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.

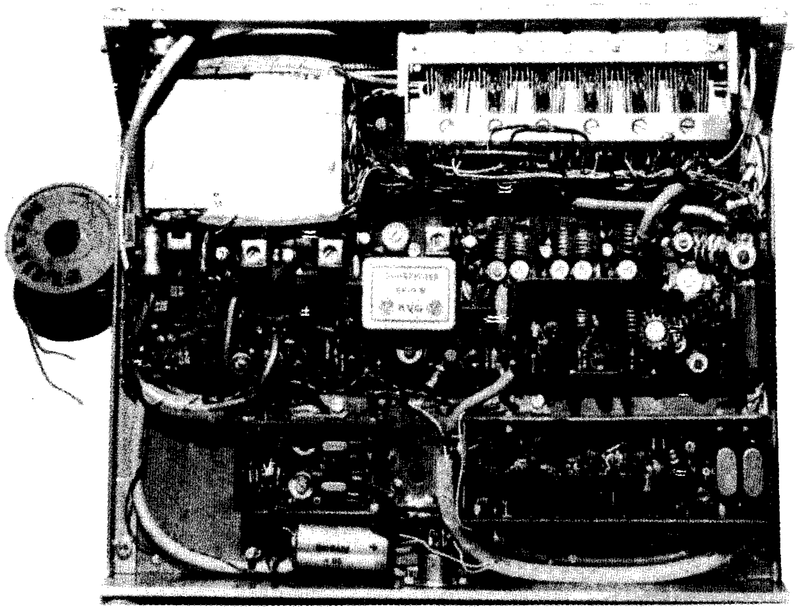


Fig. 1: Internal view of the DC 6 HL SSB-Transceiver

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:	⩾ 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:	⩽ 3 dB
Control range of the whole receiver:	⩾ 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:	⩾ 1.5 W at U_b = 12 V ⩾ 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

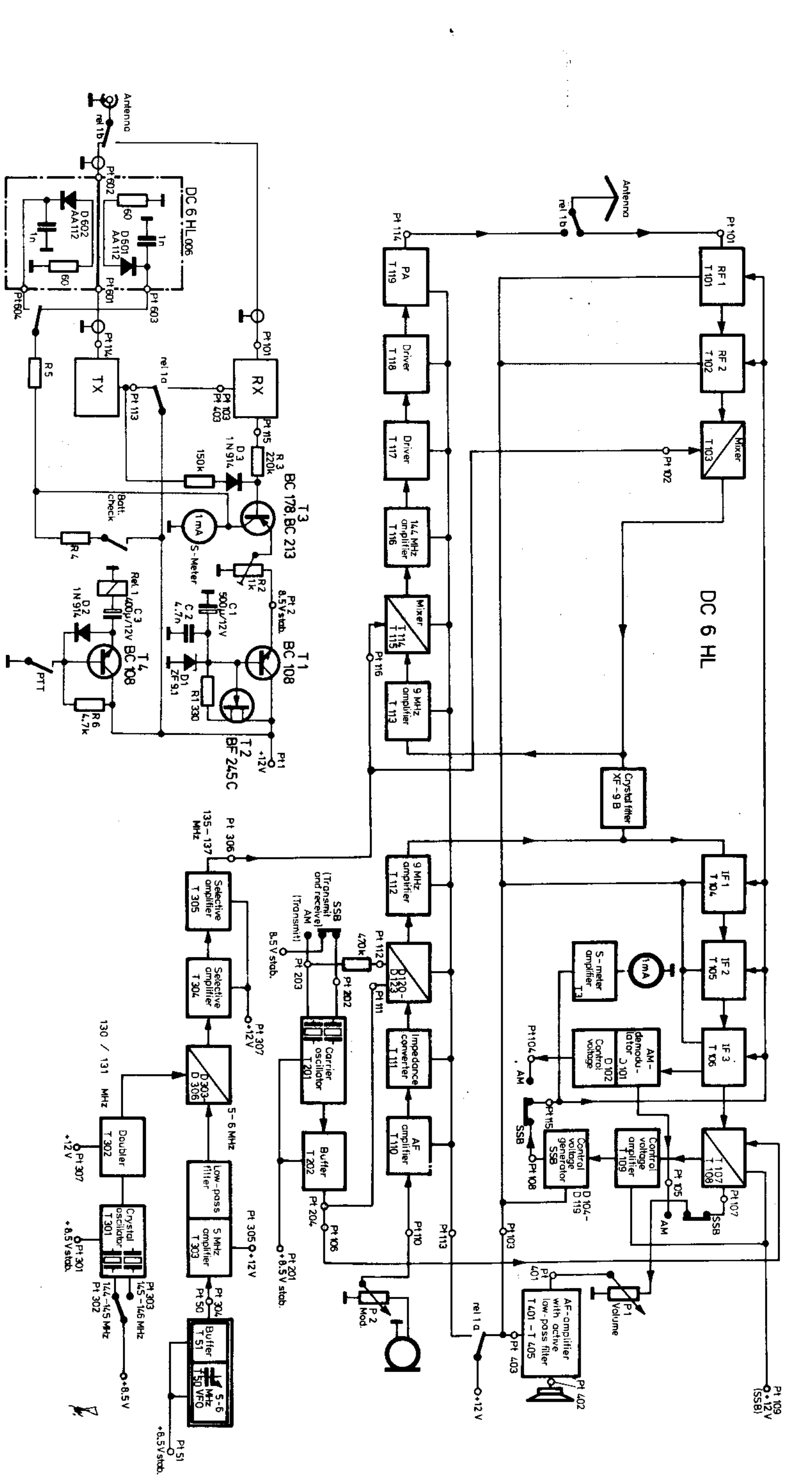


Fig. 2: Block diagram and connection details

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.

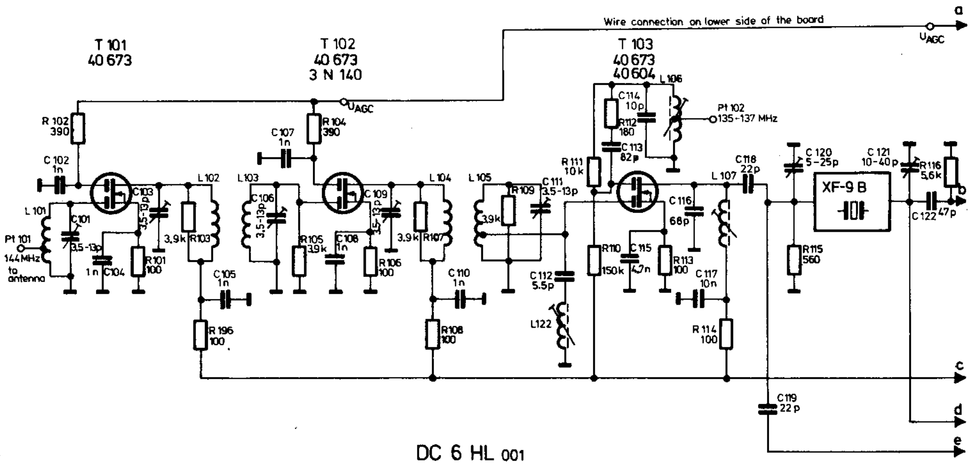
2.1.2. SINGLE SIDEBAND FILTER

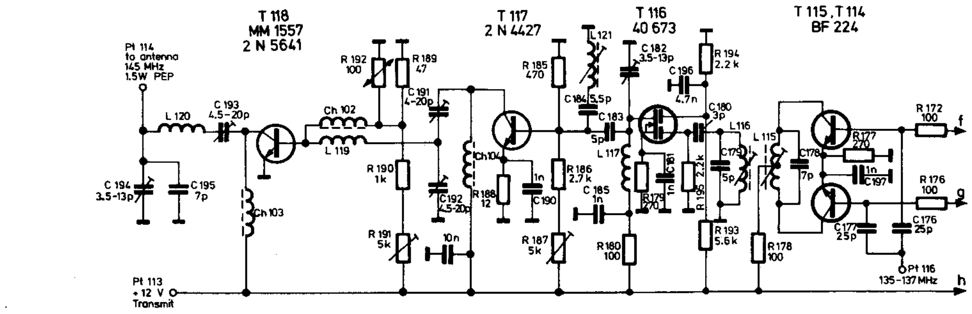
A resonant circuit for the intermediate frequency (C 116/L 107) is connected to the drain of the mixer transistor so that the crystal filter can be loosely coupled simultaneously to the receive mixer and the 9 MHz SSB transmit amplifier (T 113). The crystal filter is terminated, as prescribed by the 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.





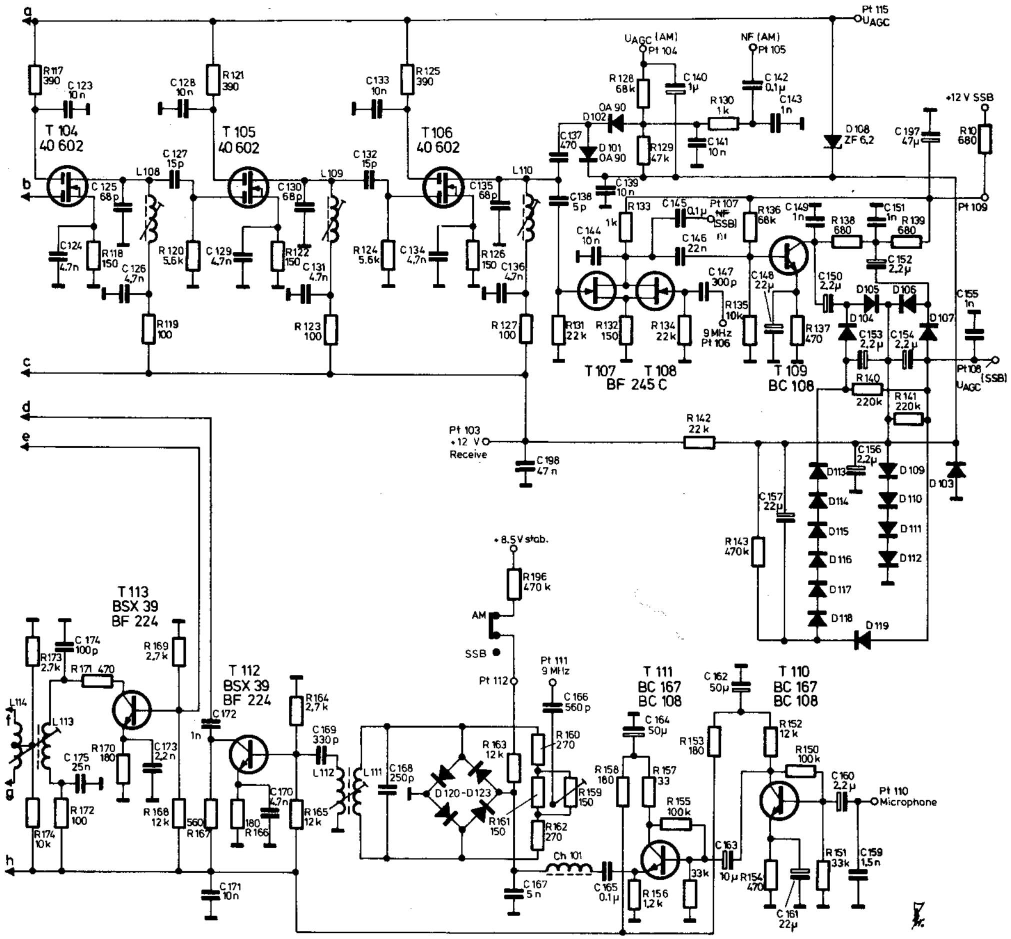


Fig. 3: Circuit diagram of the main board including the main modules of transmitter and receiver

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).

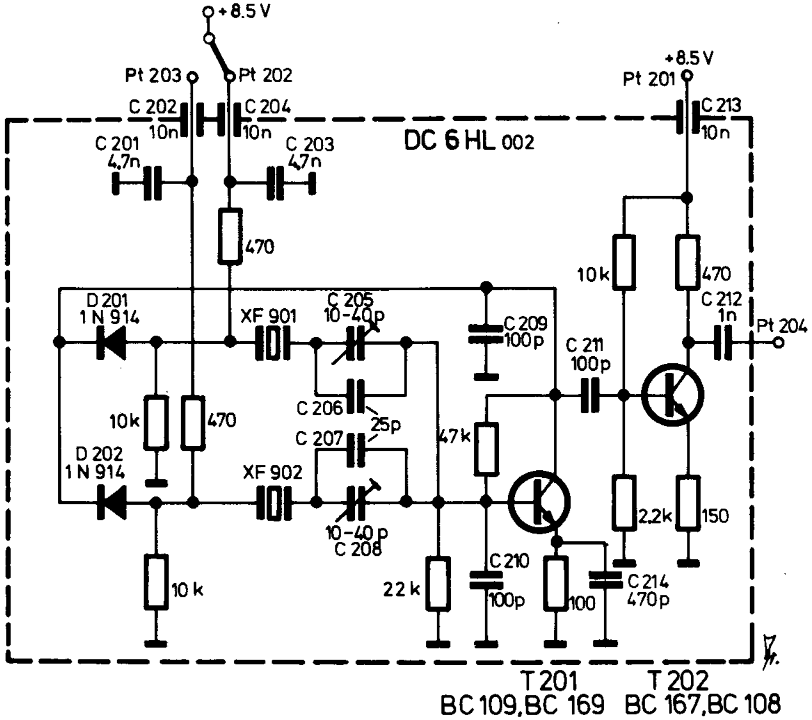


Fig. 4: Circuit diagram of the carrier oscillator

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.

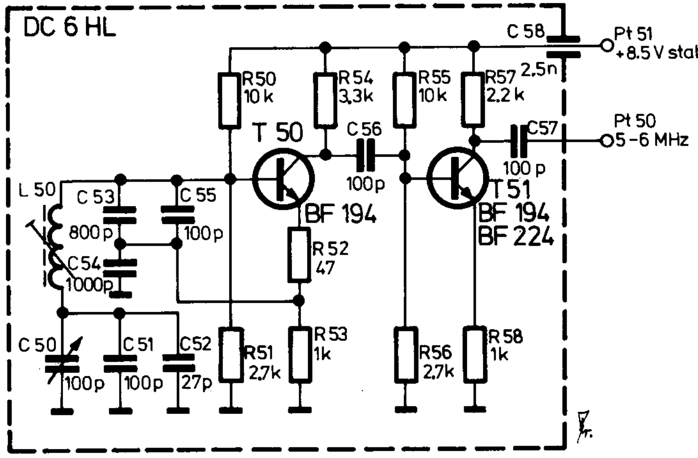


Fig. 5: Circuit diagram of the 5 – 6 MHz VFO

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.

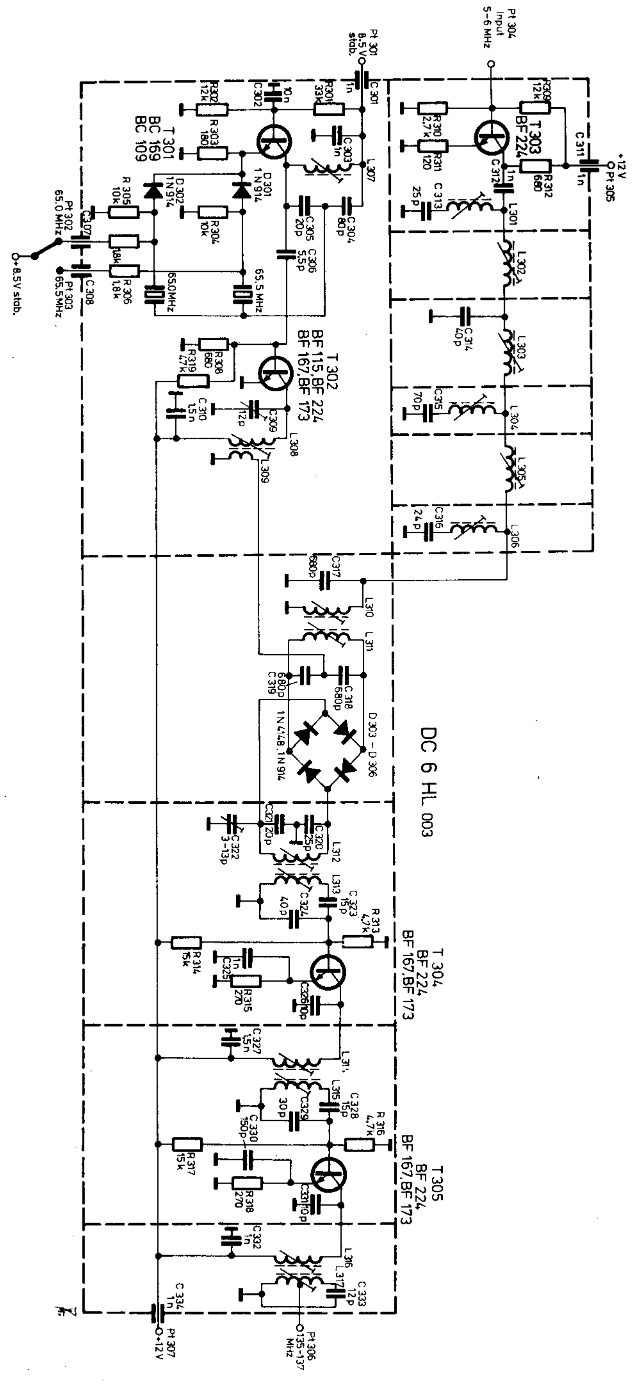


Fig. 6: Circuit diagram of the local oscillator chain

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.

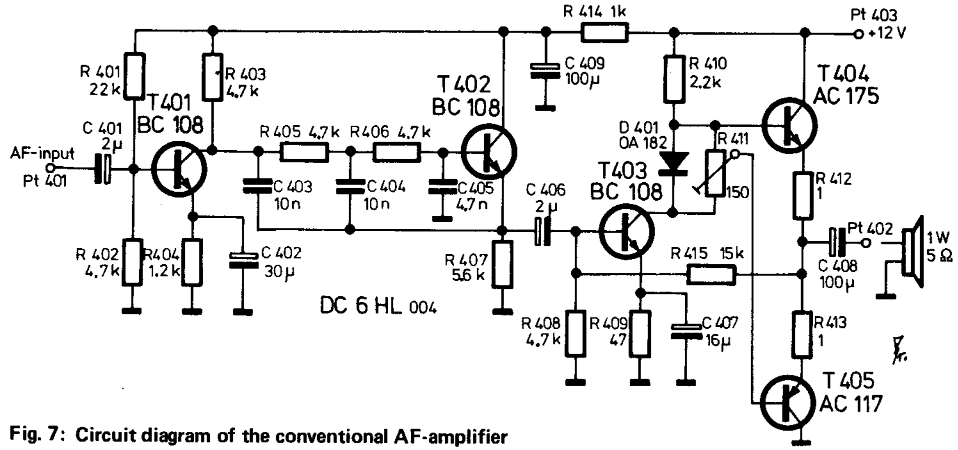


Fig. 7: Circuit diagram of the conventional AF-amplifier

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.

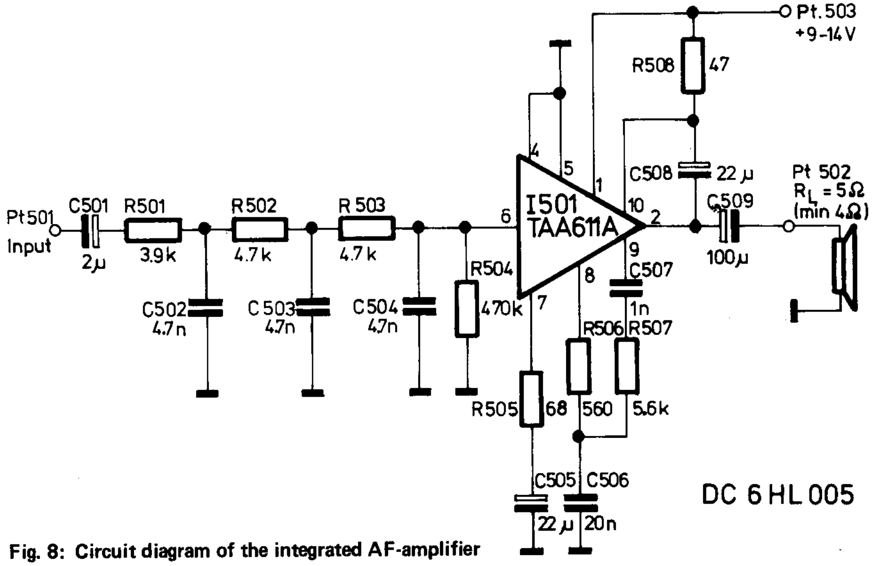


Fig. 8: Circuit diagram of the integrated AF-amplifier

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 simple 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 ϵ_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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