metal strip should be bent as far as possible from trimmer C 7. The exciter is now switched on and trimmers C 5 and C 7 aligned so that an indication is visible. These two trimmers will be virtually at minimum capacitance. After this, the idler-circuit (C 4) should be aligned for maximum output power. The resonance of this circuit is relatively wide which means that it is only necessary to correct the adjustment after completing the alignment process.

After also aligning trimmers C 1 and C 3 of the 432 MHz circuits, the most favourable position of capacitor C 6 is determined. In order to do this, the metal strip is bent slowly, millimetre for millimetre towards trimmer capacitor C 7 and all circuits with the exception of the idler-circuit tuned for maximum output power at each step. Finally, the output power will no longer increase, but will return to the old value on increasing the capacitance of C 6 and aligning for maximum. Since the increasing bandwidth of the tripler achieved in this manner far exceeds the required 2 MHz (1296 - 1298 MHz) and since the suppression of spurious signals will deteriorate, the most favourable adjustment is where the maximum power is achieved at the lowest value of C 6. In this position, the interaction between the circuits is at a minimum.

A certain hysteresis will be noticed when aligning trimmer capacitor C 5 and C 7. This means that the trimmer capacitance will not be increased or reduced constantly on rotating the spindle but will sometimes be reduced when it should increase and vice versa. The expensive glass tubular trimmers do not exhibit this effect. With wider resonant curves, this effect is not so noticeable. It should be mentioned that trimmer types equipped with invar-spindles were more reliable during a large number of alignment attempts (due to unsuitable diodes) than those types equipped with brass spindles. The type given in this description is equipped with an invar-spindle.

If an efficient reflectometer is available for 432 MHz, trimmers C 1 and C 3 should finally be aligned for the best compromise between output power and the input standing wave ratio. This is favourable for the exciter and reduces any unwanted radiation of 70 cm power by the connection cable.

5. AVAILABLE PARTS

The trimmers and a ready to operate interdigital bandpass-filter are available from the publishers. See advertising page.

6. REFERENCES

(1) Hewlett Packard: Application Note 920

(2) Motorola: Application Notes AN 147, AN 151, AN 159, AN 176, AN 177, AN 213, AN 228, AN 232, AN 416

(3) AEG-Telefunken (H. Nickel): Telefunken-Reaktanzdiode BAY 79 und ihre Anwendung in Frequenz-Vervielfachern (No. 67 A 011)

(4) S. Hamilton, R. Hall: Shunt-mode Harmonic Generation using Step-Recovery Diodes
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(5) H. J. Franke: A 10-Watt Transmitter for 70 cm
VHF COMMUNICATIONS 1 (1969), Edition 4, Pages 243-248

(6) R. Lentz: An Universal VHF-UHF Transmitter for AM and FM
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(7) R. E. Fisher: Interdigital Bandpass Filter for Amateur VHF-UHF Applications
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COAXIAL LOW-PASS FILTERS FOR VHF AND UHF

by Hans J. Dohlus, DJ 3 QC

1. INTRODUCTION

A state-of-the-art radio station will not transmit any spurious signals or harmonics. This is achieved by an effective, sometimes multiple, screening of the whole transmitter, as well as introduction of suitable filter elements between the transmitter and the antenna feeder.

Both measures must be carried out with the same care. Furthermore, attention must be paid to ensure that RF currents are not induced into the power line. In order to avoid this, the power line voltage is fed into the screened transmitter via suitable feed-through filters.

The extent of screening and filtering that is necessary depends on the circuitry and construction. This is expecially the case for VHF and UHF transmitters.

2. GENERAL MEASURES TO AVOID INTERFERENCE

In commercial stations, the previously mentioned measures are carried out with the utmost care. If this was not the case, such stations would not be allowed to operate by the postal authorities. Amateur transmitters, on the other hand, and VHF and UHF transmitters in particular do not usually conform to these requirements.

Previous to the introduction of UHF television, radio interference caused by the harmonics of VHF amateur transmitters was not to be expected. With the exception of the unavoidable direct injection into the audio stages of neighbouring broadcast and television receivers, only interference caused by the fundamental and harmonic waves of the frequency multiplier stages of VHF amateur transmitters were prevalent. Due to the avoidance of certain frequency multiplying processes and the use of bandpass filter stages, most of these interference sources could be eliminated. Since the introduction of UHF television, however, the number of cases where certain video and sound channels are interfered with by the harmonics from 2 metre transmitters has increased considerably. Since the channel usage varies according to regional considerations, the interference also differs from area to area. In some cases, interference can only be avoided, or reduced, by avoiding certain frequencies in the range of 144 to 146 MHz. Generally speaking, only suitable low-pass and bandpass filters are able to cure this problem.

The main measures against avoidable VHF amateur radio interference - powerline filters, screening, bandpass filter stages and output filters - are already necessary in order to reproduce a good, clean signal. The matching of the antenna array to the transmitter with the aid of reflectometers is also easily falsified, which is also valid for power measurements on VHF and UHF power amplifiers. This is another reason why harmonic and spurious radiation must be avoided.

As long as the whole power supply is accommodated within the screened transmitter casing, each individual pole of the power line should be fed into the transmitter via wideband feed-through filters. These filters are built up in a π-circuit comprising feed-through capacitors as capacitive components, as well as a ferrite choke as inductive link. The feed-through capacitors should be conventional suppression capacitors according to the electrical regulations. Such filters are available from a number of manufacturers. The transmitter cabinet should be grounded to the protective ground connection.

The power supply voltages of receiver and transmitter modules are usually fed via ceramic feed-through capacitors. However, nowadays feed-through filters are gaining popularity in their place. Such elements possess roughly the same dimensions. They are built up in a similar manner to the previously mentioned wideband feed-through filters, in a π-circuit. These components, which can be soldered or screwed into place, possess two ceramic feed-through capacitors and a ferrite tubular core. Such filters are dimensioned for a voltage of approx. 350 V, where a total current of 6 A is possible. For higher voltages, mica disc feed-through capacitors should be used.

It should be noted that only suitable suppression capacitors should be used when filtering the power line voltage. The use of ceramic feed-through capacitors is usually forbidden.

With resonant line PA stages (1), the most important part of the transmitter is very well screened. This means that only odd harmonics can be generated. With capacitively loaded shorted resonant lines, the resonance wave lengths of possible harmonics are not in an even relationship to the fundamental wave. This means that harmonics will not be provided with a suitable resonance condition in a resonant cavity tuned to a certain fundamental frequency; the resonant points are lower. This means that a certain degree of harmonic suppression is achieved.

The amplitude of the harmonic voltage increases with the degree of output coupling. Normally, amateur radio transmitters are very tightly coupled. When using resonant cavities and resonant line circuits, the output coupling can easily be combined with a filter if the output coupling link feeds into the inner conductor of a variable coaxial absorption circuit (bandstop filter). Such an arrangement can be extended to form a whole cascade of individual tuned coaxial filters.

Harmonic filters are usually built up as separate components having two connection sockets. After alignment, they are usually directly connected to the transmitter output. Mostly low-pass filters or bandstop filters are used, the most favourable type of filter is, however, a bandpass filter. Not very much has been written in amateur literature regarding effective harmonic filters for VHF transmitters. Of those types that have been described, most of them employed concentrated components (inductances, capacitors, etc.). Such filters are extremely easy to assemble, but are virtually impossible to align without the aid of a high quality VHF and UHF reflection or filter measuring system. It is especially in the VHF and UHF range where coaxial line components (radio frequency lines) are suitable for resonant circuits and filters. Bandpass, bandstop, high-pass and low-pass filters can be manufactured simply and effectively in the coaxial technology. Such elements can be easily calculated according to theoretical considerations. With concentrated components in the UHF range this is hardly possible since they cannot usually be assumed to possess their original qualities at these frequencies. The manufacture of fixed coaxial filters requires a certain amount of mechanical work, however, if the proved dimensions are maintained, it is not normally necessary for an alignment to be carried out.

Filters are usually only fully effective when their input and output impedance coincides with that of the connected equipment or cables. The antenna feeder should be accurately matched, which is not possible without a suitable reflectometer.

The following article describes an effective, well-proved low-pass filter for 145 MHz which guarantees a high order of harmonic suppression if the assembly is carried out carefully and the previously mentioned measures have been carried out. If the given axial dimensions are reduced by a third, the filter will also be suitable for 70 cm transmitters.

3. A COAXIAL LOW-PASS FILTER FOR 145 MHz

Figure 1 shows the assembly of a coaxial low-pass filter. Simply speaking, the low-pass filter can be assumed to be a π-filter for the operating frequency and subsequent harmonics: The short, thick conductor pieces with the impedance Z_2 effect a capacitance between inner and outer conductor. They are joined together by means of a line piece having the high impedance Z_3, which operates as an inductance. The end pieces with the impedance Z_1 only represent an extension of the line used, and serve as connections.

Fig. 1a shows an arrangement by which the required line pieces of differing impedance can be realized by altering the diameter of the inner conductor.

Fig. 1b shows a similar arrangement, however, in this case the diameter of the outer conductor is varied. This arrangement is electrically equal to that given in Fig. 1a. However, it is more difficult to construct and the voltage capabilities are lower.

Fig. 1c shows a complete equivalent circuit of the arrangement shown in Fig. 1a and Fig. 1b. It has also been taken into consideration that each line piece individually represents a π-link. The inductances L 1 to L 3 and the capacitances C 1 to C 3 are not dependent on the frequency. An accurate equivalent circuit is necessary for calculation of the filter.

It is possible, for several coaxial filters as shown in Fig. 1 to be connected in series. In this case, the connection pieces with the impedance Z_1 are deleted. Fig. 2a shows a double-link filter, Fig. 2b a triple-link filter. As is shown by the measurements, the harmonic attenuation is correspondingly greater.

Fig. 3 shows a stripline in a sandwich configuration which is especially suitable for manufacturing low-pass or other filter links. Fig. 4 represents the inner conductor of a double-link low-pass filter.

Fig. 1: Coaxial low-pass filters

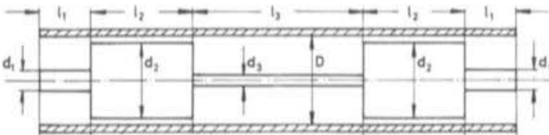


Fig. 1a: Filter with constant outer conductor diameter

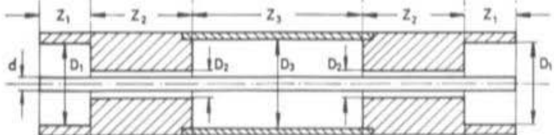


Fig. 1b: Filter with constant inner conductor diameter

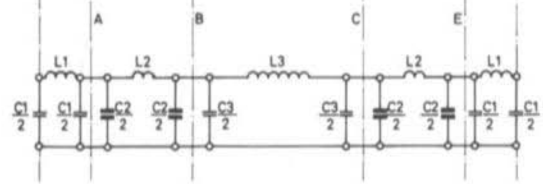


Fig. 1c: Equivalent circuit of the low-pass filter (π -circuit)

Fig. 2: Multi-link coaxial filters

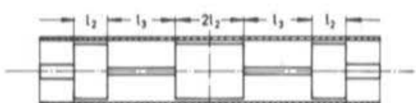


Fig. 2a: Coaxial double-link filter

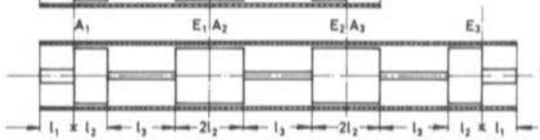


Fig. 2b: Coaxial triple-link filter

Fig. 3: RF stripline with the impedance Z

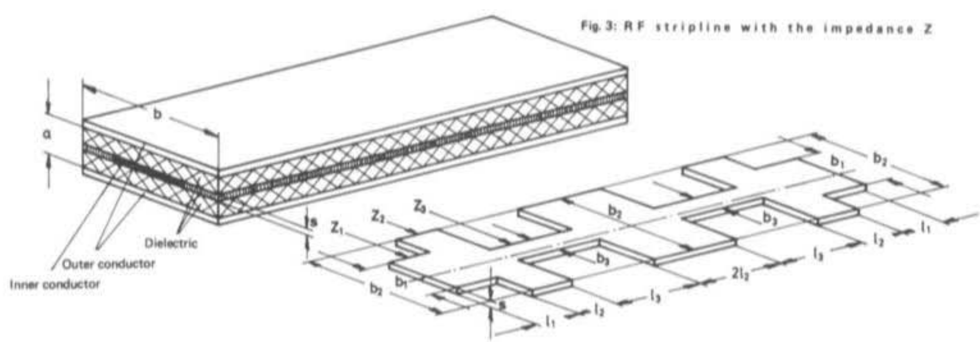


Fig. 4: Inner conductor of a stripline low-pass filter

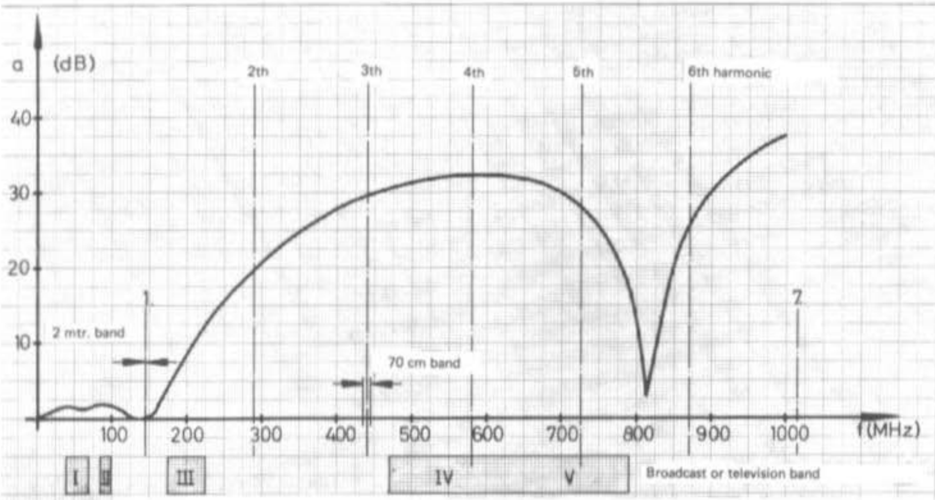


Fig. 5: Measured attenuation response of a single-link 145 MHz filter

Fig. 5 shows the measured attenuation curve on a prototype single-link coaxial low-pass filter according to Fig. 1a or Fig. 10a. All measurements and calculations are based on the conditions given in Fig. 6. The filter is connected between a RF generator having an output impedance Z_{out} of 60 Ω and a coaxial terminating resistor of Z = 60 Ω. For measurements on a filter measuring set, a directional coupler is connected between the generator and the filter as well as between the filter and the terminating resistor.

The attenuation measurements, as well as the theoretical calculation, were made up to a frequency of 1000 MHz. The measurement was commenced at 40 MHz. As is shown in Fig. 5, the single-link filter obtained a maximum attenuation pole of approx. 33 dB at 600 MHz, a minimum attenuation pole is exhibited at 810 MHz between the 5th and the 6th harmonic. The 2nd harmonic of the 145 MHz signal is attenuated by approx. 20 dB, the 3rd harmonic by approx. 30 dB, the 4th harmonic by 33 dB and the 5th harmonic by 28 dB. After the minimum attenuation pole at 810 MHz, the 6th harmonic is attenuated by approx. 28 dB. The 4th and 5th harmonic, which fall into television bands IV and V, are very effectively attenuated. The 2nd minimum attenuation pole is to be found between the 10th and 11th harmonics of 145 MHz.

The passband range for a single and multiple-link coaxial low-pass filter is given in Fig. 7. The insertion loss of the measured single-link filter in the 2 metre band was less than 0.1 dB.

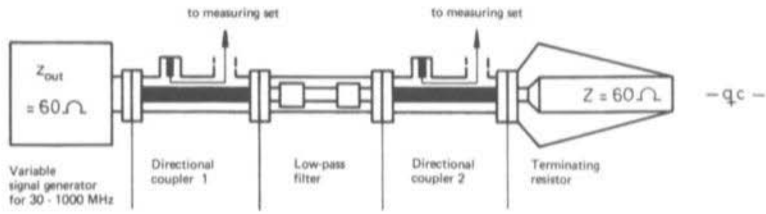


Fig. 6: Arrangement for measuring the filter response

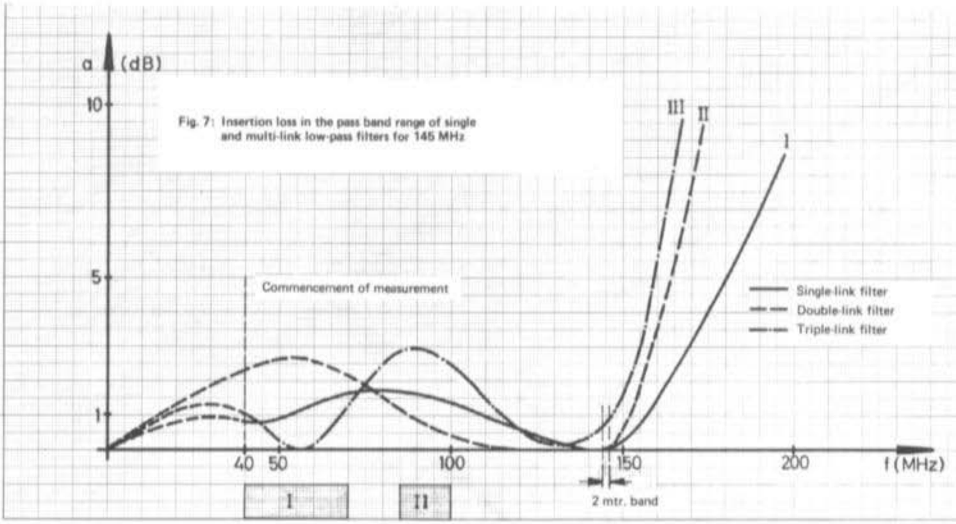


Fig. 7: Insertion loss in the pass band range of single and multi-link low-pass filters for 145 MHz

The impedance response of a single-link low-pass filter for 145 MHz was calculated up to 1000 MHz and is given in Fig. 8. The Smith Chart indicates the impedance X as a function of the frequency when compared to Z = 60 Ω which is the impedance that is exhibited to the signal generator in spite of the 60 Ω terminating resistor. In the passband range of 0 to 145 MHz, the impedance X firstly possesses a capacitive component which becomes real at 112 MHz (approx. 27 Ω) from where it runs with decreasing inductive component until X = Z = 60 Ω is achieved at 145 MHz. In the stopband range above 145 MHz, the reactive impedance becomes very rapidly capacitive and of low impedance, even real at 800 MHz. The reactive impedance then varies rapidly over the inductive range from approx. 800 MHz to approx. 834 MHz. At 834 MHz, X is once again real (approx. 54 Ω) and the filter will allow this frequency to pass. This position should correspond to the minimum attenuation pole given in Fig. 5. This pole is to be found at approx. 810 MHz, at approx. 834 MHz in Fig. 8. The deviation of 24 MHz results from the simplified calculation used for the prototype. After leaving the passband range, the filter quickly reaches the capacitive, low impedance stopband range again. In the non-covered range above 1000 MHz, the filter continues to exhibit maximum and minimum attenuation poles as in the range between 800 and 1000 MHz. A virtually infinite number of cycles occur in the low-impedance reactive impedance range.

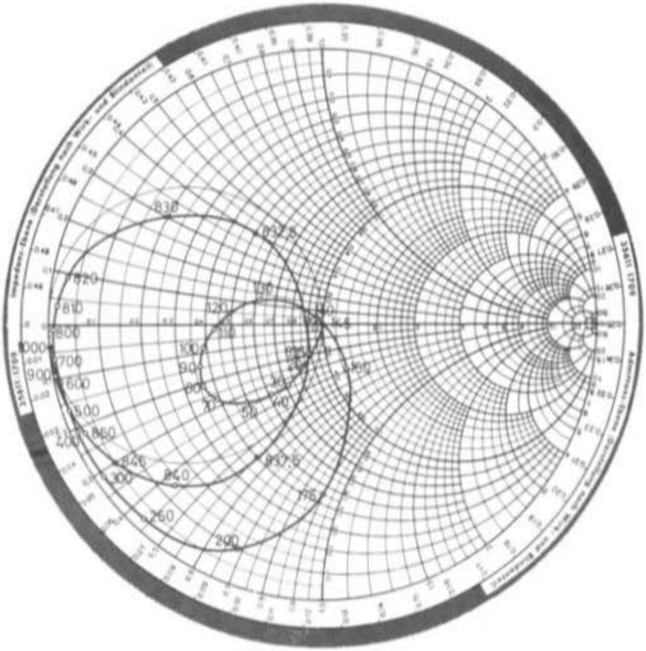


Fig. 8 a:
Calculated response of the impedance at the input of a single-link low-pass filter in the range 0 - 1000 MHz, referred to Z = 60 Ω

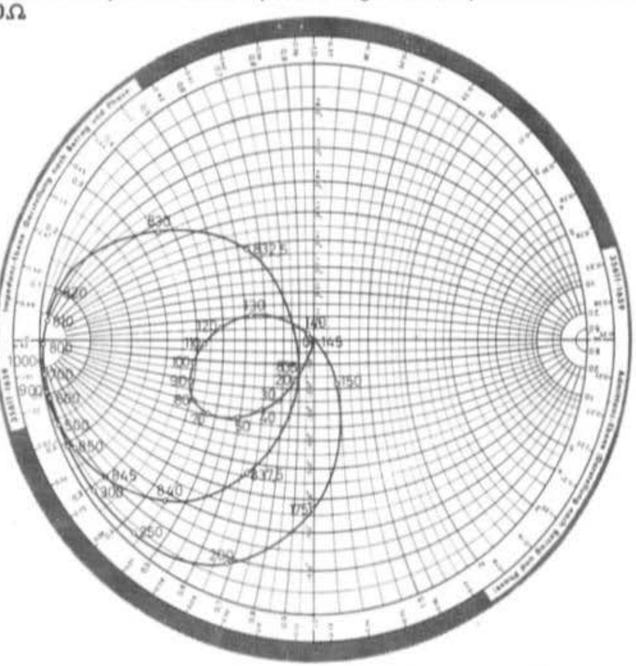


Fig. 8 b:
Calculated response of the impedance at the input of the single-link low-pass filter, according to amount and phase in the range of 0 - 1000 MHz, referred to Z = 60 Ω

The calculation of such a filter is very simple but time consuming. The impedance transformation of each individual line section must be calculated as a function of the frequency in the correct impedance and phase. If a computer is not available for this task, several days will be needed to calculate a single-link filter. Multiple-link filters, of course, require even longer. Besides displaying the impedance according to the active and reactive component (Fig. 8a) it is also favourable to display it according to amount and phase (Fig. 8b).

Figure 9 shows the measured attenuation curves of single, double and triple-link coaxial low-pass filters for 145 MHz according to Fig. 1a, or Fig. 2a and 2b. Figure 7 shows the extended passband range of these filters. Table 1 gives the attenuation values for the individual harmonics.

Power attenuation dB	Frequency MHz	145 MHz coaxial low-pass filter single-link	double-link	triple-link
1st harmonic	145	<0.1	<0.1	<0.3
2nd harmonic	290	20.5	38.0	(61)
3rd harmonic	435	30.0	52.5	(75)
4th harmonic	580	32.5	54.5	(73)
5th harmonic	725	28.0	41.0	(60)
6th harmonic	870	26.5	41.0	(62)

Table 1 Power attenuation of 145 MHz low-pass filters

As shown in Figure 9, or table 1, multi-link filters exhibit a correspondingly higher attenuation. At 800 MHz resonance points exist which, due to the reversal of the reactive impedance curve appear repeatedly. It was, on the filter measuring set used for the measurement, difficult to read off values over 60 dB. This is the reason why these values are given as dashed lines or in brackets in Figure 9 and Table 1 respectively.

4. MECHANICAL ASSEMBLY OF THE 145 MHz COAXIAL FILTER

Figures 10 and 11 show the mechanical assembly of the prototype single and double-link coaxial 145 MHz filters. The 2 metre power capacity of the given low-pass filters is several kilowatts which is far more than permissible for amateur transmissions. The upper power limit of the low-loss filters is determined by the voltage capability of the teflon insulating collars used.

The outer conductor of the filter is formed by a tube with flange as shown in Figure 10c. The length differs according to the number of links. Besides the connection pieces (part 7 or 8 in Fig. 11f, 11g) the inner conductor consists of the thick bushings (part 2 or 3 in Fig. 11a, 11b)and the thin inner conductor pieces (part 4 in Fig. 11 c). This arrangement is screwed together according to Figure 10a, 10b or 10d and soldered. Teflon collars (part 5 or 6 in Fig. 11d or 11e) are used to centre and insulate the thick inner conductor pieces. They are slid over the completed inner conductor. The inner conductor should be easily slid into the outer conductor, it should not be too loose or too tight. The protrusion of 3 mm in length and 14 mm inner diameter made in the teflon collar part 5 or 6, serves as a stop to ensure that these pieces are not shifted when placing the inner conductor within the outer conductor. These stops should be all located on the same side, namely the side that is firstly inserted into the outer conductor.

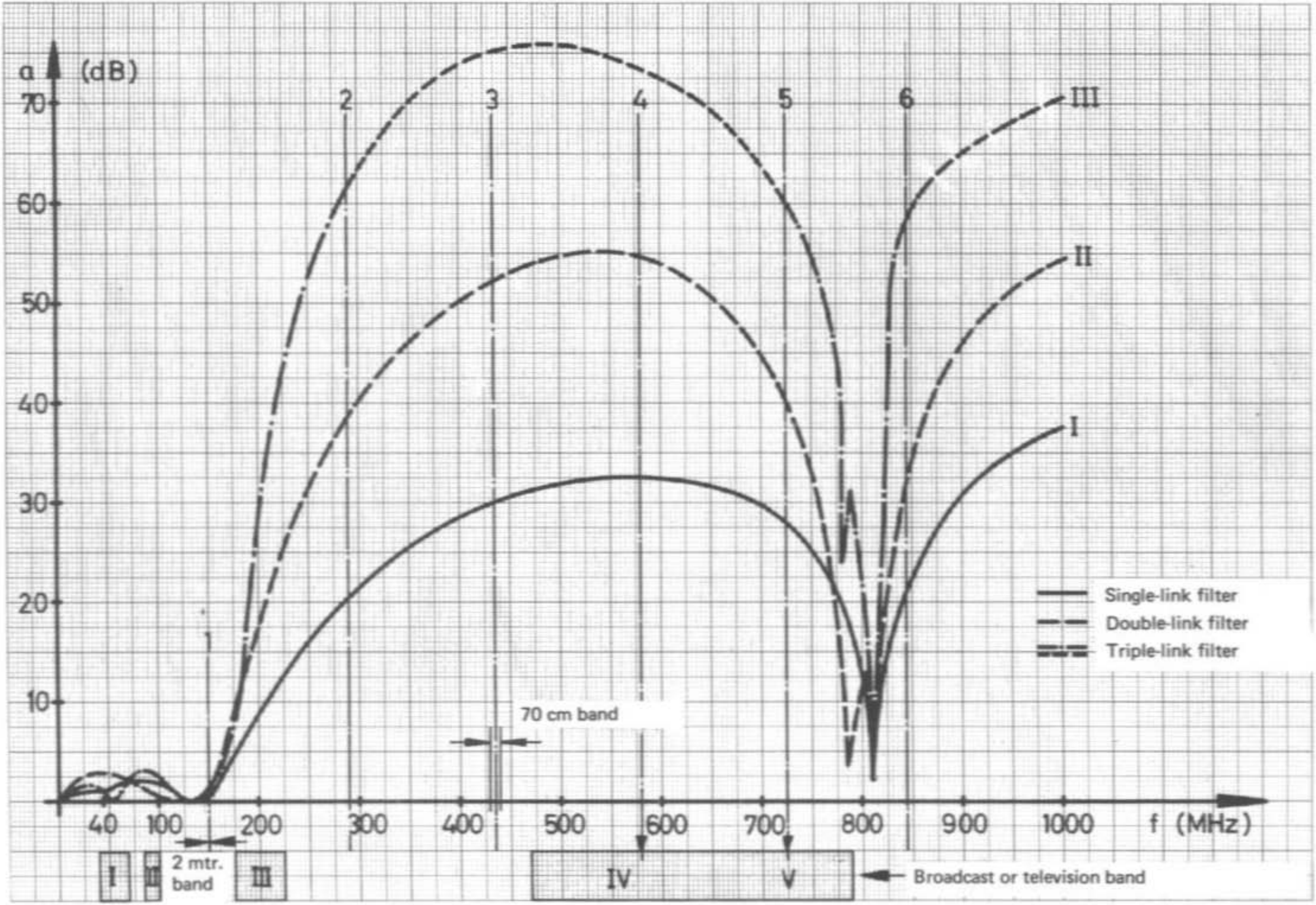


Fig. 9: Measured attenuation response of single and multi-link coaxial low-pass filters for 145 MHz

The inner conductor of the coaxial socket (part 9) is inserted into the connection piece (part 7) up to a spacing of 2 mm (Fig. 11h) and soldered. After screwing this coaxial socket to the flange of the outer conductor, the inner conductor is fixed. The coaxial socket at the upper end of the filter is inserted into the slotted or spring-loaded part 8 as shown in Fig. 11g and also screwed to the outer conductor flange. After this, the assembly process is complete.

The spacing of 2 mm between the flange surface F-F of the coaxial socket and the end of part 7 or 8 is used to compensate the point of impedance discontinuity which results from the sudden increase of the cross section. Figure 11 shows this clearly, and gives further dimensions for the flange.

The mechanical dimensions and tolerances should be maintained as accurately as possible. The teflon collars are placed on the corresponding inner pieces and lathed. The inner diameter of the outer conductor (16.00 mm) can vary by 0.1 mm. This is the reason why the dimension "a" in Fig. 11d and 11e must be made so that the collars mounted on the bushings (Fig. 11a and 11b) can be inserted easily into the associated outer conductor (part 1). These tolerances cause a slight variation of the impedance, but do not vary the filter characteristics noticeably.

At the passband frequency of 145 MHz, the impedance of the filter amounts to exactly 60 Ω. The thick portions of the inner conductor possess approximately 6 Ω, the thin portions approx. 96 Ω. The calculation was carried out with these rounded values, which coincide to a great degree. Naturally, an impedance of 6 Ω is easier to construct with a greater conductor cross section. With the design given in Fig. 1b, the very thin insulation collars would be very difficult to achieve, which is also valid when using small cross sections for the design given in Fig. 1a. Filters where foil was used instead of the teflon collars (part 7 and 8) were not tested.

The dimension "a" in Fig. 10c amounts to 301 mm for single-link filters, and 562 mm for double-link filters.
The dimension "b" in Fig. 10d amounts to 297 mm for single-link filters, and 558 mm for dluble-link filters.

By increasing the number of sections, both dimensions will be increased by 261 mm per filter link. It is not really necessary to silver-plate the brass pieces; however, if this is carried out, the silver-plating should not be more than the given tolerances.

5. PARTS LIST

The following parts are required for a single-link 145 MHz filter:

Part	Fig.	Pieces	Designation, Material, Processing, Notes
1	10c 11h	1	Outer conductor, brass tube 22 x 3 mm, flanges made from 3 mm brass plate hard soldered
2	11a	2	Inner conductors, round brass 15 mm in diameter
4	11c	1	Inner conductor, round brass 4 mm in diameter
5	11d	2	Insulating collars, teflon tube 14.00/03.50
7	11f	1	Inner conductor, round brass, 6 mm in diameter
8	11g	1	Inner conductor, round brass, 6 mm in diameter
9	(11h)	2	Coaxial sockets (60 Ω)

Fig. 10: Coaxial filter for 145 MHz (Z = 60 Ω)

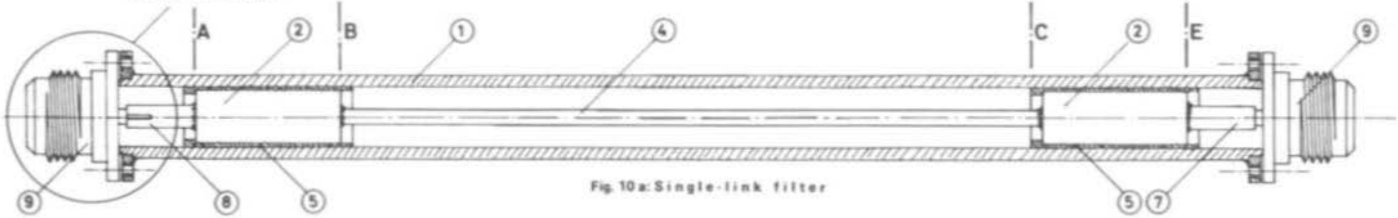


Fig. 10 a: Single-link filter

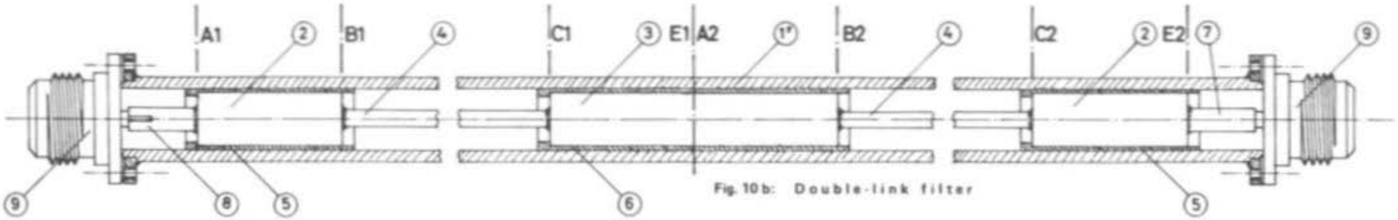
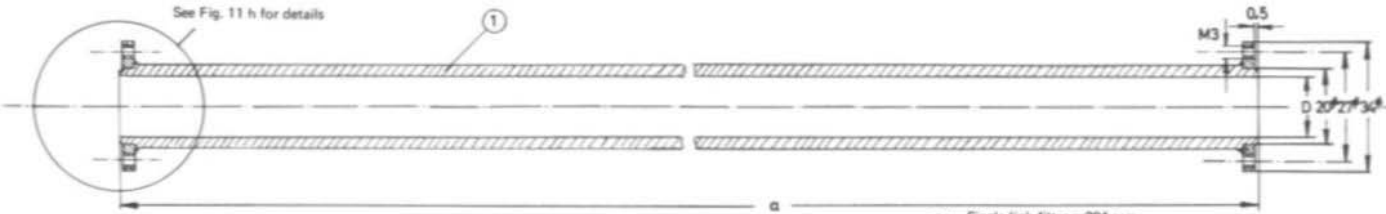


Fig. 10 b: Double-link filter



a: Single-link filter = 301 mm
Double-link filter = 562 mm

Fig. 10 c: Outer conductor of the filter

b: Single-link filter = 297 mm
Double-link filter = 558 mm

Fig. 10 d: Inner conductor of the double-link filter
(without insulating collars)

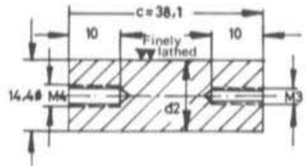


Fig. 11 a: Inner conductor, part 2

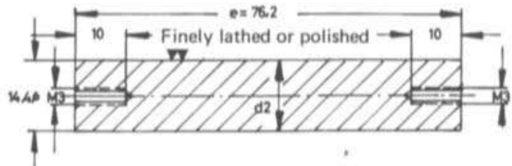


Fig. 11 b: Inner conductor, part 3

Fig. 11 c: Inner conductor, part 4

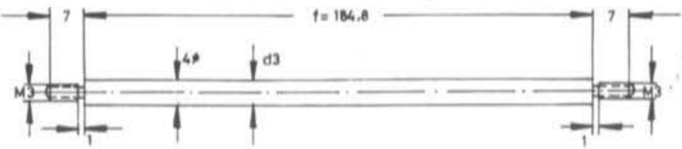


Fig. 11 d: Insulating collar, part 5

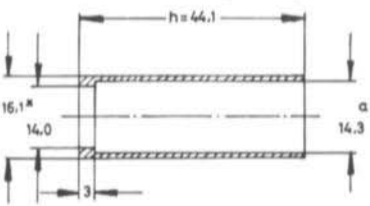
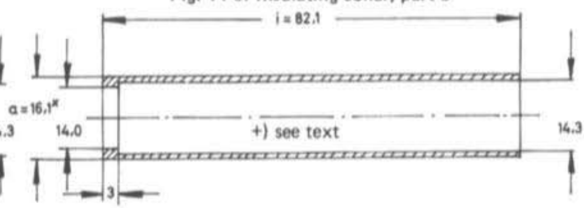


Fig. 11 e: Insulating collar, part 6



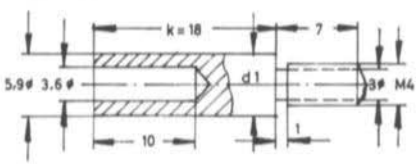


Fig. 11 f: Inner conductor, part 7

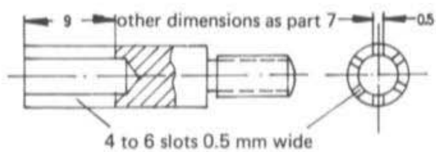


Fig. 11 g: Inner conductor, part 8

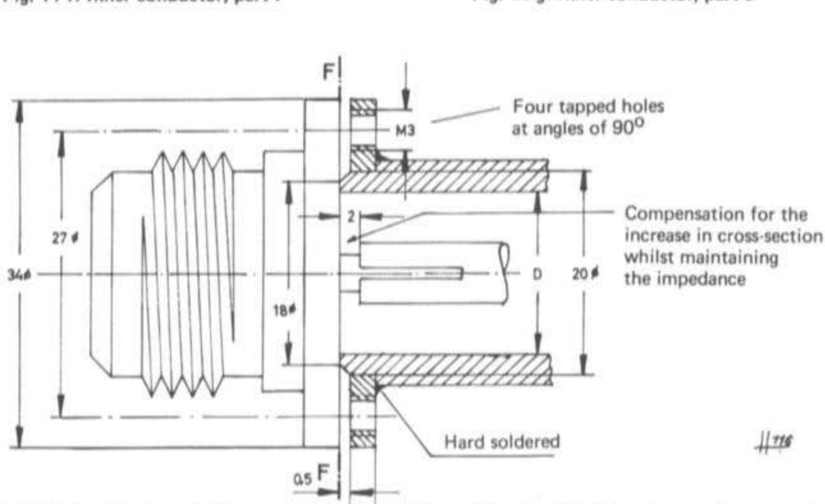


Fig. 11 h: Construction details

Fig. 11: Individual parts of the filter

For the double-link filter, a somewhat longer part 1 is required. In addition to this, one part 4, 3 and 6 will be required. Part 3 is made out of 15 mm round brass, part 6 from teflon tube.

6. RESULTS

Several filters have been tested. The efficiency of the filters can be easily tested with the 435 MHz harmonic of a two metre transmitter. For instance, if a S9 signal was present on 70 cm, this signal will no longer be audible after placing a double-link two metre coaxial filter into the 2 metre transmitter feedline.

7. NOTES

It has already been mentioned that the two metre filter represents a mis-match at frequencies lower than 144 MHz. This (bandpass) characteristic of the filter ensures that the spurious 48 MHz and 72 MHz signals from the frequency multiplier stages are effectively suppressed. Experiments made, proved this assumption.

8. REFERENCES

(1) D. Grossmann: Simple, Compact PA Stages for Two Metres
VHF COMMUNICATIONS 2 (1970), Ed. 1, p. 45-55 and Ed. 2, p. 111-122

ELECTRONICALLY STABILIZED POWER SUPPLY
WITH DC-DC CONVERTER

by K. P. Timmann, DJ 9 ZR

1. INTRODUCTION

The described power supply is especially designed for operation in conjunction with the 5 W two metre SSB-transmitter DJ 9 ZR 001 (1). However, it is also suitable for other applications.

An electronically stabilized power supply supplies an output voltage of 13.5 V. The output voltage of a DC-DC converter is added to the voltage so that a total output voltage of 28 V is available in the transmit mode.

The power supply will not be overloaded even when the output power of the SSB-transmitter is increased to 10 W. With mobile operation from a 12 V car battery, this voltage can be directly fed to connection point 18 of the printed circuit board, or into a socket connected to this connection point.

2. CIRCUIT DIAGRAM (Figure 1)

The power supply consists of a stabilized power supply and a DC-DC converter that are built up as separate units on a common printed circuit board. This means that these units can be utilized independent of another by cutting the printed circuit board.

2.1. THE STABILIZED POWER SUPPLY

The power transformer which is not shown in Figure 1 - should provide an output-AC voltage of 24 V. If the power supply is only to be used for the DJ 9 ZR transmitter with an output power of 5 W, the secondary winding must be dimensioned for an output current of approx. 1 A.

If the transmitter is to provide more than 5 W output power, or if other consumers are to be connected, it will be necessary for the secondary winding to be dimensioned for a current of approx. 2 to 3 A. Special attention should be paid that the AC inputs (connection points 3 and 4) are floating in order to be able to drive the rectifier bridge D 1 (4 x 1 N 4004) symmetrically. The DC voltage gained in this manner is filtered by a 2500 μF capacitor and fed to the stabilized power supply comprising the pass transistor T 1 (BD 106) and the two DC-voltage amplifier transistors T 2 (BC 184) and T 3 (BC 182). The reference voltage is provided by a zener diode (ZF 13) at the output. A silicon diode (1 N 914) is connected in series with the zener diode for temperature compensation. The 220 Ω resistor in the base connection of transistor T 3 maintains the current via the zener diode at approx. 3 - 5 mA. The parallel-connected 250 μF capacitor is used to filter any residual AC voltage, or voltage peaks. If the output voltage deviates from the nominal value, the current via the zener diode will also alter. This affects the voltage drop across the 220 Ω base resistor of transistor T 3 which causes an amplified variation of its collector-emitter voltage. The transistor T 2, whose base is directly connected to the collector of T 3, controls the base voltage of the pass transistor T 1 via its collector-emitter path. This causes the resistance of transistor T 1 to vary until the nominal value is re-obtained.

The DC voltage of 13.5 V provided from the stabilized power supply only fluctuates by 0.2 V between non-load and full-load, even when currents of up to 1.5 and 2 A are taken.

The output voltage is fed via the parallel-connected 1 N 4004 diodes to connection point 18 of the printed circuit board. In addition, this connection point is connected to the input of the DC-DC converter. When operating in conjunction with a 12 V car battery, the diode gate ensures that no reaction on the stabilized power supply can occur.

Connection point 7 is connected to the base of transistor T 2. For switching on and off, a switch can be connected between here and ground. When the base of this transistor is grounded, no collector current can flow and the pass transistor will not receive current. This means that no voltage is available at the output. It is possible in this manner for a power-less DC voltage switching to be made with a small on - off switch. If the switch connection is long, such as to a remote switch etc., or of RF injection is to be expected, connection-point 7 should be connected to connection point 1 or another ground point via a 25 μF capacitor.

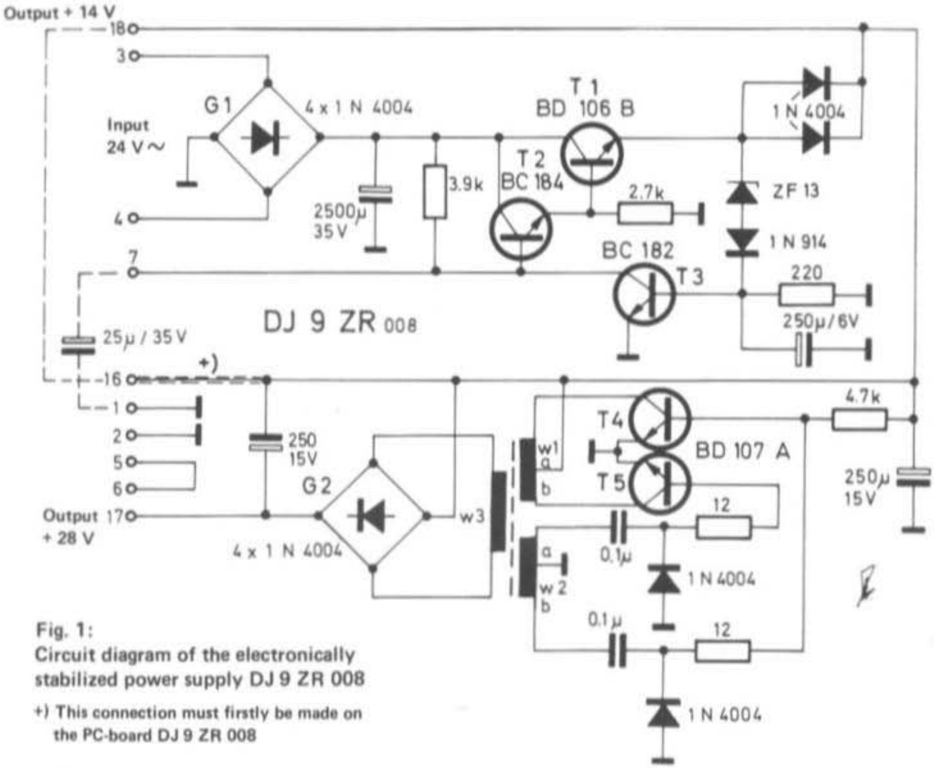


Fig. 1:
Circuit diagram of the electronically stabilized power supply DJ 9 ZR 008

+) This connection must firstly be made on the PC-board DJ 9 ZR 008

2.2. THE DC-DC CONVERTER

The input of the DC-DC converter is connected to the output of the stabilized power supply and is also connected to connection point 18. A new type of DC-DC converter circuit is used by which the full voltage from winding w 2 is not fed to the bases of the two converter transistors T 4 and T 5 (BD 107 A), but the differential voltage from winding w 2. The differential links consist of the two 0.1 μF capacitors, the diodes 1 N 4004 and the 12 Ω resistors connected in series with the internal base resistance of the switching transistors. To ensure commencement of oscillation, a 4.7 kΩ resistor is provided in the circuit. This resistor can remain in the base circuit, where it ensures that transistor T 4 is fully driven. Such a high current will then flow via this transistor that the converter will surely commence oscillation due to the magnetic flow in winding w 1.

Winding w 3 supplies the alternating voltage which is rectified in the bridge rectifier D 2 (4 x 1 N 4004). The DC voltage gained in this manner is added to the input DC voltage. This means that a voltage of 28 V is available, which is fed to connection point 17. At the input of the converter, a 250 μF capacitor is connected to ground in order to avoid the introduction of switching peaks from the converter into the operating voltage source. In order to filter the output voltage, a 250 μF capacitor is connected parallel to the output. The voltage difference between non-load and full-load is less than 1 V; the alternating voltage component at full-load is less than 0.1 V.

3. MECHANICAL ASSEMBLY

The electronically stabilized power supply and the DC-DC converter are accommodated on a common printed circuit board DJ 9 ZR 008 with the dimensions 167 mm by 84 mm. The printed circuit board is shown in Figure 2; the corresponding component location plan in Fig. 3. All connections are made in the form of a contact strip on one side of the board. This means that the PC-board can be plugged into a contact strip provided for this purpose. The heat sink for the pass transistor T 1 is mounted in the centre of the printed circuit board. The converter core is also screwed directly to the printed circuit board. No special mount is necessary. A photograph of the complete module is given in Fig. 4.

3.1. THE CONVERTER TRANSFORMER

Due to the high efficiency of 90% it was possible for a small core to be used for the converter. The heating of the transformer at full-load (0.7 A) remains within normal limits. A temperature of approx. 70 °C is achieved at full load.

A ferrite pot core manufactured by Siemens is used. The coil holder has 3 chambers:

Winding details:
- w 1: 2 x 15 turns of 0.3 mm dia. (29 AWG) enamelled copper wire, double wound
- w 2: 2 x 6 turns of 0.2 mm dia. (32 AWG) enamelled copper wire, double wound
- w 3: 18 turns of 0.3 mm dia. (29 AWG) enamelled copper wire.

With the push-pull windings, the end of winding half "a" is connected to the beginning of winding half "b" to form the centre tap. It is then only necessary for the outer ends of the winding to be exchanged if the converter does not commence oscillation due to incorrect polarity.

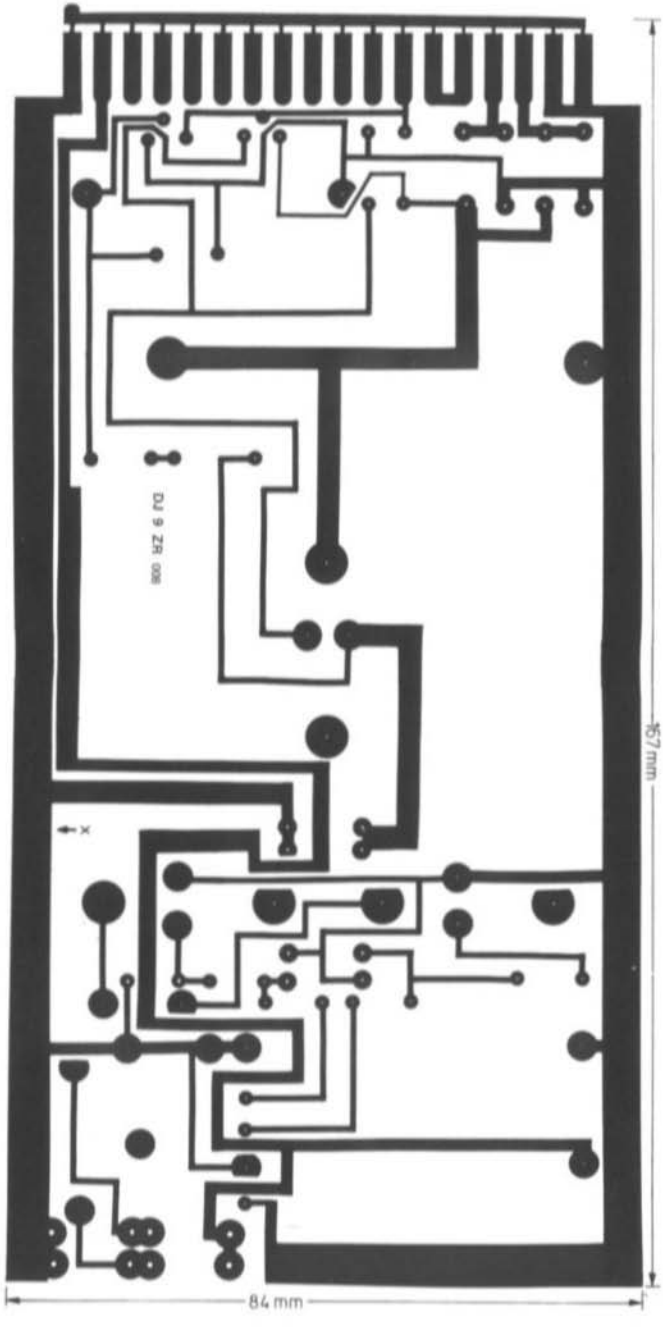


Fig. 2: Printed circuit board of the electronically stabilized power supply with DC-DC converter DJ 9 ZR 008

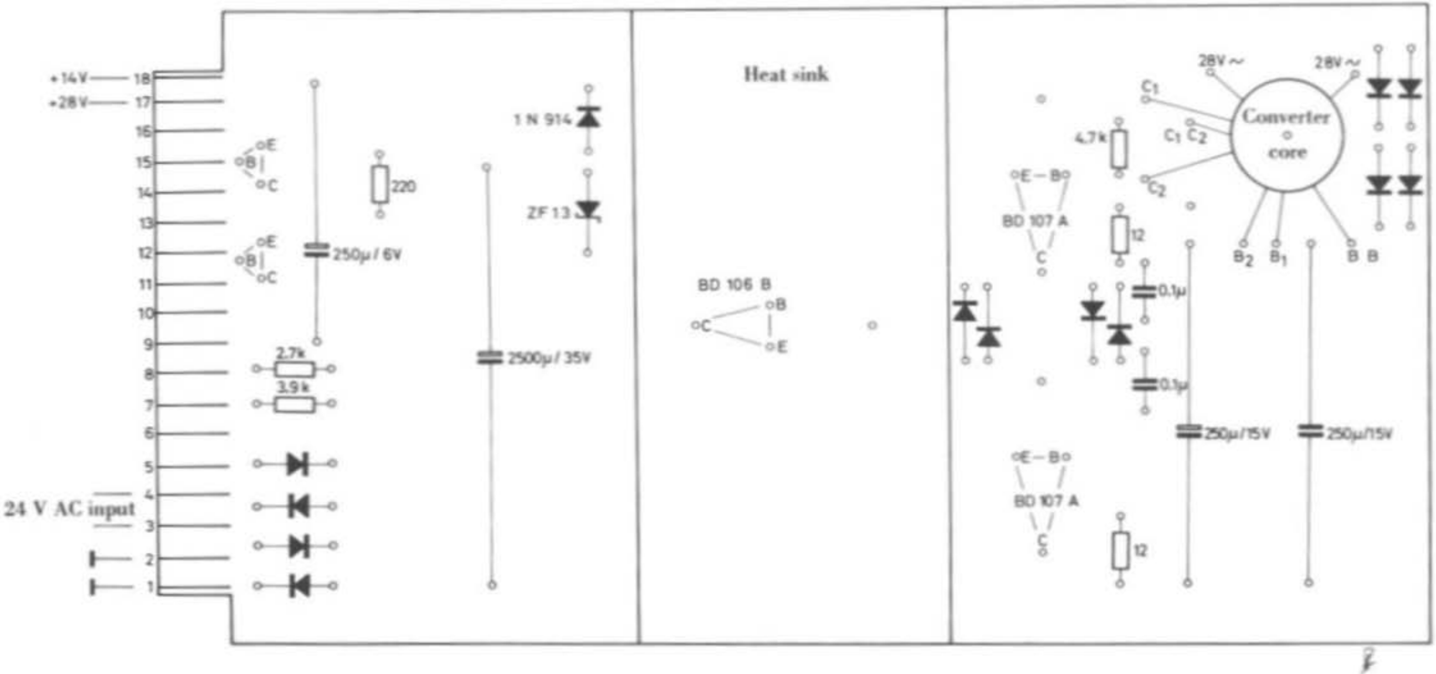


Fig. 3: Component location plan to printed circuit board DJ 9 ZR 008

3.2. COMPONENTS LIST

Resistors:	2	12 Ω	1/4 W	1	220 Ω	1/4 W;
	1	2.7 kΩ	1/4 W	1	3.9 kΩ	1/4 W;
	1	4.7 kΩ	1/4 W;			

Capacitors:
2 ceramic disc capacitors 0.1 μF 30/60 V
2 electrolytic capacitors 250 μF, 15 V
1 electrolytic capacitor 250 μF, 6 V
1 electrolytic capacitor 2500 μF, 35 V

Semiconductors: T 1: BD 106 B (ITT-Intermetall) or 2 N 3054 (RCA)
T 2: BC 184 (Texas Instruments) or
BC 109, BC 131, BC 149, BC 169, BC 173, BC 199,
BC 209, BC 239, PBC 109, SC 109, 2 N 2926
T 3: BC 182 (Texas Instruments) or
BC 107, BC 129, BC 147, BC 167, BC 171, BC 197,
BC 207, BC 237, PBC 107, SC 107, 2 N 3904
T 4, T 5: BD 107 A (ITT-Intermetall) or 2 N 3054 (RCA)

12 silicon diodes 400 V/1 A: 1 N 4004 (ITT-Intermetall) or similar
1 silicon zener diode ZF 13 (ITT-Intermetall) or 1 N 4107 or similar
1 silicon diode 1 N 914 or similar

Converter transformer:
1 Siemens ferrite pot core type B 65 671 - L 0000-R 026 consisting of two pieces
1 coil former with 3 chambers, type 65 672 - A 000 - M 003
approx. 3 metres of 0.3 mm dia. (29 AWG) enamelled copper wire
approx. 1 metre of 0.2 mm dia. (32 AWG) enamelled copper wire
1 mounting screw M 4, 25 mm long with nut

Mechanical pieces: 1 Heat sink (see photograph)
1 Printed circuit board DJ 9 ZR 008 (available from the publishers)
6 screws 8 mm long with nuts

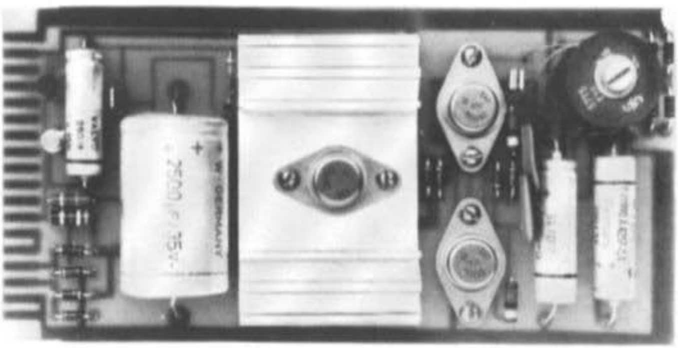


Fig. 4: Photograph of the power supply module DJ 9 ZR 008

3.3. POSSIBILITY OF USING OTHER SEMICONDUCTOR TYPES

The transistor BD 106 B and the two transistors BD 107 A can be replaced by the 2 N 3054. However, the mounting hole spacing of the 2 N 3054 is somewhat greater than the originally planned transistors. This means that the mounting hole in the printed circuit board which is adjacent to a conductor lane must be somewhat shifted. It should be noted, however, that the spacing to the neighbouring conductor lane will be reduced. In order to avoid an unwanted contact, a heat-resistant insulating washer (mica or teflon) should be placed between the printed circuit board and the mounting nut.

4. MEASURED VALUES

4.1. Stabilized power supply:
Output voltage: 13.5 V
Maximum load: 2 A
Voltage fluctuation between non-load and full-load: 0.2 V

4.2. Power supply with DC-DC converter:
Output voltage: 28 V
Maximum load: 0.7 A
Voltage fluctuation between non-load and full-load: 1 V
AC component of the output voltage: 0.1 V
Efficiency: 90%
Heating of the converter transformer at full-load: max. 70 °C

5. NOTES

When using power supplies in transceivers, it is often favourable for the power supply of the transmitter to be switched off, whereas the receiver operating voltage is still available. The described module also allows this type of operation: Since the converter input is connected to the output of the stabilized power supply, it is, however, necessary for the conductor lane to be broken at position "x" (Fig. 3). The input of the DC-DC converter is then connected to a free-connection point using a wire bridge. A transmit-receive relay can then connect the DC-DC converter input to the power supply in the transmit mode.

As was already mentioned in the introduction, the stabilized power supply and the DC-DC converter can be separated when they are not to be operated together. If the two modules are to be mechanically separated, the printed circuit board can be cut; the cutting edge is located between the diode gate and the DC-DC converter transistors.

6. AVAILABLE PARTS

The printed circuit board DJ 9 ZR 008 as well as the semiconductors and other components are available from the publishers or their national representatives (see advertising page).

7. REFERENCES

(1) K.P. Timmann: A 5 Watt Transistorized SSB Transmitter for 145 MHz VHF COMMUNICATIONS 1 (1969), Edition 2, Pages 73-82

A SIMPLE ROTARY COAXIAL-COUPLING

by P. Saffran, DC 8 OH

1. INTRODUCTION

Generally speaking, the rotators used by radio amateurs for directional antennas possess two end contacts which prevent the antenna from making more than one revolution. This measure ensures that a directly connected coaxial cable looped past the rotator will not be wound onto the mast after a number of revolutions and finally broken. Although this principle is mostly used, it is often a disadvantage not being able to rotate to the new direction in the shortest manner.

One means of avoiding this disadvantage is for a rotary coaxial-coupling to be used in the coaxial cable so that the cable can be rotated continuously.

2. DESCRIPTION

The simple rotary coaxial-coupling described here exists essentially of a socket SO 239 and the matching plug PL 259. These two connectors are, with the aid of several simple modifications and additions, joined to form a rotary coupling. The assembly is shown in the sectional drawing given in Figure 1.

The principle is based on the fact that the inner portion of the plug can be made to rotate with respect to the blocked collar (8) by adding ball bearings (2) and (9). The coupling is manufactured in the following manner:

The teeth on the socket (4) and the matching pins on the plug (5) are removed. This is achieved carefully with the aid of a file. After the inner conductor (3) of the coaxial cable (15) and the inner conductor pin (1) of the plug have been soldered, the pin of the plug is shortened to approximately 6 mm length and rounded with a file. The collar (8) is now drilled with a hole (10) of 3 mm diameter opposite the internal thread. A brass nut (11) is soldered above this hole. The collar is then removed and pushed onto the cable (15).

Approximately 30 steel ball bearings (9) of 1 mm diameter are now placed with a little grease onto the inner side of the plug behind the short threaded disc (7). After a small pressure spring (6) of approximately 4 mm in length and a ball bearing (2) of 3.5 mm diameter have been inserted into the socket, the collar is carefully pushed over the ball bearing (9), the socket and plug plugged together and the collar tightened until the plug just moves freely within the collar.

After locking the collar in this position using a grub screw (12), the rotary coupling will be ready for use.

3. EXPERIENCE

A rotary coaxial-coupling manufactured in the described manner has been used for mobile applications for a few years without noticeable wear. The standing wave ratio at 144 MHz corresponds to that of the plugged connection. The contact resistance is hardly measurable with amateur means. The maximum power capacity was not determined however, it is most certainly lower than that of the plugged connection.

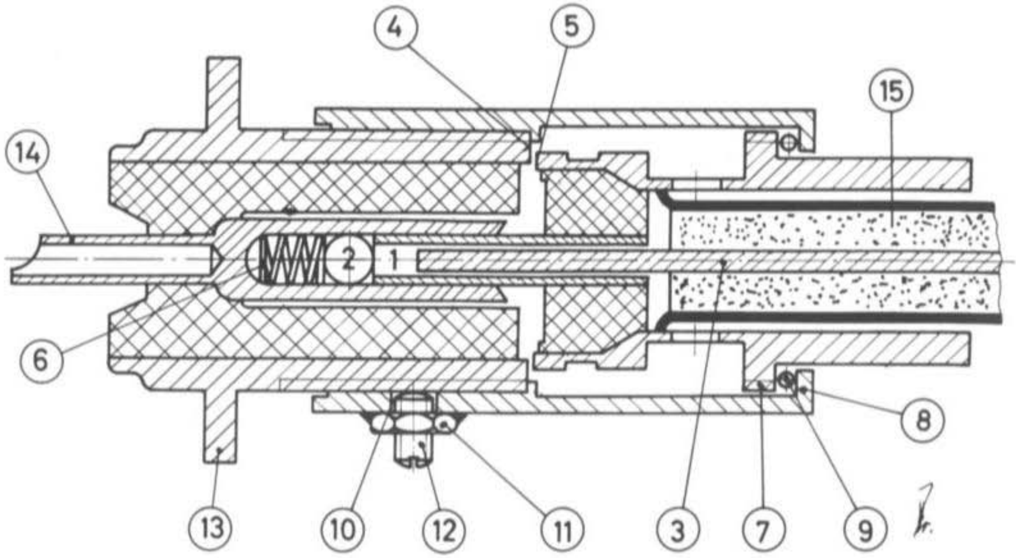


Fig. 1: Rotary coaxial-coupling manufactured from a SO 239 socket and a PL 259 plug.

1. Inner conductor pin of the PL 259 plug
2. Steel ball bearing, 3.5 mm diameter
3. Inner conductor of the coaxial cable
4. Teeth of the socket
5. Pins on the plug (2)
6. Pressure spring, 4 mm long, 3 - 3.4 mm diameter
7. Short threaded disc on the inner part of the plug
8. Plug collar
9. 30 steel ball bearings, 1 mm diameter
10. Hole drilled in the collar, 3 mm diameter
11. Nut, soldered above the hole (10)
12. Grub screw
13. Flange of the SO 239 socket
14. Inner conductor of the socket
15. Coaxial cable

MODIFICATIONS FOR THE S-METER AND CONTROL VOLTAGE CIRCUITS IN THE 9 MHz PORTION OF THE DL 6 HA TRANSCEIVER

by R. Störmer, DJ 3 FT and G. Laufs, DL 6 HA

A few improvements that can be made by modifications on the 9 MHz transceiver module DL 6 HA 002 were brought to the notice of the editorial staff by the two authors. These modifications are now to be explained.

1. THE S-METER CIRCUIT

In the original circuit (1) it was assumed that a meter having 1 mA FSD and an impedance of 1 kΩ was available. Since this is not always the case, the voltmeter circuit shown in Fig. 1 is recommended. The trimmer potentiometer of 50 to 100 kΩ allows the FSD of the sensitive meter to be adjusted. The control characteristic is determined independently by the adjustment of the emitter resistor R 33.

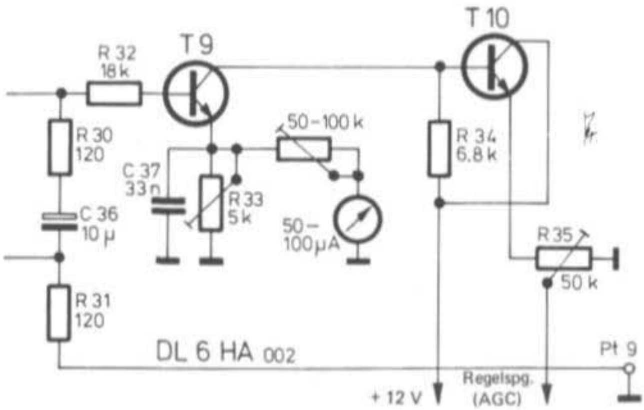


Fig. 1: Improved S-meter circuit and adjustable control voltage for the 9 MHz portion of the DL 6 HA transceiver

2. THE AUTOMATIC GAIN CONTROL

The numerous difficulties with the automatic gain control are caused by the fluctuations in the current amplification factors of the used transistors. It is therefore recommended that transistors possessing the same current amplification be selected for the transistors T 5a, T 6a, T 202, T 205. In order to be able to adjust the exact commencement of the control process, resistor R 35 (51 kΩ) should be replaced by a trimmer resistor of 50 kΩ. The conductor lane for the control voltage is broken at the emitter of transistor T 10 and connected to the wiper of the trimmer potentiometer with a wire bridge. The new circuit is also shown in Fig. 1.

For adjustment, the potentiometer wiper is firstly brought to the emitter side. After this, the receiver is tuned to a signal and the wiper slowly adjusted towards the ground side. The correct adjustment is just before a reduction of the gain is observed.

3. FURTHER EXPERIENCE

In one case, the transistor T 202 (DL 6 HA 004) caused a noticeable deterioration of the noise factor. If this effect is observed, the transistor should be replaced by another of the same type. However, in order to be able to reconstruct the equipment without problems, an electrolytic capacitor of 10 µF/12 V should be connected from the connection point Ch 201/270 Ω to ground. This capacitor will then short any possible noise voltages caused by the control transistor.

In order to improve the stability, it is advisable to make a bridge between the ground conductors of the IF amplifier, approx. in the vicinity of components R 26, R 24, C 27.

4. REFERENCES

(1) G. Laufs: A SSB Transceiver with Silicon Transistor Complement
Part 2 : The 9 MHz Transceiver
VHF COMMUNICATIONS 2 (1970), Edition 2, Pages 65-75

Centimetre / inches

		0	1	2	3	4	5	6	7	8	9	1o cm
cm	0	0	.394	.787	1.181	1.575	1.969	2.362	2.756	3.150	3.543	3.937
"	.1	.039	.433	.827	1.221	1.614	2.008	2.402	2.795	3.189	3.583	
"	.2	.079	.472	.866	1.260	1.654	2.047	2.441	2.835	3.228	3.622	
"	.3	.118	.512	.906	1.299	1.693	2.087	2.480	2.874	3.268	3.661	
"	.4	.158	.551	.945	1.339	1.732	2.126	2.520	2.913	3.307	3.701	
"	.5	.197	.591	.984	1.378	1.772	2.165	2.559	2.953	3.347	3.740	
"	.6	.236	.630	1.024	1.417	1.811	2.205	2.598	2.992	3.386	3.780	
"	.7	.276	.669	1.063	1.457	1.850	2.244	2.638	3.032	3.425	3.819	
"	.8	.315	.709	1.102	1.496	1.890	2.284	2.677	3.071	3.465	3.858	
"	.9	.354	.748	1.142	1.535	1.929	2.323	2.717	3.110	3.504	3.898	

Millimetre / inches

		0	1	2	3	4	5	6	7	8	9	10 mm
mm	0	0	.0394	.0787	.1181	.1575	.1969	.2362	.2756	.3150	.3543	.3937
"	.1	.0039	.0433	.0827	.1221	.1614	.2008	.2402	.2795	.3189	.3583	
"	.2	.0079	.0472	.0866	.1260	.1654	.2047	.2441	.2835	.3228	.3622	
"	.3	.0118	.0512	.0906	.1299	.1693	.2087	.2480	.2874	.3268	.3661	
"	.4	.0158	.0551	.0945	.1339	.1732	.2126	.2520	.2913	.3307	.3701	
"	.5	.0197	.0591	.0984	.1378	.1772	.2165	.2559	.2953	.3347	.3740	
"	.6	.0236	.0630	.1024	.1417	.1811	.2205	.2598	.2992	.3386	.3780	
"	.7	.0276	.0669	.1063	.1457	.1850	.2244	.2638	.3032	.3425	.3819	
"	.8	.0315	.0709	.1102	.1496	.1890	.2284	.2677	.3071	.3465	.3858	
"	.9	.0354	.0748	.1142	.1535	.1929	.2323	.2717	.3110	.3504	.3898	

PRICE LIST of NEW EQUIPMENT described in VHF COMMUNICATIONS 3/1970
AND MISCELLANEOUS ITEMS

For earlier equipment, see price list in edition 1 and 2/1970

DM 3.60 = US$ 1.00 - DM 8.60 = £ 1.0.0

DL 6 HA 004 — 9 MHz-14 MHz Transmit-Receive Converter

PC-board only	DM	16,--
Set of coil formers and trimmers	DM	7,70
Set of semiconductors	DM	54,30
Kit: PC-board, coil formers, trimmers, semiconductors and special drill	DM	78,--

DL 6 HA 005 — 14 MHz - 144 MHz Transmit Converter

PC-board only	DM	8,--
Set of coil formers and trimmers	DM	4,30
Set of transistors	DM	20,50
Kit: PC-board, coil formers, trimmers, transistors	DM	32,20

DL 6 HA 006 — VFO-Module

PC-board only	DM	8,--
Set of transistors	DM	6,20
Kit: PC-board, transistors	DM	14,20

DL 6 HA 007 — Low-pass Filter

PC-board only	DM	3,--
Kit: PC-board, coil formers, transistor	DM	7,30

Oscillator board for the MOSFET-Converter DL 6 HA 001/14

PC-board DL 3 YK 001 (available from November 1970)	DM	4,--

DK 1 PN — Tripler 432/1296 MHz

Set of 5 trimmers	DM	12,50

DJ 9 ZR 008 — Stabilized power supply with DC-DC converter

PC-board only	DM	14,--

Kit with semiconductors and transformer core on request

MISCELLANEOUS

NEW

Interdigital bandpass filter for the 23 cm band, similar to the description in QST, March 1968; ready to operate, with BNC connectors	DM	82,30
without connectors	DM	76,10

Plastic files (binders) for your VHF COMMUNICATIONS magazines in "VHF blue". Allow up to 12 copies to be filed neatly	DM	5,--

DJ 1 NB 004 — 2-metre transmitter

Trimmer set now complete with air-spaced trimmers		
PC-board only	DM	10,--
Set of trimmers and chokes	DM	12,60
Set of semiconductors	DM	60,20
Kit: PC-board, trimmers, chokes, semiconductors and special drill	DM	82,80
Crystal filter XF-9A with both sideband crystals	DM	106,--
Crystal filter XF-9B with both sideband crystals	DM	137,--
Crystal filters XF-9C, XF-9D, XF-9E	DM	137,--
Crystal filter XF-9M with carrier crystal	DM	106,--
Crystal discriminator XD-09-03 (matching XF-9E)	DM	78,--
Crystal 38.6667 MHz for 2 metre converters	DM	13,70
Crystal 42.0000 MHz for 4 metre converters	DM	15,60
Crystal 45.4780 MHz (HC-18/U) for VXO DJ 9 ZR 002	DM	24,50
Crystals 46.0000 + 46.333 MHz (HC-18/U) for phase-locked oscillator DJ 7 ZV / DJ 9 ZR, per set	DM	49,--
Crystals 65.0000 + 65.500 MHz for 144/14 MHz converters (DL 6 HA 001 MOSFET converter), per set	DM	33,--
Crystals 72.... MHz for 2 metre transmitters (please state your required frequency. Delivery time 4-6 weeks)	DM	21,50
Crystal 84.5333 MHz for 23 cm converters (DL 3 WR 001)	DM	21,50
Crystal 96.0000 MHz for 70 cm converters (DL 9 GU 001)	DM	21,50
Standard-frequency crystal 5.000.000 MHz	DM	25.--
Standard-frequency crystal 1.000.000 MHz	DM	20,50
Standard-frequency crystal 100.0000 kHz	DM	28,--

24 MHz VFO for Transmitter DL 3 WR 003

Set of 4 potted cores	DM	16,40
Set of 4 trimmers	DM	10,30
Set of semiconductors	DM	35,95
PC-board (DL 3 WR 007) available from November 1970		
Kit: Potted cores, trimmers, semiconductors	DM	58,--

AF Preamplifier for Transmitter DL 3 WR 003

PC-board DJ 4 BG 001	DM	3,--
Set of transistors	DM	7,80
Kit	DM	10,80

High performance equipment from Braun

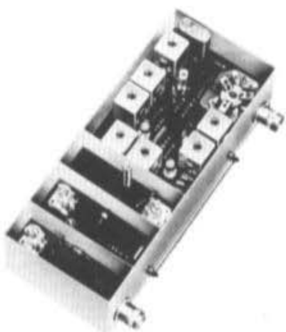
Two Metre Transceiver SE 600

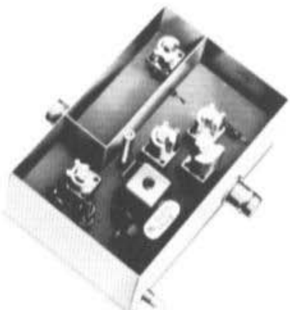
A selective two metre transceiver for all operating modes having a very low noise figure and extremely high cross-modulation rejection.
True transceiver operation or separate operation of transmitter and receiver are possible. Transmitter and receiver can be individually switched to the following modes: CW, LSB, USB, AM and FM. The separate operation and the possibility of selecting either LSB or USB make the transceiver suitable for operation with balloon carried translators or satellites.
Separate crystal filter for transmitter and receiver. True AM using plate/screen grid modulation of the PA tube. Bulit-in clipper. Crystal discriminator for FM demodulation, with IC limiting. Product detector for SSB. VOX and anti-trip. RF output and S meter. Built-in antenna relay. Power supply for 115 and 220 volt as well as a DC-DC converter for 12 volt are built in. **Price: DM 2675,—**

Two Metre Converter DGTC 22

High performance dual-gate MOSFET converter. Very high sensitivity and cross-modulation rejection. Highest possible spurious signal rejection by using a 116 MHz crystal.

Price: DM 122,—





70-cm-Converter DGTC 1702

High performance dual-gate MOSFET converter. An excellent 70-cm-converter. Variable overall gain — without effecting the other specifications — using a built-in 60 ohm T-Control so that the most optimal amplification matching can be made to the following receiver. Completely screened silver-plated brass cabinet. All 432 MHz circuits are a true stripline circuits with 10 µ silver plating. Input and output: 60 ohm BNC connectors. **Price: DM 228,—**

144 MHz/432 MHz Tripler LVV 270

A varactor tripler for input powers of up to 30 watt. For AM, FM and CW operation. High fundamental and harmonic rejection due to the built-in, selective bandpass filter at the output. Completely screened, silver-plated brass cabinet. All 432 MHz circuits are true stripline circuits with 10 µ silver plating.
Input and output: 60 ohm BNC connectors. **Price: DM 236,—**

144 MHz/432 MHz Transverter TTV 1270

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.
Input and output: 60 ohm BNC connectors. **Price: DM 142,—**

KARL BRAUN · 8500 Nürnberg · Bauvereinstraße 40 · Western Germany

Representatives:		
	France:	R. D. Electronique, 4, Rue Alexandre-Fourtanier, Toulouse
	Switzerland:	Equipel S. A. 1211 Genève 24
	United Kingdom:	M. E. L. 319 Ballards Lane, Finchley, London, N. 12
	USA and Canada:	Spectrum International, P. O. Box 87, Topsfield, MA 01983