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VHF COMMUNICATIONS

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144 MHz Transceiver for AM, FM and SSB

VHF COMMUNICATIONS

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A PUBLICATION FOR THE RADIO AMATEUR
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With this edition we commence a new volume of VHF COMMUNICATIONS. We thank you for your continuing interest in our magazine in spite of the various revaluations or devaluations of most currencies. We intend to publish some very interesting articles in 1972. However, we feel that a large number of our readers have designed some very good equipment which would be of interest to other readers. We would be very pleased to consider your designs for publication in VHF COMMUNICATIONS for readers in all parts of the world.

We know that there are a large number of VHF and UHF amateurs that do not subscribe to VHF COMMUNICATIONS. We would therefore be very grateful for any introductions you can make on our behalf. We are willing to provide any interested amateurs with a free sample copy of VHF COMMUNICATIONS. We wish all readers a very prosperous 1972 and lots of interesting reading in VHF COMMUNICATIONS.

73's

Hans DJ 3 QC	Robert DL 3 WR	Terry G 3 J VQ/DJ ∅ BQ

PORTABLE SSB TRANSCEIVER FOR 144 - 146 MHz WITH FM-ATTACHMENT

Part 1: Circuit description and specification

by G. Otto, DC 6 HL

INTRODUCTION

A two-metre SSB transceiver is to be described whose compact dimensions make it suitable for portable and mobile operation. However, the good large-signal behaviour, the special AGC-circuit of the receiver as well as the excellent stability and purity of the VFO signal make the transceiver also a very efficient home station. The output power of approximately 8 W PEP can be increased to any required level using a subsequent linear amplifier. However, 8 W PEP represents a good compromise between output power and current requirements for portable operation. The transceiver is operated from a DC-voltage source of 11.5 V to 14 V, which can be taken from three flat batteries (4.5 V each) connected in series, or from a automobile accumulator.

The title photograph of this magazine shows the author's transceiver, whose overall dimensions are 255 mm by 75 mm by 210 mm. It consists of one main module board containing the main circuits of the transceiver, as well as the five auxiliary modules: Carrier oscillator, VFO, 137 MHz local oscillator signal chain, AF-amplifier and reflectometer. The operating controls, loudspeaker, meter and a 12 V/2.6 Ah accumulator are also built in to the small cabinet. The weight of the complete transceiver including accumulator amounts to 4.2 kg. An extension of the transceiver for FM will be brought in part 3 of this description and will include a separate IF-amplifier and discriminator for this mode.

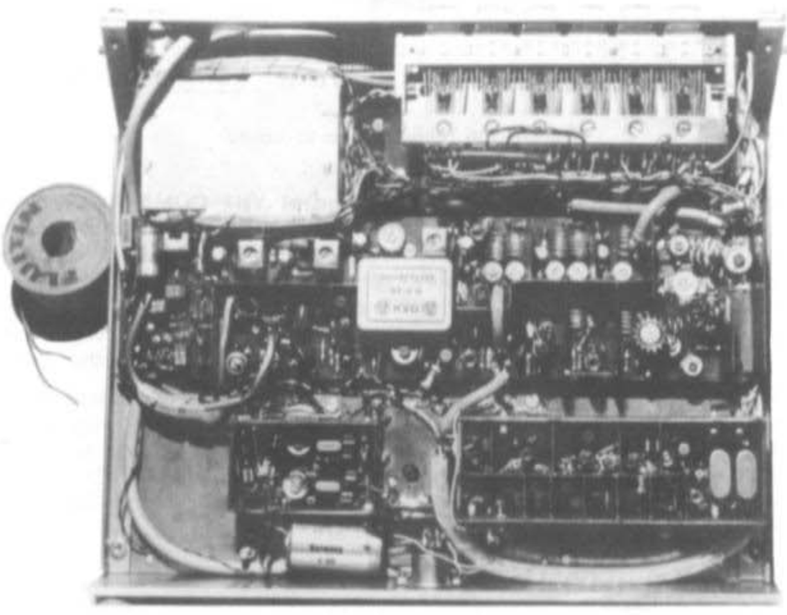


Fig. 1: Internal view of the DC 6 HL SSB-Transceiver

1. CHARACTERISTICS AND SPECIFICATIONS

The frequency processing, individual stages, as well as the characteristics and specifications are to be discussed with the aid of the block diagram given in Figure 2. The dimensioning of the circuitry is, in certain points, similar to the transceivers of K. P. Timmann (1) and G. Laufs (2). In contrast to (2), however, a single-conversion superhet principle is used with an IF of 9 MHz and a local oscillator frequency of 135 - 137 MHz. This injection frequency is not obtained from a phase-locked oscillator as in (1) but from a frequency synthesis circuit (6). The local oscillator frequency is obtained by mixing the VFO frequency of 5 to 6 MHz with a switchable crystal-controlled frequency of 130 MHz or 131 MHz. The elaborate mechanical construction of the VFO and highly effective filtering guarantee that the local-oscillator signal is just as good as that of a phase-locked oscillator.

Specifications of the local oscillator signal:

Output voltage (135 - 137 MHz):	0.7 - 1 V
Suppression of spurious signals:	$\geq$ 100 dB

Frequency stability: In the period from 5 seconds to 2 hours after switching on, the frequency variation amounted to 90 Hz. This value was not exceeded on varying the ambient temperature in the range of +10 °C to +30 °C.

Specifications of the receiver:

Input voltage for 10 dB signal-to-noise ratio:	0.15 μV
At an input impedance of 60 Ω and a bandwidth of 2.4 kHz, this value corresponds to a noise figure of:	$\leq$ 3 dB
Control range of the whole receiver:	$\geq$ 120 dB
Rise-time of the control voltage:	approx. 0.5 ms
Fall-time of the control voltage: with weak signals	approx. 1 s
with strong signals	approx. > 6 s

The receiver possesses automatic switching of the control time constants. As can be seen from the above values, the automatic circuit increases the fall-time in conjunction with strong signals. This improves the reception of weak signals and suppresses the unpleasant pumping effect in conjunction with strong signals that impairs the intelligibility considerably.

The transformer-less audio amplifier provides an output power of 1 W into a 5 Ω loudspeaker. Current requirements during reception with an AF-power of 50 mW (U_b = 12 V): 100 mA

Specifications of the transmitter:

Mean output power with single-tone modulation:	$\geq$1.5 W at U_b = 12 V $\geq$2.5 W at U_b = 14 V
Current requirements with single-tone modulation:	400 mA at U_b = 12 V 500 mA at U_b = 14 V
Mean current drain with voice modulation (U_b = 12 V):	250 mA
Current drain without modulation:	160 mA

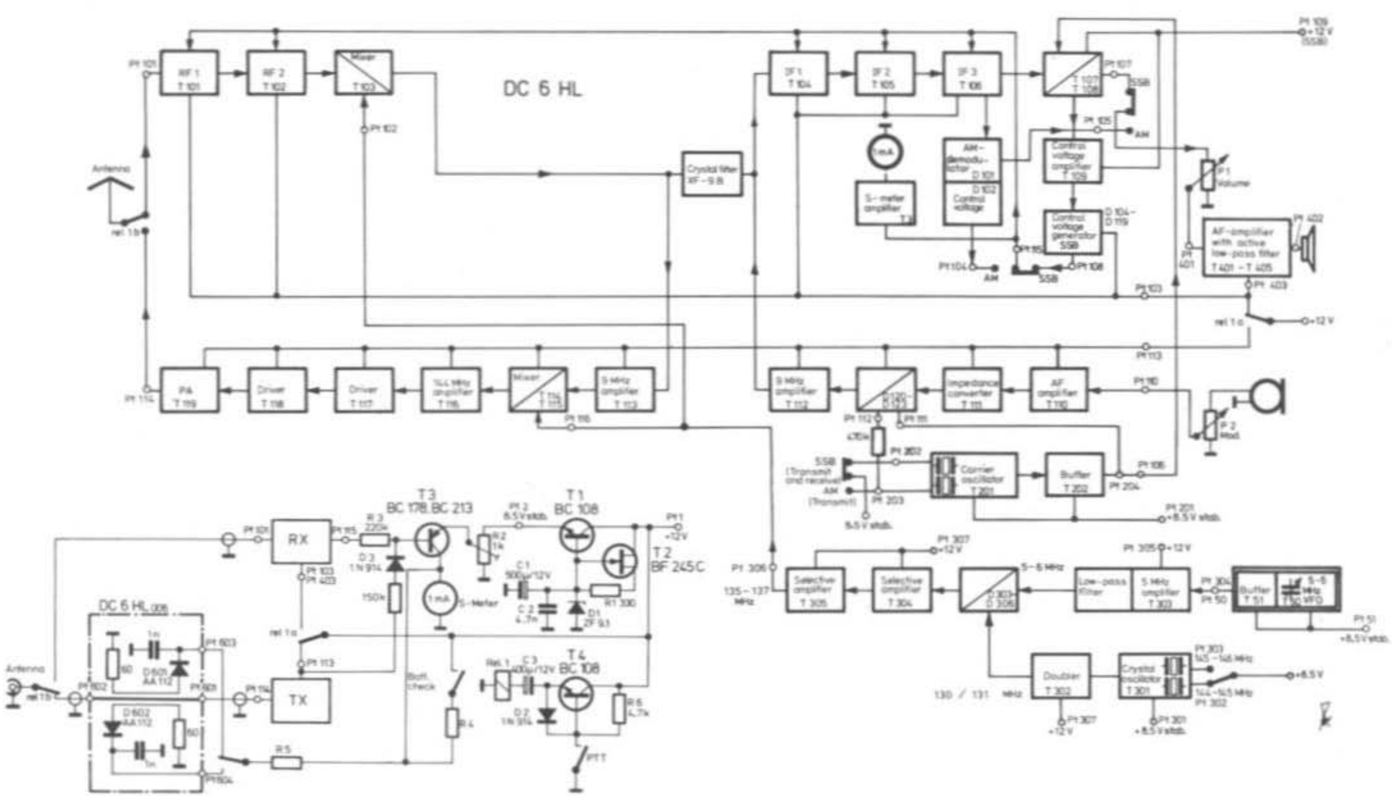


Fig. 2: Block diagram and connection details

The block diagram Figure 2 also shows the voltage stabilizer circuit which feeds the S-meter amplifier and oscillators with a stabilized voltage of 8.5 V. During transmission, the S-meter indicates the voltage from the reflectometer. This allows the RF-output and the antenna matching to be monitored. The S-meter amplifier (T 3) is blocked by diode D 3 during transmission.

In order to save battery power, a latching-type relay (Rel. 1) is used. A special circuit comprising transistor T 4 has been provided to drive the relay so that PTT-operation (push-to-talk) is possible. Of course, it is also possible for a reed-contact coaxial relay to be used as was described in (3).

2. CIRCUIT DETAILS

2.1. MAIN BOARD DC 6 HL 001

Figure 3 shows the elaborate circuit of the transmitter and receiver that are accommodated on the main board DC 6 HL 001. The most critical stages of the receiver are equipped with dual-gate MOSFET's.

2.1.1. VHF CIRCUIT

The 2m converter comprises two preamplifier stages (T 101 and T 102) that are coupled via a bandpass filter. Both stages are included in the AGC-circuit. The resonant circuits are damped by resistors R 103, R 105, R 107 and R 109 so that a sufficient VHF-bandwidth is obtained. A control range of 35 dB was obtained using a gate-protected MOSFET 40673 in the first RF-amplifier stage and a non-protected 40603 or 3 N 140 in the second stage.

An absorption circuit for the image frequency of 126 to 128 MHz is connected to gate 1 of the mixer transistor T 103. The local oscillator frequency of 135 to 137 MHz is fed to gate 2 of this transistor via the resonant circuit comprising inductance L 106.

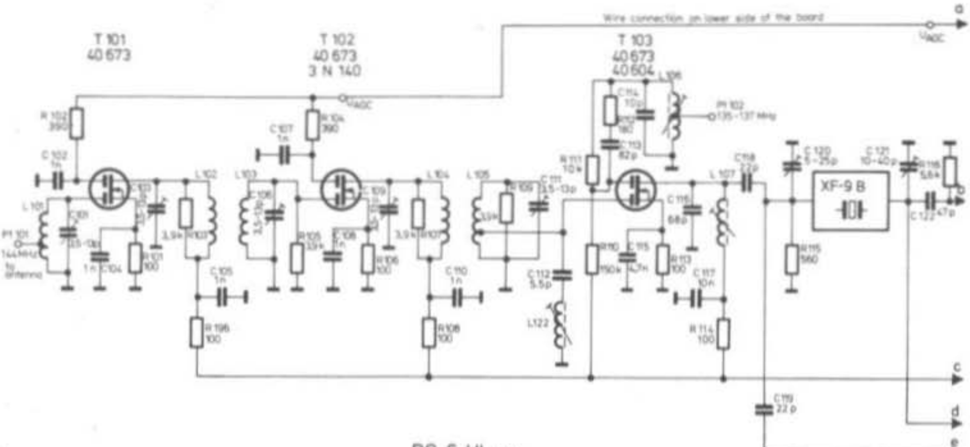
2.1.2. SINGLE SIDEBAND FILTER

A resonant circuit for the intermediate frequency (C 116/L 107) is connected to the drain of the mixer transistor so that the crystal filter can be loosely coupled simultaneously to the receive mixer and the 9 MHz SSB transmit amplifier (T 113). The crystal filter is terminated, as prescribed by the manufacturers, at input and output with 560 Ω and a trimmer capacitor which is aligned for minimum ripple in the passband range. The second 560 Ω resistor (R 167) is simultaneously the collector resistor for the double-sideband transmit amplifier with T 112.

2.1.3. IF AMPLIFIER

The crystal filter is followed, in the receiver, by a three-stage intermediate frequency amplifier. Three different circuits were tried:

1. A circuit with two integrated circuits CA 3028 A similar to the description given in (1).
2. Cascode circuits with BF 173/BC 109 as described in (4) and (2).
3. A circuit with three dual gate MOSFET's 40602.



DC 6 HL 001

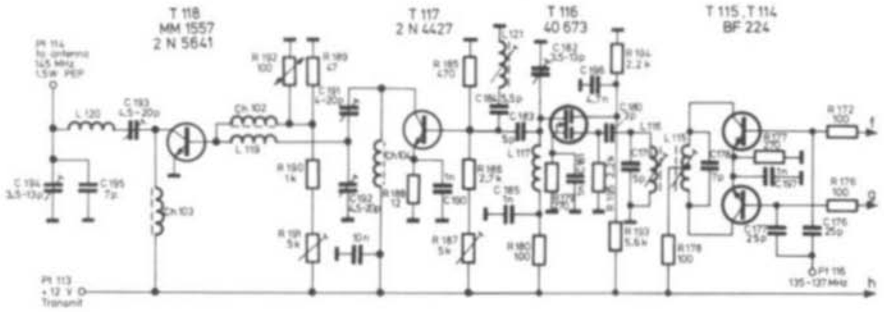
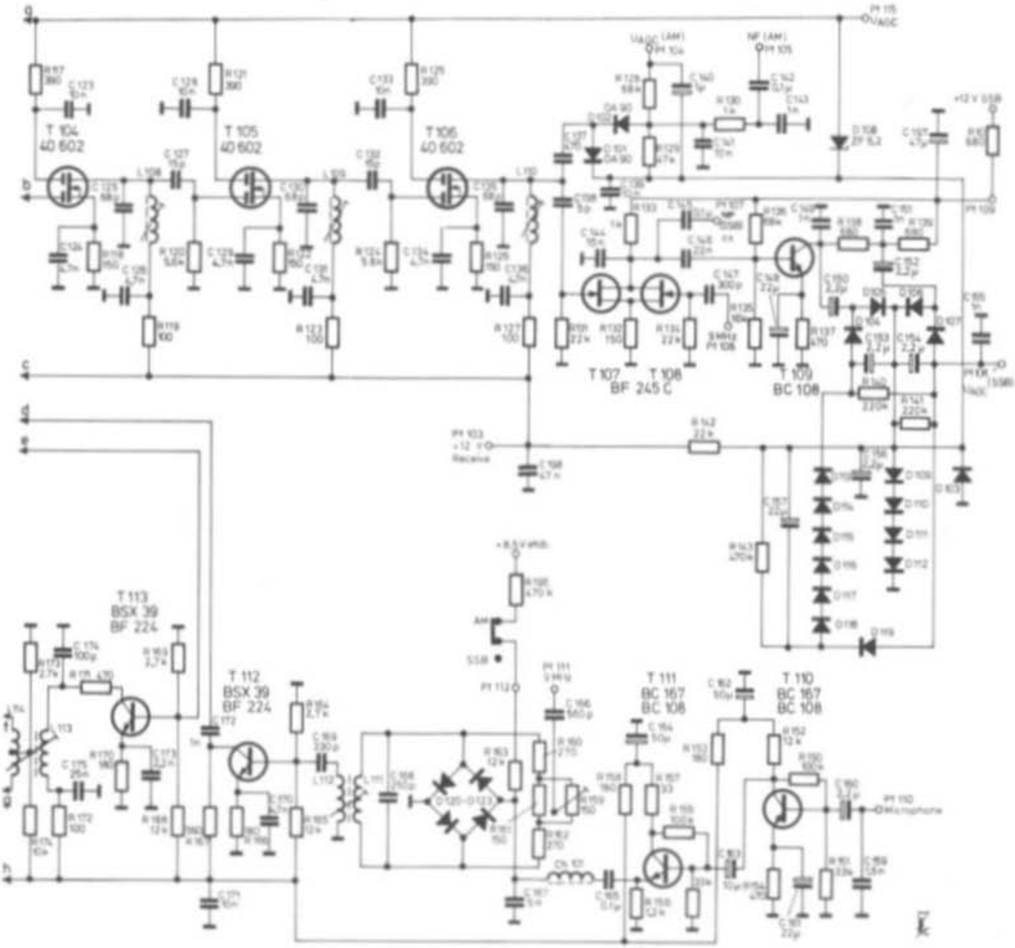


Fig. 3: Circuit diagram of the main board including the main modules of transmitter and receiver


T 104
40 602
T 105
40 602
T 106
40 602
T 107
T 108
BF 245 C
T 109
BC 108
T 113
BSX 39
BF 224
T 112
BSX 39
BF 224
T 111
BC 167
BC 108
T 110
BC 167
BC 108

Subsequent measurements showed that the first two circuits had virtually identical characteristics while the third circuit exhibited a lower gain. The control range was practically equal for all circuits at approximately 35 dB per stage.

The IF amplifier equipped with the dual gate MOSFET's exhibited better intermodulation characteristics. Considerable advantages are obtained due to the simple external circuitry and the high input impedance of the MOSFET's. This is also valid for the high-impedance control voltage source, which has been noticeably simplified using the given circuit. Since the integrated IF amplifier could not be constructed smaller and requires an equal number of miniature components, the IF amplifier equipped with the inexpensive MOSFETs has been selected. The output of each of the three stages is fed to a resonant circuit comprising a screened inductance and a 68 pF capacitor in parallel.

2.1.4. DEMODULATOR AND CONTROL CIRCUITS

The IF amplifier is followed by a separate demodulator for AM and SSB which generate a different control voltage for each mode. The appropriate control voltage is fed to the control line on switching the AF amplifier to the required demodulator. An attachment for frequency modulation is also under construction and will be described in a later edition of this magazine.

The AM demodulator comprises a voltage doubler (D 101 and D 102) whose base is maintained at a voltage of 2.2 V by the four series-connected, forward biased diodes D 109 to D 112.

The AM control voltage at connection point Pt 104 is therefore between + 2.2 V and approximately + 0.5 V. The demodulated AM signal is fed via an RF-filter to connection Pt 105.

A differential mixer comprising the two junction field effect transistors T 107 and T 108 is used as SSB demodulator. This circuit is also able to handle high input voltages and ensures a sufficient isolation between intermediate frequency and BFO. It is especially important that the two AC voltages have a ratio of about 20 : 1 to another. For this reason, the IF frequency is only injected via a 5 pF capacitor. Equally good characteristics are provided using a mixer equipped with a single, gate-protected dual-gate MOSFET, however, such transistors are still more expensive than two junction types.

The audio frequency signal is now passed to connection Pt 107 and to a single-stage control voltage amplifier comprising T 109, which in turn feeds a voltage divider so that two AF-voltages are available having a ratio of 2 : 1 to another. Both of these voltages are now rectified in a separate voltage doubler circuit (D 104/D 105 or D 106/D 107).

The lower voltage is connected directly to SSB control voltage connection Pt 108.

The higher voltage charges capacitor C 157 via the series-connected diodes D 113 to D 118. These diodes ensure that the capacitor is only charged when the signal strength (and thus the AF-voltage) exceeds a certain threshold value. If this is the case, capacitor C 157 will be discharged onto the control line via diode D 119. Since the discharge time constant of the basic control voltage has been dimensioned shorter than that of the control voltage generated at high signal strengths, this circuit provides a simple means of increasing the fall time constant of the control voltage with strong signals.

The fall time constants are mainly determined by the capacitance of the appropriate filter capacitor (2.2 μF or 22 μF), the load resistor (220 kΩ or 470 kΩ) and the resistance of the control and protective circuit. The rise time constant, on the other hand, is dependent on the AF source impedance (T 109, R 138 and R 139), coupling capacitor C 152 and filter capacitor C 154.

The transition point between the fall time constants is dependent on the value of the audio voltage and thus on the overall gain of the receiver. It can be altered to match a certain gain by increasing or decreasing the diode chain. The longer fall time should be effective with signals that are about 40 dB or more above the noise level. A basic control voltage potential of 2.2 V is also valid for the SSB control voltage under uncontrolled conditions.

In order to protect the non-gate-protected MOSFETs, diodes D 103 and D 108 have been provided in the control voltage circuit. They ensure that the control voltage does not exceed the limit values of + 20 V and - 8 V. The MOSFETs would be endangered in a high impedance control circuit even if the control line was only touched with ones finger. Positive voltages in excess of 3 V are shorted to ground via diode D 103 and the diode chain D 109 to D 112. Negative voltages are shorted to ground after the zener voltage of D 108 is exceeded. The control characteristics are not affected by the protective circuit.

2.1.5. TRANSMITTER CIRCUIT

The AF-amplifier and balanced modulator are, with the exception of a few non-important modifications, dimensioned as given in (2). Silicon planar diodes (D 120 to D 123) are used in the balanced modulator. It is true that such diodes require a higher local oscillator amplitude than germanium types, however, they keep the carrier suppression more stable during ambient temperature fluctuations. An especially high carrier suppression can be obtained if diodes having the same forward resistance are selected.

Amplitude modulation (A 3) is obtained by unbalancing the balanced modulator so that the carrier is not suppressed. This is achieved by connecting the stabilized operating voltage via resistor R 106 to the modulator so that the two upper diodes (in Fig. 3) conduct.

The balanced modulator is followed by a low-gain amplifier (T 112) for the double-sideband signal, whose load resistor R 167 is also the terminating resistor for the sideband filter. After passing the sideband filter the resulting SSB-signal is amplified in transistor T 113 and fed to the push-pull mixer stage comprising T 114 and T 115. The same local oscillator signal is used to convert the transmit signal to the 2 metre band as was used in the receiver.

The 2 m SSB-signal is now fed to a three-stage linear amplifier equipped with a dual-gate MOSFET in the first stage and includes an absorption circuit for the local oscillator frequency of 135 MHz to 137 MHz. Transistors must be used in the driver and output stages that can be driven linearly to high current values at the relatively low operating voltage of 12 V. The inexpensive overlay type 2 N 4427 has been found suitable for the driver, and the 2 N 5641 (MM 1557) for the output stage. The large linear drive range of the 2 N 5641 and its insensitivity to incorrect tuning and mismatch are the result of the integrated emitter ballast resistors at the individual emitters of the transistor system. The aligned operating point of the output transistor is stabilized by a thermistor (NTC-resistor), which is mounted on the heat sink. The output coupling is in the form a resonant transformation link.

2.2. SUB-MODULES DC 6 HL 002 to 006

The following auxiliary modules are required in addition to the main board DC 6 HL 001: Switchable carrier oscillator (002), VFO (no PC-board number), local frequency chain (003), audio frequency module with discrete components (004) or with integrated construction (005), and reflectometer (006).

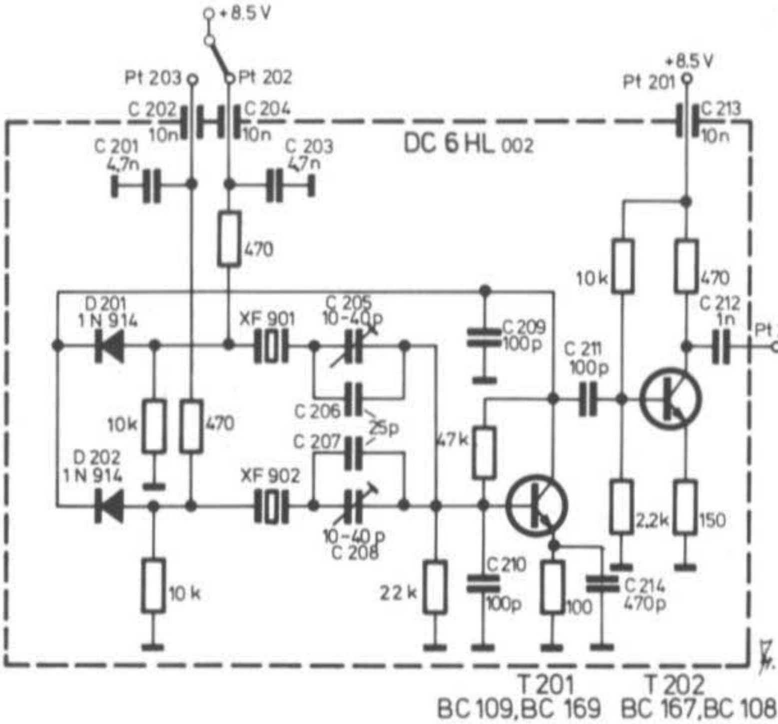


Fig. 4: Circuit diagram of the carrier oscillator

2.2.1. CARRIER OSCILLATOR CIRCUIT

Figure 4 shows the simple carrier oscillator circuit and buffer stage. It is very similar to the circuit of (2) except that the DC sideband switching with diodes is a standard feature. The connection of the 10 kΩ resistor to each diode ensures that the appropriate diode is completely blocked and that a single-pole switch can be used instead of the two-pole switch (2). The buffer amplifier is untuned.

2.2.2. VFO CIRCUIT

The actual variable frequency oscillator is the only module not built up on a printed circuit board, but on ceramic supports mounted on a 5 mm thick aluminium plate. Figure 5 shows the circuit of the VFO, which is variable between 5 and 6 MHz and possesses an untuned buffer (T 51). Since the output voltage is taken from the collector, low-reactive transistors are required.

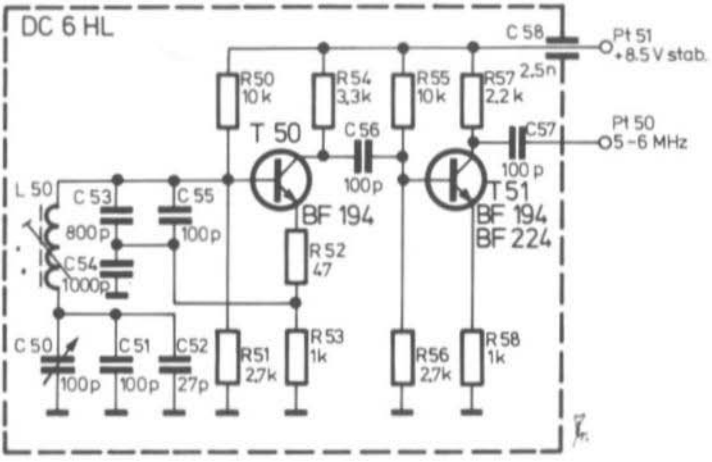


Fig. 5: Circuit diagram of the 5 – 6 MHz VFO

The author would like to advise against using a different transistor type than the given BF 194. Also the other frequency-determining components should closely adhere to the details given, since the temperature compensation has been carefully carried out for the given types and manufacturers. A very good, single-bearing variable capacitor of 100 pF (Hopt type 220), styroflex capacitors and a home-made coil on a trolitul coilformer are used. If the construction details are followed closely it will not be necessary to use an expensive ceramic coilformer with burnt-in silver windings and mica capacitors. By the way, the oscillator was designed and constructed whilst continuously checking the circuit for the lowest harmonic content indication on a 150 MHz oscilloscope.

2.2.3. LOCAL OSCILLATOR CHAIN

In order to convert the variable frequency of 5 to 6 MHz to the required local frequency range of 135 to 137 MHz, a further mixer with crystal oscillator and a tuned amplifier are required. The circuit of this module is given in Figure 6; it is accommodated on printed circuit board DC 6 HL 003.

The output signal of the VFO is passed via the low-pass filter comprising inductances L 301 to L 306 to suppress any residual harmonics. An amplifier with strong feedback (T 303) ensures a defined source impedance for the low-pass filter. The filter has been calculated for an impedance of 560 Ω. It comprises a T-section, a complete M-derived section and a half M-derived section at input and output. The calculation was based on a cutoff frequency of 7 MHz with an attenuation pole at 11 MHz (1st harmonic of the centre frequency of the VFO). After calculation, the inductance and capacitance values were combined to form the circuit shown in Figure 6. The calculated values and the practical winding data will be given in the constructional description.

In order to obtain a local frequency range of 135 MHz to 137 MHz from a frequency variation range of 1 MHz, it is necessary for two crystal frequencies to be used for frequency conversion that are spaced 1 MHz from another.

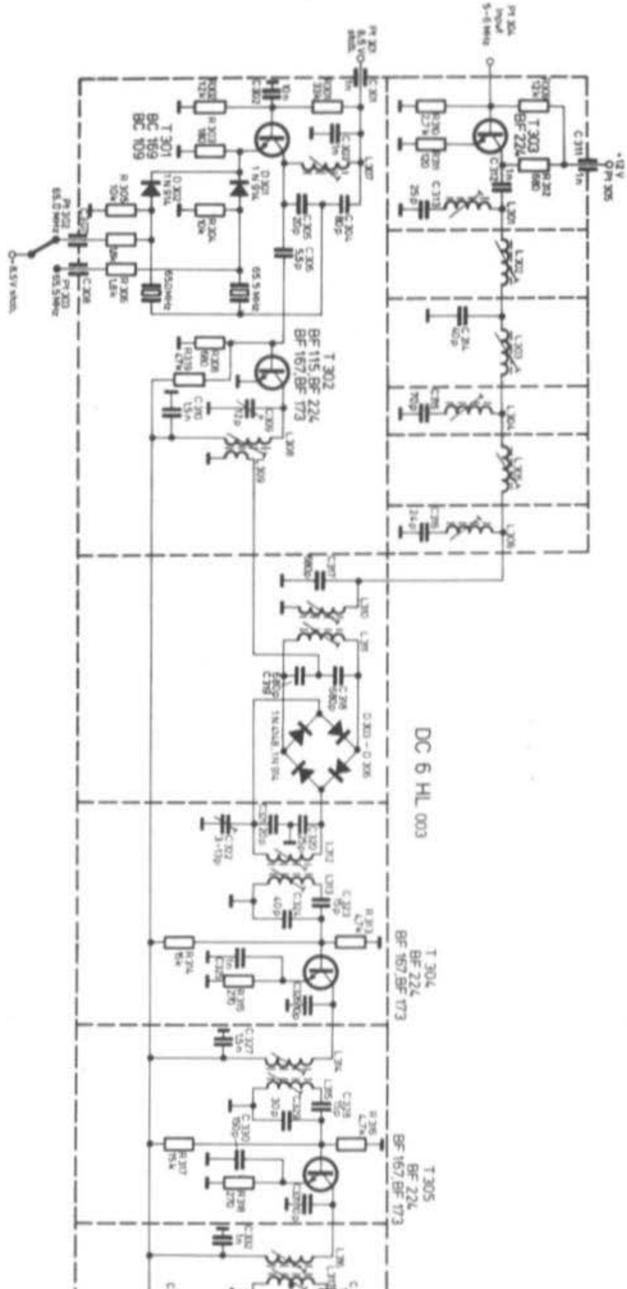


Fig. 6: Circuit diagram of the local oscillator chain

The crystal oscillator (T 301) is therefore equipped with two crystals, and the required crystal is selected with the aid of diodes. Since the crystal oscillator operates at half the required frequency, the crystal frequencies are spaced 0.5 MHz from another at 65.0 MHz and 65.5 MHz. This is followed by a frequency doubler (T 302) so that fixed crystal-controlled frequencies of 130 MHz and 131.0 MHz are obtained. Transistor T 302 operates virtually in class B, which means that it is able to work with a relatively low drive voltage.

The variable frequency (5 - 6 MHz) and fixed frequency (130 or 131 MHz) are added in the ring mixer (D 303 to D 306) to form the required local oscillator frequency of 135 to 137 MHz. Theoretically, the two original frequencies and harmonics thereof should not be present at the output; this is, unfortunately, not obtainable in practice, but the better the balance of the mixer, the greater will be the suppression of these frequencies. Trimmer capacitor C 322 is provided for compensation of any capacitive unbalance and is aligned for maximum suppression of the first harmonic of the crystal frequency.

The mixer is followed by a two-stage amplifier which is intercoupled with band-pass filters, and amplifies the local oscillator signal to the required value of approximately 1 V. The harmonic suppression (purity) of this signal is more than 100 dB down on the required frequency. The exact value could not be determined using the available measuring equipment. However, the value of 100 dB is extraordinarily good and the module is to be recommended even when the output of the transceiver is to be amplified to the highest power levels. However, the effort is well worthwhile even with low-power stations and will show itself as a freedom of spurious signals during reception and a good, clean signal whilst transmitting.

2.2.4. AUDIO AMPLIFIER WITH DISCRETE COMPONENTS

Figure 7 shows the circuit of a transformerless audio amplifier with a complementary transistor pair (T 404 and T 405) in the output stage. Such circuits are well-known. However, this circuit is provided with an active low-pass filter with a cut-off frequency of 3 kHz, which suppresses higher frequency noise components from the last mixer, audio preamplifier or FM-demodulator. This audio module is designated DC 6 HL 004.

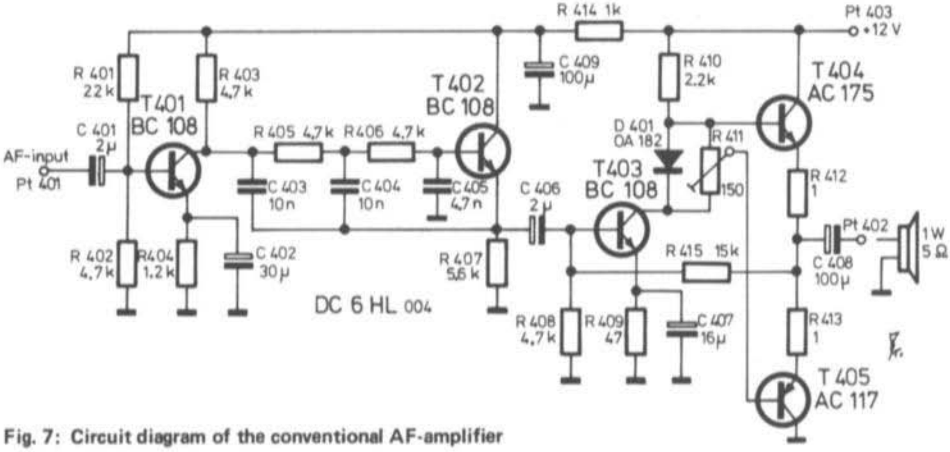


Fig. 7: Circuit diagram of the conventional AF-amplifier

2.2.5. AUDIO AMPLIFIER WITH INTEGRATED CIRCUIT

Figure 8 shows the circuit of an integrated audio amplifier using the TAA 611 A. This module is also provided with a low-pass filter. It provides an output power of 1 W into a load of 5 Ω at 12 V. Of course, this module, which is designated DC 6 HL 005, can be used instead of the previous AF-module DC 6 HL 004. Although it is somewhat more expensive, it is far more compact.

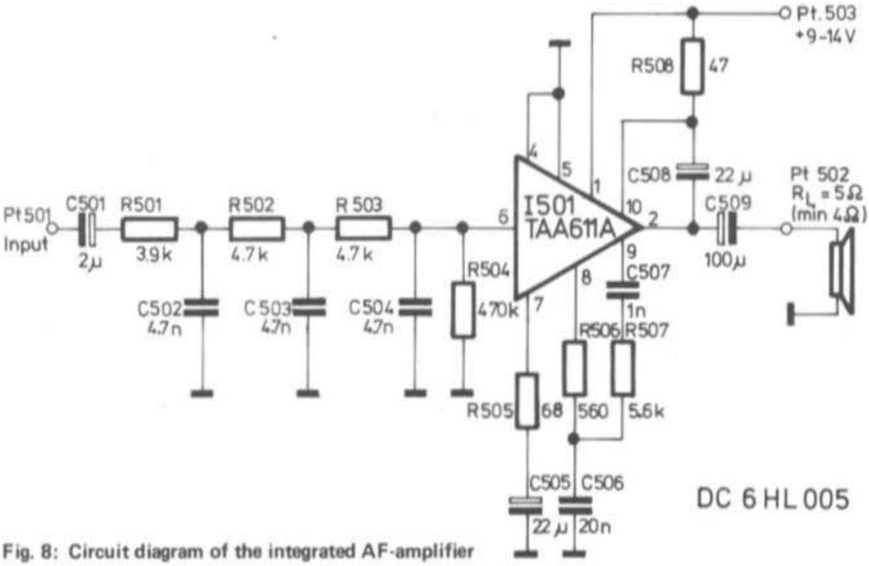


Fig. 8: Circuit diagram of the integrated AF-amplifier

2.2.6. REFLECTOMETER

The transceiver is provided with a built-in reflectometer (DC 6 HL 006). The simple circuit is shown in Figure 2. It is, in principle, very similar to the stripline reflectometers described in (5). The difference here is that short, straight striplines are used. The lines are dimensioned for an impedance of 60 Ω, and based on a dielectric constant of $\epsilon_r = 5$ and a board thickness (D) of 1.5 mm. The following formula was used to calculate the width (b) of the stripline, which can be modified for other impedances by substituting the required impedance instead of the 60 Ω used:

$$Z = \frac{60}{\sqrt{\epsilon_r}} \times \log_{10} \frac{7\,D}{b} \quad \text{or} \quad b = 7\,D \times e^{-\frac{Z \times \sqrt{\epsilon_r}}{60}}$$

The stripline width b corresponds to 1.1 mm for an impedance of 60 Ω. Since the auxiliary arms are not connected together they will each require a terminating resistor of 60 Ω.

The full constructional details will be brought in a later edition of this magazine. At the moment, a number of prototypes are being constructed by different amateurs to ensure that the circuits are as foolproof as possible before the constructional details are published, and kits of components are offered. This should be possible in the next edition of this magazine.

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CALCULATIONS FOR A LINEAR VFO

by H. Schotten, DJ 1 FO

In this article the lowest possible frequency of a variable frequency oscillator is calculated by which the maximum linearity error of a linear scale with a variable capacitor having linear capacitance characteristics does not exceed a given value. Vice versa, it is also possible to calculate the linearity error under given conditions.

In principle, the oscillator circuit to be used is of no importance here. For simplification, however, it is assumed that the variable capacitor is connected directly in parallel with the inductance of the resonant circuit. Figure 1 shows a Colpitts oscillator as an example.

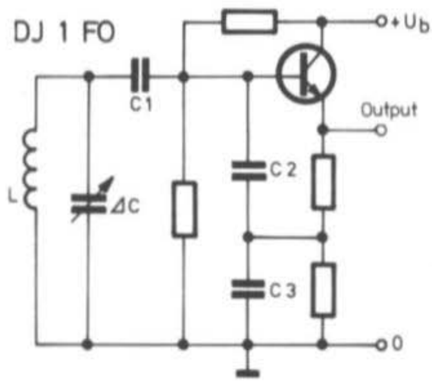


Fig. 1: Tuning an oscillator with a variable capacitor without series capacitor

The frequency range $f_1 - f_o$ that is to be covered by the VFO is designated with Δf. The linearity error, which is the difference between the actual frequency and the frequency indicated on the linear scale, is designated with δ.

The calculation is based on the resonant circuit formula:

$$f = \frac{1}{2\pi \times \sqrt{L \times C}} \qquad (1)$$

The following formulas allow the lower corner frequency f_o and the upper corner frequency f_1 to be calculated:

$$f_o = \frac{1}{2\pi \times \sqrt{L \times (C_o + \Delta C)}} \qquad (2)$$

$$f_1 = f_o + \Delta f = \frac{1}{2\pi \times \sqrt{L \times C_o}} \qquad (3)$$

C_o is the effective capacitance of the resonant circuit with the variable capacitor at minimum capacitance; this is mainly the series circuit of capacitor C 1, C 2, and C 3 and the minimum capacitance of the variable capacitor.

Formula 2 will be valid under two assumptions:

1. The tuned and indicated frequencies coincide at the band limits f_1 and f_o.
2. The greatest linearity error appears exactly at the centre frequency between f_o and f_1, e.g. at $f_o + \frac{\Delta f}{2}$

The first assumption can be fulfilled by alignment of the inductance and capacitance. The second condition is not fulfilled exactly, but represents a sufficient approximation for amateur applications.

Figure 2 shows the characteristics of the actual tuning curve (actual curve) and that of the linear scale (nominal curve), as laid down under the given assumptions.

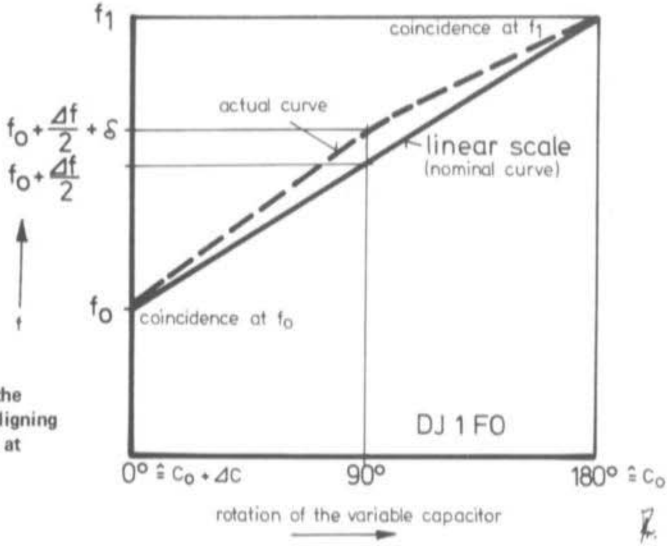


Fig. 2: Comparison between the actual and nominal curve on aligning the frequency for coincidence at both band limits

Formula 3 is now divided by formula 2:

$$\frac{f_o + \Delta f}{f_o} = \sqrt{\frac{C_o + \Delta C}{C_o}} \quad \text{or} \quad 1 + \frac{\Delta f}{f_o} = \sqrt{1 + \frac{\Delta C}{C_o}}$$

thus: $$\frac{\Delta C}{C_o} = \frac{2 \times \Delta f}{f_o} + \frac{\Delta f^2}{f_o} \qquad (4)$$

Two unknown quantities are given in equation 4: the frequency f_o and the relationship $\frac{\Delta C}{C_o}$. In order to determine these two values, it is necessary for a second equation to be considered by forming the difference between the frequency adjusted by half the capacitance variation $\frac{\Delta C}{2}$ and the frequency indicated in the centre of the linear scale. At half the capacitance of the variable capacitor (capacitance variation $\frac{\Delta C}{2}$), the resulting frequency will be:

$$f_{\frac{1}{2}} = f_o \times \sqrt{1 + \frac{\Delta C}{C_o} \times \frac{1}{2}} \qquad (5)$$

The linear scale will indicate the value:

$$f\frac{1'}{2} = f_o + \frac{\Delta f}{2} \qquad (6)$$

The linearity error δ therefore results as the difference of equations 5 and 6:

$$\delta = f_o \times \sqrt{1 + \frac{\Delta C}{C_o} \times \frac{1}{2}} - \left(f_o + \frac{\Delta f}{2} \right) \qquad (7)$$

The expression $\frac{\Delta C}{C_o}$ is now isolated from equation 7:

$$\sqrt{1 + \frac{\Delta C}{C_o} \times \frac{1}{2}} = 1 + \frac{\Delta f}{f_o} \times \frac{1}{2} + \frac{\delta}{f_o} = 1 + \frac{1}{f_o} \left(\frac{\Delta f}{2} + \delta \right)$$

$$\frac{\Delta C}{C_o} = \frac{4}{f_o} \times \left(\frac{\Delta f}{2} + \delta \right) + \frac{2}{f_o{}^2} \times \left(\frac{\Delta f^2}{4} + \Delta f \times \delta + \delta^2 \right) \qquad (8)$$

Since equations 4 and 8 are both resolved according to $\frac{\Delta C}{C_o}$, they can be equated. The resulting formula 9 only contains the unknown frequency f_o:

$$\frac{2 \times \Delta f}{f_0} + \frac{\Delta f^2}{f_0{}^2} = \frac{4}{f_0}\left(\frac{\Delta f}{2} + \delta\right) + \frac{2}{f_0{}^2}\left(\frac{\Delta f^2}{4} + \Delta f \times \delta + \delta^2\right)$$

$$\cancel{\frac{2 \times \Delta f}{f_0}} + \frac{\Delta f^2}{f_0{}^2} = \cancel{\frac{2 \times \Delta f}{f_0}} + \frac{4\delta}{f_0} + \frac{\Delta f^2}{f_0{}^2} \times \frac{1}{2} + \frac{2 \times \Delta f \times \delta}{f_0{}^2} + \frac{2\delta^2}{f_0{}^2}$$

$$\Delta f^2 = 4\delta \times f_0 + \Delta f^2 \times \frac{1}{2} + 2 \times \Delta f \times \delta + 2\delta^2$$

$$\boxed{f_0 = \frac{\Delta f^2}{8\delta} - \frac{\Delta f}{2} - \frac{\delta}{2}} \qquad (9)$$

Example 1: It is assumed that 0.5 MHz are required as frequency variation Δf; 1 kHz is permissible as linear error δ:

The lowest frequency f_o by which these values can be obtained is:

$$f_o = \frac{(0.5 \times 10^6)^2}{8 \times 10^3} - \frac{0.5 \times 10^6}{2} - \frac{10^3}{3} \approx 31 \times 10^6 \text{ Hz} \approx 31 \text{ MHz};$$

After explaining the computation process, the greatest linearity error is to be reduced by altering the characteristic of the tuning curve. This is obtained by transposing the second point of coincidence, which was aligned with the aid of a trimmer capacitor to the upper frequency limit f_1, to the frequency $f_o + 0.7 \times \Delta f$. The resulting characteristic tuning and scale curves are given in Figure 3.

The value $0.7 \times \Delta f$ must now replace Δf in the appropriate formula equation 9. The values assumed in example 1 now give:

Example 2: $\Delta f = 0.5$ MHz $\quad 0.7 \times \Delta f = 0.35$ MHz $\quad \delta = 1$ kHz

$$f_o = \frac{(0.35 \times 10^6)^2}{8 \times 10^3} - \frac{0.35 \times 10^6}{2} - \frac{10^3}{2} \approx 15.2 \text{ MHz}$$

One can see the improvement that is offered by the tuning characteristic given in Figure 3. If the frequency $f_o = 31$ MHz given in example 1 is maintained, a linearity error $\delta \approx 0.5$ kHz would result with the new characteristic.

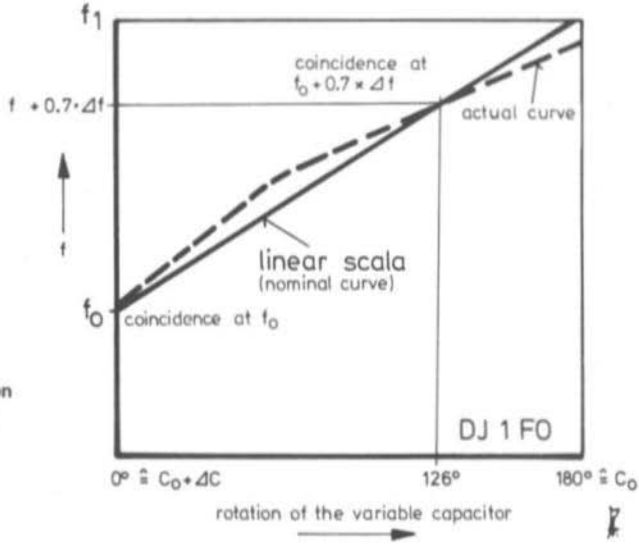


Fig. 3: Difference between the actual and nominal curve on aligning for coincidence at f_0 and $f_0 + 0.7 \times \Delta f$

Finally, an example in conjunction with Figure 3 with an extended tuning range:

Example 3: $\Delta f = 1$ MHz $\quad \delta = 1$ kHz

$$f_0 = \frac{(0.7 \times 10^6)^2}{8 \times 10^3} - \frac{0.7 \times 10^6}{2} - \frac{10^3}{3} \approx 61 \text{ MHz}$$

One can see in this example that the difficulty of achieving a linear VFO characteristic increases on increasing the tuning range.

The derived equation 9 and the calculated frequencies are only valid in conjunction with a linear capacitance variation. If the capacitance characteristics of the variable capacitor are distorted in the correct manner by bending the capacitor vanes, it is also possible for low linearity errors to be obtained at relatively low frequencies. Of course, suitable measuring equipment is required for this.

A DIGITAL CALIBRATION-SPECTRUM GENERATOR

Part 2: 1.001 MHz Accessory and Power Supply

by D. E. Schmitzer, DJ 4 BG

The first part of a modular standard frequency system, the digital calibration-spectrum generator DJ 4 BG 004, was described in Edition 4/71 of VHF-COMMUNICATIONS (1). The complete system comprises:

1. A calibration-spectrum generator that can be switched to 1 MHz, 500 kHz, 100 kHz, 50 kHz, 10 kHz, 5 kHz and 1 kHz, and whose 1 MHz crystal can be pulled with the aid of a varactor diode (1).
2. An additional calibration-spectrum generator (DJ 4 BG 005) equipped with a 1.001 MHz crystal which allows the order of the harmonic in question to be determined. This module also contains the power supply of the system. This unit is to be described here.
3. A fixed-frequency receiver for the 200 kHz longwave transmitter Droitwich with a synchronizing circuit (DJ 4 BG 010) for the calibration-spectrum generator. The frequency accuracy and stability of the system is that of the Droitwich signal at the receiving site. This is especially favourable for UHF/SHF amateurs as well as for EME, MS or satellite experiments. The Droitwich receiver and synchronizing unit can also be used to synchronize any 1 MHz crystal oscillator incorporating a varactor diode for pulling the frequency.

1. CIRCUIT DETAILS OF THE 1.001 MHz OSCILLATOR

Figure 1 gives the circuit diagram of the 1.001 MHz oscillator and subsequent pulse shaper for harmonic generation. The crystal oscillator is the same as that of the main calibration spectrum generator.

A low-pass filter section (R 12/C 7) has been added to the feedback link since it has been found that some crystals oscillate at an overtone. If required, this RC-link should also be included in the feedback link of the main calibration-spectrum generator. It should be pointed out that none of the supplied crystals manufactured by KVG have exhibited any such effects.

The crystals, which are manufactured for parallel resonance with a load capacitance of 30 pF, require a series capacitance of approximately 35 pF in the series resonance circuit used here. A part of this capacitance is formed by the varactor diode D 3 so that it is possible for a fine alignment of the frequency to be made with the aid of a potentiometer, without opening the unit (connection Pt 1). The fine alignment should be made exactly at a higher harmonic since the signal must be very accurately aligned for a spacing of 1 kHz from the 1 MHz frequency of the main calibration-spectrum generator. For instance, an error of only 6.9 Hz would cause the 144th harmonic to be indicated instead of the 145th. The varactor diode allows the frequency to be pulled by ± 10 Hz. The correct frequency spacing can be aligned at a known harmonic (e.g. 144 MHz). If varactor diode D 3 is used, the parallel capacitance C 3 can be deleted. If D 3 is not used, C 3 should have a capacitance of approximately 15 pF.

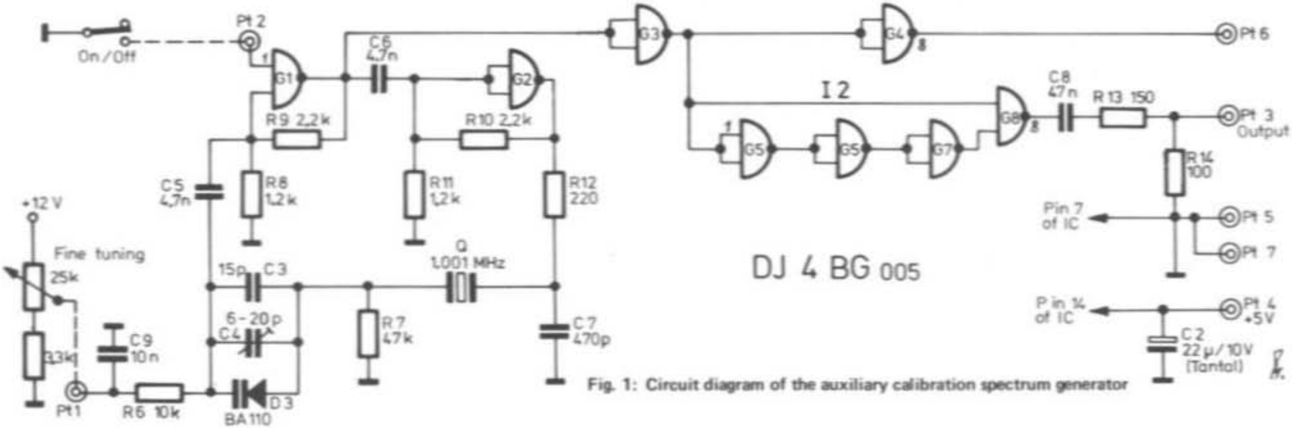


Fig. 1: Circuit diagram of the auxiliary calibration spectrum generator

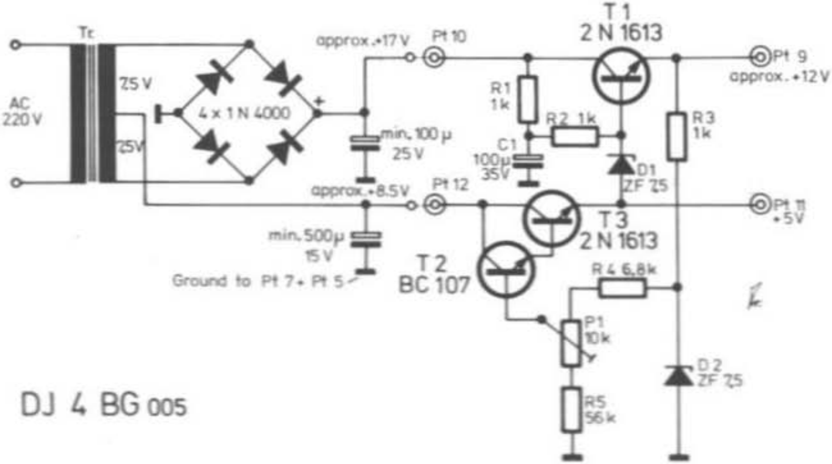


Fig. 3: Stabilizer circuit

The crystal oscillator and pulse shaper are built up using eight gates of two integrated circuits. The construction is extremely simple. If this attachment (DJ 4 BG 005) is only to be used in conjunction with the calibration-spectrum generator DJ 4 BG 004, the pulse-shaper circuit on PC-board DJ 4 BG 005 is not necessary. The oscillator signal at connection Pt 6 of DJ 4 BG 005 could be connected to Pt 4 of DJ 4 BG 004 and the pulse-shaper circuit of the main calibration-spectrum generator also used for forming the harmonics of the 1.001 MHz signal.

Of course, module DJ 4 BG 005 can also be operated on its own as a simple calibration-spectrum generator if a 1.000 MHz crystal is used. The pulse-shaper (G 5 - G 8) must, naturally, be used in this case.

2. FREQUENCY MEASUREMENT USING THE CALIBRATION SPECTRUM GENERATOR AND 1.001 MHz ATTACHMENT

The harmonics of the auxiliary 1.001 MHz oscillator are spaced further from the harmonics of the 1 MHz calibration-spectrum generator the higher the frequency multiplication factor is. This means, for instance, that the 145th harmonic of the 1.001 MHz oscillator will be 145 kHz higher than the 145th harmonic of the 1 MHz calibration-spectrum generator at 145 MHz.

If, for example, a receiver is tuned to an unknown frequency range where the harmonic of the 1.001 MHz oscillator is heard 39 kHz above the harmonic of the calibration spectrum generator, this will indicate that a frequency of 39 MHz has been tuned.

The method of determining unknown frequencies is now to be explained with the aid of an example. Figure 2 shows the frequency scale.

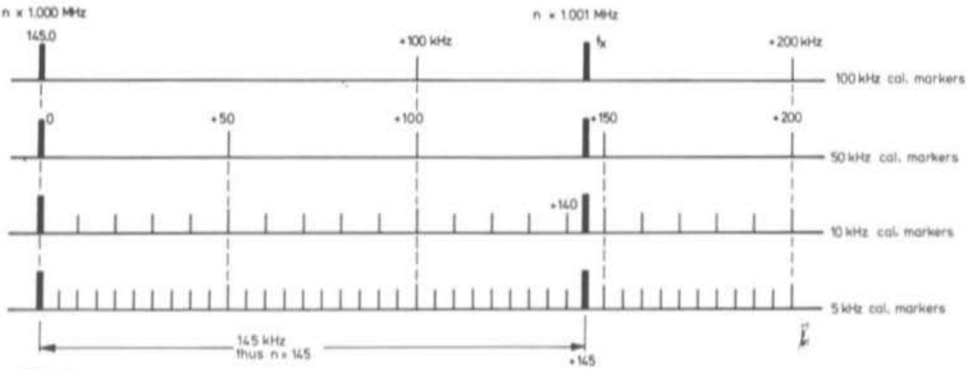


Fig. 2:

The 1 MHz harmonic of the calibration-spectrum generator is firstly marked, in our example 145 MHz. After this, the 1.001 MHz calibration-spectrum is switched on and the receiver tuned somewhat higher until this signal is heard. This position should also be marked on the scale. The 1 MHz calibration-spectrum generator is switched on again and switched to its 100 kHz spectrum. In our example it can be seen that the harmonic (f_x) of the 1.001 MHz generator is approximately in the centre between the two 100 kHz signals appearing above the reference frequency. This means, the frequency is between

145.1 MHz and 145.2 MHz (first line in Fig. 2). The 50 kHz spectrum is now selected. It will be seen that the frequency f_x is just below the third calibration marker above the reference signal (e.g. just below 145.15 MHz in this example). See the second line in Figure 2. The 1 MHz calibration-spectrum generator is now switched to the 10 kHz spectrum which indicates that the frequency f_x is between 140 kHz and 150 kHz above the reference frequency (3rd line). The final determination of the frequency is made with the 5 kHz or 1 kHz spectrum. This is 145.145 MHz in the example (fourth line in Fig. 2).

The resulting frequency spacing of 145 kHz indicates that the 145th harmonic of the 1 MHz calibration spectrum generator has been tuned, e.g. 145 MHz.

The combination of the 1 MHz and 1.001 MHz calibration-spectrum generators is also very useful for carrying out other types of measurements on receivers: For example, two harmonics of the 1.001 MHz calibration signal are heard at 145 kHz and 127 kHz above a reference frequency of 145 MHz on a single-conversion receiver having an IF of 9 MHz. The 127 kHz signal indicates that a signal of 127 MHz is being received which will be the image frequency of the receiver. This will indicate a poor selectivity of the input stages.

A second example in conjunction with a 2 metre converter and 10 metre receiver: Two harmonics of the 1.001 MHz spectrum are heard at 29 kHz and 87 kHz above the reference frequency in addition to the true harmonic at 145.145 MHz. By calculating the reasons for this, it is found that the 10 metre receiver is receiving the 29th harmonic of the calibration signal directly (29 MHz marker), and that the image frequency suppression is also poor. A frequency of 87 MHz is the required frequency of 145 MHz minus twice the intermediate frequency (2 x 29 = 58 MHz).

3. STABILIZED POWER SUPPLY

Figure 3 shows the two voltage stabilizer circuits that are coupled together to generate stabilized voltages of + 5 V for the TTL integrated circuits and + 12 V for the varactor diodes and 200 kHz receiver. A small transformer with two 7.5 V windings for approximately 150 mA can be used.

The bridge rectifier provides two voltages:
The two diodes whose anodes are grounded rectify the voltage in push-pull so that approximately 8.5 V are available for the first stabilizer comprising transistors T 2 and T 3. The whole bridge rectifier provides a DC voltage of approximately 17 V for the second stabilizer comprising transistor T 1.

The reference diode D 2 of the 5 V stabilizer is fed with the 12 V stabilized voltage. The 12 V stabilizer utilizes the 5 V stabilizer since its reference diode consists of the 5 V stabilized voltage, and diode D 1. The interaction between the two stabilizer circuits ensures a very good stabilization of input voltage fluctuations in a simple manner. It is therefore possible for the input voltages of the stabilizer, at connections Pt 10 and Pt 12, to have a high residual hum level (up to 1.5 V_{pp}) or, on the other hand it is possible for low capacitance values to be used for the filter capacitors. However, less than 100 and 500 μF respectively, should not be used; larger values provide an additional filtering.

The operating voltage for the TTL integrated circuits is adjusted under load to exactly 5.0 V with the aid of potentiometer Pt 1. The higher stabilized voltage can be in the range of 11.5 V to 12.5 V due to the tolerance of the zener diode D 1. However, the exact voltage value is not critical. The circuit is dimensioned for the currents required for the whole calibration spectrum generator system. Since the system represents a precision frequency standard, it is not advisable for the voltages to be used for feeding other equipment. Attention should be paid that output Pt 11 (5 V) of the stabilizer is always loaded ($\lesssim$ 680 Ω). During no-load conditions, it is not possible for the current to flow to ground via R 1, R 2 and D 1, which means that the stabilization of both voltages does not take place. It is sufficient to continuously connect the 1.001 MHz oscillator to the output (connections Pt 3 and 11 connected). The 1.001 MHz oscillator can be switched off with a switch which grounds the NAND gate G 1 via connection Pt 2.

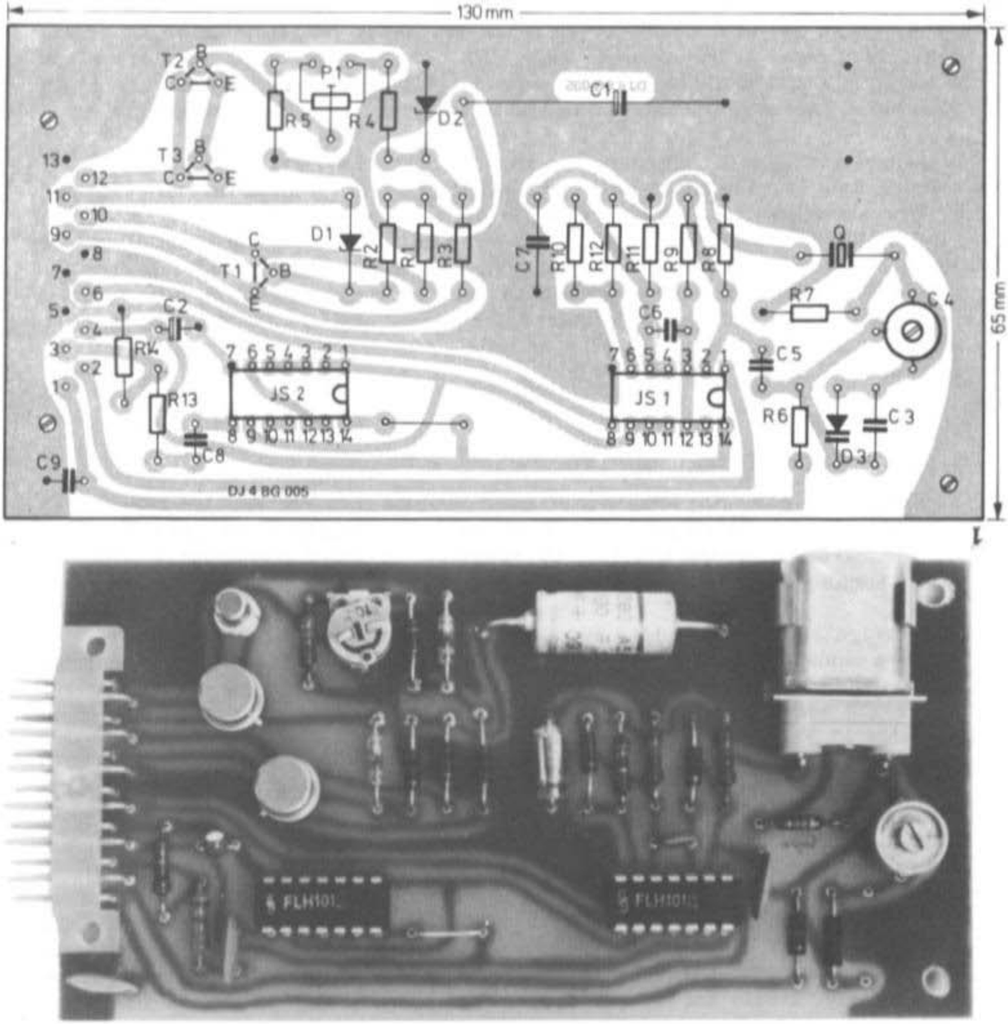


Fig. 4: Printed circuit board DJ 4 BG 005

4. SPECIAL COMPONENTS

I 1, I 2: SN 7400 N (Texas Instruments), FLH 101 (Siemens)

T 1, T 3: 2 N 1613, 2 N 2218, BSX 46-10, BSY 53
or similar NPN-types in TO 5 cases.
T 2: BC 108, BC 183, 2 N 2926 or similar NPN AF-types

D 1, D 2: ZF 7.5 (ITT-Intermetall), BZY 85/C7V5 (AEG-Tfk),
1 N 711, 1 N 4100
D 3: BA 110 (ITT-Intermetall), BA 121 (AEG-Tfk), 1 N 5462 A

Rectifier: Four 1 N 4001 or similar silicon diodes for 50 V/0.5 A
1 crystal: 1.001 000 MHz HC-6/U, parallel resonance 30 pF (KVG)
1 crystal holder for horizontal mounting
1 trimmer capacitor 3-20 pF, 7 or 10 mm dia.,
1 tantalium electrolytic capacitor 22 µF/10 V
1 trimmer potentiometer for horizontal mounting, Spacing 10/5 mm
1 13-pole connector (Siemens).

5. CONSTRUCTION

The described 1.001 MHz attachment and voltage stabilizer are accommodated on printed circuit board DJ 4 BG 005 whose dimensions are 65 mm by 130 mm (Fig. 4). The module can be mounted in a Teko-case type 4A. If used in conjunction with the 13-pole connector, a screened plug-in module is obtained. The advantages of such a modular construction were described in (2). The whole standard-frequency system comprises three such modules. The 200 kHz receiver will be described in one of the next editions of VHF COMMUNICATIONS.

6. AVAILABLE PARTS

The printed circuit board DJ 4 BG 005 semiconductors, crystal, trimmer capacitor and connectors are available from the publishers or their representatives. Please see advertising page.

7. REFERENCES

(1) D.E.Schmitzer: A Digital Calibration-Spectrum Generator
VHF COMMUNICATIONS (3) 1970, Edition 4, Pages 194-205

(2) D.E.Schmitzer: Plug-In Modular Equipment
VHF COMMUNICATIONS (3) 1970, Edition 2, Pages 107-109

MODIFYING A 27 MHz WALKY-TALKY TO 2 m

by E. Ritter, DJ 4 OG

Nearly every VHF-amateur must have thought at one time or another whether it would be possible to modify a 27 MHz or 28 MHz walky-talky for operation on the 2 m band. The appearance, size, weight and ease of operation of such walky-talkies would make them very attractive. Since the circuit does not differ greatly from that of 2 m equipment, the modification is not costly and the alignment relatively simple.

The author modified such a 28 MHz walky-talky to 144 MHz. The first thing that was necessary was for a 2 m converter to be installed into the unit for converting the input frequency of 144 to 146 MHz to the first intermediate frequency of 27 MHz (or 28 MHz). In the author's case, the printed circuit board of the 28 MHz transmitter could also be used for the 2 m transmitter and only had to be re-equipped. In the meantime, a great number of such units have been modified. The following article is to describe the modification of a TELEMASTER 28 MHz walky-talky. However, the modification is also valid for other walky-talkies having a similar circuit.

1. CHARACTERISTICS

The modified walky-talky possess the following specifications:

Width x height x depth (in mm):	70 x 180 x 63
Weight (including batteries):	900 g
Power requirements:	8 x 1.5 V miniature cells
Current drain during reception according to volume:	15 - 25 mA
Current drain during transmit:	17 - 19 mA
Mean carrier power (aligned for optimum AM):	25 mW
Peak output power:	100 mW

Receiver: Double-conversion superhet, 1st IF: 27 MHz, 2nd IF: 455 kHz
Converter tuning: with varactor diode (position "V"), can be switched to a fixed, crystal-controlled frequency (position "F")
Noise figure: better than 9 dB.

The image frequency (90 - 92 MHz or 88 - 90 MHz) will be audible when greater than 40 μV at the antenna socket. The first intermediate frequency (27 or 28 MHz) will only be heard when the signal is stronger than 1 mV. Several spurious signals will be heard between 85 MHz and 88 MHz when the input voltage is greater than 70 μV.

The built-in loudspeaker is used as microphone, as was the case in the original circuit. A small S-meter was built in which will also indicate the battery voltage in the "F" position of the sliding switch located on the side of the walky-talky.

2. MODIFICATION

After preparing the case, the transmitter board is modified and a 2 m converter built into the unit. A modified 27 MHz walky-talky is shown in Figure 1.

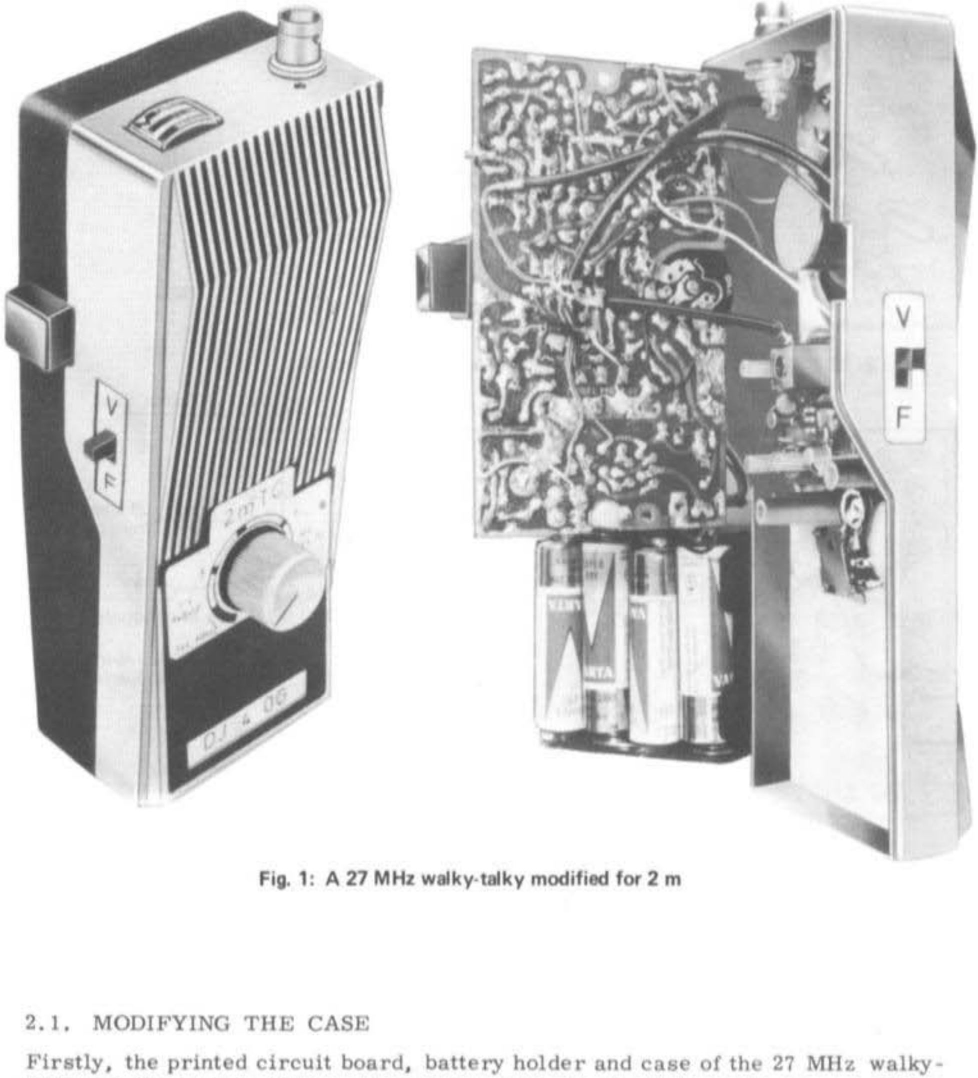


Fig. 1: A 27 MHz walky-talky modified for 2 m

2.1. MODIFYING THE CASE

Firstly, the printed circuit board, battery holder and case of the 27 MHz walky-talky are dismantled. This is followed by removing the telescopic antenna, which is no longer required. An antenna connector is now installed in its place. Enough room is available adjacent to the antenna socket for a small μA-meter which can be used as S-meter and for checking the battery voltage. Such a meter can be installed after cutting a hole of sufficient size into the case.

The calling-tone button in front of the walky-talky is removed and a 8 mm diameter hole is drilled 20 mm below it for the tuning potentiometer. No further modifications are required to the case.

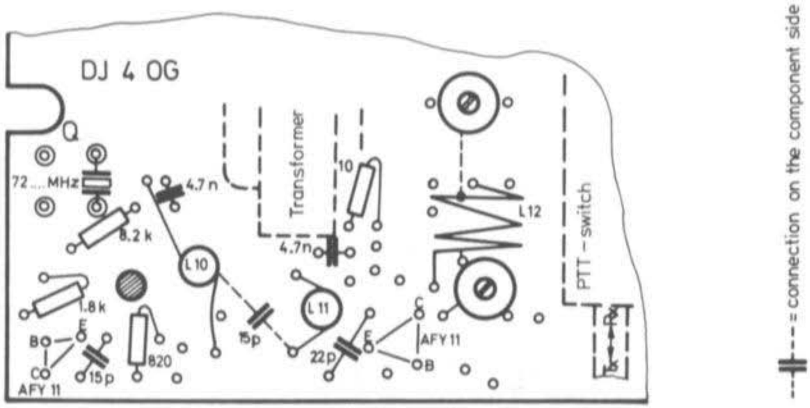


Fig. 2: Modifications made to the original PC-board of the transmitter

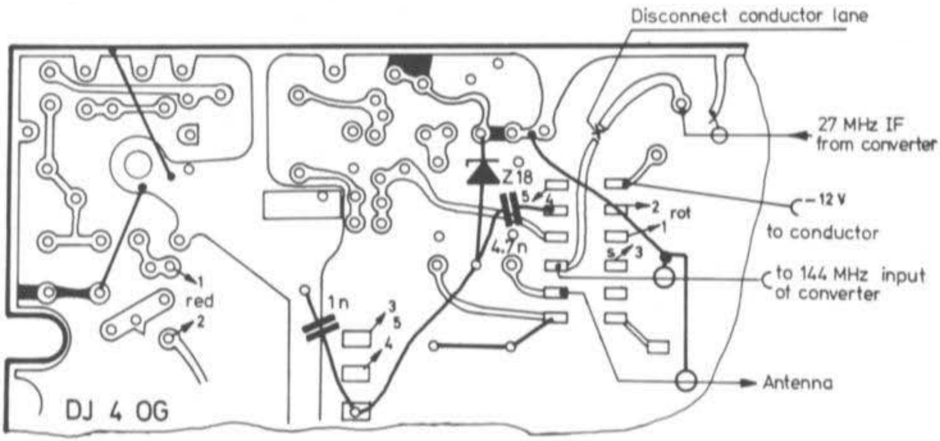


Fig. 3: Components, and interconnections on the conductor side of the PC-board

2.2. MODIFICATION OF THE TRANSMITTER

The printed circuit board of the 27 MHz transmitter can be used, without modification, for the 2 m transmitter. However, all components of the transmitter including the transistors must be removed. The transmitter portion of the printed circuit board is then re-equipped as shown in Figure 2 and Figure 3. The circuit diagram is given in Figure 4, and Figure 5 shows a photograph of the modified transmitter.

A 2-stage transmitter is also used on 2 metres. The crystal frequency is 72 MHz. It was not possible for NPN-transistors to be used on the original printed circuit board. For this reason, high-gain germanium-mesa transistors of type AFY 11 were used.

Since the modulator has not been modified, it is possible for the transmitter to be aligned and operated directly after modification. The 18 V Zener diode ensures that the relatively efficient AF amplifier does not cause overmodulation.

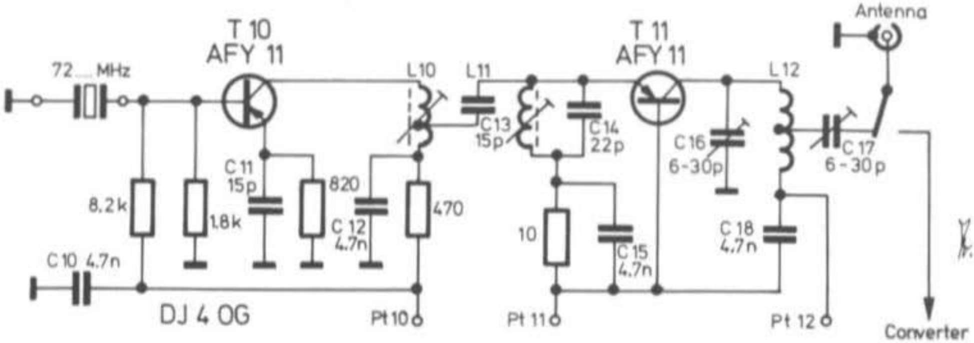


Fig. 4: Circuit diagram of the 2-m transmitter

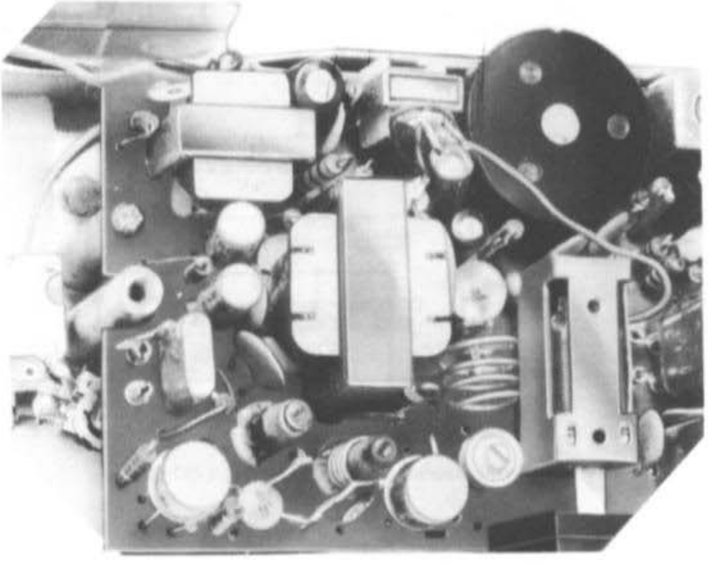


Fig. 5: The modified transmitter

2.2.1. SPECIAL COMPONENTS FOR THE TRANSMITTER

T 10, T 11: AFY 11 (Siemens), 2 N 705 (TI), 2 N 711 (Mot.)
1 Crystal 72.... MHz (HC-18/U) or HC-25/U

C 16, C 17: 6 - 30 pF, ceramic trimmer, 7 mm in diameter

L 10: 13 turns of 0.3 - 0.5 mm dia. (26 AWG) enamelled copper wire wound on a 4.3 mm dia. coilformer with core. Tap 3 turns from the cold end.
L 11: 4.75 turns of 0.8 - 1 mm dia. (19 AWG) silver-plated copper wire on on a 4.3 mm coilformer with VHF core.
L 12: 4 turns wire as for L 11, coil tap 1 turn from the cold end, wound on a 8 mm former, self-supporting.

1 Zener diode ZF 18 (ITT-Intermetall), BZY 85/C 18 (AEG-Tfk).

2.3. THE CONVERTER

The circuit diagram of the simple but stable and sufficiently sensitive converter is given in Figure 6. It possesses a common-base amplifier T 1 which is followed by the mixer transistor T 2. The oscillator can be tuned with the aid of a varactor diode between approximately 58 and 59 MHz. The VFO frequency is doubled and fed to the mixer.

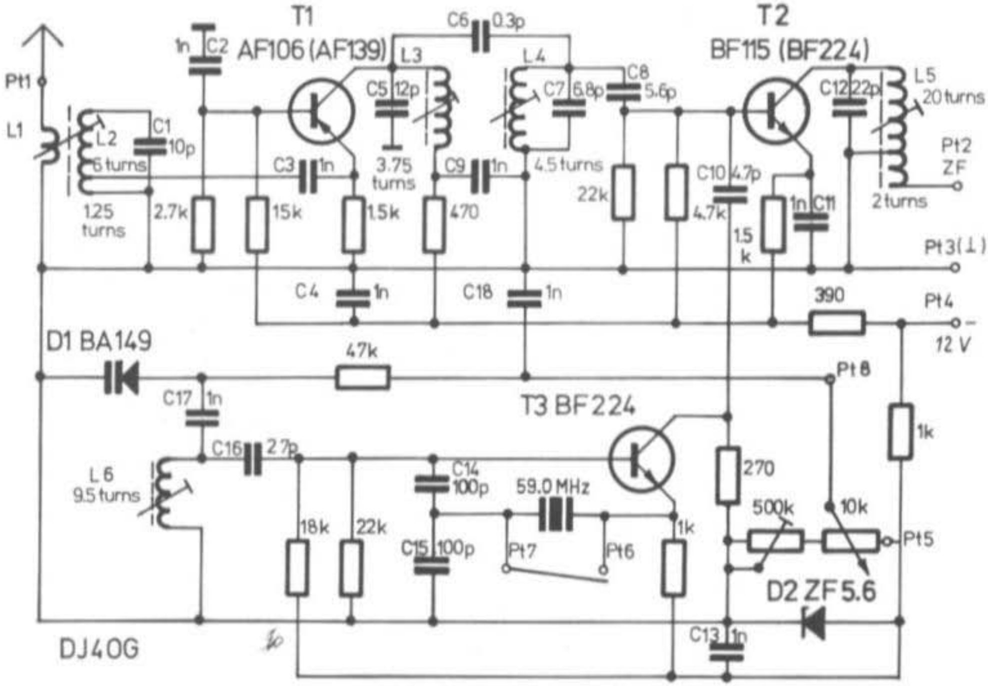


Fig. 6: Circuit diagram of the converter DJ 4 OG 001

A fixed, crystal-controlled frequency can be selected instead of the variable frequency so that the receiver can be tuned to one crystal-controlled frequency on the 2 m band. The crystal-controlled frequency is selected using the channel switch originally used for switching the two crystals of the transmitter. The wiring to this sliding switch on the side of the case would be too long for the 72 MHz crystals used for the transmitter. However, it can be used for switching the receive crystal. If the intermediate frequency is in the 27 MHz band, very inexpensive crystals are available with a frequency spacing of 10 kHz.

Both germanium-Mesa and silicon transistors are used for the converter. It is accommodated on a small printed circuit board having the dimensions 58 mm x 35 mm. The printed circuit board has been designated DJ 4 OG 001 and is given in Figure 7. A photograph of the completed converter is given in Figure 8.

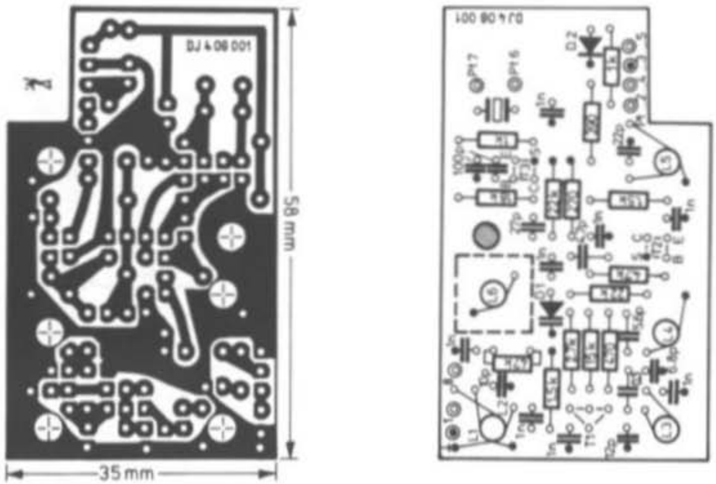


Fig. 7: Printed circuit board of the converter (DJ 4 OG 001)

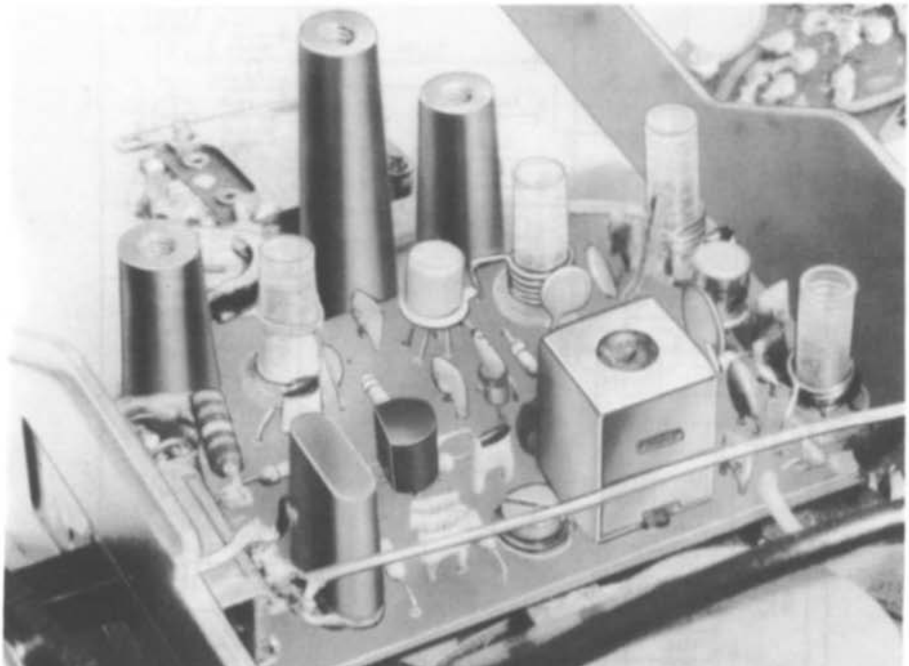


Fig. 8: Enlarged photograph of the built-in converter

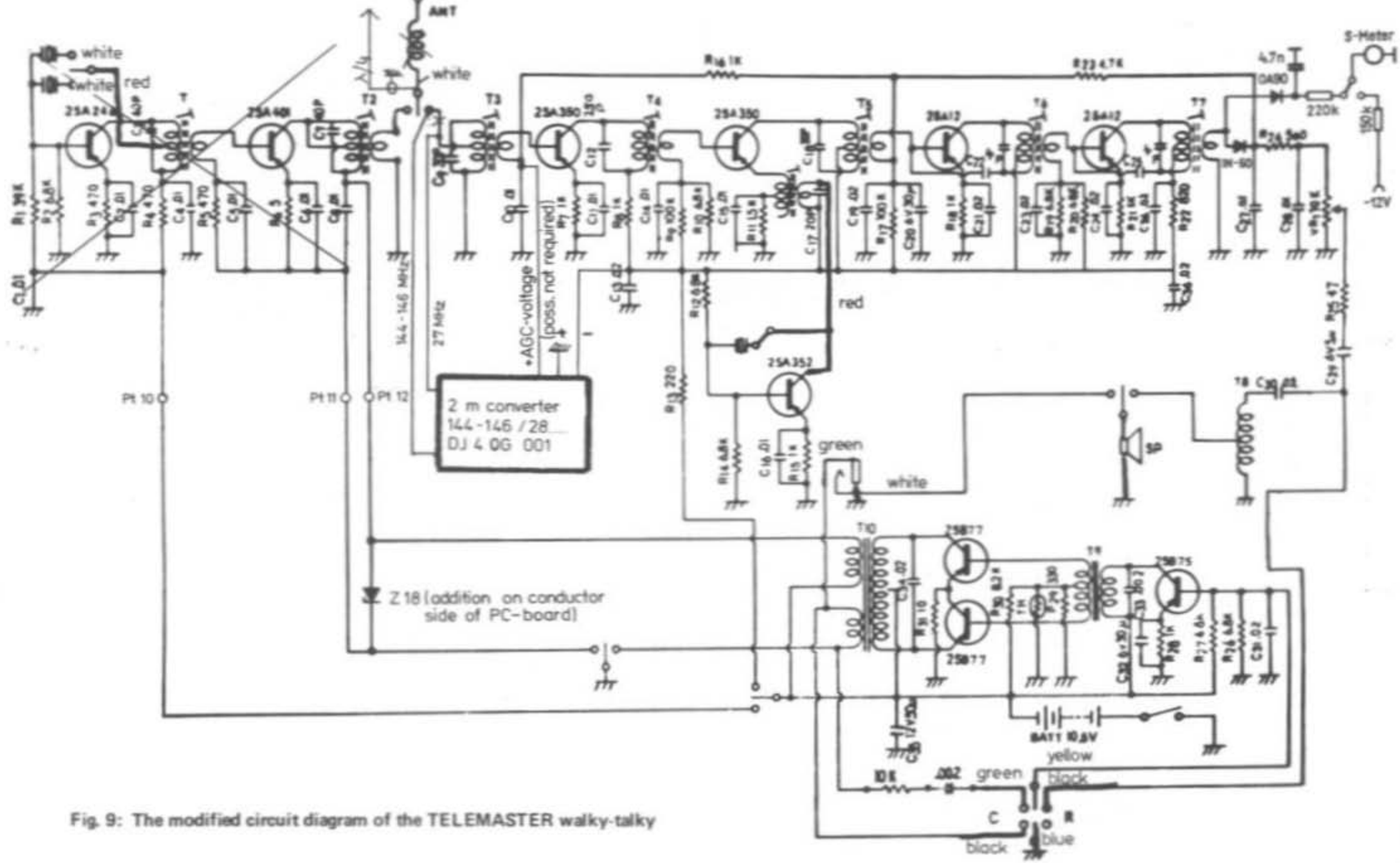


Fig. 9: The modified circuit diagram of the TELEMASTER walky-talky

2.3.1. SPECIAL COMPONENTS FOR THE CONVERTER

T 1: AF 106 or AF 139, 2 N 2494 or 2 N 3399
T 2: BF 115, BF 224 (Texas Instruments Germany), 2 N 780
T 3: BF 224 (TI), 2 N 3932

D 1: BA 149 varactor diode (AEG-Tfk), 1 N 5461 A (Mot)
D 2: ZF 5.6 (ITT-Intermetall), BZY 85/C5V6 (AEG-Tfk), MZ 4626 (Mot)

All inductances are wound on coil formers of 4.3 mm diameter, with core.

L 1: 1.25 turns of insulated copper wire wound into the cold end of L 2.
L 2: 6 turns of 0.3 - 0.4 mm dia. (27 AWG) enamelled copper wire, coil tap 1 turn from the cold end, VHF core (brown)
L 3: 3.75 turns, wire and core as for L 2
L 4: 4.5 turns, wire and core as for L 2
L 5: 22 turns, coil tap at 2 turns from the cold end, wire as for L 2, RF-core (red dot)
L 6: 9.5 turns, wire as for L 2, RF core (pink dot)
Screening can for L 6: 10 x 10 mm

1 crystal 59.0 MHz (HC-25/U)

All resistors are 0.1 W types. All capacitors have a spacing of 2.5 mm.

2.4. ALIGNMENT AND CONNECTION OF THE CONVERTER

The VFO of the receiver is firstly aligned so that the intermediate frequency is in the range of 28 - 30 MHz. The alignment can thus be made in conjunction with another 2 metre receiver before installing the converter into the walky-talky case. It is sufficient for the converter to be tuned for best reception of a relatively weak receive signal.

Figure 8 shows how the converter is installed into the case. The input is connected to the transmit-receive switch and the output is fed to the first 27 MHz (28 MHz) stage of the walky-talky. The conductor lane that originally connected this stage to the transmit-receive switch should be disconnected. Two further connections are required to the power supply. The connections are given in Figure 3. These connections are also shown in the complete circuit diagram of the walky-talky given in Figure 9, which also shows where the S-meter and modified transmitter are to be connected. After installation, the converter is aligned to the final intermediate frequency.

EDITORIAL NOTES

This article is designed to indicate how any 27 or 28 MHz walky-talky can be modified for 2 m operation. Although the description describes the modification of a TELEMASTER walky-talky, the converter can be used in any such unit and it is only necessary for the transmitter to be modified in a similar manner to that given.

AVAILABLE PARTS

Since the components for the transmitter will vary between the different walky-talky types only material for the converter is available. The printed circuit board DJ 4 OG 001, semiconductors and other components are available from the publishers or their representatives. Please see advertising pages.

AN INTEGRATED AF-AMPLIFIER AND VOLTAGE STABILIZER

by D. E. Schmitzer, DJ 4 BG

The AF-amplifier and the voltage stabilizer of the modular receiver system (1) are now to be described. In order to utilize the whole of the printed circuit board and Teko-case (2), these two separate modules have been combined. The voltage stabilizer is very similar to the universal power supply described in (3). As was mentioned in (1), all of these modules are independent from another and can be used for other applications.

1. CIRCUIT DETAILS

1.1. AF AMPLIFIER

The AF-amplifier, which is equipped with the integrated circuit PA 237 is shown in Figure 1. It is dimensioned for an operating voltage of 18 V. As was mentioned in (4), it can also be used at other operating voltages. In order to avoid any alignment, the balance circuit recommended in (4) was not used since the reduced drive range was found to be sufficient. However, if it is found that the voltage between connection 7 of the PA 237 and ground deviates by more than 1 V from half the operating voltage due to components tolerances, this can be compensated for by altering the value of resistor R 7. The other characteristics of the circuit will not be influenced greatly by this.

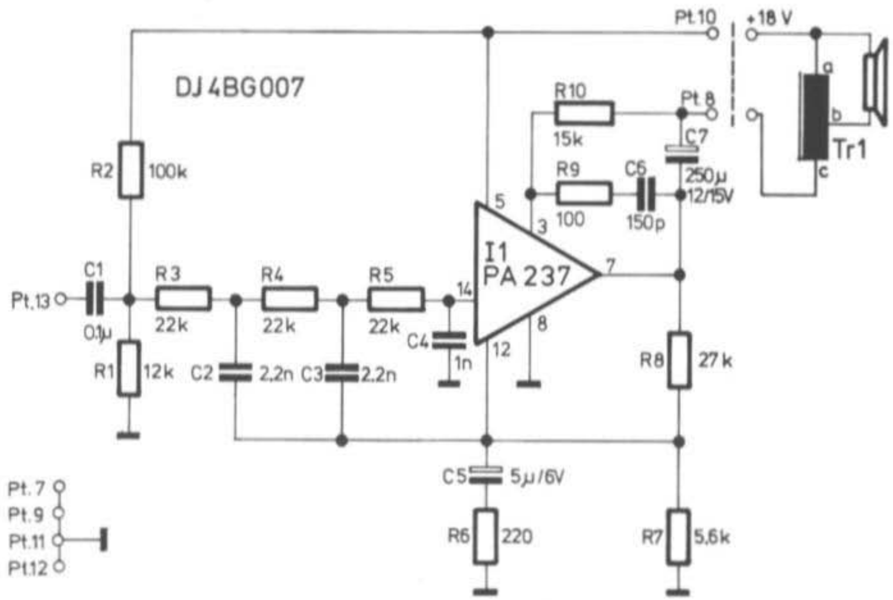


Fig. 1: Circuit diagram of the integrated AF-amplifier

With the aid of an active lowpass filter according to (5), the frequency response is limited to that required for voice transmission. This is especially important for the reception of frequency-modulated transmissions. The frequency response of the AF-amplifier is given in Figure 2.

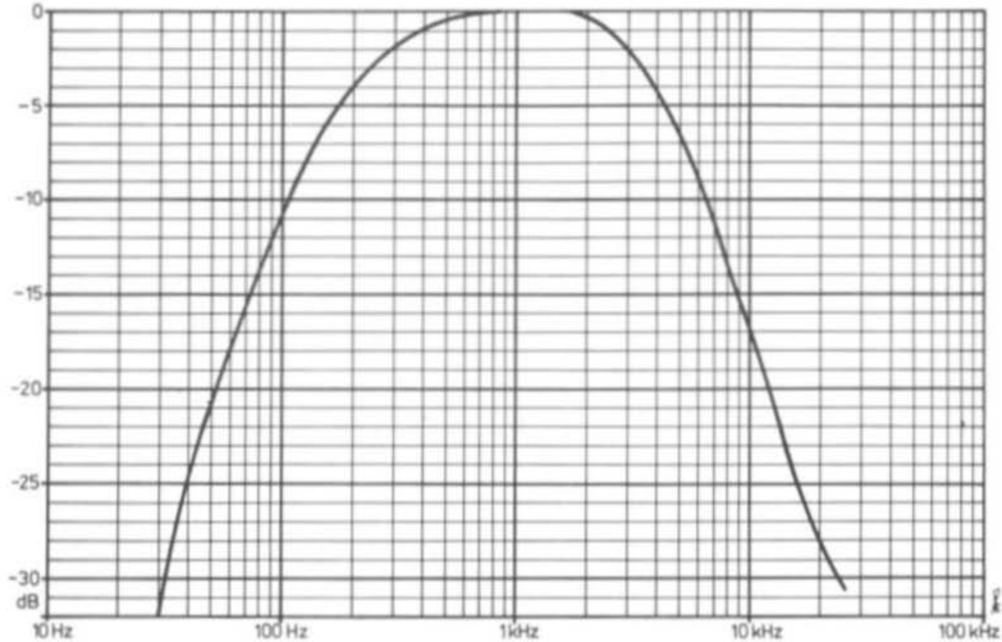


Fig. 2: Frequency response of the AF-amplifier

The bypassing of the input described in (4) is realized in this circuit with the aid of capacitor C 4. A filter capacitor in the operating voltage line is not provided here because connection Pt 10 is normally directly connected to connection Pt 2 (input of the voltage stabilizer) where such a capacitor is present.

For the application in question, the AF-amplifier should provide an output power of 0.5 to 1 W. The integrated circuit should, in this case, be provided with a load impedance of 20 to 30 Ω (4). Since loudspeakers of this value are difficult to obtain, a small transformer is provided for matching 4 to 8 Ω loudspeakers. In order not to limit the wide application range of the amplifier, the transformer is not accommodated on the printed circuit board but mounted externally. Further details regarding the winding of this transformer are given in Section 2.1.

The following table gives the most important specifications of the AF-amplifier:

Operating voltage (nominal value):	18 V
Output power (including transformer):	580 mW into 8 Ω, 1 W into 4 Ω
Input voltage for full output:	120 mV (340 mV_{pp})
Input impedance:	approx. 10 kΩ
Required source impedance (due to the AF-filter):	≤ 1 kΩ
Limit frequencies (-3 dB):	230 Hz and 3400 Hz
Quiescent current (according to IC):	approx. 9 mA
Current at full drive:	56 mA at 8 Ω, 100 mA at 4 Ω
Efficiency (including transformer) at full output:	57.5% at 8 Ω, 55.6% at 4 Ω

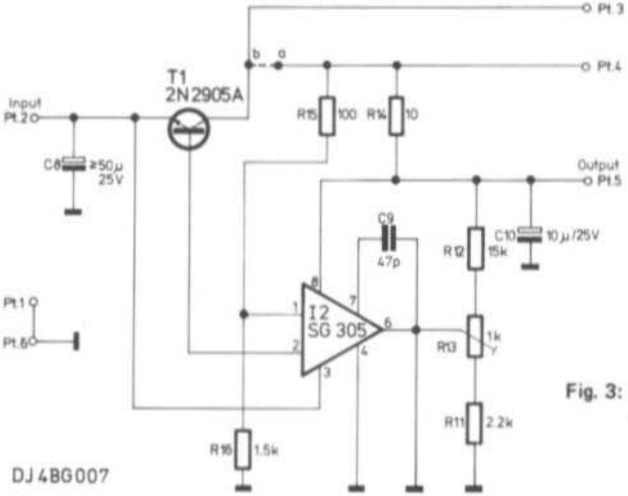


Fig. 3: Circuit diagram of the DC-voltage stabilizer

1.2. DC-VOLTAGE STABILIZER

The circuit diagram of the integrated voltage stabilizer is given in Figure 3. It is designed for an output voltage of 12 V, but any required voltage between 5 V and 24 V can be adjusted with the aid of resistor R 12. The circuit was described in more detail in (3).

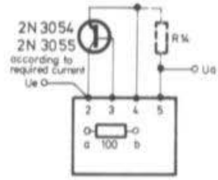
The current limiting is adjusted to 100 mA. This value, however, is altered if the circuit is modified to other output voltage values. For this reason, the following table gives orientation values which, however, can fluctuate somewhat due to the tolerances of the integrated circuit, and resistors R 11, R 12 and R 13.

R 12 (kΩ)	Adjustment range with R 13	Quiescent current (centre position of R 13)	Current limiting adjusted to approx. 100 mA at 12 V
4.7	4.2 - 6 V	4.5 mA	-
5.6	4.6 - 6.5 V	4.8 mA	-
6.8	5.5 - 7.7 V	5.4 mA	70 mA at 6.0 V
8.2	6.3 - 8.8 V	5.9 mA	-
10	7.2 - 10.2 V	6.6 mA	88 mA at 9 V
12	8.1 - 11.6 V	7.4 mA	88 mA at 9 V
15	10.0 - 14.0 V	8.6 mA	105 mA at 12 V
18	11.6 - 16.2 V	9.8 mA	125 mA at 15 V
22	13.5 - 19.2 V	11.0 mA	125 mA at 15 V
27	17.0 - 23.6 V	13.4 mA	145 mA at 18 V
33	19.9 - 28.1 V	15.7 mA	180 mA at 24 V

The input voltage of the circuit must be at least 3 V higher than the required output voltage under full load conditions.

If higher current values are to be stabilized, the circuit can be extended as given in Figure 4 using an external transistor. The required connections for this are available on the 13-pole connector. The resistor required by the power transistor between base and emitter can be mounted on the board instead of the bridge between points a and b. The resistor R 14 for adjusting the current limiting should be reduced in order to allow higher current values. Normally, R 14 is accommodated on the PC-board; however, at higher current values, it should be mounted externally and connected to connections Pt 4 and Pt 5.

Fig. 4: Extending the stabilizer for higher current values



2. CONSTRUCTION DETAILS

As has been previously mentioned, the AF-amplifier and DC-voltage stabilizer are accommodated on a common printed circuit board. The single-coated board has the dimensions 90 mm x 65 mm and has been designated DJ 4 BG 007 (see Fig. 5). It is advantageous for a small, thin brass heat sink to be provided for the integrated AF-amplifier. The dimensions of this can be taken from the component location plan given in Figure 5. As was described in (2), the printed circuit board is provided with a 13-pole connector and is built into a Teko-type 3A case.

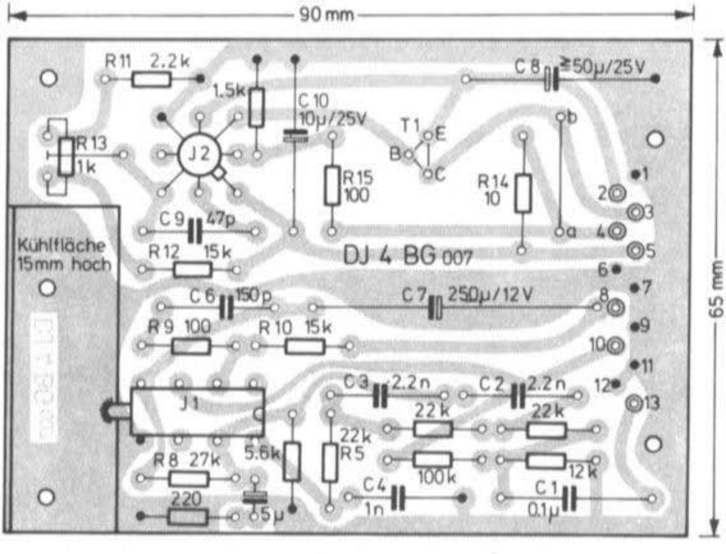
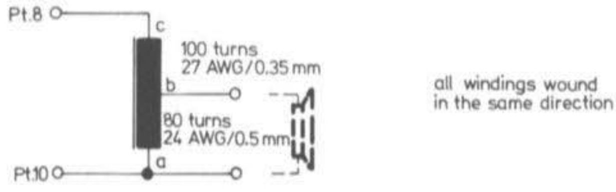


Fig. 5: Printed circuit board DJ 4 BG 007

2.1. SPECIAL COMPONENTS FOR THE AF-AMPLIFIER

I 1: PA 237 (General Electrics)
C 2, C 3, C 4: Plastic foil capacitors if possible
C 5 and C 7 : Tantalium capacitors 6/8 V or 12/15 V
Output transformer: Type C transformer,
Core material: TRAFOPERM N2, wound as an auto-transformer as follows:



2.2. SPECIAL COMPONENTS FOR THE DC-VOLTAGE STABILIZER

I 2: LM 305 (National Semiconductor) or SG 305 (Silicon General)
T 1: 2 N 2905 A, BC 160 (Silicon PNP; 0.6 A) with cooling fins
R 13: 1 kΩ trimmer potentiometer for horizontal mounting, spacing 10/5 mm

3. FURTHER DETAILS

Since no interconnections exist between the AF-amplifier and the stabilizer with the exception of the ground connection, both circuits can be used completely independent of another. The AF-amplifier is usually operated from an unstabilized operating voltage, whereas critical stages, such as oscillators and AF-preamplifiers, are fed from the voltage stabilizer. This means that no coupling can occur between the various stages over the operating voltage line.

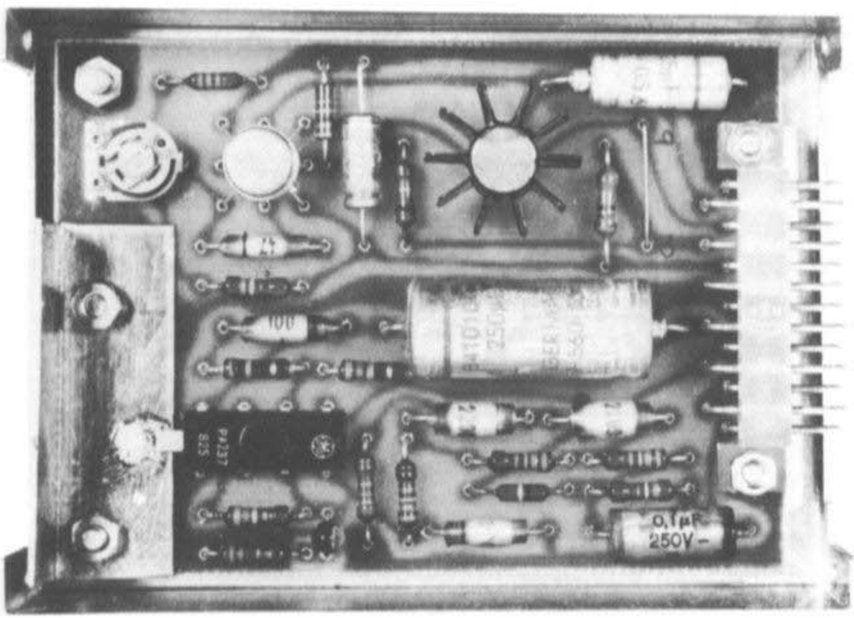


Fig. 6: Photograph of the module DJ 4 BG 007 in a Teko-case

4. REFERENCES

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VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 110-114

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VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 107-109

(3) H. J. Franke, H. Kahlert: Universal Power Supply Using an Integrated DC-Voltage Stabilizer
VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 121-126

(4) D. E. Schmitzer: An Integrated Audio Amplifier Using the PA 237
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(5) D. E. Schmitzer: Active Audio Filters
VHF COMMUNICATIONS 1 (1969), Edition 4, Pages 218-235

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	

A 9 MHz IF-MODULE FOR FREQUENCY MODULATION

by D. E. Schmitzer, DJ 4 BG

This module represents another unit of the modular receiver system (1) using screened, plug-in modules. An FM IF-amplifier and discriminator is now to be described. The module exhibits a high sensitivity which means that limitation takes place at input voltage levels of less than 30 μV. Construction is extremely simple, as is the alignment. This has been achieved by the use of integrated circuits and a crystal discriminator. The use of the crystal discriminator means that no alignment of this critical stage is required. Since the FM-signal is processed at 9 MHz (or 10.7 MHz), it is not necessary for a further conversion to be made to a lower intermediate frequency which means that no further spurious conversion products are generated.

1. CIRCUIT DETAILS

The circuit diagram of the FM IF-amplifier and discriminator is given in Fig. 1. The dimensioning of the whole modular system (1) is based on the fact that the whole selectivity is provided by a crystal filter that is located directly after the first mixer in the input stages of the receiver. This arrangement has the advantage that the input of the filter and the output of the amplifier are accommodated in two different modules that are screened from another. This means that it is virtually impossible to bypass the filter so that the ultimate attenuation of the crystal filter is maintained at more than 90 dB.

The IF-input is terminated with a 62 Ω resistor (may be replaced by 56 Ω or 51 Ω). This has been made so that a defined input impedance of 50 Ω to 60 Ω is present. This results in a high degree of neutralization in spite of the high IF-gain. Coaxial cable can be used for interconnecting the modules.

The integrated circuit type CA 3042 used comprises an RF- and an AF-amplifier. The RF part provides a wideband gain of approximately 60 dB. The resonant circuit (comprising L 1, C 8 and C 9) at the output suppresses the wideband noise voltage from the amplifier and is used for matching the subsequent final amplifier. An integrated circuit type CA 3028 A is used as IF-output stage. This is used as limiter and exhibits a good AM-suppression even at low input levels.

As already mentioned, these two amplifiers are followed by a crystal discriminator which is used as FM-demodulator. At an intermediate frequency of 9 MHz, the crystal discriminator type XD 09-03 is used in conjunction with the crystal filter XF-9E; crystal discriminator type XD 107-01 is suitable for use at intermediate frequencies of 10.7 MHz. Only the value of capacitor C 9 is dependent on the actual intermediate frequency. The value is 56 pF at 9 MHz and 33 pF at 10.7 MHz.

The AF voltage obtained from the output of the discriminator is decoupled in field effect transistor T 1 and passed on to connection Pt 2 of the 13-pole connector. The DC-component can be taken from connection Pt 1 via a dropper resistor of at least 100 kΩ and fed to a meter so that any deviation of the tuned frequency from the centre frequency of the discriminator is indicated. The meter should be a centre-zero type with a full scale deflection of $\pm$5 μA (or $\pm$ 10 μA). Meters for higher current will require a balanced DC amplifier.

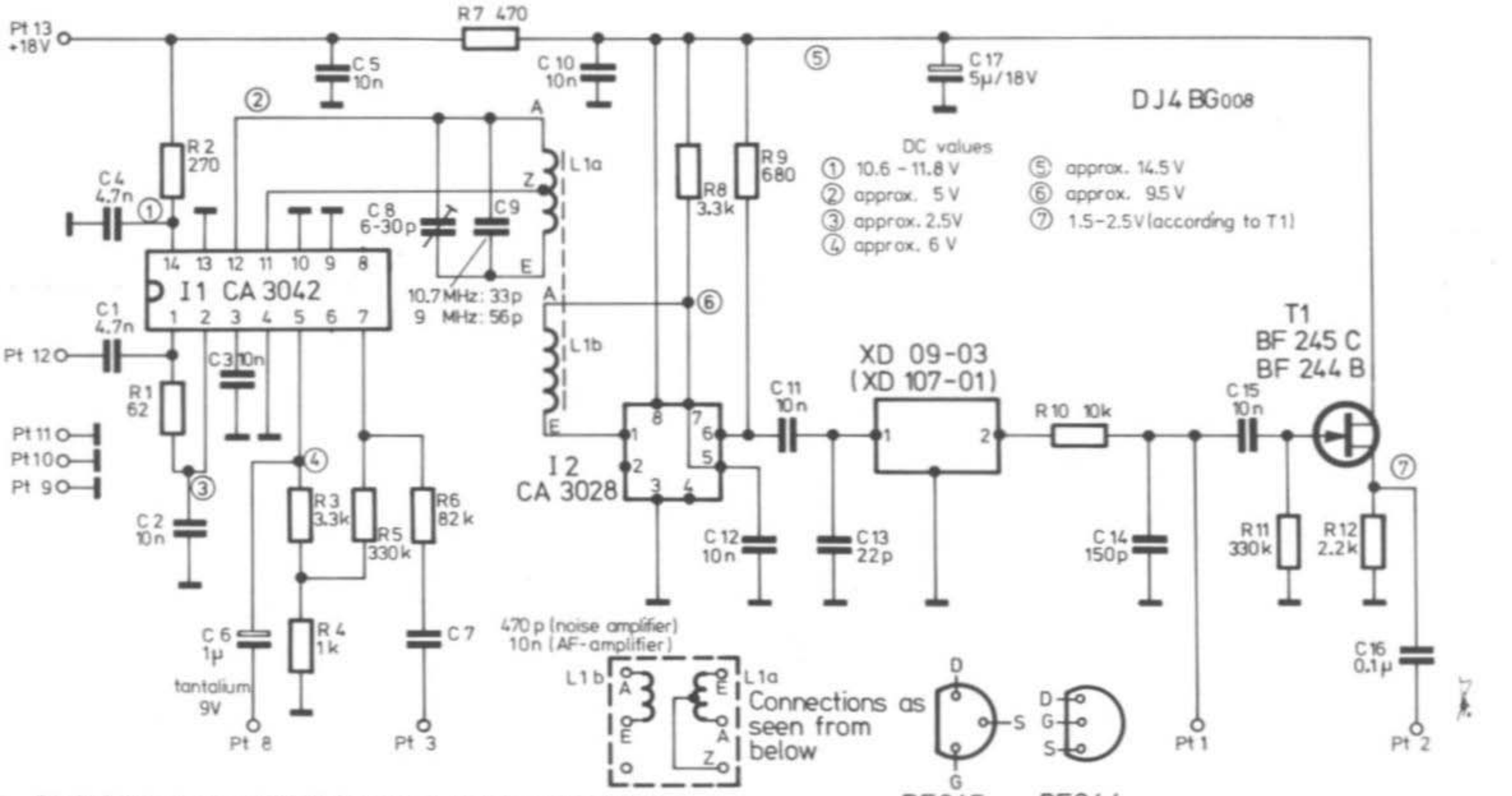

Fig. 1: Circuit diagram of the FM IF-amplifier with crystal discriminator

Since a popular, transformerless AF-amplifier, or the AF-amplifier module (3) belonging to the modular receiver system can be directly driven from the audio output voltage, the AF-amplifier in the CA 3042 is not used for increasing the audio gain but as a squelch circuit. Its output voltage is available at connection Pt 8 of the 13-pole connector which can be used to feed a rectifier circuit and a DC-amplifier (Darlington circuit) for actuating a relay.

This relay will interrupt the AF-line between the discriminator and the AF-amplifier as soon as a high level of noise appears, e.g. when no input signal is present. This squelch circuit is given in Figure 2. The Darlington amplifier is blocked by a high noise level and the relay will no longer be energized. As soon as a signal of sufficient strength is received, the noise component will be decreased so that the amplifier is able to energize the relay. The time constant of the squelch switching can be matched to the required value by altering the capacitance of C_T.

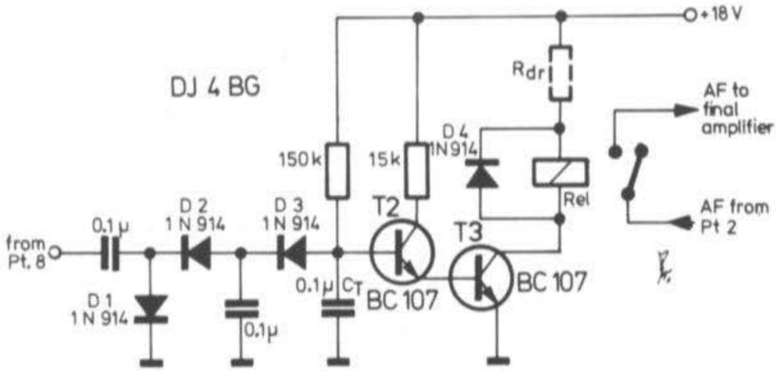

Fig. 2: Squelch circuit

1.1. SPECIFICATIONS AND MEASURED VALUES

The specifications of the FM IF-amplifier given in the following tables are valid for an operating voltage of 18 V.

Operating voltage, nominal		18 V
Current requirements		approx. 33 mA
Input impedance		50 - 60 Ω
Limiter threshold	typical 16 μV	≦ 33 μV
AF-voltage at Pt 2 for XD 09-03: 90 mV/kHz; for XD 107-01:		130 mV/kHz
Source impedance at Pt 2 (dependent on T 1)		approx. 500 Ω
AM-suppression at input voltages between 30 μV and 50 mV	typical 48 dB	≧ 45 dB
Value of C 7 for noise amplification		470 pF
for AF-amplification		10 nF

If the AF- amplifier in the CA 3042 is not to be used as squelch circuit but as AF-amplifier:

Input impedance of the amplifier	100 kΩ
Gain	19 dB
Output voltage with full drive and no load	8 V_{pp}
Output impedance (earphones or transformer)	≧ 2 kΩ

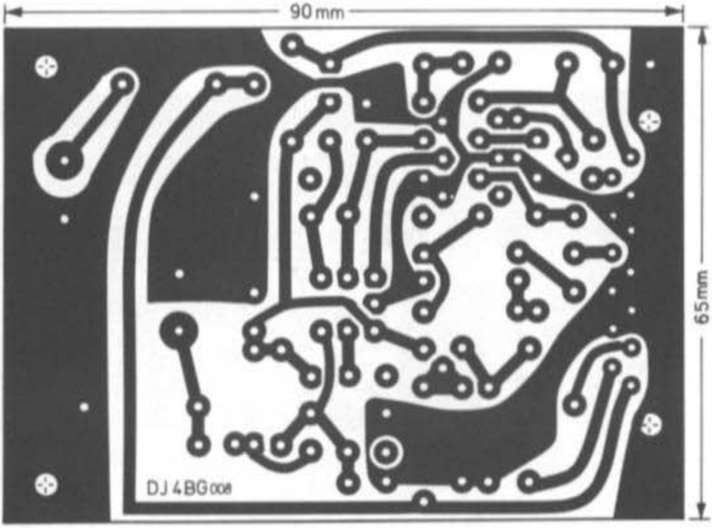

Fig. 3: Printed circuit board DJ 4 BG 008

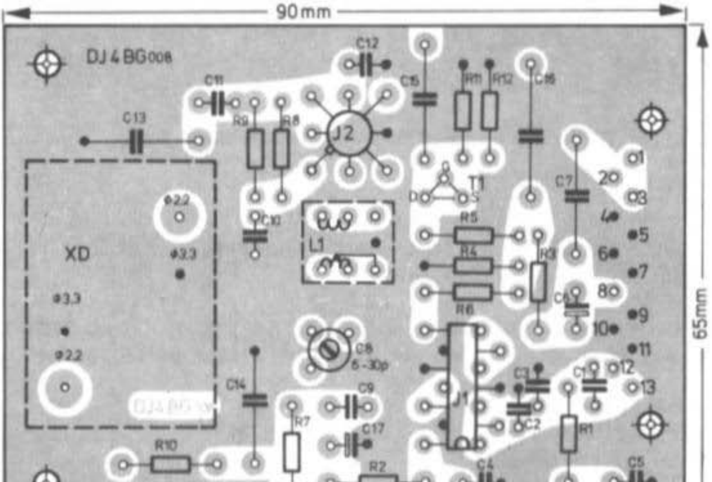

Fig. 4: Component location plan to DJ 4 BG 008

2. CONSTRUCTION DETAILS

The described FM IF-circuit is built up on a printed circuit board. This makes the construction very straight forward. The printed circuit board is double-coated and provided with through contacts, which ensures a high degree of neutralization. It is designated DJ 4 BG 008 and is shown in Figure 3. The component location plan is shown in Figure 4. The dimensions of the printed circuit board are 90 mm x 65 mm and can be accommodated together with a 13-pole connector in a Teko-case type 3A as described in (4). There is no room for the squelch amplifier within this module. It can be built up separately, e.g. on a Vero-board. The construction is not critical. Figure 5 shows a photograph of the module enclosed in a Teko-case.

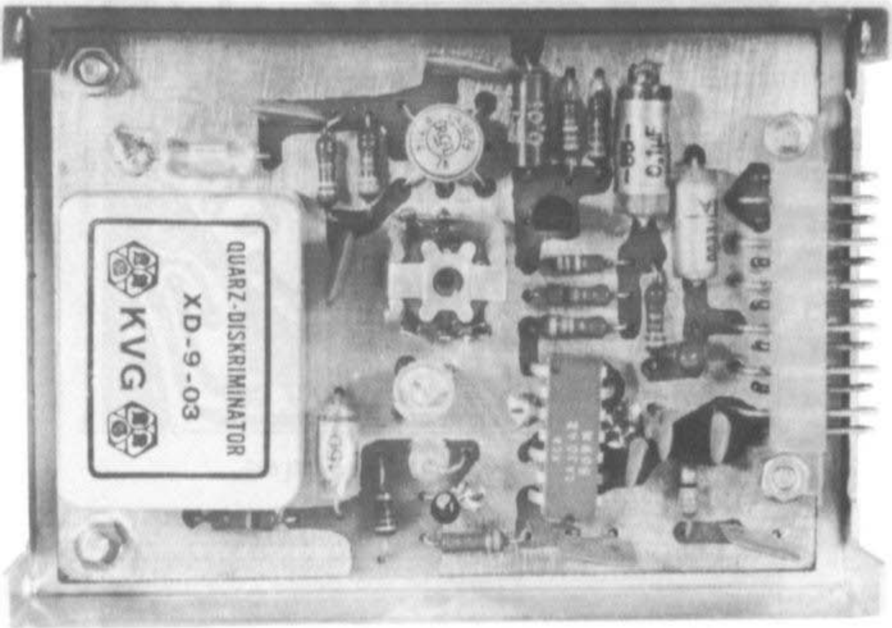

Fig. 5: Photograph of the FM IF-amplifier in a Teko-case

2.1. SPECIAL COMPONENTS

I 1: CA 3042 (RCA); I 2: CA 3028 A (RCA)
T 1: BF 245 C, BF 244 B or C (Texas Instruments), 2 N 5284
T 2, T 3: BC 107, BC 182, 2 N 3903
D 1 - D 4: 1 N 914 or any silicon diode
Crystal discriminator: XD 09-03 (9 MHz), XD 107-01 (10.7 MHz) (KVG)
L 1a: 15 turns of 0.35 mm diameter (27 AWG) enamelled copper wire, tapped at 5 turns
L 1b: 6 turns, wire as for L 1a
Core: Siferrit potted core 11 x 7, 20K12, AL = 16 with coil former and holder (Siemens)
C 8: 6 - 30 pF ceramic miniature trimmer 7 mm diameter.

3. OPERATION

If the FM IF-module is not to be used with the modular receiver system, the RF voltage level required by this module must be taken into consideration during the design of the rest of the receiver. The rough calculation should be based on the limiting threshold of at maximum 30 μV. The limiting should take place with signals that just allow voice intelligibility. With the normal sensitivity values of modern converters and a bandwidth of 12 kHz, this will correspond to an input voltage of approximately 0.5 μV. This means that a gain of approximately 40 dB is required which must be provided by the VHF converter and the subsequent RF-amplifier.

The FM IF-module is operated from an (unstabilized) operating voltage of 18 V. The only alignment that is required consists of aligning the resonant circuit comprising the potted inductance L 1 and capacitors C 8 and C 9 to the intermediate frequency. This is made by injecting an IF-voltage of approximately 5 μV to input Pt 12 and aligning inductance L 1 for maximum with the aid of a wideband oscilloscope, or VTVM with RF-probe, connected to the output of the discriminator (C 13). If no measuring equipment is available, a FM IF-signal is loosely coupled to the input of the module so that it is heard under high noise conditions. Inductance L 1 is then aligned for maximum volume. It us usually sufficient for trimmer capacitor C 8 to be provisionally adjusted to the centre of its range. If inductance L 1 has been carefully wound so that no large deviations exist from the nominal value, this will mean that the nominal frequency will be tuned with sufficient accuracy.

4. REFERENCES

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(3) D. E. Schmitzer: An Integrated AF-amplifier and Voltage Stabilizer
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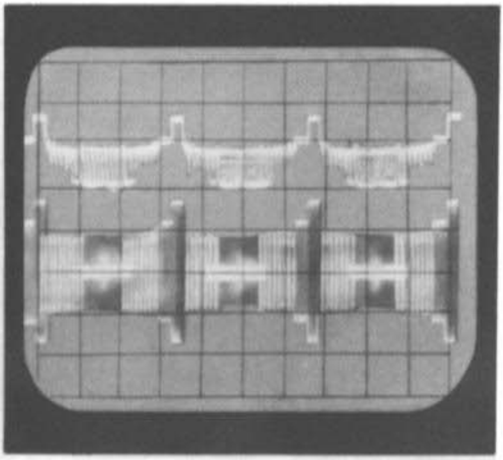

TRANSISTORIZED LINEAR AMPLIFIER FOR 2 METRES

by E. Berberich, DL 8 ZX

A competition was carried out in the German publication UKW-BERICHTE to find the best design of a transistorized linear amplifier. The following design won first prize. The linear amplifier can be used with an operating voltage of up to 26 V and will provide an output power of 10 - 12 W at a drive power of 400 - 500 mW. Three built-in reed contacts switch the linear amplifier into circuit during transmit and bridge it in the receive mode. This means that no external transmit-receive relay is necessary. The relays are controlled via the coaxial cable and a built-in control circuit.

This means that the linear amplifier can be located with it's power supply separate from the transmitter, possibly in the vicinity of the antenna. The linear amplifier is accommodated in a Tekocase type 3B and only weighs 300 g. Figure 1 shows a photograph of the author's prototype. The heat sink is used for cooling both transistors and is grounded. Of course, BNC or other connectors can be used instead of the SO 239 used by the author.

Fig. 1: Prototype of the linear amplifier

1. MEASURED VALUES AND DETAILS REGARDING OPERATION

The following values were measured by the publishers on a prototype:
Bandwidth for a gain reduction of 1 dB: 6 MHz.
The measured output power values as a function of the input power with different operating voltages are given in Figure 2. It will be seen that a linear relationship between input and output power exists up to 8 W at an operating voltage

U_b of 26 V. This means that the peak value of the input power should not exceed 150 mW. An SSB-drive signal should therefore not exceed 150 mW PEP, and an AM-drive signal should not exceed a mean carrier power of 40 mW (an increase of the feedback resistor R 3 to 2.2 Ω will increase the linearity, but will cause a reduction of output power and efficiency).

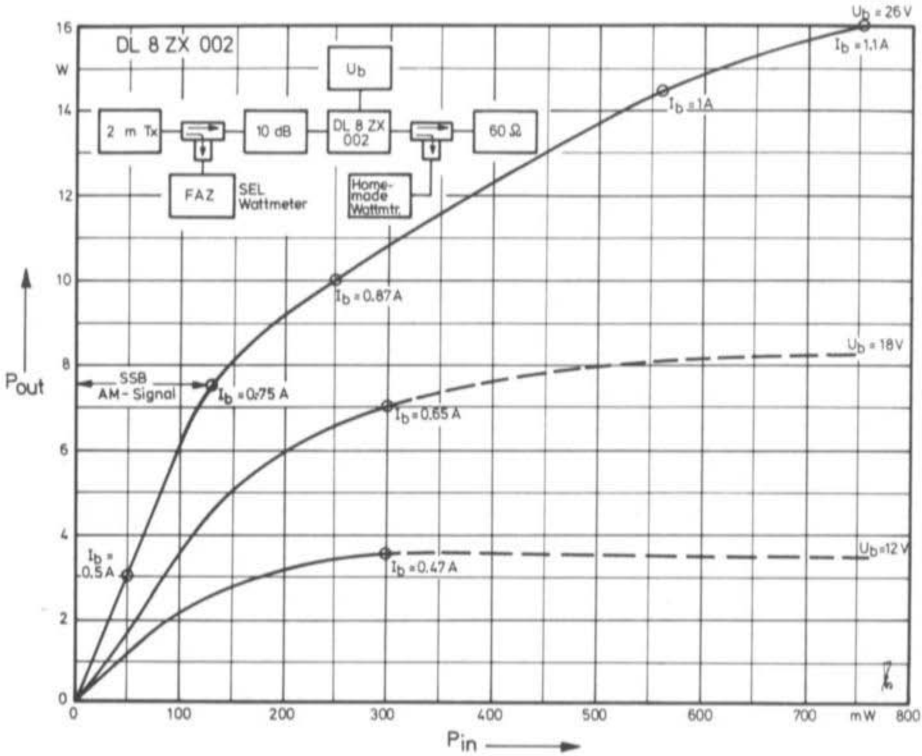

Fig. 2: Diagram of the input and output power characteristics at various operating voltages

At operating voltages of less than 26 V, the linear range is considerably smaller as will be seen in Figure 2. This means that the drive power should be correspondingly decreased. The actual values fluctuate somewhat due to tolerances of the transistors and alignment. However, they indicate the order of magnitude of the linear drive range.

In the FM or CW-mode, the drive power can be increased to approximately 700 mW which will increase the output power to a maximum of 16 W.

The output stage can also be amplitude-modulated. In this case, the operating voltage should not exceed 12 - 14 V. At a drive power of approximately 300 mW, a carrier power level of 3.5 - 5 W is obtained, which corresponds to a peak power level of 14 - 20 W during amplitude modulation peaks. This will require an audio amplifier that is able to provide 5 W into a load of approximately 24 Ω. Suitable circuits for this are given in (1). The peak modulation voltage should be limited to a value between 27 V and 33 V with the aid of a Zener diode. The diode is connected between connection Pt 7 and ground.

If transmitters with a higher output power level are to be used, it is possible for the linear amplifier to be used without driver stage. In this case, the permissible input power values will be increased by approximately 10 times, e.g. to approximately 1.5 W PEP for linear operation.

2. CIRCUIT DETAILS

Figure 3 shows the circuit diagram of the linear amplifier. Overlay transistors for high operating voltages are used for both amplifier stages T 1 and T 2.

In order to avoid switching the operating point for the different modulation modes, the operating point is adjusted to class AB_1 as a compromise. This means that the quiescent current values are relatively low (T 1: 35 mA, T 2: 65 mA). They can be altered with the aid of resistors R 2 or R 4. The silicon diodes D 1 and D 2 provide a temperature stabilization of the operating points. For instance, a quiescent current variation from 90 mA to 100 mA was measured after a long period of continuous operation by which the heat sink increased in temperature to approximately 60 ^oC.

The emitters of both transistors are bypassed more than once in order to achieve a negligible inductance between emitter and ground at the very low impedance present at these points. The emitter resistors can - if the 5.1 Ω or 1.1 Ω values are not available - be made from two or three parallel-connected resistors of higher values (e.g. 2 x 10 Ω, 3 x 15 Ω, or 2 x 2.2 Ω, 3 x 3.3 Ω). Homemade, wire-wound resistors are not to be recommended here.

The inductances L 3 and L 6 have a very low inductance in order to avoid parasitic oscillation. The connections Pt 5 and Pt 6, or Pt 7 and Pt 8 allow the collector currents to be measured. If they are to be monitored continuously, the shunt resistors required for the meter should be connected instead of the bridges.

The output of the linear amplifier is provided with an additional lowpass filter (C 17, L 5, C 18) in order to improve the harmonic suppression. The operating voltage for the amplifier is also fed into the linear amplifier via a low-pass filter (C 20, Ch 3, C 21) in order to ensure that no RF-voltage is able to leave the amplifier in this manner.

Connection Pt 10 is continuously connected to an operating voltage of 12 V or more. A positive voltage for energizing the relays is fed to the amplifier via the coaxial cable, Pt 1, R 5 and D 3 to a Schmitt-trigger comprising transistors T 3 and T 4. Transistor T 3 therefore takes over the current of T 4, and relays A and B will be energized, whereas relay C is deenergized and the amplifier will be switched on. The same process can be actuated via Pt 9, R 6 and D 4 with the aid of a pushbutton or switch. Diodes D 3 and D 4 isolate the two control inputs from another.

The positive relay voltage of 12 V (to 24 V) is present at the collector of transistor T 4 during transmit. It is therefore possible for a DC-DC converter (12 V to 26 V) to be switched on via connection Pt 11 and a switching stage (Figure 4). For mobile operation in conjunction with a 6 V accumulator, it will be necessary for the DC-DC converter to be switched on separately, since relay A to C will not operate at 6 V.

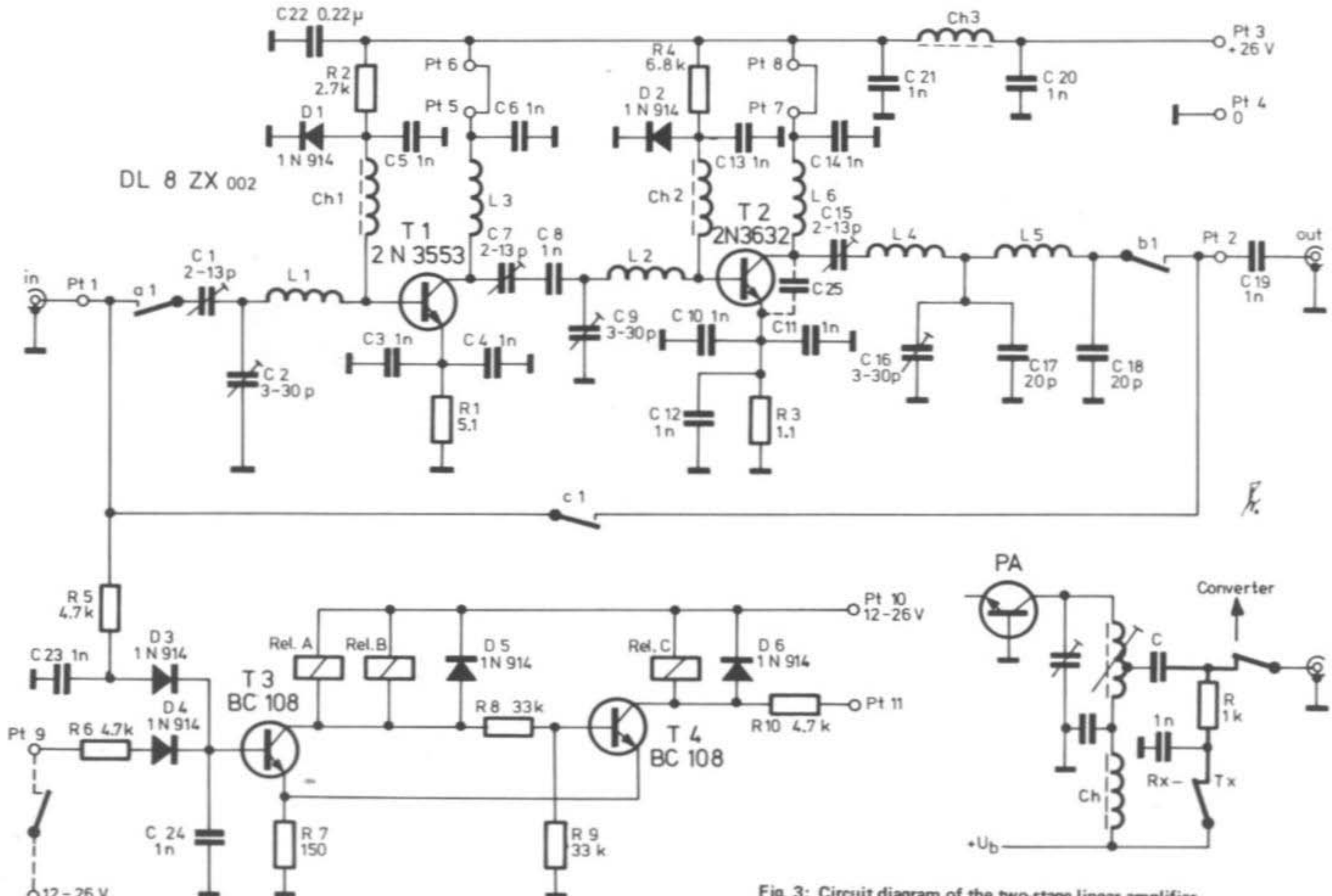

Fig. 3: Circuit diagram of the two-stage linear amplifier

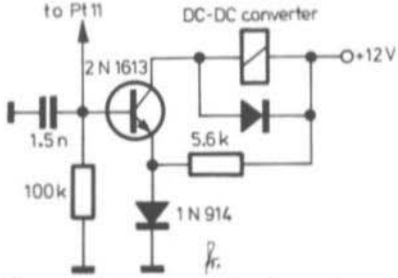

Fig. 4: Circuit for switching on a DC-DC converter

If no positive control voltage is present at the base of transistor T 3, T 4 will once again take current and relay C will be energized, the control voltage for the DC-DC converter will be reduced to the low residual value and relays A and B will be deenergized. The linear amplifier will thus be bridged and the antenna will be directly connected to the transmitter or receiver.

The transmitter-receiver must be provided with a RC-link for the remote control that is also given in Figure 3. The output connector of the transmitter will receive a positive operating voltage in the transmit mode from the transmit-receive switch and resistor R. The isolating capacitor C is usually already available in the transmit circuit. This means that the operating voltage will be present on inner conductor of the antenna cable in the transmit mode. However, resistor R ensures that the voltage is of such a high impedance that no adverse effects occur if it is short-circuited, e.g. by the antenna.

3. SPECIAL COMPONENTS

T 1: 2 N 3553; T 2: 2 N 3632
T 3, T 4: BC 108, 2 N 3704 or similar silicon NPN audio transistors
D 1 - D 6: 1 N 914, 1 N 4148 or similar silicon planar diodes

Relays A, B, C: Energizing coils with 16 000 turns/4600 Ω
Contacts a 1, b 1, c 1: miniature reed contacts (Hamlin type MRR-2)
All inductances are self-supporting:

L 1, L 2: 3 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 5 mm former, coil length 7 mm

L 3, L 6: 2 turns of 0.5 mm dia. (24 AWG) enamelled copper wire wound on a 5 mm former

L 4: 6 turns of 1.5 mm dia. (15 AWG) silver-plated copper wire wound on a 6 mm former, coil length 11 mm

L 5: 4 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 5 mm former, coil length 7 mm

Ch 1, Ch 2: 3 turns of 0.25 mm dia. (30 AWG) enamelled, silk-covered copper wire pulled through a ferrite bead.

Ch 3: Wideband ferrite choke (Philips 4312 020 36700)

C 1, C 7, C 15: 2 - 13 pF air spaced trimmer for PC-board mounting
C 2, C 9, C 16: 3 - 30 pF air spaced trimmer for PC-board mounting

Heat sink: 100 mm long, 70 mm wide, see Figure 1

For T 1: Heat sink T x P 0503 b, manufactured by Souriau

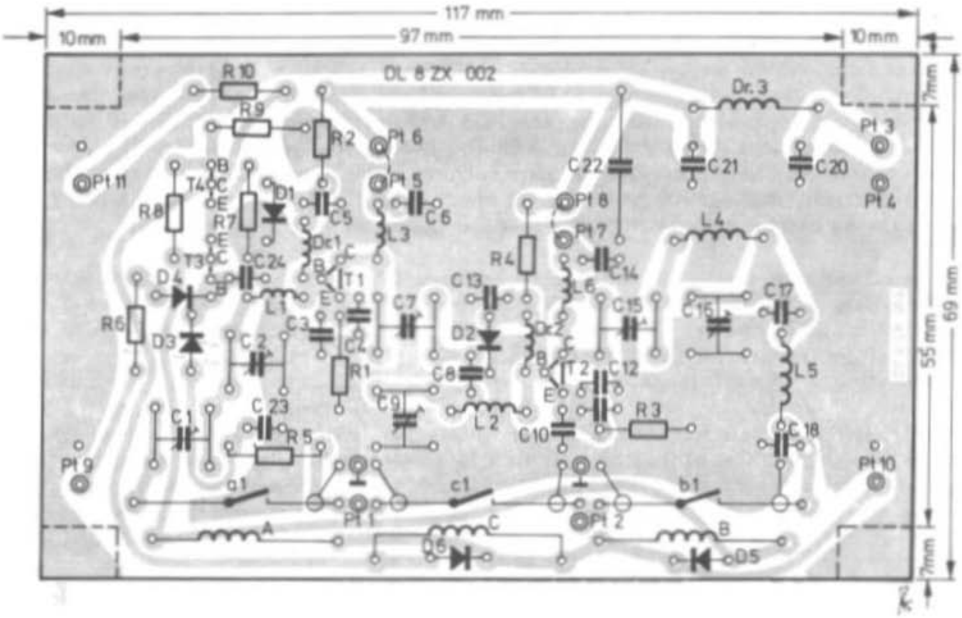

Fig. 5: Printed circuit board DL 8 ZX 002

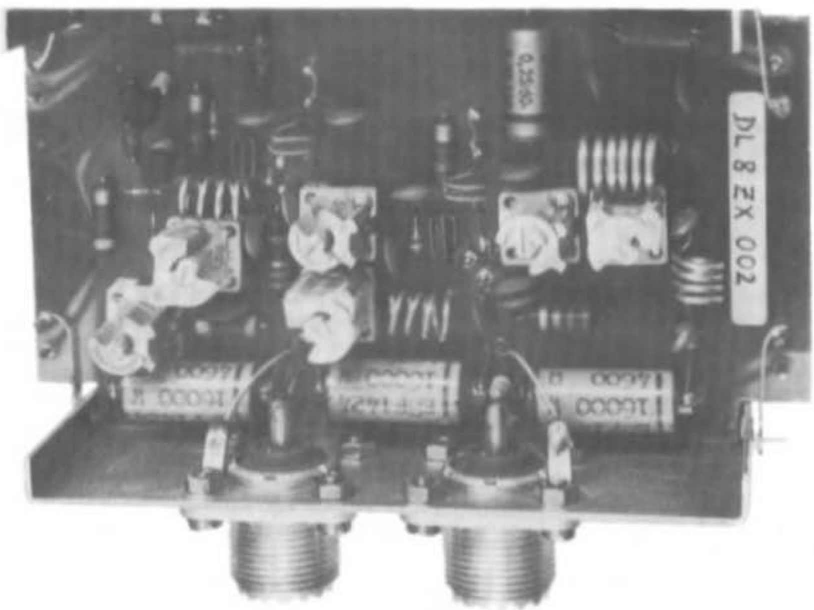

Fig. 6: Internal view of the prototype

4. CONSTRUCTION

The two-stage linear amplifier, the relays and corresponding switching stage are accommodated on the printed circuit board DL 8 ZX 002. The dimensions of this PC-board are 115 mm x 70 mm. Figure 5 shows the printed circuit board and the component location plan. Since the printed circuit board is to be accommodated in a Teko 3 B case with the connection points protruding from the casing, it is necessary for the four corners to be removed. After this, the printed circuit board will protrude by approximately 10 mm on both sides of the case as can be seen in Figure 6 that the connections are accessible.

A small hole is drilled in the centre between the emitter, base and collector connections of transistors T 1 and T 2 in order to subsequently mark the holes for the mounting bolts. The emitter, base and collector holes for these two transistors are provided with small hollow rivets that are soldered to the corresponding conductor lanes.

The energizing coils for the relays are self-supporting and not wound on coil formers. Copper foil of the correct size is covered with Sellotape in order to avoid short-circuits and placed inside the coils. The copper foil is necessary to ensure that the reed contacts are of approximately the correct impedance. Without copper foil, the coupling attenuation between input and output of the amplifier would not be sufficiently great.

The completed printed circuit board is placed temporarily into the casing and the two holes for the mounting of transistors T 1 and T 2 are marked through the small previously made holes. After removing the printed circuit board, the holes are drilled out to a diameter of 5.2 mm. The corresponding holes in the heat sink are marked through these holes and are subsequently drilled. The heat sink is mounted only with the aid of the threaded bolt of transistor T 2.

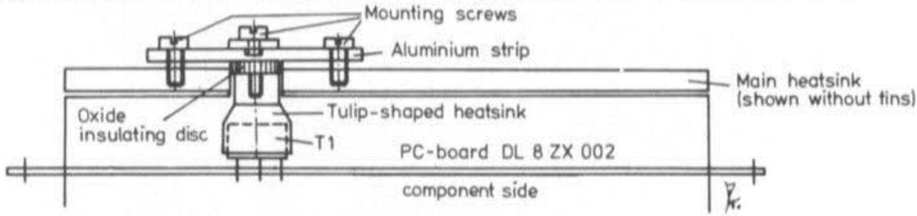

Fig. 7: Mounting and cooling of T 1

The cooling of transistor T 1 is somewhat problematic due to the height of conventional heat sinks. For this reason, a special type of heat sink with an insulating disc of Beryllium or aluminium-oxide is used. This heat sink is shown in Figure 7, which also shows the installation in a sectional drawing. The case and the heat sink are drilled out at the mounting position of transistor T 1 to approximately 11 mm. The special heat sink can then be placed into the hole without touching the sides. A strip made from aluminium plate holds it in an insulated manner and is screwed to the main heat sink.
The casing and the four pertinax strips required for holding the printed circuit board are now prepared. Figure 8 gives the dimensions of these pieces.

The four pertinax strips are now glued into the case and transistor T 2 is loosely screwed into the case and heat sink. The printed circuit board is now inserted and the power transistor rotated until the connections are opposite the hollow rivets in the printed circuit board. The connections are then placed

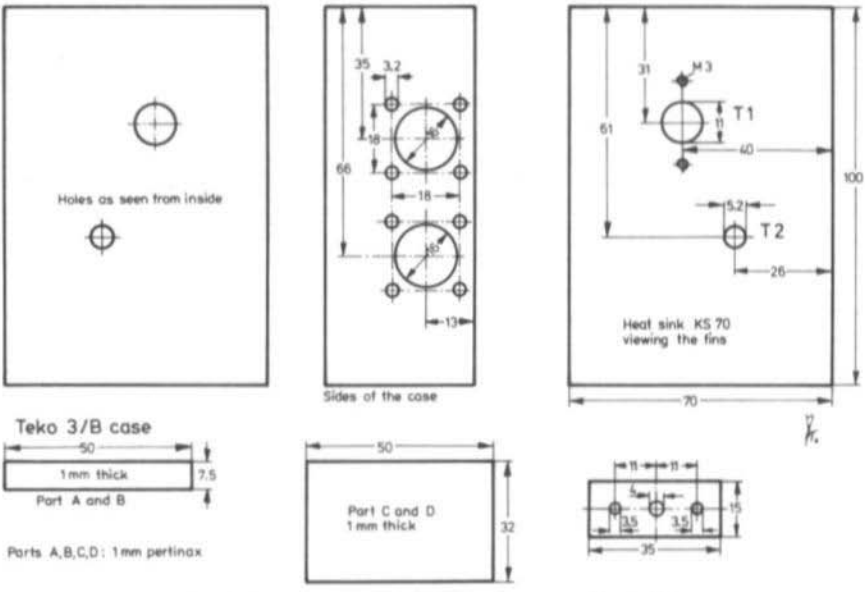

Fig. 8: Dimensional drawings for the case and heat sink

through the printed circuit board, after which the transistor is screwed tightly and the connections soldered. The hollow rivets ensure a good contact to the conductor lanes. After this, T 1 should be mounted as shown in Figure 7.

After mounting the coaxial sockets, it is possible for the input and output to be connected. The input is connected to connection point Pt 1 with the aid of a short piece of silver-plated copper wire. Isolating capacitor C 19 provides the connection between the output of the printed circuit board and the connector. In addition to this, it is necessary for six ground connections to be made. One each to a soldered tag connected to the four mounting screws of the case and two thick wires in the vicinity of connection Pt 1 and Pt 2 to the mounting screws of the coaxial connectors. A wide strip of copper foil as shown in Figure 5 between two widely spaced ground surfaces of the printed circuit board will improve the neutralization. The linear amplifier is ready for alignment after completing the ground connections.

5. ALIGNMENT

Connect the output via an output indicator, e.g. a reflectometer as described in (2), to a terminating resistor. Feed a voltage of 12 to 26 V to connections Pt 9 and Pt 10. Relays A and B are energized. Place a resistor of approximately 30 Ω betwen Pt 5 and Pt 6 instead of the bridge and approximately 15 Ω between Pt 7 and Pt 8. Connect these resistors via a mA-meter to 26 V. Adjust the quiescent current of transistor T1 to approximately 35 mA if necessary, by exchanging R 2. The output stage should be adjusted to approximately 65 mA with the aid of R 4.

Place all trimmer capacitors to their central position. Drive the amplifier with an input power of aproximately 50 mW. Adjust all trimmers carefully for maximum output power. Repeat the alignment on a number of occasions. Vary the drive: The output power should also vary.

Bridge connections Pt 5/Pt 6 and Pt 7/Pt 8 or connect with low value resistors. Measure the current flowing to Pt 3. Several measured values for I_b are given in Figure 2. Drive the input of the linear amplifier again and improve the alignment. Trimmer capacitor C 15 should not reach its minimum value since the output stage would then virtually be unloaded. If preferred, a capacitor of approximately 10 pF can be connected between the collector and emitter of transistor T 2. As explained in (1), this will ensure that transistor T 2 will not be destroyed due to incorrect alignment or incorrect load.

It is possible to determine whether the cut-off frequency of the low-pass filter is high enough by inserting a VHF core and aluminium core into inductance L 5. With the iron dust core, the output power should be reduced and should not increase when the aluminium core is inserted. If necessary, the turns of inductance L 5 should be bent until both conditions are fulfilled. The same effect can be obtained by altering the capacitance of C 17 and C 18.

6. AVAILABLE PARTS

The printed circuit board DL 8 ZX 002 and various components for the linear amplifier are available from the publishers or their representatives. Please see advertising page.

7. REFERENCES

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CIRCULATORS AND ISOLATORS

by R. Lentz, DL 3 WR

Circulators and isolators are used extensively in the professional microwave technology and are also becoming more and more popular at UHF and VHF. These components are expensive and are difficult for the radio amateur to obtain or manufacture, and are therefore only used where inavoidable, e.g. in parametric amplifiers.

The number of radio amateurs experimenting at UHF and SHF is increasing and they will be very interested in circulators that can be manufactured inexpensively according to a new process.

1. TERMS

Circulators are electromagnetic components having three or more connections (ports) in which RF-energy circulates in one direction from one port to another whereas a relatively high attenuation occurs in the opposite direction. In addition to this, all ports are matched. The clear symbols for three-port and four-port circulators are given in Figure 1.

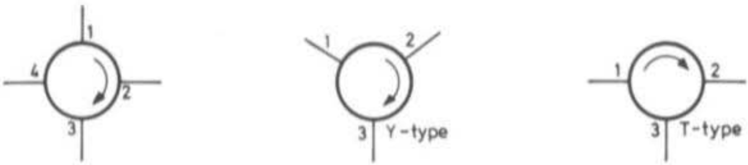

Fig. 1: Symbols of four-port and three-port circulators giving the circulation direction

FOUR-PORT CIRCULATORS operate according to the principle of a non-reciprocal phase-shifter, and also according to Faraday's law of induction.

They are used in the radar technology due to their high power capability. THREEPORT CIRCULATORS have become popular for microwave applications, especially simple Y-type JUNCTION CIRCULATORS. Such junction circulators use a symmetrical Y (or T) junction of three lines (waveguide or stripline). A ferrite body is arranged at the centre of the junction.

LUMPED-CONSTANT CIRCULATORS (with lumped inductances and capacitors) are used for frequencies below approximately 1 GHz since they are smaller than junction types for such low frequencies. Both lumped-constant and junction circulators are to be described in more detail in the following sections.

The (low) attenuation in the circulation direction is called INSERTION LOSS and is in the order of tenths of a dB. The attenuation in the opposite direction is called ISOLATION and amounts to 20 dB or more. Further important magnitudes are the STANDING WAVE RATIO (VSWR) at the ports and the BANDWIDTH of the circulator. Due to the method of operation, resonance effects occur which limit the useful frequency range. However, even the types with the lowest bandwidth have a bandwidth sufficient for amateur applications.

ISOLATORS have only two ports: The insertion loss is very low in the forward direction whereas a high isolation (attenuation) is present in the opposite direction. A MAGNETICALLY BIASED FERRITE CORE which absorbs the electromagnetic waves in one direction but does not influence them in the other direction, is the most important part of both isolators and circulators.

2. APPLICATION

The most popular application of the circulator is probably as duplexer where a transmitter and receiver are connected to a common antenna. In this manner, the circulator decouples the receiver from the transmitter and provides a low-loss path from the transmitter to the antenna and from the antenna to receiver (Fig. 2). Normally, transmitter and receiver operate at different frequencies within the bandwidth of the circulator so that additional filters can be used to increase the limited isolation in order to allow simultaneous transmission and reception.

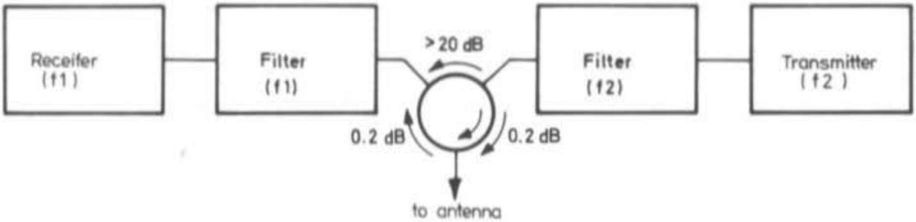

Fig. 2: Application of a three-port circulator as a duplexer

Circulators are also used as coupling links between a signal source and the input of circuits possessing a negative impedance (e. g. tunnel diodes or parametric amplifiers) as well as for passing the amplified signals to the output (Fig. 3). Since the circulator separates the reflected signal from the forward signal and feeds the former to a terminating resistor where it is absorbed, the result is an amplifier with only two ports which operates very constantly even when the source and load impedances vary. In principle this arrangement would also be possible with a three-port circulator, however, without terminating resistor.

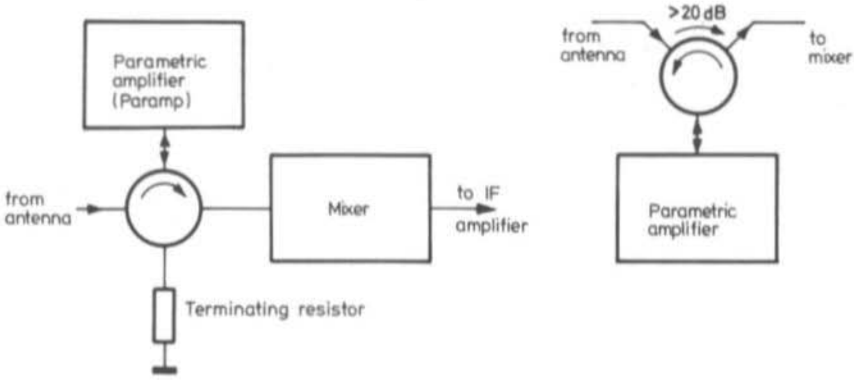

Fig. 3: Parametric amplifiers are provided with a defined input and output impedance by addition of a circulator

A further application for terminated circulators is to decouple two or more transmitters connected to the same antenna in order to reduce intermodulation distortion (Fig. 4). If no such isolation was present, the energy from one transmitter could pass to the other and cause third-order intermodulation products. The terminating resistor need only be dimensioned for the reflected power and can be deleted completely in some cases. The transmitters can either operate at different frequencies within the bandwidth of the circulator or on the same frequency (for doubling the power).

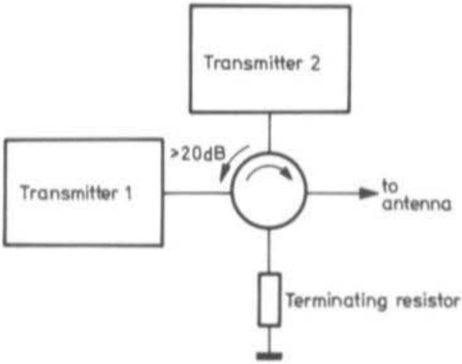

Fig 4: Two transmitters working into a common antenna via a circulator

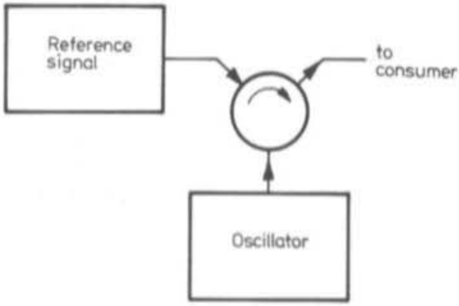

Fig. 5: Phase-locking of an oscillator

Circulators can also be used for phase-locking a diode oscillator to an external signal (Fig. 5). In this case, the oscillator is connected to the first port, the reference signal (at lower power) to the second; the synchronized output signal is then taken from the third port.

The application of an isolator need not be explained in any great detail: it is used where the reaction of a fluctuating load is to be isolated from an oscillator or amplifier. A typical example of this is the pump oscillator (e.g. klystron) of a parametric amplifier which must be kept from the fluctuating impedance of the varactor diode (Fig. 6). However, an isolator will not be required in this position if the power level of the pump oscillator allows a basic attenuation setting of 6 dB or more for the attenuator.

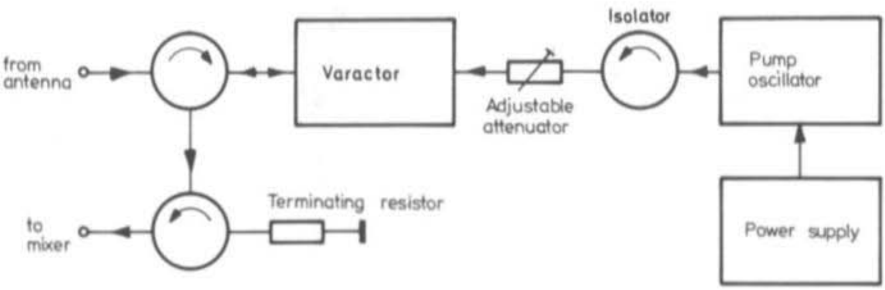

Fig. 6: Block diagram of a parametric amplifier using two three-port circulators and an isolator

3. PRINCIPLE AND CONSTRUCTION

3.1. LUMPED-CONSTANT CIRCULATORS

The construction and basic relationships can be shown in conjunction with lumped-constant circulators usually used for frequencies below approximately 1 GHz (1), (2), (9). This type represents a balanced, non-reciprocal configuration of three inductances that are arranged on a ferrite disc so that the magnetic fields of the three inductances are shifted by 120° to another (Fig. 7). Additional discs for the return path of the RF-field are arranged above and below the main disc. In addition to this, a magnetic DC field exists perpendicular to the surface of the disc for providing magnetic bias up to the vicinity of the ferromagnetic resonance ω_0. This frequency has the following dependence on the magnetic bias H_A:

$$\omega_0 = \gamma \times H_A - 4\pi \times M_S$$

Where: γ = gyromagnetic relationship; a constant that is usually assumed to be 2.8 MHz/Oe (Oersted).

$4\pi M_S$ = magnetic saturation of the associated ferrite material in G (Gauss).

If the ferrite disc were not present, the voltage fed to port 1 would induce equally great voltages into inductances 2 and 3 due to the balanced arrangement. Due to the magnetic disc, the magnetic coupling between the inductances becomes non-reciprocal; this means that the coupling from inductance 1 to inductance 2 is not the same as from inductance 2 to inductance 1. This non-reciprocality causes the attenuation or isolation in the reverse direction.

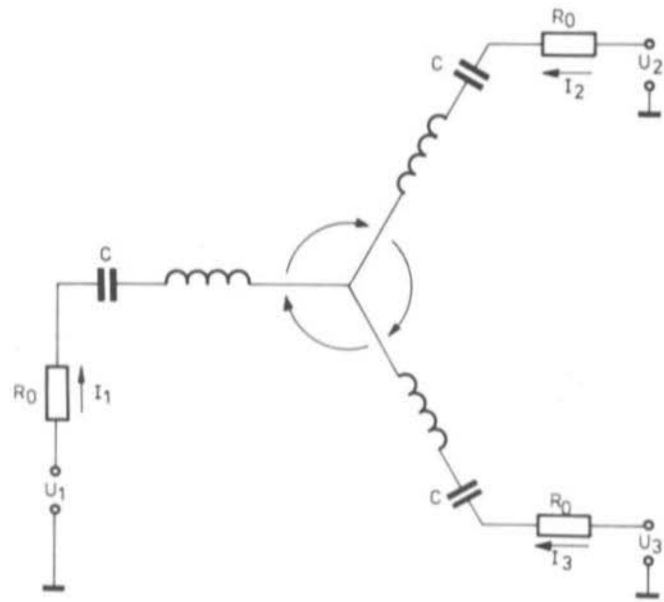

Fig. 7: Equivalent circuit diagram of a three-port circulator with lumped elements. The inductances are brought to resonance with the aid of series capacitors

An analysis of the lumped constant circulator shows that the following conditions must be fulfilled for ideal circulation at any frequency ω:

$$\frac{\omega \times L_o}{2} \quad (\mu^+ - \mu^-) \; = \; \frac{R_o}{\sqrt{3}}$$

$$\frac{\omega \times L_o}{2} \quad (\mu^+ + \mu^-) \; = \; \frac{1}{\omega C}$$

Where: R_o = characteristic impedance
C = capacitance in series with each inductance
L_o = proportional to the inductivity of one of the coils without ferrite
μ^+, μ^- : effective scalar permeabilities of the ferrite material for positive or negative circular polarized fields.

A typical run (1) of the characteristic curves as a function of the magnetic bias is shown in Figure 8.

These equations still do not provide the complete basis for practical construction. For example, the difference between μ^+ and μ^- at a given frequency depends on the degree of magnetic bias in the ferrite material. The value required for L_o is dependent on this. The operating point of the magnetic bias can by no means be selected at will since the bandwidth and also the insertion loss will increase when the magnetic bias approaches resonance. On the other hand, the insertion loss also increases when the operating point is far from resonance.

The equations therefore only guarantee a good isolation at one certain frequency. At other frequencies, the isolation is reduced and the insertion loss increased. The bandwidth over which the isolation is greater than 20 dB - the minimum for practical purposes - is given by the following equation:

$$\frac{\Delta\omega_{20\ \mathrm{dB}}}{\omega} \; = \; 0.2 \times \sqrt{3} \; \left(\frac{\mu^+ - \mu^-}{\mu^+ + \mu^-} \right)$$

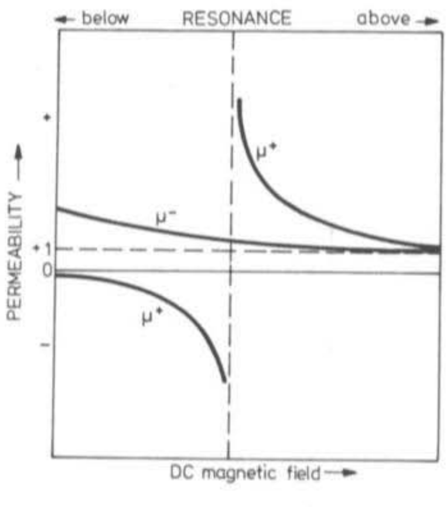

Fig. 8: The magnitude of the magnetic bias determines isolation, insertion loss and bandwidth

The lowest operating frequency in MHz for practical purposes can be expressed as: $f_u \approx 2.8 \times 4\pi M_S$

The magnetic saturation value $4\pi M_S$ for normal materials (3), (4) is between 100 and approximately 5000 Gauss.

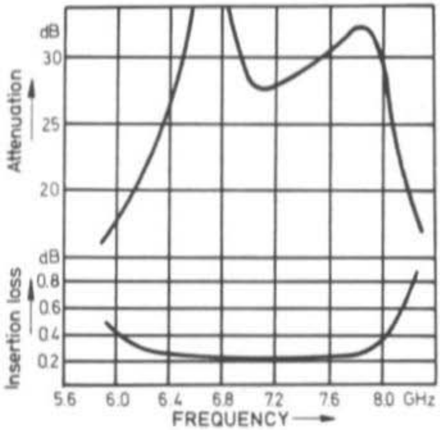

Fig. 9: Extension of the bandwidth of an X-band stripline circulator using double-tuned circuits

It will be seen that a material with a low magnetic saturation is required for VHF applications. A value of less than 100 Gauss is required for operation below 280 MHz. Unfortunately these materials tend to have very poor temperature characteristics.

Circulators with magnetic operating points above the ferromagnetic resonance exhibit lower bandwidths and require considerably greater magnetic bias for higher frequencies.

Good operating points are to be found just below the resonance frequency. The isolation tends to drop here but the insertion loss as well. The final selection of the operating point naturally depends on the special demands of the application in question. Lumped-constant circulators have been constructed and operated at frequencies from UHF down to 35 MHz. The insertion loss is 0.3 dB at 400 MHz, over 0.8 dB at 100 MHz and up to 2 dB at 35 MHz. Typical values are given in Fig. 9 for a double-tuned micro-stripline circulator.

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- to be concluded -

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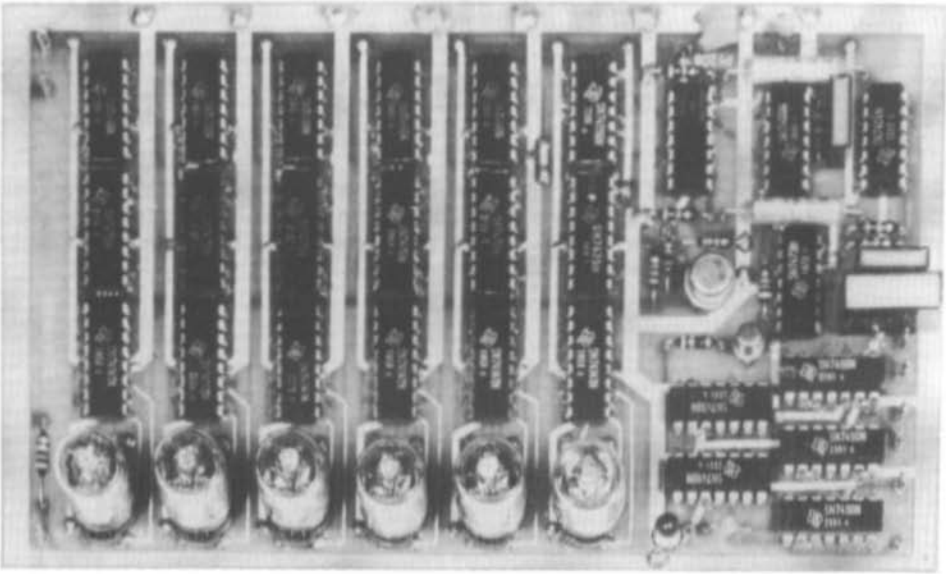

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Crystal	67.3333 MHz	(HC- 6/U)	for 70 cm / 10 m convert.	DM	22.--
Crystal	66.5000 MHz	(HC- 6/U)	for synthesis VFO (DJ 5 HD)	DM	22.--
Crystal	65.7500 MHz	(HC- 6/U)	) for TX + RX con-	DM	22.--
Crystal	65.5000 MHz	(HC- 6/U)	) verters 130 / 130,5 /	DM	22.--
Crystal	65.2500 MHz	(HC- 6/U)	) 131 / 131,5 MHz	DM	22.--
Crystal	65.0000 MHz	(HC- 6/U)	....	DM	22.--
Crystal	64.3333 MHz	(HC- 6/U)	for ATV converter (DJ 5 XA)	DM	22.--
Crystal	62.0000 MHz	(HC- 6/U)	for synthesis VFO (DJ 5 HD)	DM	22.--
Crystal	57.6000 MHz	(HC-25/U)	....	DM	33.50
Crystal	57.6000 MHz	(HC- 6/U)	....	DM	22.--
Crystal	38.9000 MHz	(HC- 6/U)	for DJ 4 LB 001 ATV-TX	DM	25.--
Crystal	38.6667 MHz	(HC- 6/U)	for 2-m-converters ..	DM	17.--
Crystal	1.4400 MHz	(HC- 6/U)	for synthesizer	DM	22.50

STANDARD FREQUENCY CRYSTALS

Crystal	1.0000 MHz	(XS 6002)	....	DM	26.--
Crystal	1.0000 MHz	(XS 0605)	for 75° ovens	DM	50.--

Crystal oven	XT-2 (12 V) 75°C	....	DM	82.--
Ceramic filter	455 D for FM IF-strip	DC 6 HL 007	DM	70.--
Crystal socket	for HC- 6/U	horizontal mounting	DM	5.--
Crystal socket	for HC-25/U	horizontal mounting	DM	5.--
Crystal socket	for HC-25/U	vertical mounting	DM	1.50

Crystals 72..... MHz (HC-25/U) .... DM 33.--

Following frequencies available as long as stock lasts:
72.025 / 72.050 / 72.075 / 72.100 / 72.125 / 72.150 / 72.175 / 72.200 / 72.225 / 72.250 / 72.275 / 72.300 / 72.325 / 72.350 / 72.375 / 72.400 / 72.425 / 72.450 / 72.475 / 72.500 MHz

Sideband crystal	XF-901	8.9985 MHz	DM	15.--
Sideband crystal	XF-902	9.0015 MHz	DM	15.--