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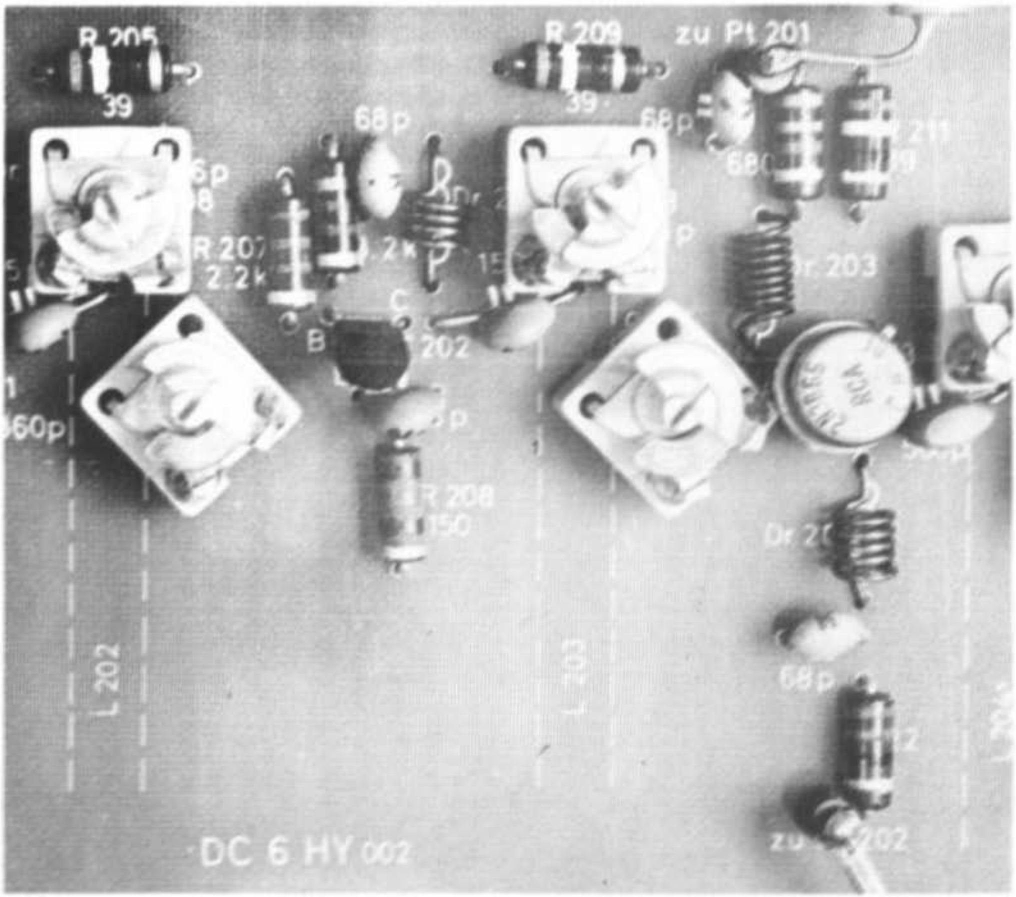

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A SSB TRANSCEIVER WITH SILICON TRANSISTOR COMPLEMENT

Part 4: Power Supply and AF Amplifier

by G. Laufs, DL 6 HA

INTRODUCTION

The fourth part concludes this description of a complete SSB transceiver (1-3). Part 1 described the VHF converter, Part 2 the 9 MHz transceiver, Part 3 the 9 MHz - 14 MHz transmit-receive converter, the 5 MHz VFO with low-pass filter as well as the transmit mixer 14 MHz - 144 MHz with linear amplifier. Part 4 is now to describe a stabilized power supply and a ferrous-free audio amplifier.

The block diagram of the whole transceiver is given in Figure 1.

1. THE STABILIZED POWER SUPPLY

The input Pt 602/Pt 603 of the power supply shown in Figure 2 is connected to a transformer providing an output voltage of 12.6 V to 15 V at approximately 150 mA (providing that the power supply is only to feed the described transceiver).

The rectified but unstabilized voltage is available at output Pt 601 (for instance for the operation of relays). In addition to this a stabilized voltage is available at output Pt 604 whose value can be varied with the aid of potentiometer Pt 601. Finally, output Pt 605 provides a fixed stabilized voltage of 9 V. Since the pass transistor T 604 is fed with the already stabilized voltage from output Pt 604, the voltage at Pt 605 is extremely stable and hum-free. In order to allow this double stabilization to be effective, the single-stabilized voltage at Pt 604 must be adjusted to at least 11 V. Generally, a value of 12 V will be used.

The voltage value at output Pt 605 is determined by the zener diode D 602. The zener voltage must be approximately 1 V above the required output voltage.

The author used the transistor type BSY 86 as pass transistor (U_{CEo} = 64 V/ I_{Cmax} = 1 A). For the higher loaded output Pt 604, two transistors were connected in parallel. Since the dissipation power requires cooling fins in spite of the parallel connection and since no measures have been taken to ensure an

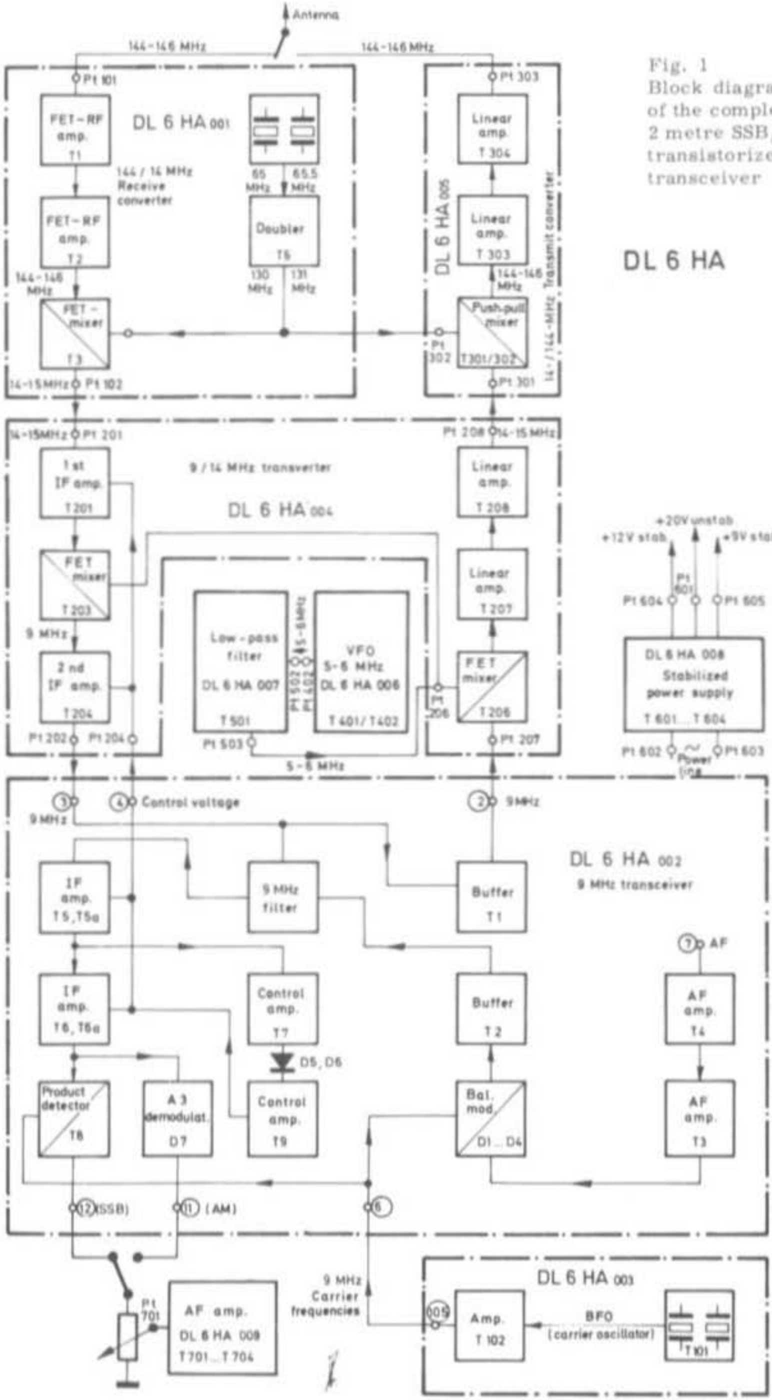


Fig. 1
Block diagram
of the complete
2 metre SSB/AM
transistorized
transceiver

DL 6 HA

equal current distribution onto the two transistors, it is probably more advisable to use one single transistor having a higher power dissipation. Several equivalent types are given in Section 1.2. for this reason. The pass transistor T 604 for output Pt 605 is not loaded to a great degree since the double stabilized voltages is usually only used for the operation of the oscillators.

The transistors used in the power supply exhibit very high transit frequency values. It is therefore possible that they can be driven to incorrect operating points by induced RF voltages. This is the reason why all inputs and outputs are blocked by wideband chokes.

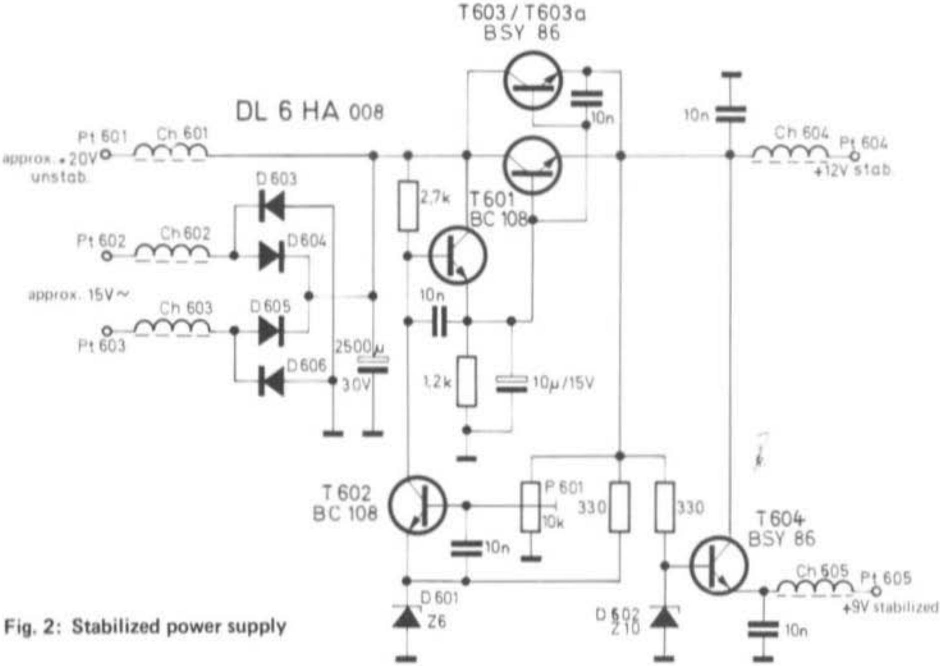


Fig. 2: Stabilized power supply

1.1. ASSEMBLY

With the exception of the power transformer, the power supply is accommodated on the printed circuit board DL 6 HA 008 shown in Figure 3. The corresponding component location plan is given in Figure 4. The dimensions of the printed circuit board are 114 mm by 62 mm. Further details can be seen in the photograph given in Figure 5.

1.2. SPECIAL COMPONENTS FOR DL 6 HA 008

T 601, T 602: 2 N 2926, BC 107, BC 108 for similar AF types
T 603: 2 x BSY 84, 2 x BSY 86 (ITT-Intermetall), 2 x 2 N 2219
2 x 40 347 (RCA) or 1 x 2 N 3054, 2 x 40250 (RCA)
T 604: BSY 84, BSY 86, 2 N 2219, 40 347

D 601: ZF 62 (ITT-Intermetall), BZY 85/C6V2 (AEG-Tfk), 1 N 753
D 602: ZG 10, ZF 10 (ITT-Intermetall), BZY 85/C10 (AEG-Tfk), 1 N 758
D603-D 606: BY 103, 1 N 4002, or similar silicon power rectifiers
Ch 601-Ch 605: ferrite wideband chokes (4312 020 36701 from Philips) or enamelled copper wire passed several times through ferrite beads.

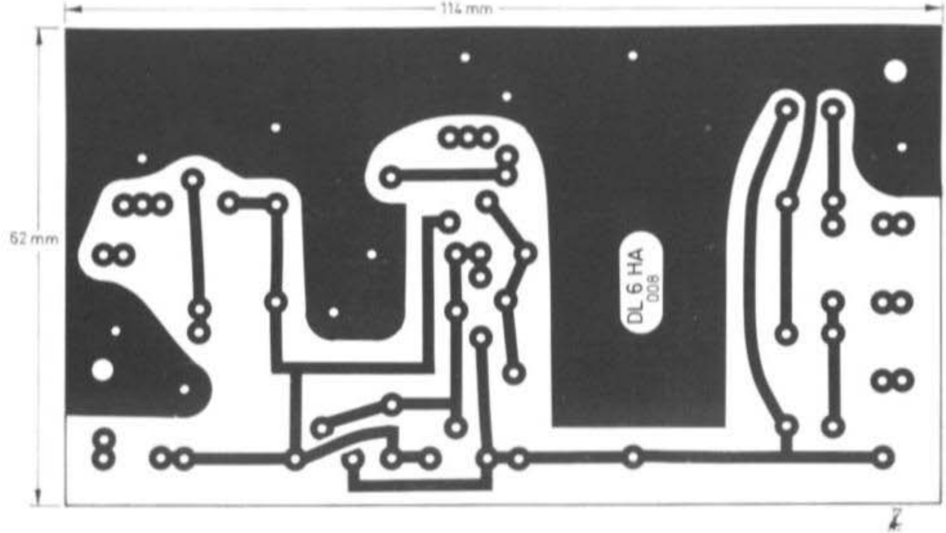


Fig. 3: Printed circuit board DL 6 HA 008

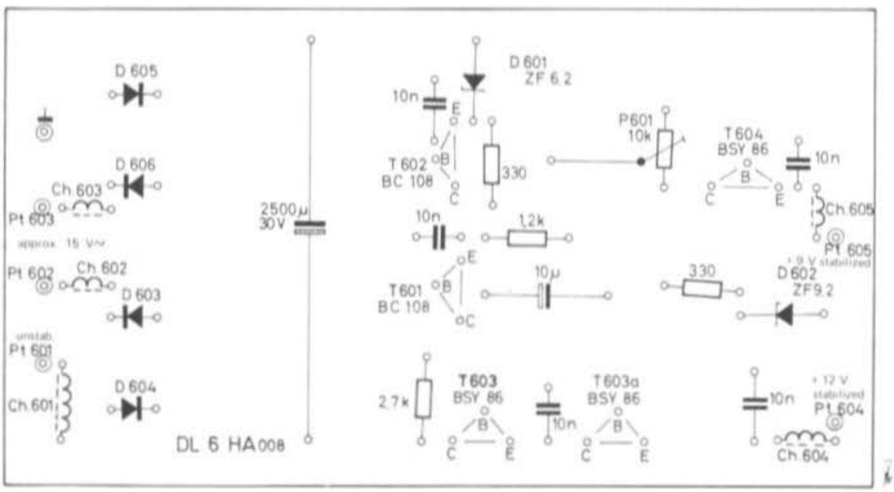


Fig. 4: Component location plan to PC-board DL 6 HA 008

Fig. 5: Photograph of the stabilized power supply

2. AUDIO AMPLIFIER

A transformer-less (ferrous-free) amplifier is used for the AF outputs of the described transceiver. The circuit diagram is given in Figure 6. The amplifier provides an output power of 2 W into a 5 Ω loudspeaker at an operating voltage of 12 V. This AF power is also sufficient during mobile operation in noisy vehicles. The amplifier exhibits a low harmonic distortion factor and a flat frequency response up to over 20 kHz. This means that it is even suitable for music. However, if such applications are planned, the output electrolytic capacitor must be increased from 500 µF to a value of more than 1000 µF (1 mF) in order to improve the bass response. The circuit of the amplifier is dimensioned so that the transistors are not endangered if the loudspeaker should be accidentally shorted.

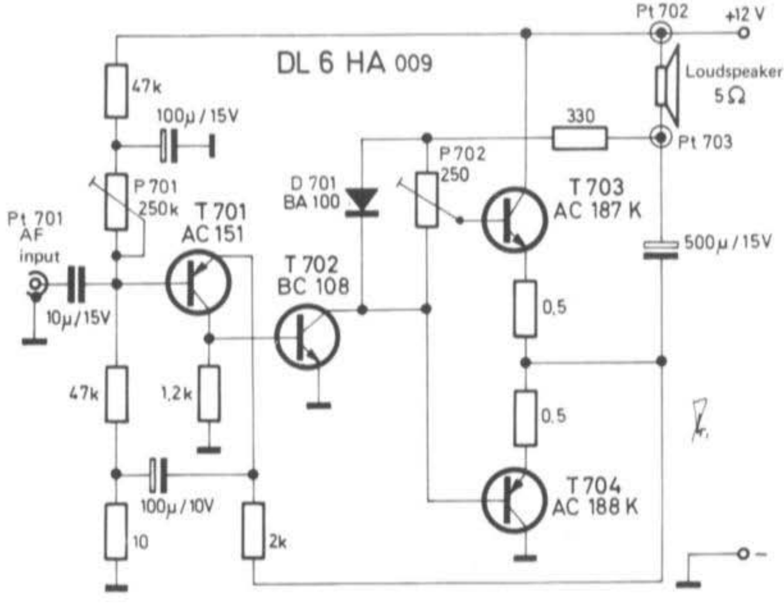


Fig. 6: Circuit diagram of the transformerless AF amplifier

2.1. ASSEMBLY

The complete AF amplifier is accommodated on the printed circuit board DL 6 HA 009. The dimensions of this PC-board are 14 mm by 62 mm. A diagram of the printed circuit board is given in Fig. 7; the corresponding component location plan in Fig. 8. The amplifier is equipped with a 30 mm high heat sink whose location is given in the component location plan. The heat sink is connected, with short wire connections, across the PC-board. The output transistors T 703 and T 704 are screwed onto this heat sink. It should be mentioned that two separate sets of holes are provided for the 2 trimmer potentiometers P 701 and P 702. This allows the use of larger or smaller potentiometer types.

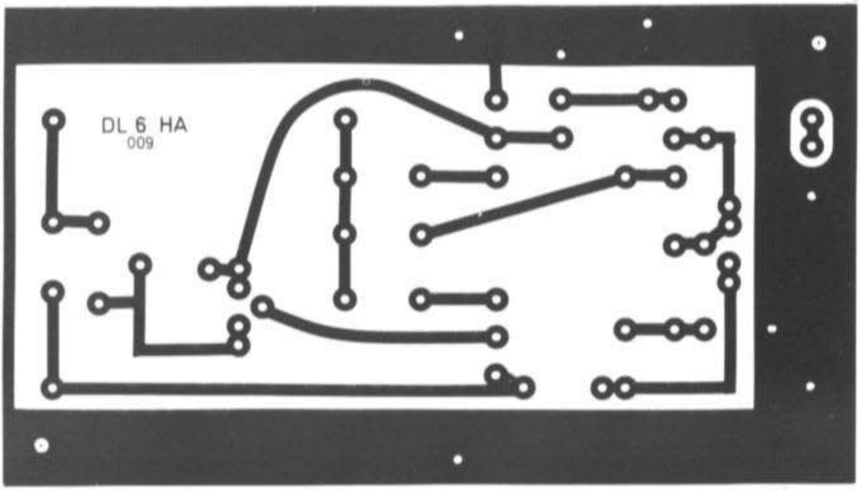


Fig. 7: Printed circuit board DL 6 HA 009

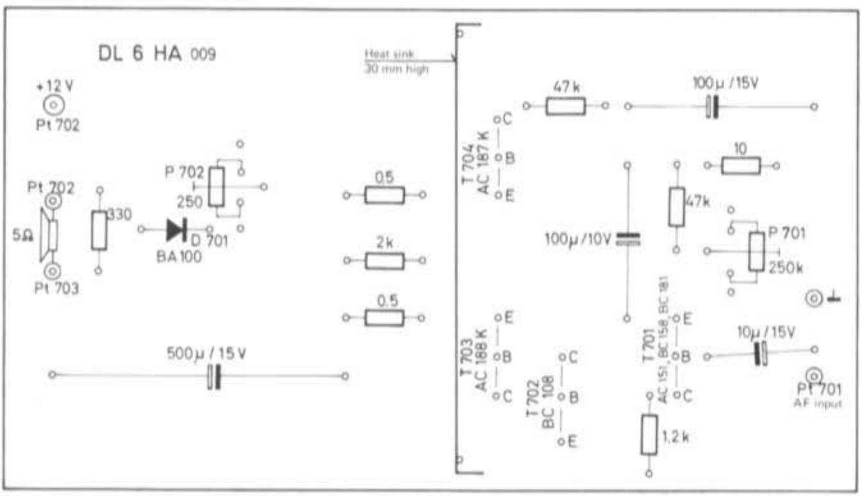
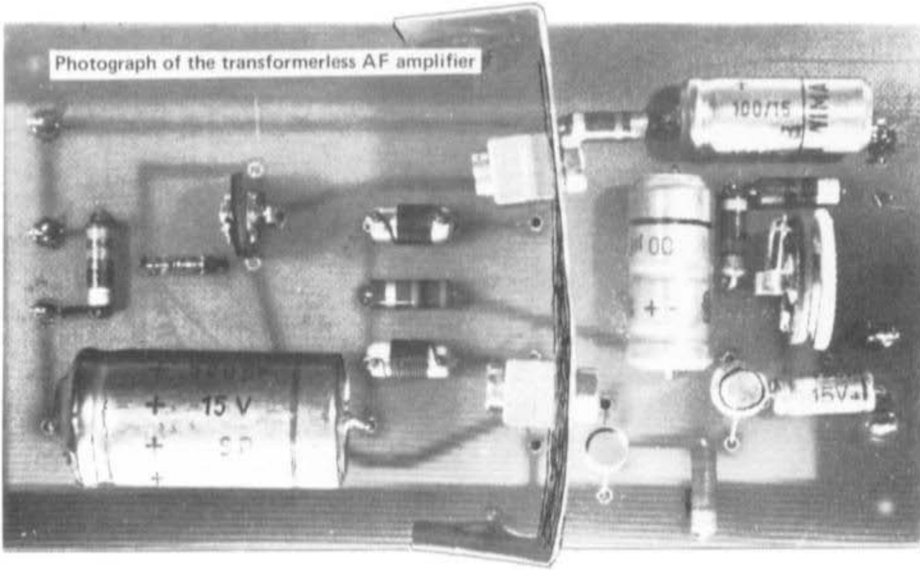


Fig. 8: Component location plan to DL 6 HA 009

2.2. SPECIAL COMPONENTS FOR DL 6 HA 009

T 701: AC 151 (Siemens), AC 122 (AEG-Tfk), BC 158, BC 181,
2 N 2907, 2 N 3906
T 702: 2 N 2926, BC 108, or similar AF transistor types
T 703: AC 187 K (Siemens, AEG-Tfk), AC 175 (AEG-Tfk)
T 704: AC 188 K (Siemens, AEG-Tfk), AC 117 (AEG-Tfk)
D 701: BAY 87 or any silicon diode (such as 1 N 914)

The two emitter resistors of 0.5 Ω can easily be made from two equal lengths of resistance wire. The absolute resistance value is not critical.



2.3. ADJUSTMENT OF THE OPERATING POINTS

In order to adjust the operating point of the AF amplifier, a 5 Ω loudspeaker or terminating resistor is connected to points Pt 702, Pt 703 and a DC-voltmeter connected between the minus pole of the output electrolytic capacitor and ground. The operating voltage of 12 V is now connected and the meter reading reduced to half the value (6 V) with the aid of potentiometer P 701.

After this, it is only necessary to adjust the total current load to a value of 30 mA with the aid of potentiometer P 702. After this, the AF amplifier is ready for operation.

3. GENERAL INSTRUCTIONS

In the converter module DL 6 HA 001, the injection of the voltage from the separate crystal oscillators to the base of transistor T 4 is made with a very short coaxial cable. The local oscillator signal for the push-pull mixer of the transmit converter module DL 6 HA 005 is taken from inductance L 9. To do this, a 100 pF capacitor is soldered onto the inductance approx. 1 turn from the cold end. The intermediate coaxial cable should be as short as possible.

In order to switch off the receive portion in the transmit mode, the plus bar should be broken on printed circuit board DL 6 HA 001 near to the connection point and connected to a transmit-receive switch. The oscillator portion must remain in operation. The voltage supply for this can, for instance, be soldered to the corresponding conductor lane beside resistor R 13. Transistor T 2 will also receive voltage during transmit, but this is not important.

When assembling the whole transceiver, the receive and transmit mixers should be accommodated in separate screened casings. However, it is advisable for these modules to be mounted side by side in order to keep the indermediate cables to the crystal oscillator as short as possible.

A printed circuit board has now been developed to accommodate the two crystal oscillators of the DL 6 HA 001/14 MOSFET converter. This board under the designation DL 3 YK 001 is described in this edition of VHF COMMUNICATIONS.

4. AVAILABLE PARTS

The printed circuit boards of all modules of the SSB transceiver as well as the coil formers, trimmer capacitors etc. are available from the publishers or their national representatives. In addition to this, all modules are available in kit-form. Please see advertising page.

5. REFERENCES

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

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

(3) G. Laufs: A SSB Transceiver with Silicon Transistor Complement
Part 3: The 9 MHz-14 MHz Transmit-Receive Converter, 14 MHz-144 MHz Transmit Converter VFO and Low-Pass Filter
VHF COMMUNICATIONS 2 (1970), Edition 3, Pages 129-146

PRINTED CIRCUIT BOARD FOR THE TWO CRYSTAL OSCILLATORS OF THE 145 MHz-14 MHz MOSFET CONVERTER USED IN THE DL 6 HA SSB TRANSCEIVER

by H. Kahlert, DL 3 YK

1. INTRODUCTION

A double-gate MOSFET converter for the 2-metre band was described in (1) which could be combined to form a complete transceiver (2), (3), (4). Since the frequency synthesizing of the transceiver uses a variable first intermediate frequency of 14 to 15 MHz, two crystal frequencies (130 and 131 MHz) are required to cover the range of 144 to 146 MHz. The printed circuit board of the converter DL 6 HA 001) only allows the provision of one crystal oscillator. The author therefore developed a small additional board onto which the two required oscillators could be accommodated. Fig. 1 shows a photograph of the completed module.

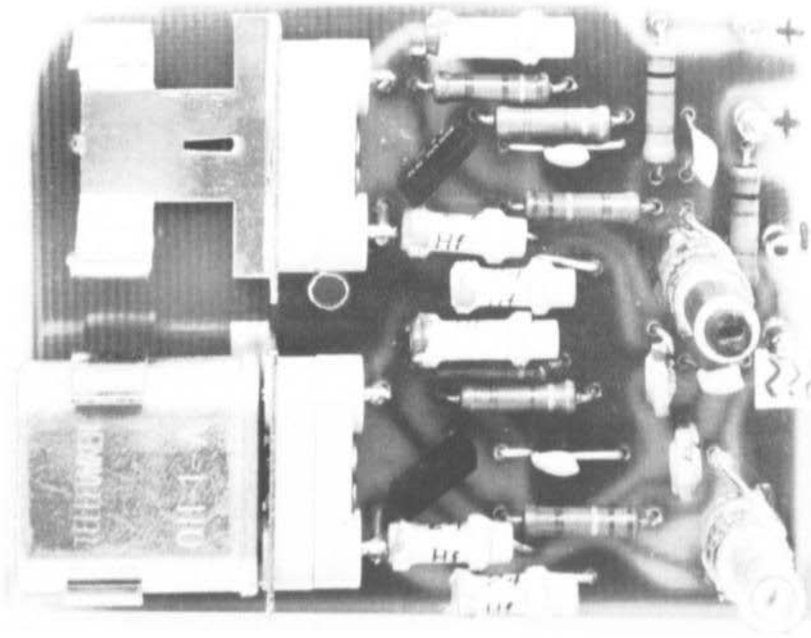


Fig. 1

Fig. 2 shows the circuit of the 144 MHz - 14 MHz receive converter with the two crystal oscillators. The crystals oscillate at half the output frequency (65.0 MHz and 65.5 MHz); the original 38.666 MHz crystal oscillator stage equipped with transistor T 4 serves here as buffer amplifier. If the converter is used with another SSB unit only having 500 kHz bandwidth (14.0 - 14.5 MHz), four crystal frequencies will be required to cover the whole 2-metre band (130.0/130.5/131.0/130.5 MHz). If this is required, this additional board should be provided twice.

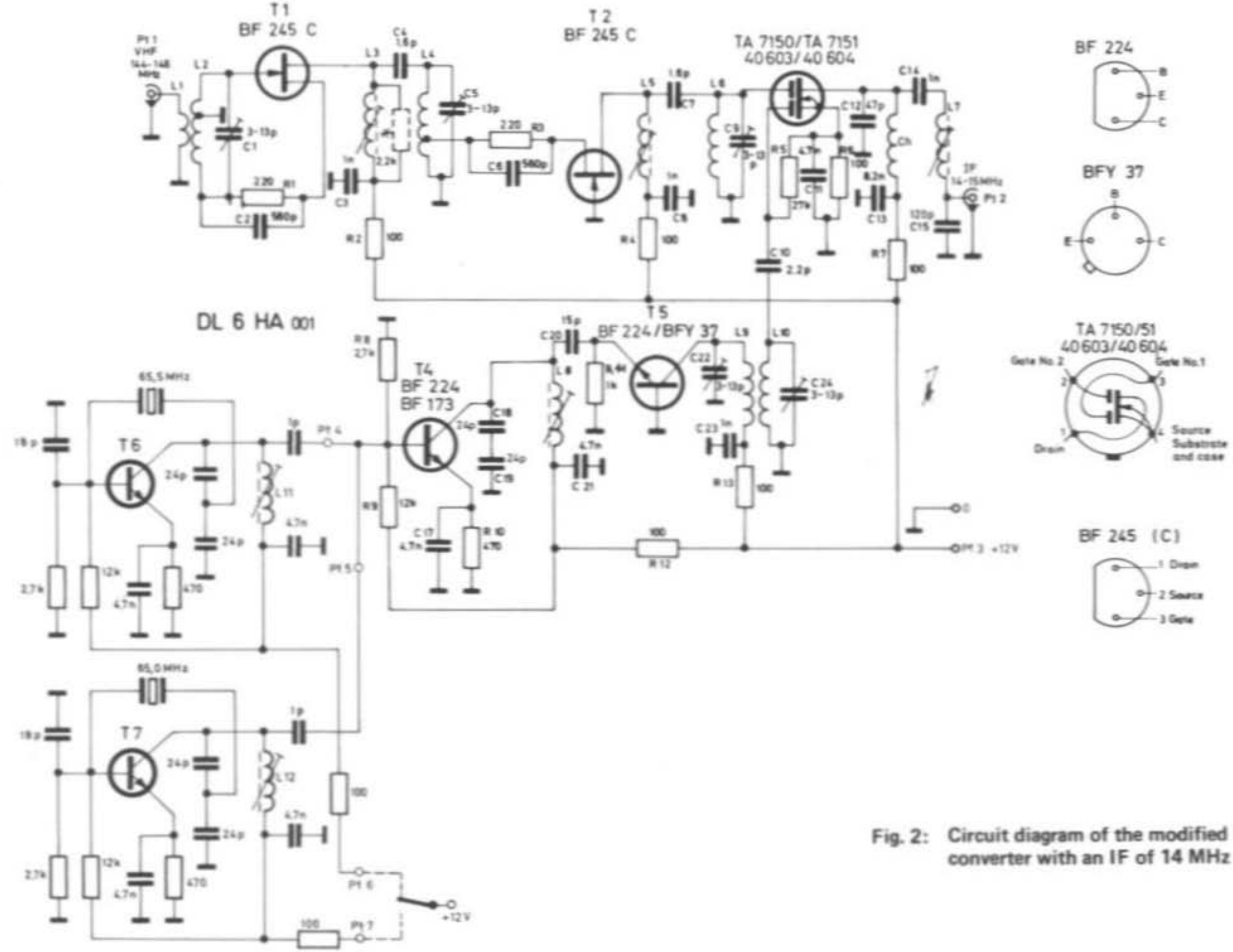


Fig. 2: Circuit diagram of the modified converter with an IF of 14 MHz

2. CONSTRUCTION

The two crystal oscillators, including their crystal holders for horizontal mounting and the two 1 pF coupling capacitors, are accommodated on the additional printed circuit board that has been designated DL 3 YK 001. Fig. 3 shows the 65 mm x 50 mm PC-board; Fig. 4 the corresponding component location plan. The assembly is not critical since sufficient room has been left for the components. Horizontal crystal holders were used in order to decrease the height of the module. As a deviation from the original description, 0.5 mm (24 AWG) diameter silk-covered enamelled copper wire was used for the inductances and not 0.3 mm (29 AWG). This enabled the Q of the inductances to be improved greatly. When fed to an oscilloscope (Tektronix, type 454/15 MHz) the oscillator was found to have a very exact sinusoidal voltage.

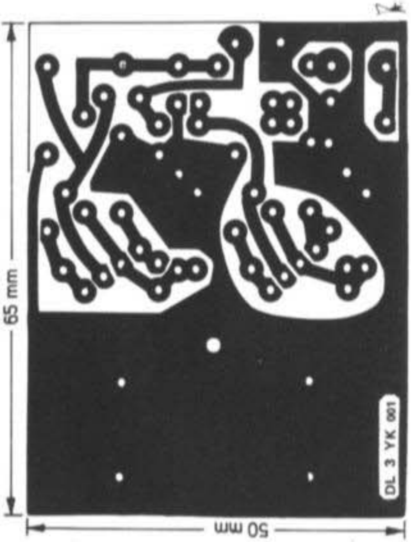


Fig. 3

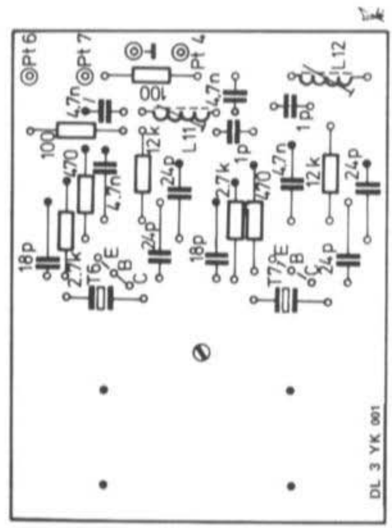


Fig. 4

A great number of transistor types are suitable; attention should only be paid that the transit frequency amounts to at least 150 to 200 MHz (the transistor types 2 N 918 and BF 224 were found to be very suitable). However, the author also used types BSW 59 and BFY 38 with success.

According to the individual transistors, the level of the output voltage is between 500 and 800 mV (peak-to-peak). It should be noted that the frequency stability can be improved by using metal-film resistors. However, carbon resistors are sufficient for the proposed application. It may be useful though to remember this for other applications.

2.1. COIL DATA

L 1: 2 turns of 1 mm dia. (18 AWG) insulated copper wire wound onto the cold end of L 2.

L 2: 6 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound onto a 5 mm former, coil length 10 mm, coil tap 2.5 turns from the gate connection, self-supporting.

L 3: 6.5 turns of 1 mm dia. (18 AWG) of silver-plated copper wire wound onto a 6 mm coil former with core, coil length 15 mm

L 4: as L 2 (coil tap 2.5 turns from the cold end)

L 5: as L 3

L 6: 5 turns, wire and former as for L 2, coil length 10 mm

L 7: 16 turns of 0.5 mm dia. (24 AWG) silk-covered enamelled copper wire on a 4.3 mm former with core, close wound

L 8: 15 turns, otherwise as L 7

L 9, L 10: 7 turns, otherwise as L 2 (without coil tap)

L 11, L 12: as L 8

Ch: 50 turns of 0.1 mm dia. (38 AWG) enamelled copper wire wound on a SW ferrite core of 5 mm diameter or ferrite choke of approx. 50 μH.

3. AVAILABLE COMPONENTS

The printed circuit board DL 3 YK 001 and DL 6 HA 001, semiconductors, trimmers, coil formers and crystals, as well as a complete kit are available from the publishers or their national representatives. See advertising page.

4. REFERENCES

(1) G. Laufs: The 144 MHz Converter with Dual-Gate MOSFET Mixer
VHF COMMUNICATIONS (1970), Edition 1, Pages 1-11

(2) G. Laufs: A SSB-Transceiver with Silicon Transistor Complement
Part 2: The 9 MHz Transceiver
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(3) G. Laufs: A SSB-Transceiver with Silicon Transistor Complement
Part 3: 9 MHz/14 MHz Transmit-Receive Converter
VHF COMMUNICATIONS (1970), Edition 3, Pages 129-146

(4) G. Laufs: A SSB-Transceiver with Silicon Transistor Complement
Part 4: Power Supply and AF Amplifier
VHF COMMUNICATIONS (1970), Edition 4, Pages 193-199

SYNTHESIS VFO FOR 24 MHz

by R. Lentz, DL 3 WR

The author described the transistor transmitter DL 3 WR 003 and matching 24 MHz VFO in (1). Due to the numerous requests for further constructional information regarding this VFO with FM attachment, the author would like to publish a modified printed circuit board.

1. PRINTED CIRCUIT BOARD DL 3 WR 007

The printed circuit board DL 3 WR 007 is shown in Figure 1. The dimensions are 75 mm by 60 mm. The two feedthrough capacitors and the output socket or feedthrough are located in the cutout of this board. Figure 2 shows the corresponding component location plan. With the exception of the actual variable oscillator and FM attachment, all circuitry is accommodated on this board.

Further details as to the components, circuit and alignment, as well as measured values are given in (1).

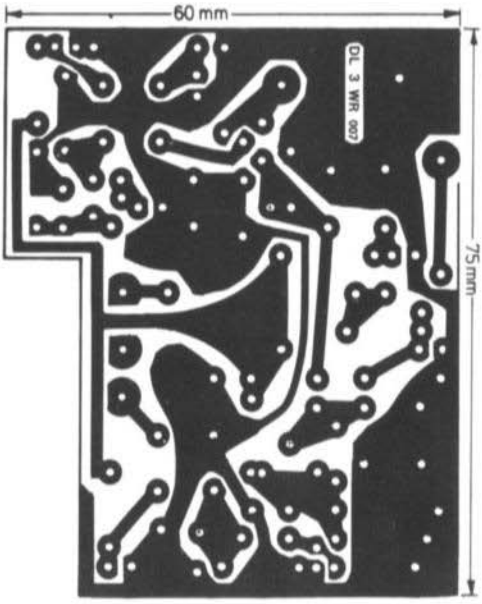


Fig. 1

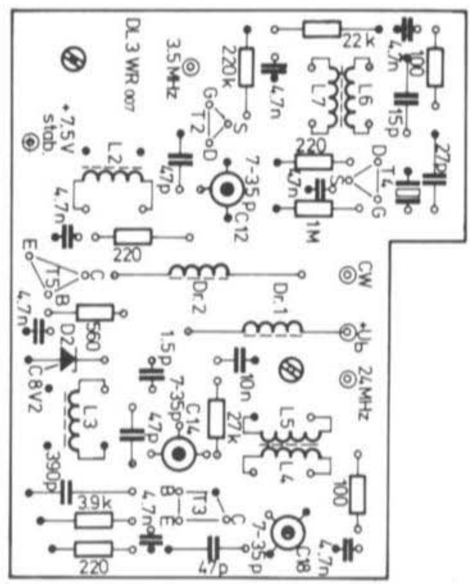


Fig. 2

2. AVAILABLE PARTS

The PC-board DL 3 WR 007, various components and a kit of parts are available from the publishers or their national representatives. see advertising page.

3. REFERENCES

(1) R. Lentz: A Universal VHF-UHF Transmitter for AM and FM
VHF COMMUNICATIONS 2 (1970), Edition 2, Pages 87-102 and
Edition 3, Pages 153-159

A SIMPLE 72 MHz-VFO FOR FM TRANSMITTERS

Further development of a circuit from B. Dietrich, DJ 8 PG

INTRODUCTION

A simple varactor tuned VFO is to be described that oscillates at 72 MHz and is provided with a buffer stage for 72 MHz which can also be used as a frequency doubler to obtain an output frequency of 144-146 MHz. This VFO can be assembled together with the buffer stage on a printed circuit board of only 55 mm by 40 mm. Of course, this oscillator does not satisfy the highest demands with respect to frequency stability. However, since the varactor diode can be simultaneously used for frequency modulation of the oscillator, it is extremely simple to construct NBFM transmitters for 144 MHz, and after redimensioning, for 70 MHz. The VFO could also be used as a simple sweep frequency oscillator for numerous amateur applications.

1. CIRCUIT DESCRIPTION

Fig. 1 shows the circuit diagram of the variable frequency oscillator. The VFO consists of an oscillator and a buffer stage. The buffer is necessary to provide a certain amount of isolation in order to reduce any reaction from the subsequent transmitter onto the actual oscillator. Since the RF voltage is taken from the collector of both stages, the reaction is greatly dependent on the collector-base capacitance of the transistors used. The lower the capacitance, the lower will also be the reaction. The capacitance of the recommended transistors is 0.23 pF for the BF 173 and 0.15 pF for the BF 167. It is possible that the transistor type BF 224 (0.3 pF) could also be used.

Transistor T 1 together with inductance L 1, capacitors C 1, C 2, C 3, C 5, C 6 and C 7 as well as diode D 1 form a clapp-oscillator-circuit. The very high circuit capacitance ensures that the transistor capacitances, which are dependent on the operating point, have practically no effect. Trimmer capacitor C 1 is used to select the centre frequency of the band.

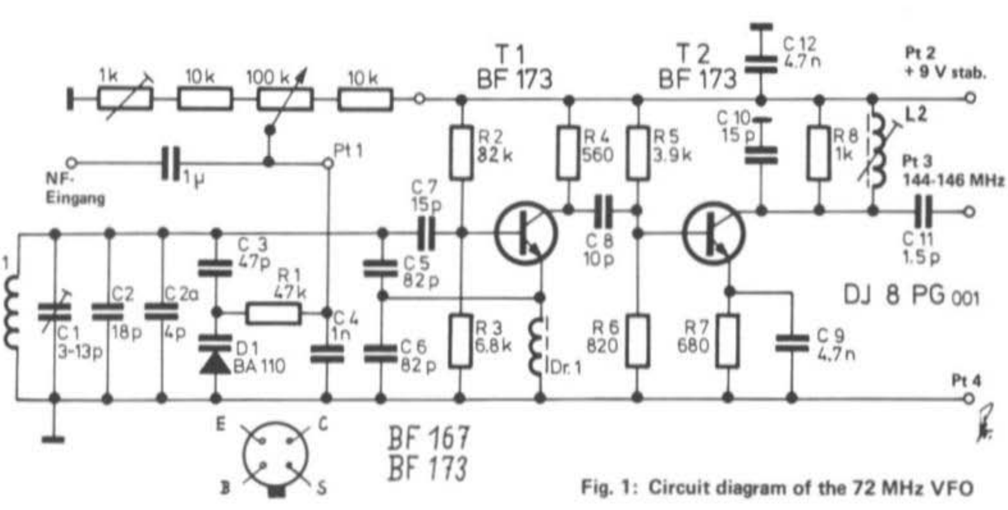


Fig. 1: Circuit diagram of the 72 MHz VFO

The actual tuning is made by varactor diode D 1. The tuning voltage is varied with the aid of a 100 kΩ potentiometer having a logarithmic (+ log) characteristic so as to obtain a virtually linear frequency scale. In order to ensure smooth frequency tuning, a potentiometer type should be used that possesses a relatively long resistor lane, e.g. a higher power type. Miniature potentiometers and wire-wound potentiometers cannot be used. A second potentiometer having a value of about 1 kΩ is used for the fine frequency adjustment. Resistor R 1 and capacitor C 4 represent a filter link to ensure that no RF voltages are induced into the tuning voltage supply of the varactor diode.

The audio frequency voltage is fed to the diode via a 1 μF isolating capacitor. The two 10 kΩ resistors are used to ensure that the AF voltage is not shorted by the voltage supply in the extreme positions of the tuning potentiometer. In addition to this, these resistors ensure that the DC voltage fed to the varactor diode does not fall below approx. 0.8 V.

The circuit has been dimensioned so that a voltage variation of 7.5 V allows a frequency range of 72 to 73 MHz to be tuned with a slight overlap at both ends. After frequency doubling, a frequency range of 144 to 146 MHz is obtained. This means that a voltage variation of 7.5 mV across the diode results in a frequency variation of 2 kHz.

The frequency variation of 2 kHz at a voltage variation of only 7.5 mV, however, shows that the tuning voltage for the diode must be very well stabilized. The operating voltage of the complete VFO should be stabilized using a zener diode (or even better using a zener diode and pass transistor) and the same voltage used for tuning.

However, the tuning voltage can be individually stabilized and fed to the tuning voltage divider. For the VFO itself, a simple zener diode stabilization is sufficient. If higher demands are to be placed on the frequency stability, the temperature-compensated stabilizing circuit given in (1) is advisable, however, the tuning voltage divider must be of high impedance. Of course this amount of work is only worth while if the other frequency determining components of the VFO are also temperature compensated. Enclosing the VFO in styrene foam and a thick metal casing will compensate for short-term stabilization. However, it is doubted whether long-term stability can be achieved using amateur means of temperature compensation mainly due to the fact that the temperature coefficient of the varactor diode alters as a function of the tuning voltage. However, this VFO is not designed for applications where a high frequency stability is required but more for FM applications where a simple oscillator is sufficient.

The collector resonant circuit of the buffer stage (inductance L 2) is tuned to 72 MHz and damped by resistorR 8 in order to obtain the required bandwidth. When using the described high-impedance coupling, the capacitance of the (short) cable to the transmitter must also be considered during the tuning process. It may be necessary for the circuit capacitance to be reduced. However, it is possible for a coupling link of one turn to be wound between the turns of inductance L 2. The RF voltage is, however, somewhat lower than in the described case where a voltage of 2 V was available. Of course, it is possible to align the output resonant circuit to 145 MHz. To do this, it is necessary to reduce the circuit capacitance; the output voltage at the directly coupled high impedance output then amounts to approx. 1.4 V (RMS).

The described VFO can be directly modulated using a dynamic microphone. If high-level (high Z) microphones are to be used, the matching can be made using a voltage divider consisting of two fixed resistors. A clipper is to be described (2) that is very suitable for operation with this VFO.

2. ASSEMBLY

The variable frequency oscillator is built up on the relatively small circuit board DJ 8 PG 001 having the dimensions 55 mm x 40 mm. Fig. 2 shows an enlarged photograph of the VFO.

Fig. 3 shows the conductor side of the printed circuit board; Fig. 4 the corresponding component location plan. As can be seen, the voltage divider for the varactor and the stabilized power supply are not accommodated on the small PC board; they should be connected to the corresponding connection points.

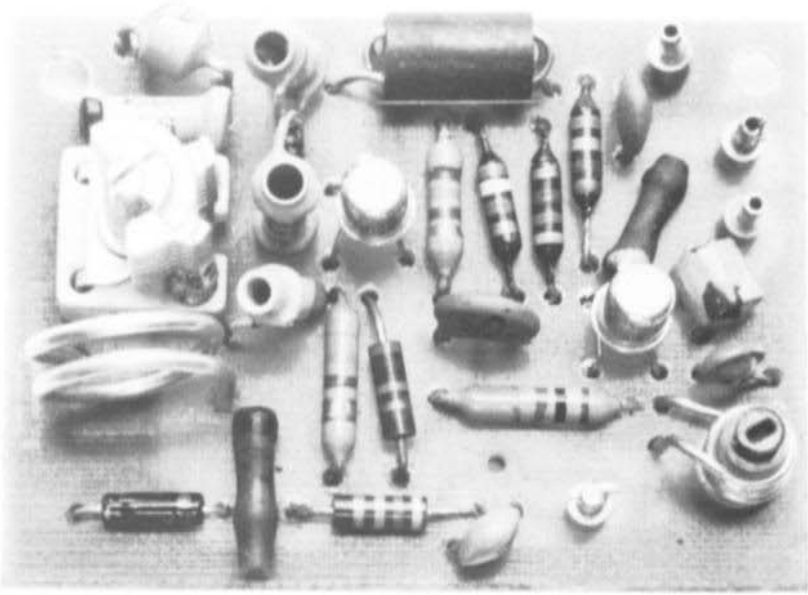


Fig. 2: Photograph of the VFO

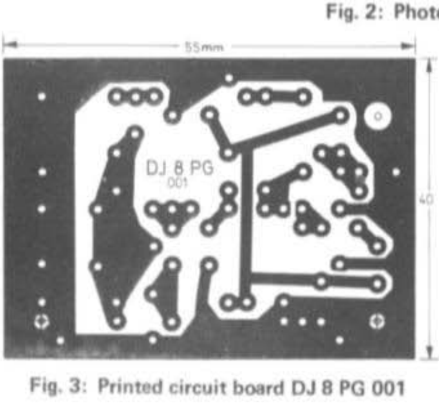


Fig. 3: Printed circuit board DJ 8 PG 001

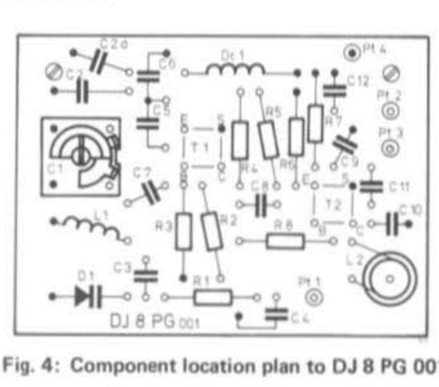


Fig. 4: Component location plan to DJ 8 PG 001

3. SPECIAL COMPONENTS

T 1, T 2: BF 167, BF 173 (Valvo, AEG-Tfk)

D 1: BA 110 (ITT-Intermetall), BA 121 or BA 149/8 (AEG-Tfk), 1 N 5462 A (Motorola)

L 1: 2 turns of 1.3 mm diameter (16 AWG) silver-plated copper wire wound on a 10 mm former, self-supporting, with short connections.

L 2: 4 turns of 0.8 mm diameter (20 AWG) silver-plated copper wire wound on a 4 mm diameter coil former with VHF core.

Ch 1: Ferroxcube wideband choke (Philips 4312 020 36701)

C 1: 2 - 13 pF air spaced trimmer

4. AVAILABLE PARTS

The printed circuit board DJ 8 PG 001, trimmer capacitor, coil former, ferroxcube choke and the semiconductors as well as a kit of parts are available from the publishers or their national representatives. Please see advertising page.

5. REFERENCES

(1) H. J. Franke: Stable Reference Voltages
VHF COMMUNICATIONS 2 (1970), Edition 2, Pages 76-86

(2) D. E. Schmitzer: Speech Processing
VHF COMMUNICATIONS 2 (1970), Edition 4 and vol. 3 (1971), Edition 1

STEEP-SKIRTED ACTIVE AUDIO FILTERS

by D. E. Schmitzer, DJ 4 BG

INTRODUCTION

In the descriptions (1) and (2), the author described active audio filters that were built up from resistors, capacitors and transistors in a simple manner on a printed circuit board (DJ 4 BG 001). Active low-pass, speech-weighted and bandpass filters for voice and telegraphy were described for various applications in transmitters and receivers, as well as the practical assembly of same. The author carried out further experiments based on these filter circuits in order to achieve a steeper transposition from the passband to the stopband range. Such steep filter skirts are required, for instance, for a low-pass filter following an amplitude limiter. Several circuits are now to be described which can be combined to form a filter whose cut-off frequency is 2.9 kHz, and that already exhibits an attenuation pole of approx. 80 dB at 6.4 kHz.

2. COMBINATION OF SEVERAL FILTERS HAVING DIFFERENT ATTENUATION CURVES

As was shown in several examples in (2), steep-skirted filters can be achieved by series-connection of two or more basic low-pass filters. In order to obtain a fast transition from the passband into the stopband range, a certain amount of overshoot must be tolerated. However, the overshoot can add itself to non-permissible high values when connecting a number of filters in series. However, a very steep sum curve together with a permissible overshoot can be achieved by combining a filter having a large overshoot and a filter with a more gradual transition. Figure 1 shows a circuit developed according to this principle. The frequency response curve of this circuit is given in Figure 2.

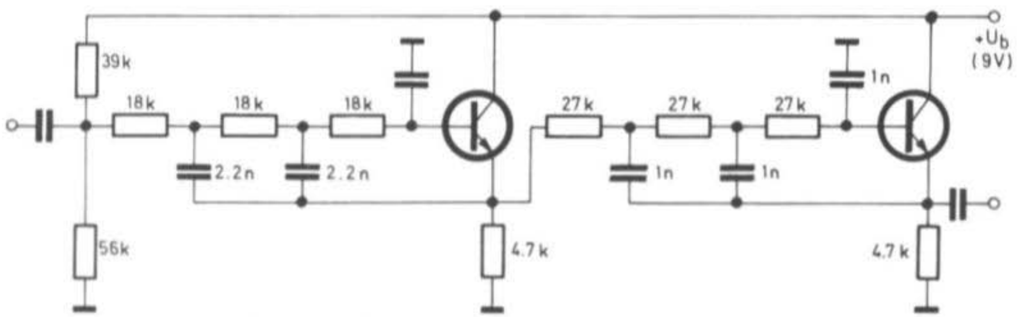


Fig. 1: Steep-skirted low-pass filter made up from separate individual filters

3. COMBINATION OF A RC-LINK AND AN ACTIVE FILTER

A steep attenuation curve together with a low overshoot can be achieved by combining a basic low-pass filter link having a large overshoot and a simple RC-link having a correspondingly low cut-off frequency. Frequency response curves result with the correct dimensioning that already exhibit a certain ripple in the passband range and have a steep transition into the stopband range.

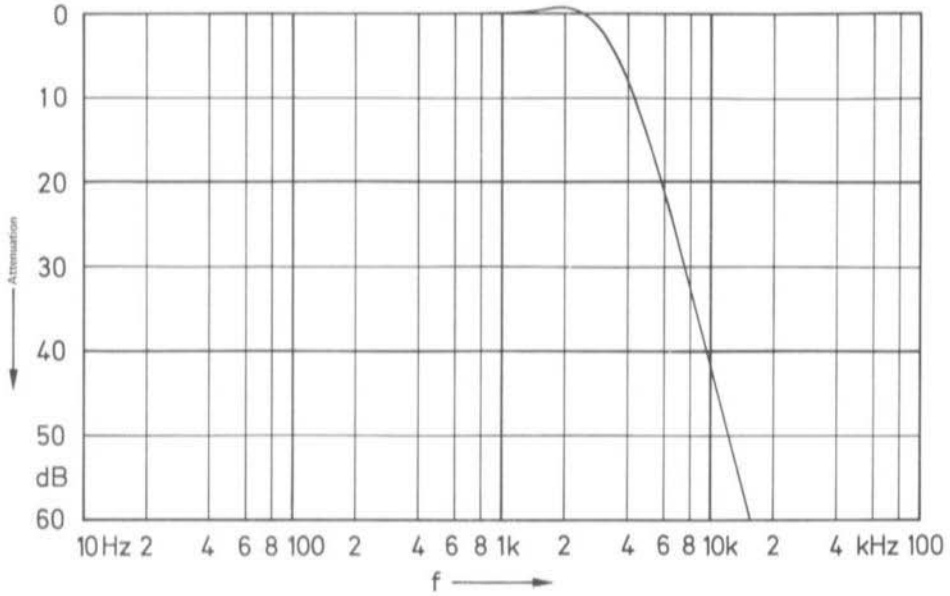


Fig. 2: Frequency response of the circuit given in Fig. 1

In addition to this, the pulse behaviour of this arrangement, e.g. the distortion of a squarewave pulse due to the transient behaviour, is very favourable. Chebyshev-filters exhibit a very similar behaviour, which means that this arrangement can be assumed to be an active RC-Chebyshev-filter. A diagram of such a circuit is given in Fig. 3 and the frequency response in Fig. 4.

The oscilloscope trace given in Fig. 5a illustrates how the original squarewave pulse appears after passing through the described RC-Chebyshev-filter. As a comparison, the output pulse of a simple, active low-pass filter as given in Fig. 1a in (2) is given in Fig. 5b. It can be seen that the simple basic low-pass link generates a large overshoot (Fig. 5b). The active RC-Chebyshev-filter on the other hand has a somewhat longer "ringing" but the resulting peak does not protrude so far from the crest of the squarewave pulse as is the case with the basic low-pass filter circuit. This is very important for some applications.

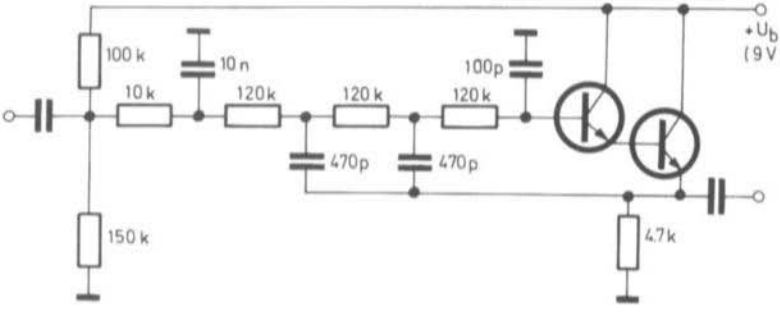


Fig. 3: Low-pass filter with preceeding RC-link

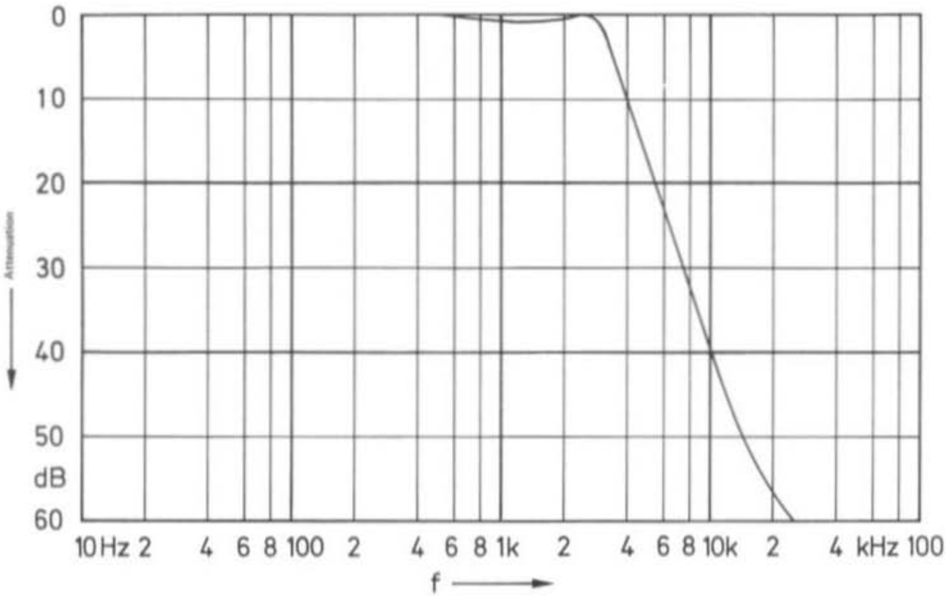


Fig. 4: Frequency response of the circuit given in Fig. 3

4. AN ACTIVE FILTER WITH ATTENUATION POLE

Since three RC-links are included in the basic low-pass circuit, the output voltage will exhibit a phaseshift of 180° with respect to the input signal at a certain position in the attenuation range, that is at a certain frequency. The anti-phase condition at this frequency will cause a more or less complete attenuation of the signal. This fact means that an attenuation pole will result in the frequency response whose amplitude is dependent on the relationship of the output voltage to input voltage.

This attenuation pole can be placed in the vicinity of the cut-off frequency, which results in a very steep attenuation skirt. A disadvantage for some applications is the fact that the attenuation of frequencies above the attenuation pole decreases again since the phaseshift of 180° only occurs for one certain frequency.

Such a circuit is given in Fig. 6, the resulting frequency response curves are given in Fig. 7. The parameter for the various curves is the value of the feedback resistor R_f. In order to obtain the required dimensioning of such a filter, the required value of R_f to obtain a certain, required frequency of the attenuation pole f_p can be taken from Fig. 8.

Due to the relatively high-impedance feedback path, it is important that a high impedance stage is connected to the output of this active filter. A common-collector configuration (emitter follower) or a field effect transistor stage are suitable.

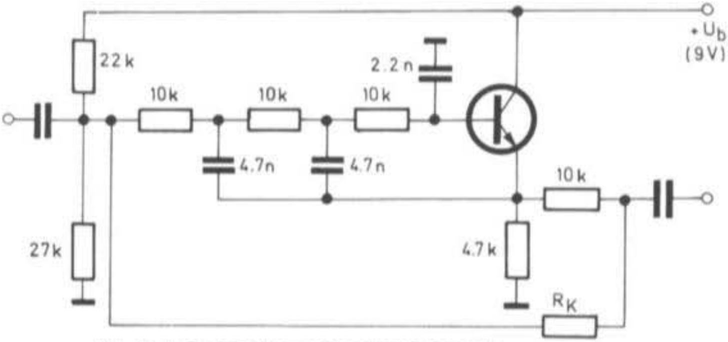
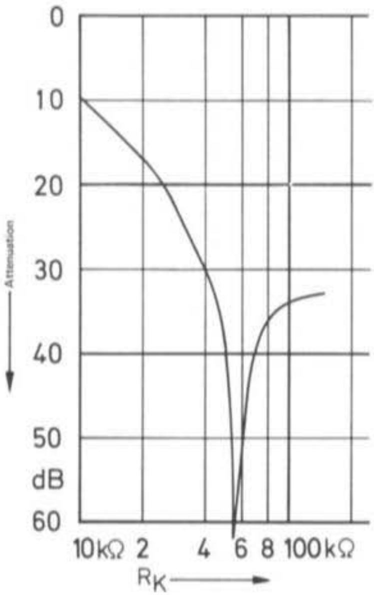


Fig. 6: Low-pass filter with attenuation pole



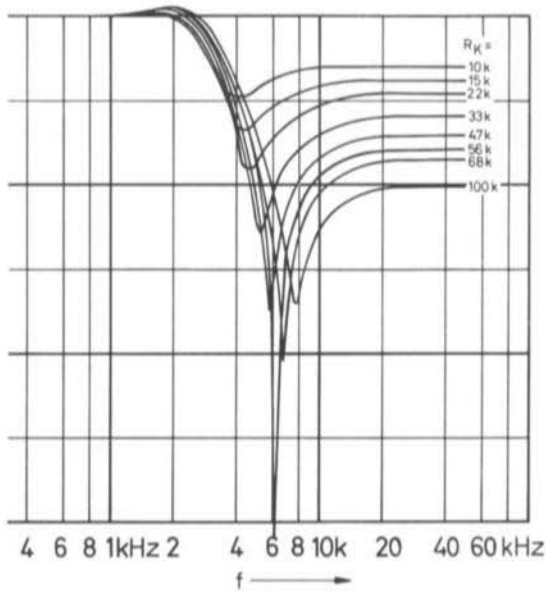


Fig. 7 a: Value of the attenuation pole as a function of Rf

Fig. 7 b: Frequency response of the circuit given in Fig. 6 according to the value of Rf

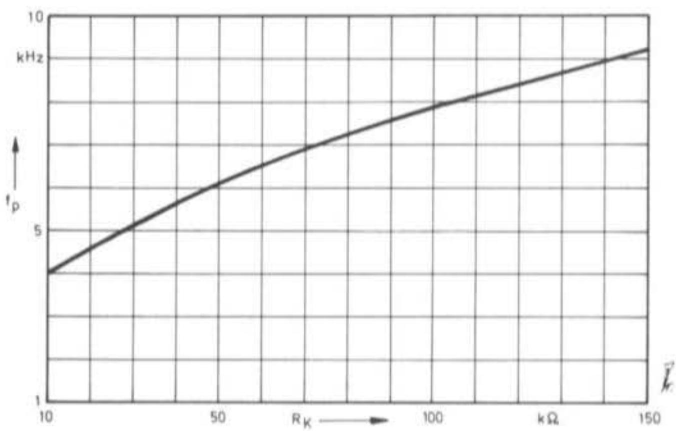


Fig. 8: Frequency of the attenuation pole as a function of Rf

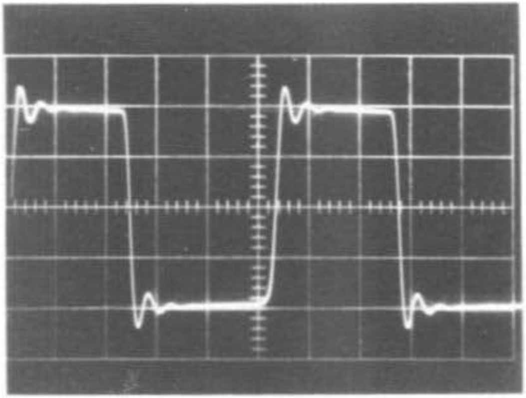


Fig. 5 a: Pulse behavior of the circuit given in Fig. 3

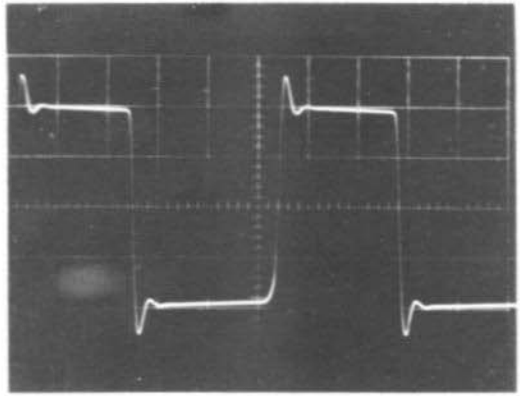


Fig. 5 b: Pulse behavior of the low-pass filter according to Fig. 1 a in (2)

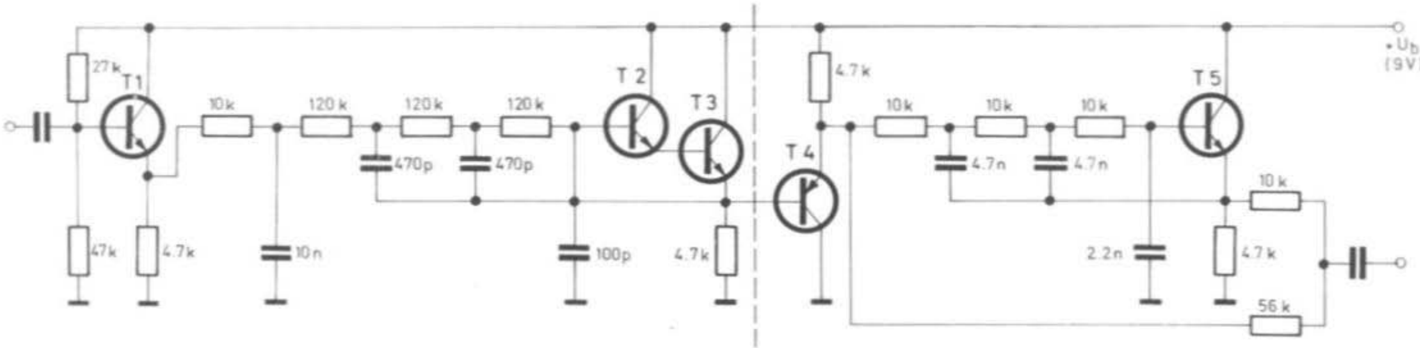


Fig. 9: Combined filter to obtain a steep skirt

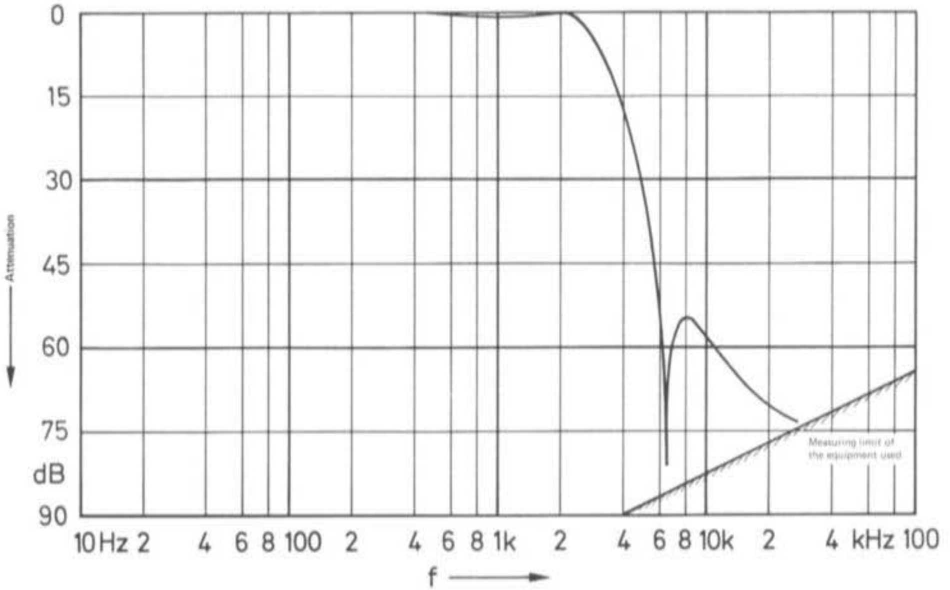


Fig. 10: Frequency response of the circuit given in Fig. 9

5. COMBINATION FILTER WITH ATTENUATION POLE

A combination of the measures, described in the previous sections 3. and 4. results in a very effecient filter. The circuit (Fig. 9) is still relatively simple, however, the resulting attenuation curve (Fig. 10) would require an extremely extensive construction with LC-filters. The circuit is provided with a common-collector stage at the input which guarantees a defined, low source impedance for the following filter. This is followed by a RC-link (10 kΩ - 10 nF) after which a triple-link, high impedance low-pass filter is provided which is high-impedance connected to a Darlington-circuit (T 2, T 3). The buffer stage (T 4) was found to be necessary in order to avoid any reaction between the two filter portions. Such reactions are observed as deviations of the measured frequency response from that calculated. In addition to this, the buffer stage shifts the DC voltage operating point of the circuit by at least 0.6 V in the positive direction. This is important for the drive range of the arrangement, since transistor T 1, T 2, T 3 and T 5 each shift the operating point by 0.6 V in the negative voltage direction.

The buffer stage is followed by a further (but low-impedance) triple-link low-pass filter which is equipped with the feedback configuration described in section 4. The frequency of the attenuation pole is adjusted with the aid of a feedback resistance of R_f = 56 kΩ to somewhat more than 6 kHz. As can be seen in Fig. 10, the measured curve corresponds to the calculated values.

Due to the very high attenuation values that are obtainable with this filter, it is important that a very high degree of decoupling is made between the output and the input. If this was not the case, it would be especially the high frequencies of the input voltage that would bypass the filter and induce themselves into the output. This would mean that the high attenuation value would not be obtained.

Further details on how to achieve attenuation poles are given in (3). However, the circuit given in this publication was not examined.

6. NOTES

The most important reason for the development of this filter is to obtain a steeper attenuation skirt on increasing the frequency. This is why only low-pass filters have been considered. Of course, it is also possible for the resulting steep frequency response curves to be extended to form speech-weighted or bandpass filter configurations. This can be achieved by series connection of one of the high-pass filter circuits that were described in (2). In addition to this, it is possible for one stage to be modified to provide a certain amount of gain. Further details regarding this were given in section 1.2. in (2).

If an amplifier stage is to be used, or when the active filter is to be used in conjunction with an amplifier, special attention must be paid to obtain a high degree of decoupling in the operating voltage supply. It is advisable to insert a series resistance of approx. 680 Ω and for a capacitor of approx. 10 μF to be connected from this point to ground. When using transistors having a very high current gain and high transit frequency (e.g. BC 109 C), a certain instability was observed sometimes due to unfavourable assembly. This was sometimes only observed as a distortion of the original sinusoidal voltage. In order to avoid this, a resistor (R_b) of approx. 100 Ω to 1 kΩ directly connected to the base connection of the transistor in question, has proved to be successful. Figure 11 shows a diagram of this.

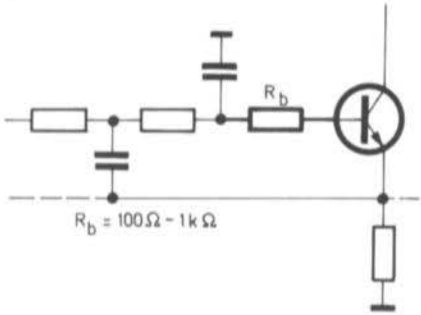


Fig. 11: Stabilization of the amplifier stage using an additional base resistor R_b

All silicon NPN transistors having a current gain B of at least 100 are suitable for these circuits. It is only the buffer stage given in Fig. 9 that must be equipped with a silicon PNP transistor (e.g. BC 213, 2 N 2907). The same points are valid with respect to the drive range of the filter circuits as were given in section 3. and (2).

7. REFERENCES

(1) D.E.Schmitzer: Active Audio Filters
VHF COMMUNICATIONS 1 (1969), Edition 3, Pages 218-225

(2) D.E.Schmitzer: Active Audio Filters
VHF COMMUNICATIONS 1 (1969), Edition 3, Pages 226-235

(3) W. Farrer: A Simple Active Filter With Independent Control Over the Pole and Zero Locations
Electronic Engineering 1967, April, Pages 219-222

SPEECH PROCESSING

by D. E. Schmitzer, DJ 4 BG

INTRODUCTION

The effect of speech processors was demonstrated very clearly by the author at the 14th VHF Conference in Weinheim, West Germany, in September 1969.

During contests etc., one often observes overmodulated signals that cause splatter and other interfering effects to adjacent stations. Of course, one is greatly tempted to overmodulate since the maximum distance that can be covered under given conditions (path attenuation, antenna gain, transmit power, receiver sensitivity), is mainly dependent on the mean depth of modulation. If no special measures are made, the mean depth of modulation for voice transmissions is less than 30% if the modulation power has been adjusted so that the voice peaks are just below 100%. This is the reason why the modulation power is very often adjusted to too high a level without considering that voice peaks overmodulate the transmitter and cause a certain amount of splatter.

There are, however, various means of increasing the mean depth of modulation without increasing the peak modulation level. When correctly used, they allow a noticeable increase in volume and thus coverage without causing splatter.

1. DYNAMIC COMPRESSION

The mean depth of modulation can be increased using a so called dynamic compressor. Such circuits operate according to the AGC principle where a control voltage is taken from the output voltage of a AF amplifier and used to control the gain so that the output voltage is virtually constant. With an ideal circuit no additional distortion should occur. However, it is very difficult to dimension such circuits to obtain the required rise time and fall time constants for the required application, namely the highest possible modulation of transmitters. The time constants of the majority of described dynamic compressor circuits are generally too long so that they are only able to compensate for volume variations in the same manner as an automatic gain control circuit. In the intermediate periods between words, the circuit increases the gain so that a great amount of background noise is observed as well as echo effects. When commencing the next word, the first sound will cause an overmodulation condition since the too long rise time will not immediately generate a new control voltage to compensate for this. Furthermore, weak tones subsequent to strong ones, will not be heard due to the large fall time constant of the control voltage.

On the other hand, the rise time constants should not be small enough to control the lowest modulation frequency to be transmitted. For this reason, it is necessary that dynamic compressors be provided with an audio filter so that the control circuit is not fed with frequencies which it would control.

This description of dynamic compressors it, of course, not complete; the previously mentioned overmodulation caused by the first tone after an intermission could be avoided by providing an additional limiting circuit. However, it can be seen that dynamic compressors must be dimensioned within very narrow limits if they are to provide more than an automatic gain control circuit.

2. AMPLITUDE LIMITING

In this mode, the audio amplitudes are limited to a certain value. This process is often called "clipping". Since the voltage peaks that were responsible for the overmodulation of the transmitter are eliminated in this manner, it is possible for the modulation level to be increased so that the limited voltage modulates the transmitter to a depth of 100%. The more the voltage peaks are limited, the higher will be the mean depth of modulation.

If no further measures are used, however, distortions will be exhibited by the limiting process which can have such an adverse effect on the intelligibility that it has no advantage whatsoever. Under such conditions, the limited signal will contain a large number of audible harmonics which will cause the modulated RF signal to be increased in bandwidth.

The various methods of limiting the amplitude and circuits for maintaining the speech intelligibility are to be covered in the following sections.

2.1. PREREQUISITES AND POSSIBILITY OF UTILIZATION

It is assumed that the whole aim is to obtain the highest possible intelligibility during voice transmissions, especially under high-noise conditions. No importance is to be paid to a natural reproduction of the individual voice and a certain amount of distortion is to be allowed when it does not noticeably affect the speech intelligibility.

Attempts with amplitude limited voice and measurements of the syllable intelligibility as a function of the degree of limiting were already carried out and described in 1936 by F. Strecker (1). Since then, the international telecommunications authorities have experimented with such limiting circuits which has led to their use in the telephone and even broadcast technology. These and other experiments led to accurate values that could be used for comparison of the various types of limiting.

During amplitude modulation (AM), correct amplitude limiting of the AF voltage allows the mean depth of modulation to be increased without causing overmodulation. On the other hand, with frequency modulation (FM) the mean frequency deviation is increased without exceeding the maximum permissible frequency deviation, and thus the maximum bandwidth. Increasing the mean frequency deviation has the same effect for FM as increasing the mean depth of modulation with AM.

Although the AF amplitude limiting circuits described here can, in principle, also be used for single sideband modulation, a certain amount of caution must be paid in practice. Most SSB transmitters are equipped with final amplifier tubes where the modulation peaks overload them in pulse-type operation, but the mean value of the drive during voice transmissions is so low that the maximum permissible plate dissipation of the tube is not exceeded. A noticeable increase of the mean depth of modulation with the aid of a limiter would also increase the mean plate dissipation and could endanger the final amplifier tubes - and also the somewhat under-dimensioned power supply.

It should be noted that a correctly dimensioned and well adjusted amplitude limiting can increase the mean power by at least 6 dB which is equivalent to increasing the power by 4 tumes. This shows that if all possibilities are to be utilized to the full, generously dimensioned stages are necessary with power supplies and tubes or transistors that are not only capable of operating at a mean amplitude level of 20 to 30% but up to 80% of the peak modulation level.

2.2. AMPLITUDE LIMITING METHODS

Of the various methods of achieving amplitude limiting, only two are to be dealt with here. These are the RF and AF type of limiting. Both methods are capable of providing good results according to the measurements given in (2) - even if the required equipment differs greatly. The demands on both systems are the same: The frequency band must not be wider than that necessary to transmit the required information. For voice transmissions, this corresponds to a frequency range of approx. 0.3 to 3 kHz.

2.2.1. LIMITING AT RF LEVEL

With so called RF clippers, the AF band is transposed into a RF suppressed carrier signal. This is followed by a sideband filter which suppresses the non-required sideband. Up to this point, the arrangement corresponds to a normal SSB transmitter circuit. The single sideband signal is now amplitude-limited. Since the harmonics resulting from this are at multiples of the carrier frequency and not of the audio frequency, they can easily be suppressed using an additional filter. In addition to this, conversion products of the individual frequencies which fall in the vicinity of the carrier frequency, are also suppressed by the second single sideband filter. In the case of a SSB transmitter, the limited signal is then processed in the conventional manner (7). However, it is also possible for a second mixer to be used which transposes the RF signal back transmitter. The principle of such a circuit is given in Fig. 1.

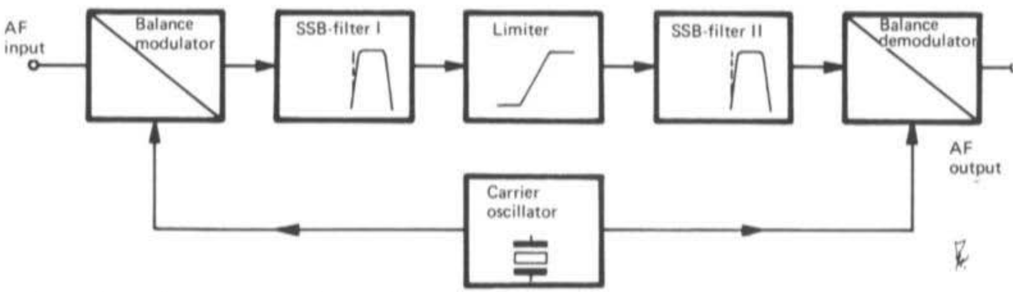


Fig. 1: Block diagram of a RF limiter circuit

The advantage of limiting at RF level is that the harmonics generated during the limiting process do not fall into the usable frequency range. They are therefore extremely easy to suppress and have virtually no adverse effect on the intelligibility even at high degrees of limiting. The disadvantage is the high amount of filtering, which makes this method actually only seem advisable for SSB transmitters where one of the two filters is already present.

2.2.2. LIMITING AT AF LEVEL

The so-called audio clippers also offer excellent results at practical degrees of limiting when correctly dimensioned. The noticeable simplification in the build-up allows a great versatility. For this reason, various practical circuits for this system are to be discussed in detail.

Without additional measures, a noticeable decrease in intelligibility will occur even at very low degrees of limiting of a voice signal. Fig. 2 shows a diagram of the spectral energy distribution of the human voice, indicating possible improvements.

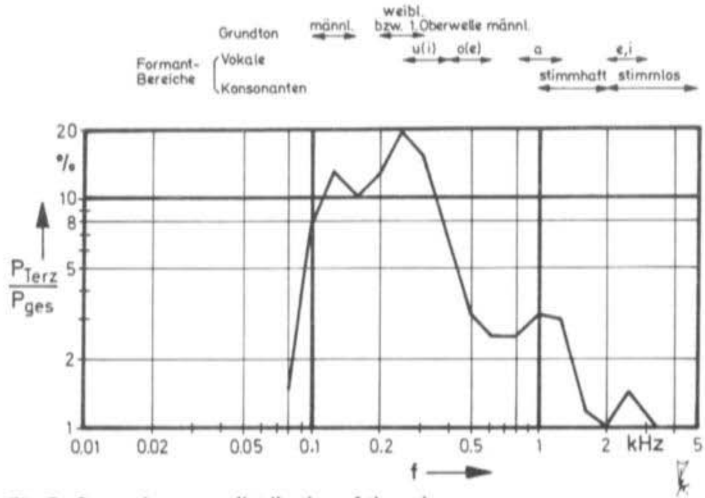


Fig. 2: Spectral energy distribution of the voice

2.3. MEAN ENERGY DISTRIBUTION OF THE HUMAN VOICE

Normal,unfiltered speech contains the highest energy components in the range between 100 and 300 Hz. The energy decreases continuously towards higher frequencies as can be seen in Fig. 2 which was taken from (3). It can be seen that the range in the vincinity of 1 kHz is important to the intelligibility, whereas the components between 100 and 300 Hz affect the naturalness of the voice but are not essential for the intelligibility. Frequencies in excess of approx. 3 kHz are only necessary for recognizing certain letters, e.g. f and s. However, they form combination tones during the limiting process (difference frequencies during intermodulation distortions) which fall into the important lower frequency range where they increase the distortion factor.

It can be seen that an increase of intelligibility whilst limiting can be achieved when the voice frequency range is limited to 300 to 3000 Hz and the frequency range around 1 kHz is accentuated. In addition to this, the bass rejection has the advantage that fewer as well as weaker harmonics of the lower frequencies fall into the required frequency range. Only harmonics of the medium and higher frequencies can be suppressed using a 3 kHz low-pass filter subsequent to the limiter.

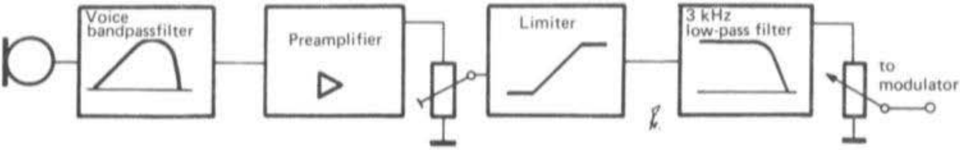


Fig. 3: Block diagram of an AF amplitude limiter

3. DIMENSIONING OF THE AF LIMITER

The block diagram given in Fig. 3 shows the principle of the described filters in a limiter circuit. The following sections describe the dimensioning of the individual circuit groups.

3.1. PREAMPLIFIER AND SPEECH FREQUENCY FILTER

The required accentuation of the frequencies around 1 kHz is obtained simply by attenuating the lower frequencies. This is achieved easily by decreasing the value of a coupling capacitor at a suitable position in the preamplifier until the capacitor together with the input impedance of the subsequent stage results in a cut-off frequency of approx. 4 to 5 kHz. In this manner, a speech-frequency bandpass filter characteristic is obtained. Of course, an active audio filter with a certain amount of gain as described in (4) or (5) is especially suitable.

3.2. LIMITER

By correctly dimensioning the actual limiter stage, the amount of unavoidable distortion can be reduced. The basis of this is the fact that a squarewave signal whose duty cycle is exactly 1 : 1 only consists of the fundamental wave and odd harmonics. Even the slightest unbalanced condition causes additional even harmonics to be generated that cause a reduction of the intelligibility.

This means that if the limiter stage is exactly balanced, half of the possible harmonics will be avoided. Due to the difference between their characteristic curves, any two anti-parallel connected silicon diodes will not represent a symmetrical limiter circuit. The balance can be improved by the use of a bias voltage which reduces the differences between the forward voltages with respect to the total voltage.

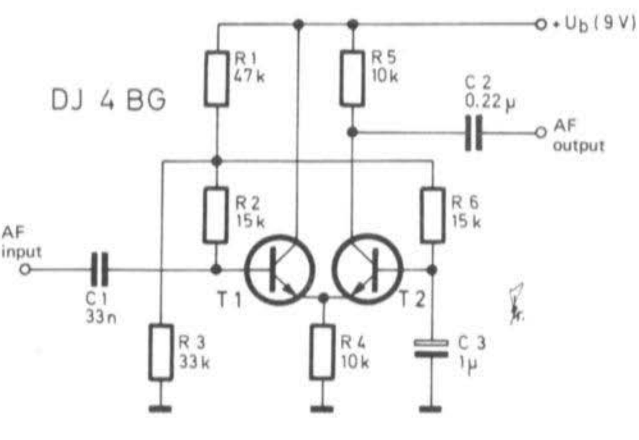


Fig. 4: Differential amplifier as limiter

However, it is more favourable to use a differential amplifier for limiting, especially because the limiting process is, in this manner, combined with a considerable voltage gain. Fig. 4 shows a differential amplifier dimensioned for such an application. When driving with a peak-to-peak voltage of 100 mV, the output voltage (peak-to-peak) will just be limited at 3 V.

This stage therefore exhibits a voltage gain of 30. The preamplification must be dimensioned so that the weakest sounds (e.g. l, m, n, v) are fed to the limiter at a peak-to-peak voltage of 100 mV.

The required balance virtually appears as a by-product if a double transistor or a corresponding integrated circuit is used instead of individual transistors.

Since the individual transistor systems of double transistors or integrated circuits are manufactured at exactly the same time and are only located a fraction of a millimetre from another, they exhibit practically the same behaviour. A limiter built up in this manner mostly exhibits a very high degree of balance which is meintained even when the operating voltage fluctuates.

The following table contains several examples of double transistors and suitable integrated circuits as well as information with regard to the off-set voltage, which is the difference of the base voltages for the same collector current of both systems.

Type	Manufacturer	Off-set voltage, measuring conditions
BFY 81	SGS	max. 10 mV at U_{CE} = 5 V and I_C = 0.1 mA
BFY 85	AEG-Tfk	max. 10 mV at U_{CE} = 5 V and I_C = 0.1 mA
BFY 86	AEG-Tfk	max. 5 mB at U_{CE} = 5 V and I_C = 0.1 mA
CA 3018	RCA	max. 5 mV
CA 3018 A	RCA	max. 2 mV
CA 3046	RCA	max. 5 mV, typ. 0.5 mV

With the integrated circuits, two individual transistor systems should be used.

3.3. LOW-PASS FILTER

It is important that the harmonics caused by the limiting process are attenuated as far as possible. If a simple, steep-skirted 3 kHz low-pass filter is used, the squarewave oscillation generated in the limiter will cause a large "overshoot". This effect is exhibited by both LC and active low-pass filters and will cause overmodulation if the drive level is not decreased by the value of the overshoot. However, the overshoot can be suppressed using a special filter circuit as used in the pulse technology (e.g. in video amplifiers).

So-called Chebyshev filters are suitable for this application. In its simplest form, a normal active low-pass filter - which has been dimensioned for an especially large overshoot - is combined with an RC link. This arrangement it is true, possesses a certain amount of ripple in the passband range, however, it has a far steeper transition inti the stopband range together with a very small overshoot.

Such filter combinations are described in (6) so that they need not be mentioned in detail here.

4. SUMMARY

AF amplitude limiting is a very suitable means of obtaining the highest possible depth of modulation, avoiding, at the same time, any overmodulation. By varying the output level of the preamplifier, the behavior of the circuit can be varied so that only the peaks are limited, or by increasing the level, that an extreme compression of the speech is achieved. However, in order to ensure that no unnecessary loss of intelligibility occurs, it is necessary that the audio bandwidth previous to the limiter is not greater than approx. 0.3 to 3 kHz whereby the bass frequencies are attenuated by 6 dB per octave within the range.

The limiter whould be as symmetrical as possible in order to suppress even harmonics. The limiter should at least be followed by a low-pass filter in order to suppress all harmonics of over 3 kHz. The low-pass filter should, in addition, exhibit the lowest possible overshoot when driven with a squarewave signal. The combination tones in the range below 300 Hz generated during the limiting process are more easily filtered out at the receive end.

5. COMPARISON BETWEEN VARIOUS LIMITING SYSTEMS

A very informative comparison between the various limiting systems with respect to the intelligibility was given in (2). The diagram given in Fig. 5 is taken from this paper. It shows the intelligibility of amplitude-limited speech as a function of the degree of limiting for various limiting systems.

Whereby:

Curve 1 : AF limiter with a 160 to 3200 Hz filter previous to the limiter together with a 3200 Hz low-pass filter subsequent to the limiter.

Curve 2 : As curve 1 except with preemphasis (6 dB/octave) previous to the limiter and the corresponding deemphasis subsequent to the limiter.

Curve 3 : RF single sideband limiting.

Curve 4 : A system where the speech-band is distributed into several channels.

The degree of limiting l is defined in (2) as follows:

$$l = 20 \times \log \frac{\text{peak-to-peak value of the unlimited AF voltage}}{\text{peak-to-peak value of the limited AF voltage}}$$

The diagram shows that the RF single sideband limiting system (curve 3) and the AF limiting with preemphasis (curve 2) exhibit practically the same intelligibility at a degree of limiting between 20 and 30 dB. At higher degrees of limiting, the RF limiting is superior; however, speech limiting by more than 30 dB is not usually used due to the fact that it is difficult to suppress spurious sounds, such as echo, during the punctuation pauses.

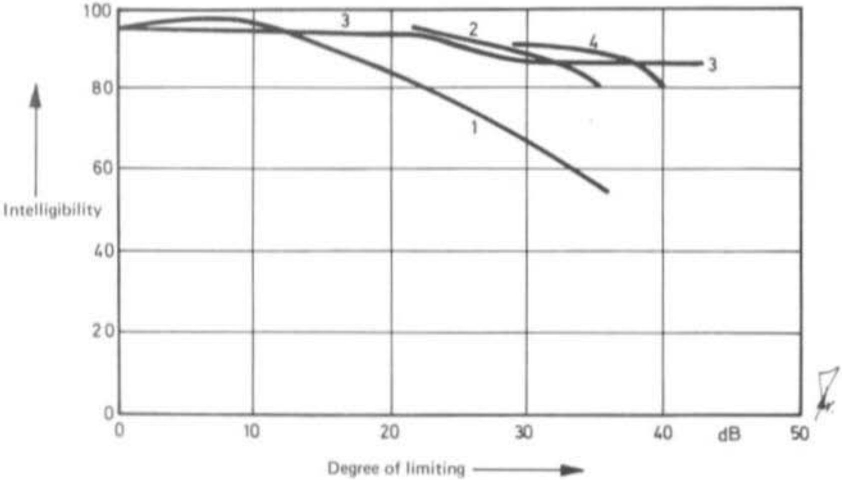


Fig. 5: Intelligibility as a function of the degree of limiting for various limiting systems

It is surprising that the RF limiter is not better with respect to the intelligibility, although no distortion can be heard on the signal limited in this manner in contrast to AF limited systems. The RF clipper can, however, be improved when a bass attenuation of 6 dB per octave is also used, which was not the case in the paper given in (2).

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STRIP-LINE TRANSVERTER FOR 70 cm

by K. Eichel, DC 6 HY

1. INTRODUCTION

A transverter is to be described that converts the transmit signal from a 2 metre transmitter to the 70 cm band and, in the receive mode, converts 70 cm signals to the 2 metre band. It should be noted that the frequency multiplication to the 70 cm band can be achieved both by tripling or by mixing with a frequency of 288 MHz.

The tripling process, which is undesired in this transverter, cannot be completely avoided. The auxiliary frequency has therefore been chosen so that the required signal is spaced at least 500 kHz from the tripled spurious signal. If the frequency range of 432.0 to 433.45 MHz, recommended in IARU-Region I for telegraphy and voice transmissions is used, the spurious wave will be in the range of 433.5 to 437.85 MHz. Under full drive conditions, the spurious signal is approx. 40 dB weaker than the required signal.

Assumung an intermediate frequency of 144 to 146 MHz and an auxiliary frequency of 287.5 MHz, the frequency range of the transverter will be 431.5 to 433.5 MHz.

The receive converter is equipped with a simple high-pass filter at the input which attenuates signals lower than approx. 300 MHz.

The transverter consists of three modules (Fig. 1). One module comprises the receive converter with auxiliary oscillator, the second the transmit converter and the third a linear amplifier. A block diagram of the complete transverter is given in Figure 2. The linear amplifier is equipped with the EC 8020 tube and supplies approx. 5 W of output power. If a tube was not to be used for the linear amplifier, a two-stage transistor amplifier would be necessary, which would mean that the cost of the whole transverter would be doubled.

All other modules are transistorized and operate with a supply voltage of 12 V (minus to ground). Each of these modules is built up on a epoxy printed circuit board, and all UHF resonant circuits are in the form of capacitively shortened, printed striplines. It is easy to construct and is very inexpensive. In spite of this, it represents a high quality transverter for the 70 cm band.

2. THE RECEIVE CONVERTER

If the suppression of VHF signals by the high-pass filter is not sufficient, an additional filter consisting of 2 of the described stripline circuits can be made. They possess the same dimensions and should be enclosed in a casing.

2.1. CIRCUIT DETAILS

The circuit diagram of the receive converter given in Fig. 3 shows the already mentioned high-pass filter (C 101, L 101, C 102), two single-circuit RF preamplifier stages (T 101, T 102), the mixer stage (T 103) with the IF-bandpass filter and a two-stage oscillator (T 104, T 105) together with a stripline bandpass filter.

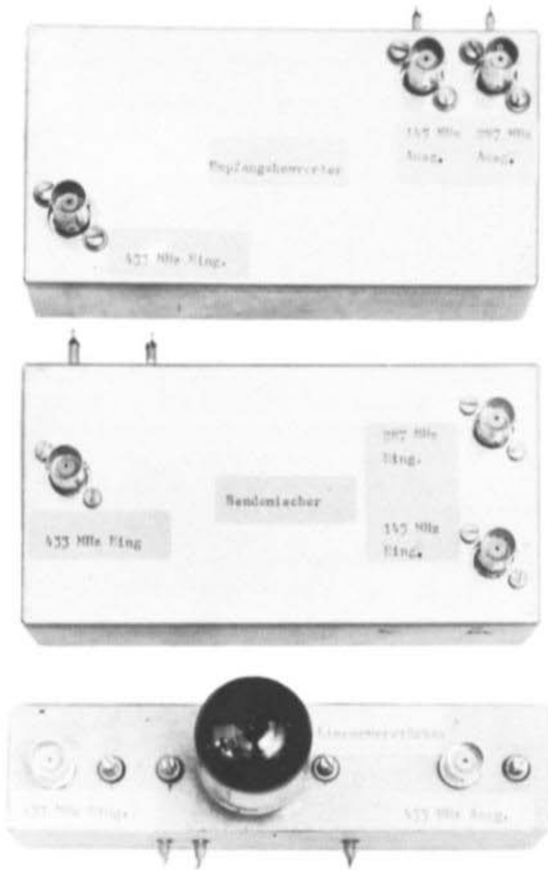


Fig. 1: Photograph of the complete DC 6 HY transmit-receive converter

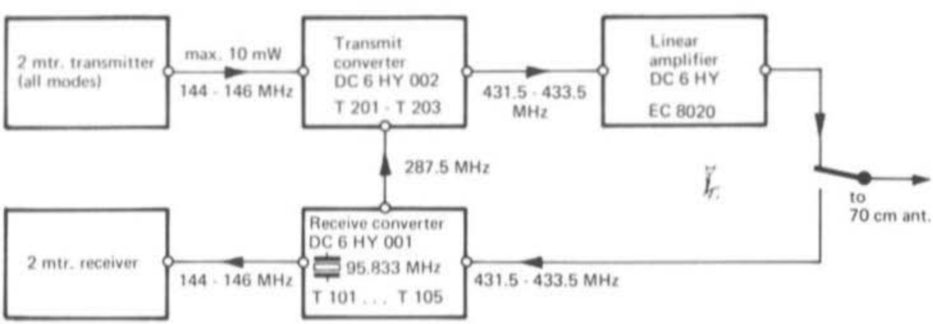


Fig. 2: Block diagram of the 144-432 MHz transverter

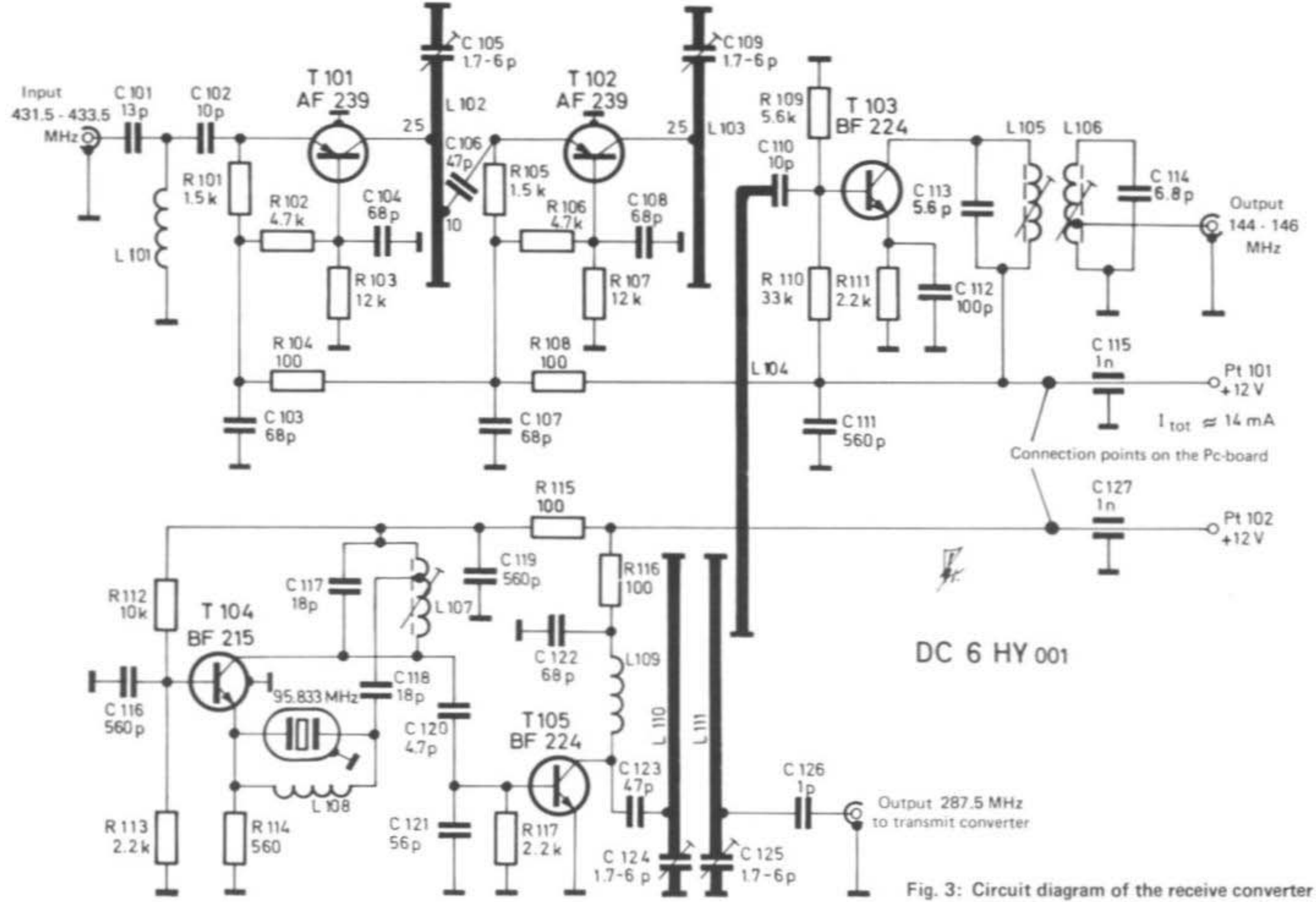


Fig. 3: Circuit diagram of the receive converter

Germanium-Mesa transistor type AF 239 have been selected for the preamplifier stages because equally low-noise silicon transistors are still far more expensive. Since no input circuit is able to increase the noise figure, and since the conversion noise does not have a great effect due to the high preamplification by the first two stages, the sensitivity of the converter is practically only determined by the noise figure of the transistor type used. This amounts to 4 to 5 dB for the AF 239.

The two common-base preamplifier stages are built up identically and the transistors are connected to tapping points on the stripline circuits in order to achieve a high Q. The UHF bandwidth amounts to approximately 10 MHz; it is therefore suitable for amateur television.

An UHF silicon transistor BF 224 is used in the mixer stage. This transistor exhibits a high conversion gain of 10 dB. The inductive coupled output filter is built up in a conventional manner and adjusted to a bandwidth of approx. 3 MHz.

The coupling link L 104 is coupled to 2 stripline circuits; the oscillator voltage is taken from the circuit comprising L 111 (287.5 MHz) and the signal voltage from stripline L 103. Both voltages are fed to the base of the mixer transistor.

The crystal oscillator operates at a third of the auxiliary frequency (95.833 MHz). The used overtone circuit with the transistor in a common base circuit is well-known and need not be described here. The crystal oscillates at the 5th overtone. In order to increase the stability and to simplify the alignment, it is necessary for the crystal holder capacitance to be neutralized. This is made using inductance L 108. It should be dimensioned so that the oscillator ceases oscillation on varying the core of L 107 in both directions.

The subsequent tripler stage (T 105) is matched via a capacitive voltage divider (C 120, C 121) and operates in class C. A resonant line filter selects the 3rd harmonic. In order to be able to connect the cold end of the primary circuit of this filter to ground, a parallel feed via choke L 109 was selected for the transistor (T 105). The auxiliary frequency output (287.5 MHz) is coupled to the secondary circuit via the small capacitor C 126. A power of approx. 1 mW is available at this position for the transmit mixer.

The operating voltage of 12 V is fed to two connection points: Pt 101 only receives the operating voltage during reception; Pt 102 is continuously connected to the operating voltage because the oscillator is required both for receive and transmit.

Furthermore, it should be noted that a great deal of care has been paid to achieve an effective blocking. The values of the filter resistors, coupling and bypass condensers should be maintained.

2.2. SPECIAL COMPONENTS

T 101, T 102: AF 239 or if not available AF 139
T 103, T 105: BF 224, BF 173, possibly 2 N 918 (C_{re} = 0.3 pF; f_T = 700 MHz)
T 104: 2 N 918, BF 115, if not available BC 108.

All conventional inductances are made from 0.5 mm dia. (24 AWG) enamelled or silver-plated copper wire.

L 101: 3 turns, close wound on a 3 mm former, self-supporting
L 102, L 103: printed inductance, tapping points 10 mm or 25 mm from the cold end respectively
L 104: enamelled copper wire bent and soldered as given in the component location plan, spaced 1 mm from the PC-board
L 105: 5 turns close wound onto a 4.3 mm dia. coil former, with VHF core
L 106: as L 105 except coil tap 3/4 turn from cold end
L 107: 4 turns spaced one wire diameter and wound onto a 4.3 mm dia. coil former with VHF core (light green marking), coil tap 1 1/2 turns from cold end
L 108: 8 turns wound onto a 5 mm former. Coil length 8 mm, self-supporting
L 109: 15 turns close wound onto a 3 mm dia. former, self-supporting
L 110, L 111: Printed inductance, coupling point at the hot end direct at the trimmers

C 105, C 109, C 124, C 125: 1.7 - 6 pF or, if not available, larger values up to a maximum of 2.0 - 13 pF with rotor and stator connections for printed circuit boards with standard spacing can be used.

C 115, C 127: 1 nF feed-through capacitors, screw-fitting

Crystal: 95.833 MHz, 5th overtone, series resonance, HC-18/U for soldering

3 BNC-sockets with teflon insulation: UG 290/U (Amphenol)

1 metal casing 141 mm long, 71 mm wide and 28 mm high, with matching cover.

2.3. ASSEMBLY OF THE RECEIVE CONVERTER

The receive converter is accommodated on the printed circuit board DC 6 HY 001 having the dimensions 135 mm x 65 mm. The printed circuit board is illustrated in Fig. 4; the corresponding component location plan in Fig. 5. BNC-sockets should be soldered to the three marked positions with the flange on the conductor side. Attention should be paid that the tapped holes are not filled with solder. After all components have been mounted onto the printed circuit board, the preliminary alignment of the module can be made. After this, the printed circuit board is fixed to the casing by passing the BNC-sockets through the holes made for this purpose, passing screws through the smaller holes, providing them with 2.5 mm thick spacer bushings and screwing them into the threaded holes of the BNC-sockets. This ensures that the printed circuit board is provided with defined ground points and that the stripline circuits maintain a spacing of 5 mm to the ground area.

The two feed-through capacitors are now mounted and connected. The position of these, and a general idea of the construction can be seen in the photograph given in Fig. 6.

Finally, it should be pointed out that it is not advisable to operate the receive converter in a common casing with the transmit converter when adequate screening is not provided. Since the crystal oscillator is sensitive to RF injection, the SSB signal could cause a frequency modulation of the crystal oscillator. If the described construction is adhered to, this is not possible. However, it is important that the casing of the crystal is grounded. The solder connection should not be made at the solder joint of the casing but somewhat higher.

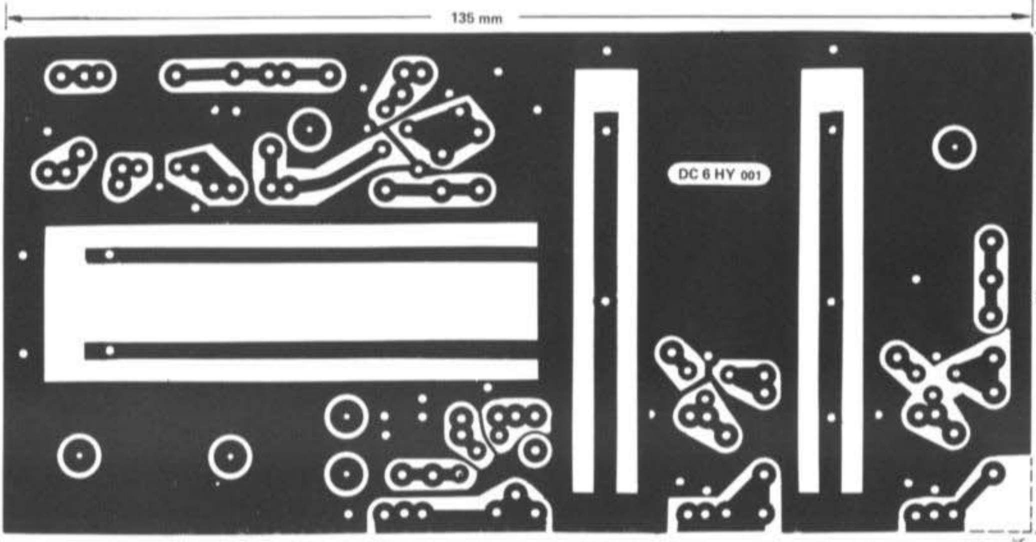


Fig. 4

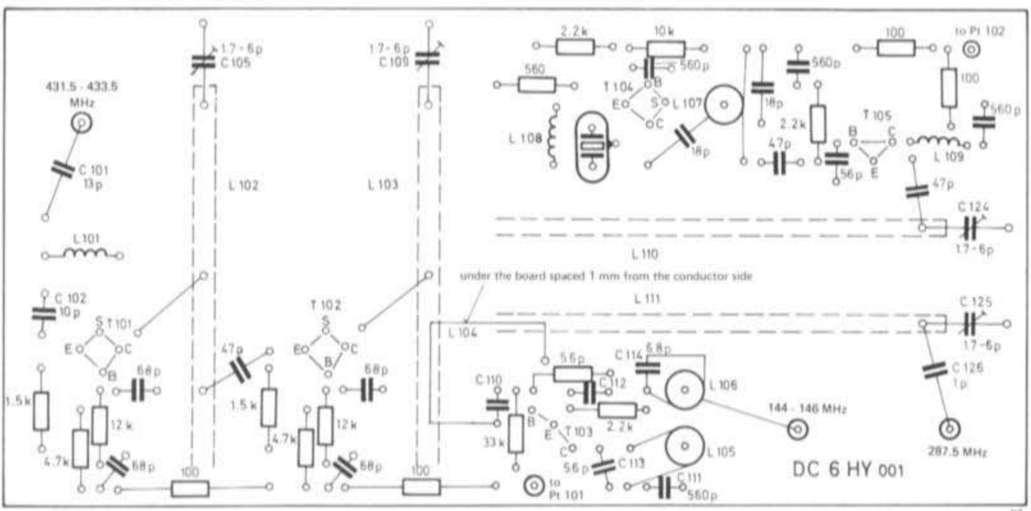


Fig. 5: Component location plan to DC 6 HY 001

2.4. ALIGNMENT OF THE RECEIVE CONVERTER

As has already been mentioned, it is possible for the receive converter to be tested before mounting it into its casing. This is done by connecting the ground, as well as an operating voltage of +12 V to point Pt 102. The crystal oscillator can be monitored on a VHF-FM broadcast receiver (channel 29 +). The commencement of oscillation is indicated by an increase in current of approx. 2 mA.

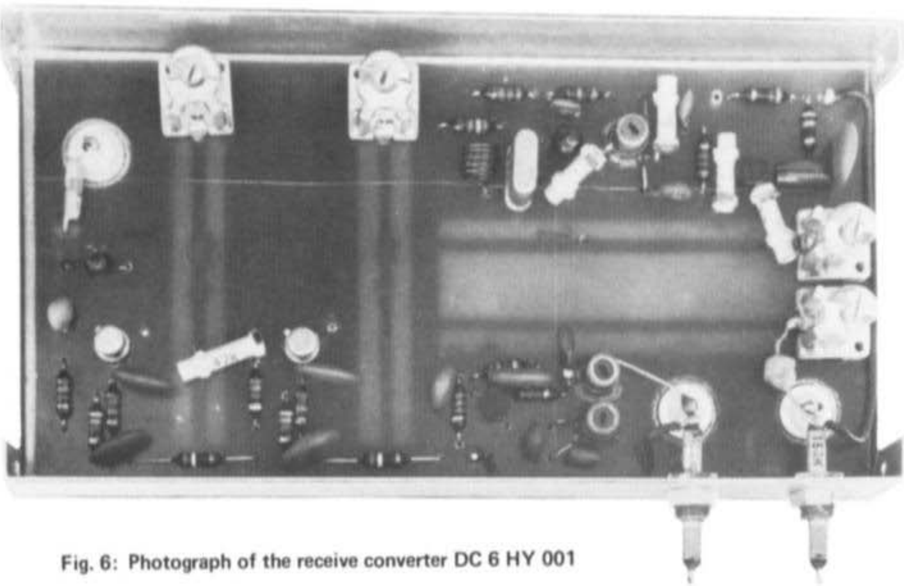


Fig. 6: Photograph of the receive converter DC 6 HY 001

Point Pt 102 is then provided with voltage and the emitter current of the mixer transistor should be measured as the voltage drop across the emitter resistor R 111. The 287 MHz filter is now adjusted with the aid of trimmer capacitors C 124 and C 125 for maximum voltage drop across R 111.

A 2 metre receiver is now connected to the IF output socket and the following circuits aligned for maximum noise in the order given: L 105, L 106, C 109 and C 105. The S-meter of the receiver should indicate approx. 35 dB more (noise voltage) than without converter.

If all stages have been aligned it is now possible for the module to be built into the casing as has already been explained. After insertion, it is only necessary for the stripline circuits to be corrected. This is done by slightly increasing the capacitance of trimmers C 105, C 109, C 124 and C 125. Noise matching is not necessary with this converter since the cable impedance of 50 - 60 Ω coincides well enough with the input impedynce of the transistor in a common-base circuit.

2.5. MEASURED VALUES FOR THE RECEIVE CONVERTER

All voltages were measured using a valve voltmeter (VTVM) at an operating voltage of 12 V. (x) without oscillator signal, this voltage is 1.1 V)

Voltage at connection	E	B	C
T 101	8.8 V	8.5 V	-
T 102	8.8 V	8.5 V	-
T 103	x) 1.25 V	1.75 V	12 V
T 104	1.55 V	2.05 V	11.6 V
T 105	-	0.05 V	12 V

Total current flow: Approx. 13 mA when the oscillator is in operation
Approx. 11 mA when the oscillator is not in operation.

3. THE TRANSMIT CONVERTER

The transmit converter consists of the mixer and two linear amplifier stages. The auxiliary (local oscillator) signal of 287.5 MHz at 1 mW is obtained from the receive converter. The output signal of a 2 metre transmitter at approx. 3 mW is converted into the frequency range of 431.5 to 433.5 MHz. The transistor output stage provides an RF power of approx. 250 mW at an operating voltage of 12 V.

3.1. CIRCUIT DETAILS

As can be seen in the circuit diagram of the transmit converter shown in Figure 7, the mixer transistor T 201 receives the 2 metre (SSB) signal via the voltage divider comprising R 201 and a portion of L 201 at its base. With 56 Ω, resistor R 201 virtually represents a terminating resistor for the exciter. For this reason, only a small portion of the drive voltage is fed to the mixer transistor. The input circuit is in resonance for the 287.5 MHz auxiliary signal. The printed inductance L 201 is tuned with capacitors C 202 and C 203. The voltage of the signal is therefore increased according to the resonant impedance of this circuit and passed to the base of the mixer transistor. The mixer supplies approx. 2 mW at the output frequency.

The three transistors of the transmit converter all feed into printed stripline circuits having an impedance of approx. 35 Ω. These circuits are also capacitively shortened in the same manner as for the receive converter. The input and output coupling is made at, or near the hot end of the circuits. The operating voltage is fed to the transistors in parallel with the resonant circuits using chokes. All transistors operate with fixed operating points in class A or AB. The last stage has not been provided with a stabilizing emitter resistor in order to achieve the highest possible output power at a quiescent current of 1 mA.

If the final transistor is operated from 24 V, it is advisable for an emitter combination to be used, the operating points to be re-adjusted and for the transistor to be provided with cooling fins.

It is also important for the transmit converter that the values of all chokes, coupling and bypass capacitors are maintained if the high-gain converter is to operate in a stable manner without self-oscillation. Especially the final amplifier equipped with the Overlay-transistor type, easily breaks into self-oscillation at lower frequencies due to the increasing gain. The chokes and capacitors should therefore not form any high-Q resonant circuits. This is the reason why the base-chokes have not been bypassed and why the coupling capacitor at the collector of transistor T 203 is 560 pF. However, certain measures are also necessary for the driver stage with its power gain of approx. 10 dB: the emitter capacitor C 211 of 15 pF together with the inductivity of the transistor is just at series-resonance. We assume that the connections of this (as well as the other) transistor are made in the shortest possible manner to the PC-board.

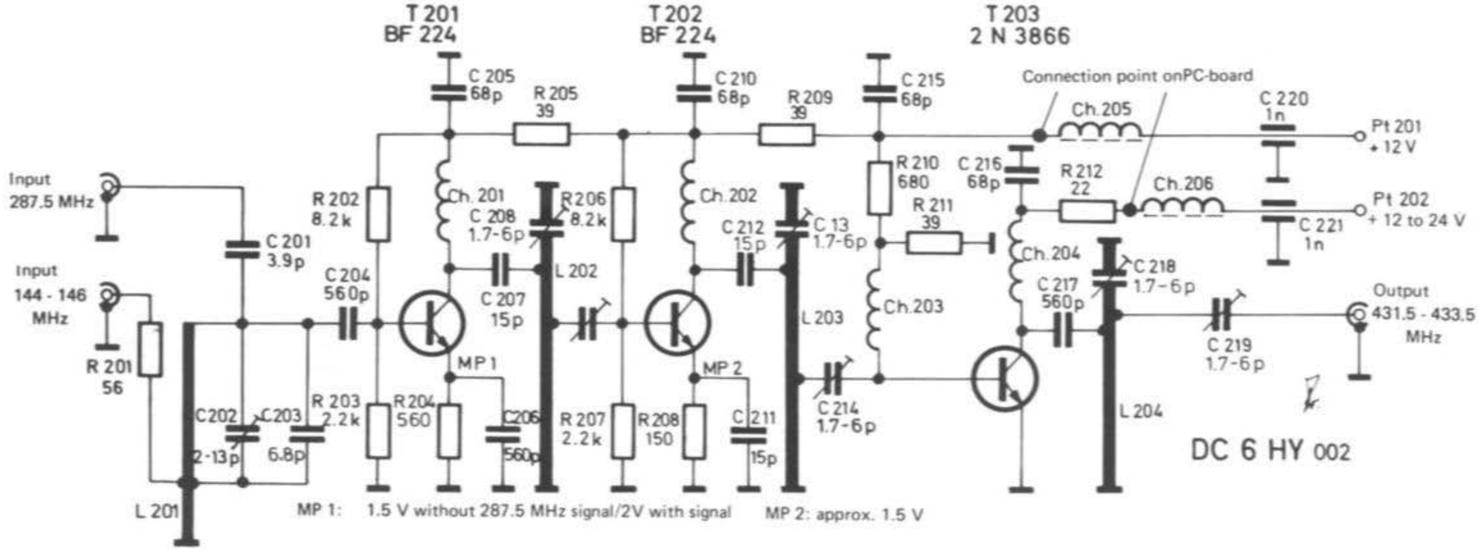
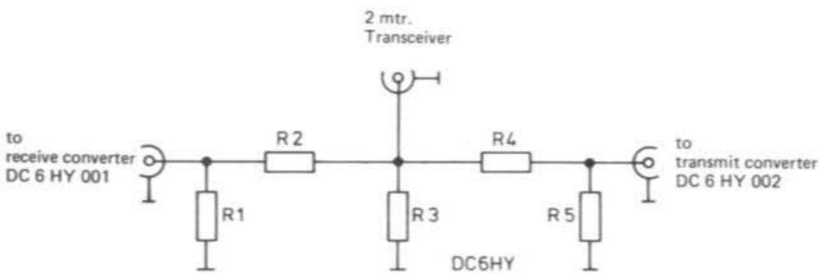


Fig. 7: Circuit diagram of the transmit converter DC 6 HY 002



R 1 : 82 Ω/0,5 W
R 2 : 560 Ω/0,5 W
R 3 : 82 Ω, cap. of transmit power
R 4 : 680 Ω/0,5 W at a power of 5 W
1 kΩ/0,5 W at a power of 10 W
R 5 : 82 Ω/0,5 W

Fig. 13: Attenuation network for connection of the transmit-receive converter to a 2 mtr. transceiver

3.2. SPECIAL COMPONENTS

T 201, T 202: BF 224 (BF 173); (C_{re} = 0.3 pF; f_T = 700 MHz)
T 203: 2 N 3866

L 201: printed inductance, tapping for R 201:
as given in the component location plan
L 202, L 203, L 204: printed inductance, injection is made at the hot end of trimmer capacitor C 208, C 213 and C 218; signal tapped off 7.5 mm from the hot end of stripline L 202 and L 203, as well as directly from the hot end of L 204.

All chokes are wound from 0.5 mm dia. (24 AWG) enamelled copper wire onto a 3 mm former. Coil spacing: half wire diameter, self-supporting.

Ch 201, Ch 202: 4 turns; Ch 203: 8 turns; Ch 204: 5 turns;
Ch 205, Ch 206: 1 or 2 ferrite beads

All trimmer capacitors should have rotor and stator connections with standard spacing for printed circuit boards.

C 202: 2 - 13 pF air-spaced miniature trimmer
C 208, C 209, C 213, C 214, C 218, C 219: 1.8 - 6 pF or larger up to max. 2 - 13 pF air spaced miniature trimmers
C 220, C 221: 1 nF feed-through capacitors, screw fitting

3 BNC-sockets, Teflon (PTFE) insulated; e.g. UG 290/U (Amphenol)
1 metal casing as for the receive converter.

3.3. ASSEMBLY OF THE TRANSMIT CONVERTER

The transmit converter is built up in the same way as the receive converter. The printed circuit board possesses the same dimensions (135 mm x 65 mm) and is designated DC 6 HY 002 (Fig. 8). This printed circuit board is completed according to the component location plan given in Fig. 9, prealigned and finally, as was also the case with the receive converter, connected into the casing with the 6 screws which are screwed into the three sockets. Due to the 2.5 mm high spacer bushings and the 2.5 mm thick flange of the BNC-sockets, the conductor side of the printed circuit board is spaced 5 mm from the casing.

The author's prototype is shown in the photograph given in Fig. 10. The location of the feed-through capacitors can also be seen in this illustration.

3.4. ALIGNMENT OF THE TRANSMIT CONVERTER

A power meter or a reflectometer with load resistor (well-matched antenna) is required for the alignment. The connection points Pt 201 and Pt 202 are fed with +12 V (Pt 202 via a mA-meter) and the emitter voltages of the two first stages checked at measuring points MP 1 and MP 2. Any multimeter is suitable for this measurement.

Whilst the voltage is measured at MP 1, the connection to the 287.5 MHz output of the receive converter is made and the input circuit aligned with trimmer capacitor C 202. The value indicated at MP 1 should increase by approx. 0.5 V. A quiescent current of approx. 1 mA flows through the final transistor; this current is variable with the voltage at Pt 201.

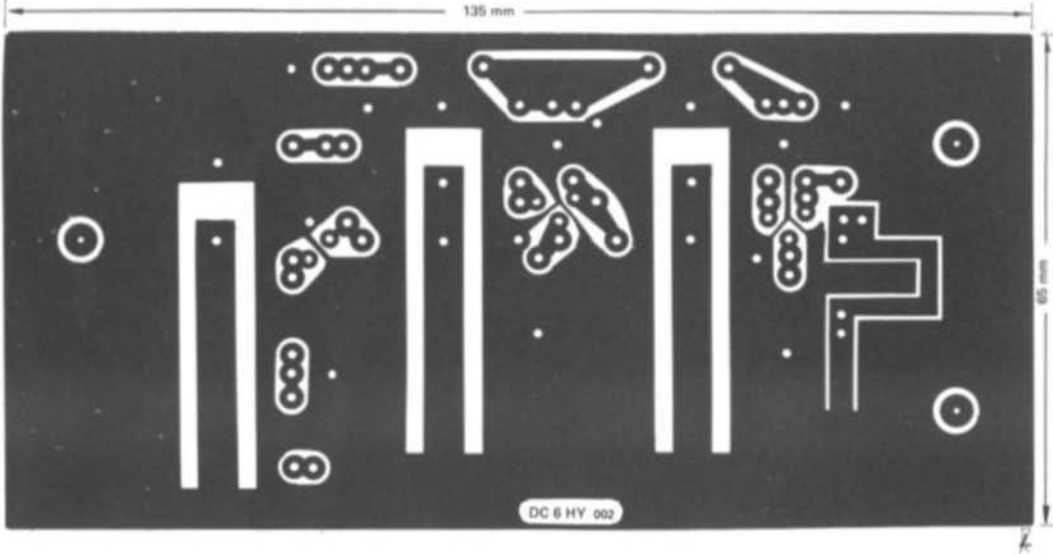


Fig. 8: Printed circuit board DC 6 HY 002

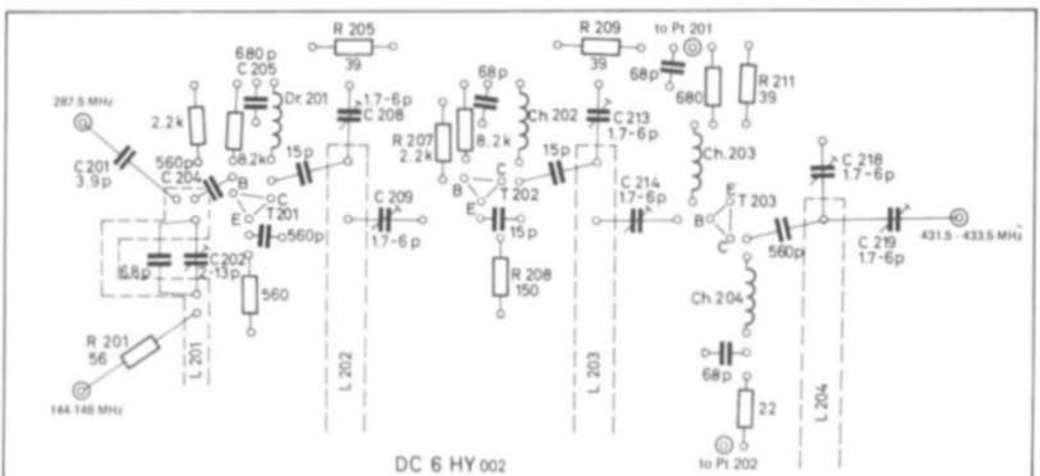


Fig. 9: Component location plan to DC 6 HY 002

Finally, the 144 MHz input is also driven. The most favourable means of providing the required 3 mW depends on the available VHF equipment. It is, however, not advisable to reduce the AF gain of the SSB exciter (Mike gain) further than its normal position since this would mean that the sideband to carrier ratio would deteriorate (carrier suppression). If it is not possible to tap off the power from a driver stage of the exciter, an attenuation link must be provided.

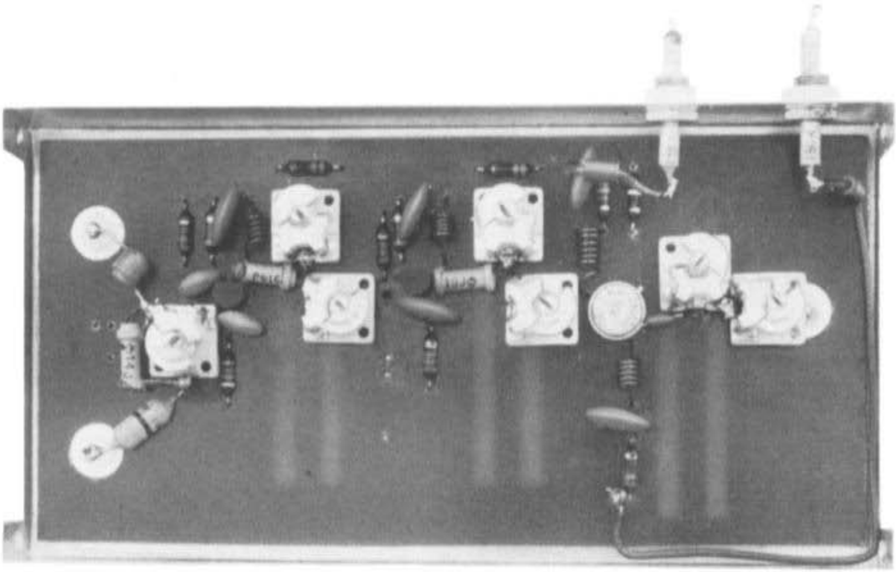


Fig. 10: Photograph of the transmit converter DC 6 HY 002

A suitable attenuation network is given in Fig. 13. It allows the various pieces of equipment to remain connected so that the receive converter need not be disconnected from the transceiver whilst transmitting. The power induced into the receive converter in the transmit mode is approx. 50 mW, and thus well below the maximum permissible power dissipation of the transistors. In addition to this, the attenuation network comprising resistors R 1 to R 5 reduces the very high overall gain of the receive converter. The transistor equipment described in (1) and (2) is favourable for feeding this transverter.

The transmit converter is now fed with a single-tone signal and trimmer capacitors C 209, C 214 and C 219 should be at a quarter of their maximum capacitance after which trimmer capacitors C 208 and C 213 are aligned for maximum collector current via transistor T 203. Before the drive is finally aligned, the output circuit (C 218) and the output coupling (C 219) must be aligned for maximum output power. After this, all trimmer positions can be carefully corrected by repeating the whole alignment process several times. The final amplifier stage (T 203) should pass a maximum of 50 mA, the two previous stages a total of 30 mA (at 12 V operating voltage).

The module can now be installed in the casing and the final alignment correction made. Usually, it is only necessary for trimmers C 202, C 208, C 213 and C 218 to be aligned by slightly increasing their capacitance.

4. THE LINEAR AMPLIFIER

As was already mentioned in the introduction, the 70 cm final amplifier stage is equipped with an EC 8020. This modern grounded grid triode is especially suitable for amplifier stages in the 70 cm band because it possesses the lowest reactive capacitance in this frequency range and because it operates at very high efficiency and exhibits high power gain (3). A photograph of the linear amplifier is given as the lowest module in Fig. 1.

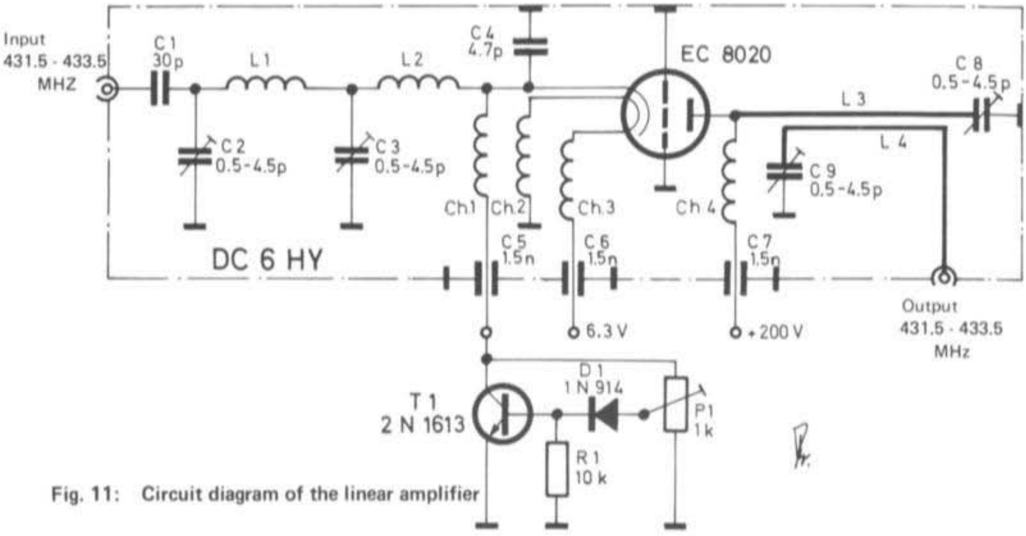


Fig. 11: Circuit diagram of the linear amplifier

4.1. CIRCUIT DETAILS

The circuit diagram of the linear amplifier stage is given in Fig. 11. The drive power of approx. 250 mW from the transmit converter is fed via a double-link low-pass filter to the cathode of the tube. The grid bias voltage of approx. 2.8 V is generated across a constant voltage two-pole in the cathode lead. This transistor circuit possesses a low differential impedance (approx. 5 Ω), which is far less than 4 silicon diodes in series. In addition to this, the bias voltage is adjustable with the aid of the trimmer potentiometer P 1. A resistor cannot be used at this position since it would greatly reduce the power gain of the tube. This is because increasing drive and thus increasing plate current would shift the operating point from class AB into class C.

The two heater voltage connections of the tube are provided with chokes in order to ensure that the drive power is not shorted by the cathode-heater capacitance of the tube. The heater voltage is 6.3 V and is grounded at one side.

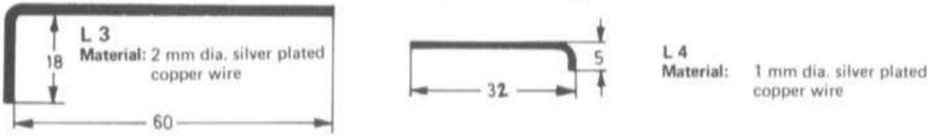
The plate of the tube is connected to a $\lambda/2$ coaxial circuit (L 3). This circuit is shortened at one side by the plate-grid capacitance, and at the other side by the trimmer capacitor C 8. The plate voltage is fed to the point of minimum voltage (max. current) of the circuit via a choke. This point is located approx. in the centre, however, if the choke is effective, the position will not be critical.

A parallel wire loop (L 4) represents the output coupling. A further trimmer capacitor (C 9) adjusts the wire loop to series resonance at the required frequency. Since this arrangement is coupled with the magnetic field, the loop is placed where the most magnetic field lines are present, in the vicinity of max. current.

The plate voltage should be in the range of 180 to 220 V. Without drive, approx. 40 mA will flow, approx. 80 mA at full-drive. Output power at this point is at least 5 W. This means that the linear amplifier stage provides a power gain of at least 13 dB at the given drive power.

4.2. SPECIAL COMPONENTS

Tube: EC 8020 (AEG-Telefunken)
T 1: 2 N 1613, 2 N 2219 or any silicon NPN transistor in TO-5 casing
D 1: 1 N 914 or any silicon diode
L 1, L 2: one turn of 1 mm diameter (18 AWG) silver-plated copper wire wound on a 4 mm former, self-supporting
L 3: resonant line made out of 2 mm diameter (12 AWG) silver-plated copper wire prepared according to drawing
L 4: resonant line made from 1 mm diameter (18 AWG) silver-plated copper wire prepared according to drawing



Ch 1 - 4: 10 turns of 0.4 mm diameter (26 AWG) wound onto a 3 mm former, self-supporting

C 2, C 3, C 8, C 9: 0.8 - 6.8 pF ceramic tubular trimmers
C 5 - C 7: 1.5 nF feed-through capacitor, for solder fitting

4.3. ASSEMBLY OF THE LINEAR AMPLIFIER

The linear amplifier is built up in the conventional, cavity principle with a screening plate between the cathode and plate chamber. The base plate is shown in Fig. 12 together with the dimensions; further details are shown in the photograph given in Fig. 14. Single-coated pertinax was used by the author because it can easily be prepared and soldered.

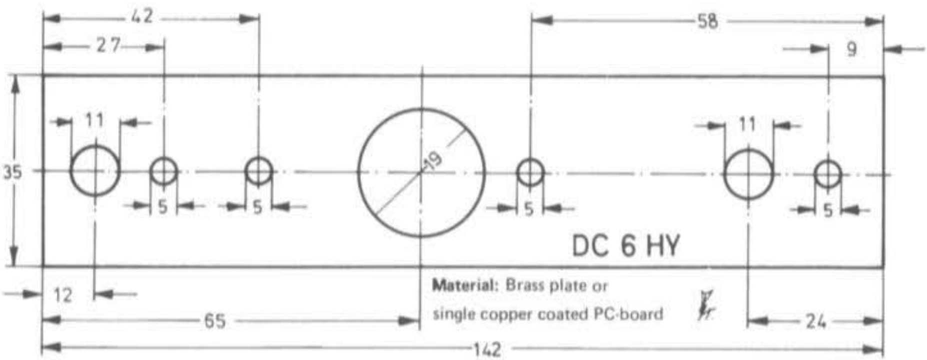


Fig. 12: The base plate of the linear amplifier

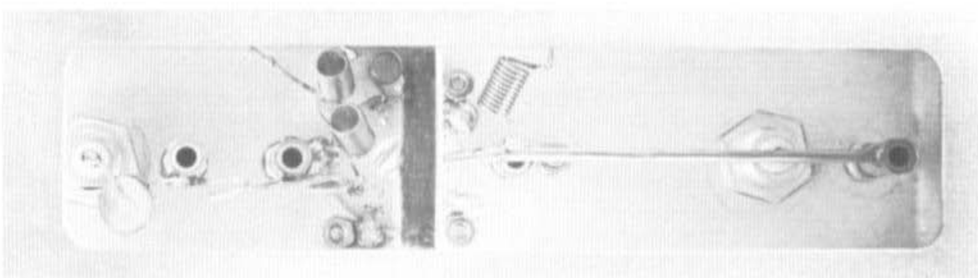


Fig. 14

The only possible type of tube socket is the UHF type of the Pico 9 base. Only the actual socket, which is made from a glass fibre material, is used. The socket is firstly mounted by soldering all grid connectors to the ground area after which it is glued into place using a two-component adhesive (such as UHU-Plus). The screening plate is installed so that as little room as possible is left below the tube base after it has been mounted. The input and output socket can also be soldered into place, however, the ground connection must be made internally. The trimmer capacitors must also be soldered onto the inside of the chambers. When assembling the linear amplifier from brass plate, these instructions will, of course, not be necessary. All further details regarding the construction can be seen in the photograph.

4.4. ALIGNMENT OF THE LINEAR AMPLIFIER

After connecting the heater and plate voltage, the quiescent plate current is adjusted to 40 mA using the trimmer potentiometer P 1. After this, the input is connected to the transmit converter and the trimmer of the input filter tuned for maximum plate current. The plate circuit and the output coupling loop are now tuned for maximum output power. The plate current should now amount to approx. 80 mA. After correcting the whole alignment procedure at the band limits, the alignment procedure is completed.

5. EDITORIAL NOTES

In some cases, the high gain of the two preamplifier stages in the receive converter led to self-oscillation. These difficulties were overcome by mounting a 15 mm high screening plate between striplines L 102 and L 103. This screening plate, which is placed from one side to the other between trimmer capacitors C 105 and C 109, is grounded by means of a short wire through the PC-board and to the case of transistor T 102. A capacitor of 470 pF connected between the connection point of the two 100 Ω resistors R 104 and R 108 concludes the neutralization process.

It was also reported that the oscillator voltage was too low for the transmit mixer. If this is observed, transistor T 105 can be altered into class B (56 kΩ from base to $+U_b$) and the output coupling capacitor C 126 increased from 1 pF to 2.7 pF. This will ensure that sufficient oscillator voltage is available.

6. AVAILABLE PARTS

The printed circuit boards, semiconductors, special components as well as complete kits of parts and complete ready-to-operate, modules are available from the publishers or their national representatives. See advertising page.

7. REFERENCES

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

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

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

A SIMPLE VHF-UHF CALIBRATION-SPECTRUM GENERATOR

by K. Eichel, DC 6 HY

The calibration-spectrum generator described in the following article consists of a simple and uncritical circuit without resonant circuits built up on a small printed circuit board (Fig. 1). The circuit, which is equipped with a 100 kHz crystal, supplies calibration points with a spacing of 100 kHz. The calibration signals are audible at 5 dB above the noise on a 23 cm converter (noise factor = 7 dB). In the 70 cm band, the calibration signals are approx. 25 dB, in the 2 m band approximately 50 dB above the noise.

The circuit can also be used without modification with a 1 MHz crystal; the calibration points, which are then spaced 1 MHz, are approx. 20 dB stronger on all bands than when using a 100 kHz crystal.

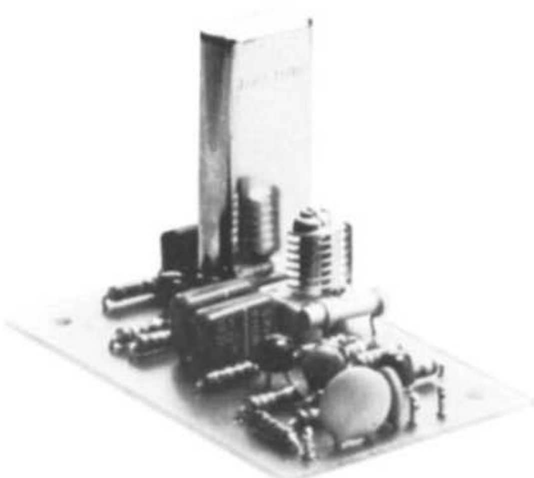


Fig. 1: Photograph of the calibration-spectrum generator

1. CIRCUIT DESCRIPTION

The circuit of the calibration-spectrum generator is shown in Fig. 2. The crystal oscillates in a Buttler-circuit (T 1 and T 2) at the fundamental frequency. The frequency can be pulled to the nominal value with the aid of trimmer capacitor C 1. This is achieved by monitoring the beat frequency between a calibration frequency transmitter (e.g. WWV) and the corresponding harmonic of the calibration frequency generator. The 100 kHz crystal exhibits a positive temperature coefficient (TC) up to a temperature of approx. 35 ^oC. The capacitor C 2 with a TC of $+100 \times 10^{-6}/^oC$ is used for coarse compensation. It should be mentioned that 1 MHz crystals exhibit lower temperature coefficients.

The crystal frequency is tapped off at the collector of transistor T 2 and is fed via the emitter-follower buffer (T 3) to a Schmitt-trigger (T 4, T 5).

The Schmitt-trigger is a flip-flop circuit which supplies a square-wave pulse whenever the input voltage increases above a predetermined value. The sinusoidal voltage from the crystal oscillator exceeds this value each half-wave which means that the Schmitt-trigger generates a square-wave oscillation at the same frequency as that of the crystal oscillator.

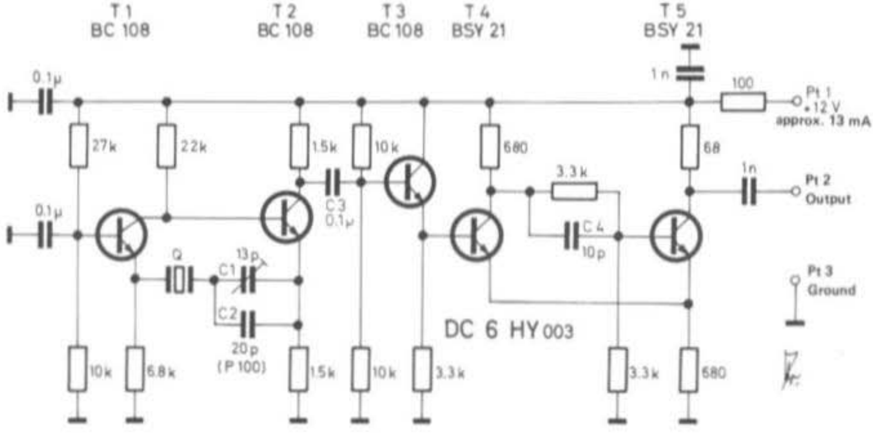


Fig. 2: Circuit diagram of the VHF - UHF calibration spectrum generator

Due to the dimensioning of the circuit and the use of very fast switching transistors, the pulses are very steep (rise time and fall time approx. 7 ns). The steep pulses mean that the square-wave signal will contain extremely high frequency harmonics. The duty cycle of the Schmitt-trigger (that is the relationship of pulse length to the intermediate period between pulses) is never 1 : 1 due to a sort of hysteresis-effect. This unsymmetrical state is further increased by the circuit dimensioning which means that the even harmonics are also contained in the harmonic spectrum. This would not be the case with a duty cycle of 1 : 1, because a symmetrical square-wave oscillation is only comprised of a sinusoidal fundamental wave and odd sinusoidal harmonics (Fourier-analysis).

The whole spectrum is passed to an isolating capacitor; the output impedance amounts to approx. 60 Ω which means that a (short) coaxial cable can be connected. The output voltage is approx. 350 mV (peak-to-peak).

The calibration-spectrum generator requires a stable and well-filtered operating voltage of 12 V. Residual hum on the operating voltage will generate a phase modulation on the slope of the output signal (Jitter) which means that the harmonics in the UHF range will possess a very high hum content.

2. SPECIAL COMPONENTS

T 1 - T 3: BC 108, 2 N 708 or any silicon NPN transistors
T 4 - T 5: BSY 21, 2 N 914 (2 N 706, 2 N 708, BF 224 are not suitable because they hardly generate harmonics in excess of 300 MHz).
Crystal: Parallel resonance with C_p = 30 pF; 100 kHz in HC - 13/U case, or 1 MHz (or 2 MHz, 5 MHz, 10 MHz) in HC - 6/U holder
C 1: 2 - 13 pF air spaced trimmer; C 2: 20 pF ceramic tubular capacitor;
C 3: 0.1 µF/100 V plastic foil capacitor.

3. ASSEMBLING THE CALIBRATION-SPECTRUM GENERATOR

The complete circuit is built-up on a printed circuit board with the dimensions 65 mm x 40 mm. Figure 3 shows the printed circuit board which has been designated DC 6 HY 003. The corresponding component location plan is given in Figure 4. The crystal casing is grounded; the connection should not be made to the soldered sealing of the crystal but somewhat higher. The connection of transistors T 4 and T 5 should be as short as possible. The photograph of the prototype (Fig. 5) gives an idea of the assembly.

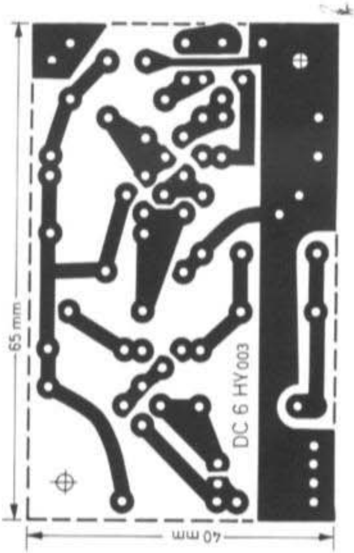


Fig. 3: Printed circuit board DC 6 HY 003

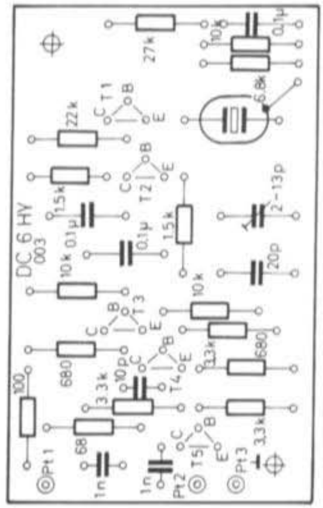


Fig. 4: Component location plan for DC 6 HY 003

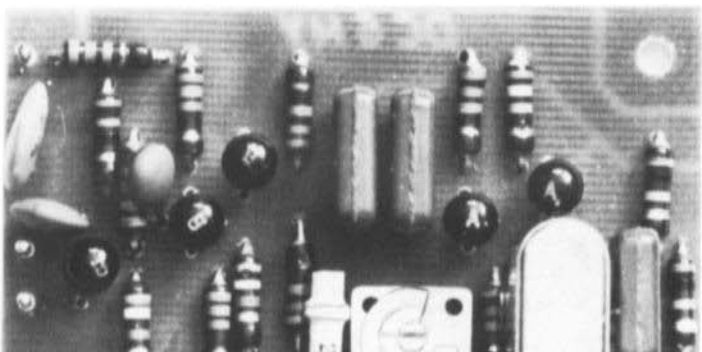


Fig. 5: Photograph of the calibration-spectrum generator from above

4. AVAILABLE PARTS

A kit consisting of printed circuit board, transistors, trimmer, crystal (1 MHz or 100 kHz) is available from the publishers. See advertising page.

5. EDITORIAL NOTES

One of the editorial staff built up a calibration-spectrum generator on a printed circuit board provided by the author. It was found to provide signals approx. 20 dB above the noise on the 23 cm band. In the 10 m band, the signal was so strong that the S-meter indicated FSD (S 9 + 40 dB). These values were obtained both with a 1 MHz crystal as well as with a 100 kHz crystal. Already available transistors type 2 N 2222 were used for the Schmitt-trigger.

6. REFERENCES

(1) H. Götting and D. E. Schmitzer: A Transistorized Calibration Spectrum Generator for Two Metres
VHF COMMUNICATIONS 2 (1970), Edition 1, Pages 41-44

NOTES TO OUR READERS

As you will see in our new material price list, we have had to streamline our material sales programme. This has meant that we have been able to reduce some of the kit prices. Please study the list for qour requirements.

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In spite of rising prices, we have been able to maintain the subscription price at DM 12.00. We hope that you have enjoyed reading VHF COMMUNICATIONS and invite you to re-subscribe for 1971. Wishing you a very Happy Christmas and a prosperous 1971, we remain,

73s
The Publishers DJ 3 QC, DL 3 WR and DJ Ø BQ / G 3 JVQ

NEUTRALIZATION OF THE DL3XW/DJ4BG CALIBRATION SPECTRUM GENERATOR

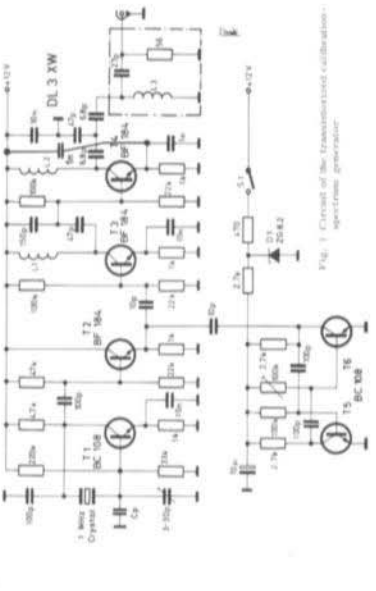
by D. E. Schmitzer, DJ 4 BG

The calibration spectrum generator developed by H. Goetting was accommodated on the printed circuit board DJ 4 BG 003 (1). It has been found that the output transistor T 4 tends to break into self-oscillation. Experiments have shown that this can be neutralized by providing a decoupling capacitor between two of the conductor lanes. Figure 1 shows this additional capacitor in the original circuit diagram given in (1); Fig. 2 shows the position of the capacitor in the component location plan of DJ 4 BG 003. The 1 nF disc capacitor should be soldered to the position shown in Fig. 2.

The author would like to point out that the multivibrator used as frequency divider does not always lock itself correctly to the tenth subharmonic. The author had to find a transistor for T 6 that possessed a moderate current gain (BC 108 A, BC 183 A, 2 N 2921), whereas T 5 required a transistor exhibiting an especially high current gain (BC 108 C, BC 183 C, 2 N 2926). In addition to this, the base resistor of transistor T 5 was increased from 100 kΩ to 120 kΩ. These tips are not to be regarded as a completely fool-proof method but more to indicate what can be tried should the multivibrator not lock-in correctly.

REFERENCES

(1) H. Goetting and D. E. Schmitzer:
A Transistorized Calibration-Spectrum Generator for Two Metres
VHF COMMUNICATIONS 2 (1970), Edition 1, Pages 41-44



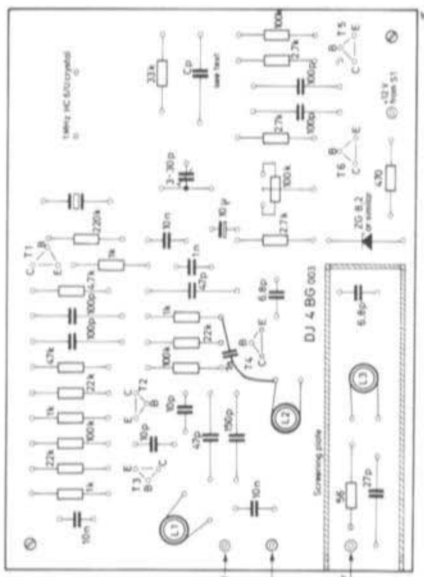


Fig. 2

TWO CIRCUITS FOR AUTOMATIC BAND SCANNING
Part 1: A Simple Band Scanner

by E. G. Hoffschildt, DL 9 FX

1. INTRODUCTION

When the amateur is busy with construction, repair or measurements in the shack, he does not usually have the opportunity to simultaneously tune over the two metre or other band. A similar state of affairs exists during mobile operation where the driver must pay full attention to the traffic. The amateur has often wished for an automatic device to carry out the tuning process for him.

Several pieces of equipment have been developed as a result of these considerations that allow an automatic scanning using the station receiver (1). It is normally possible to install such automatic scanning circuits into the receiver at a later date.

Two circuits are to be described which allow the band to be continuously observed. The common factor of both circuits is the multivibrator circuit which sweeps the oscillator frequency by means of a varactor diode. The first circuit has been kept simple which shows that an efficient band scanning circuit can be assembled easily.

The second circuit, which will be described in detail in part 2 of this article, is based on two multivibrator circuits, the first of which (astable) continuously sweeps the frequency of the receiver oscillator, whereas the second (monostable) stops the sweep of the oscillator for a short period of time in the presence of a signal to enable the signal to be identified. By depressing a push button, the search-oscillator can be stopped at any position in the band until the receiver is adjusted to this frequency using the normal tuning. A meter (switched S-meter) which varies together with the control voltage indicates the momentarily tuned frequency so that the signal of interest can be located quickly.

The described circuits can also be used for scanning the UHF-bands if the UHF-converter uses the 2 metre band as the first intermediate frequency.

2. A SIMPLE AUTOMATIC BAND SCANNER

A photograph of the simple band scanner (DL 9 FX 001) is given in Fig. 1 and the circuit diagram in Fig. 2.

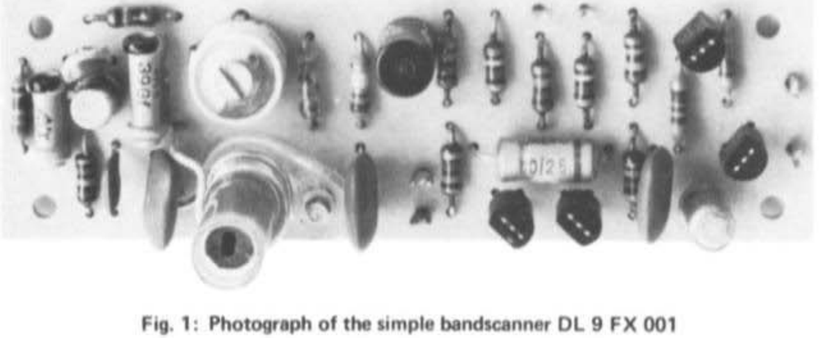


Fig. 1: Photograph of the simple bandscanner DL 9 FX 001

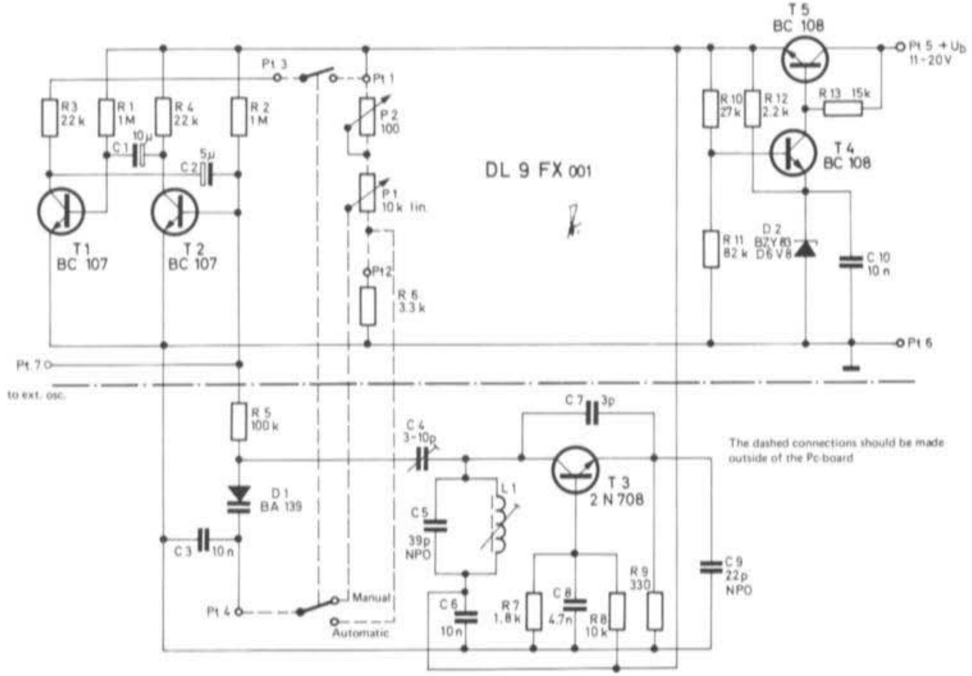


Fig. 2: Circuit diagram of the simple band scanner (DL 9 FX 001)

2.1. CIRCUIT DETAILS

Transistors T 1 and T 2 together with the associated components R 1, R 2, R 3, R 4, C 1 and C 2 form an astable multivibrator circuit. The operation of such a circuit is well-known and need not be described here. At all times, one transistor is conducting and the other is blocked. Whereas the voltage at the collectors has a squarewave form, the voltage characteristic at the base of these transistors has a sawtooth waveform. The maximum voltage value is approximately equal to the operating voltage of the multivibrator and is negative with respect to the emitter in the case of the NPN transistors used here. This voltage controls the varactor diode D 1 which sweeps the frequency of the receive oscillator. The scanning time t 2 for one scan is determined by the values of the base resistor R 2 and capacitor C 2. It is: t 2 0.7 x R 2 x C 2.

The time constant 1 = R 1 x C 1 determines the stay time t 1. During this period, the oscillator will not be swept, transistor T 2 will be conducting, and the voltage at its base will have fallen to the threshold voltage of approximately 0.6 V. The stay time and the scanning time can be varied within wide limits by correspondingly dimensioning R 1, C 1 and R 2, C 2. Attention must only be paid that the required base current for the driving of the transistors is able to flow. This is guaranteed when R 1 R 3 x B 1 and R 2 R 4 x B 2, where B 1 and B 2 are the current gain values of transistors T 1 and T 2.

The given dimensioning results in approximately 3.5 s for the scan period t 2 and 7 s for the stay time t 1. The stay time has been especially dimensioned somewhat longer than the scanning time in order to clearly establish whether several stations are active on the band.

The varactor diode D 1 is coupled to the resonant circuit of the oscillator via the trimmer capacitor C 4. The required frequency deviation can be adjusted at this point. In order to ensure that the diode is not influenced by the RF voltage of the oscillator at low voltage levels so that it would not be controlled directly, the cathode is connected to approx. 2 V.

It is, of course, easy to use the search oscillator for the tuning of the receiver. When using electronic tuning, the potentiometer P 1 used for tuning can be located at any position in the receiver. In addition to this, it is possible using a second potentiometer P 2 to achieve a fine tuning so that the extensive requirements of a mechanical vernier are not necessary.

Prerequisite for a high degree of frequency stability when using a varactor diode for tuning is a very high voltage stability (2). If one assumes that a tuning voltage of approx. 8 V is available for the tuning range of 144 to 146 MHz, a voltage fluctuation of 100 mV will already cause a frequency variation of over 20 kHz. A simple voltage stabilizing circuit using a zener diode is not sufficient for these demands. For this reason, a control circuit is used which exhibits a low temperature dependence and sufficient stabilization. The control circuit, consisting of diode D 2 and transistors T 4 and T 5, feeds all stages of the automatic band scanner.

The injection of the oscillator voltage is made in the normal manner (2) via a small capacitor to either the base or emitter of the mixer transistor of the converter. The injection to the emitter has the advantage that the connection can be made at low impedance using a coupling coil and a coaxial cable. This means that the automatic band scanner need not be located directly adjacent to the mixer stage.

It is often favourable to convert a receiver designed for variable capacitor tuning for varactor tuning and to equip the receiver with an automatic band scanner. To do this, the variable capacitor of the oscillator is replaced by a varactor diode and the frequency range adjusted to the nominal value by correcting the alignment components. If the alignment range is not sufficient, the required frequency variation can be obtained by providing a small capacitor in series with the varactor diode or by exchanging the present series capacitance.

The printed circuit board of the simple band scanner (DL 9 FX 001), as shown in Fig. 3, is dimensioned so that the oscillator portion can be separated at the dotted line. This means that the length is decreased by approximately half and that the board can be mounted into the receiver even when very little space is available. Fig. 4 gives the component location plan of the previously mentioned printed circuit board DL 9 FX 001. Further assembly instructions are not required for this simple module.

2.2. SPECIAL COMPONENTS FOR DL 9 FX 001

T 1, T 2, T 4, T 5: BC 108, BC 183, 2 N 2926, 2 N 3903 or similar
T 3: 2 N 708, 2 N 918, BF 224, BF 173, BF 115 or similar

D 1: BA 139, BA 121, BA 110, 1 N 5462 A (Motorola), approx. 16 pF at 2 V)
D 2: 1 N 754, 1 N 4099, BZY 85/C6V8, ZF 6.8 (6.8 V zener diode)

C 4: 3.5 - 13 pF ceramic micro disc trimmer
L 1: 3.25 turns of 1 mm diameter (18 AWG) silver plated copper wire wound onto a 7 mm diameter coil former with VHF core.

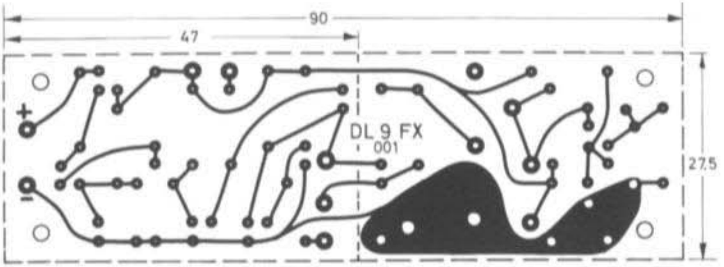


Fig. 3:
Printed circuit board
DL 9 FX 001

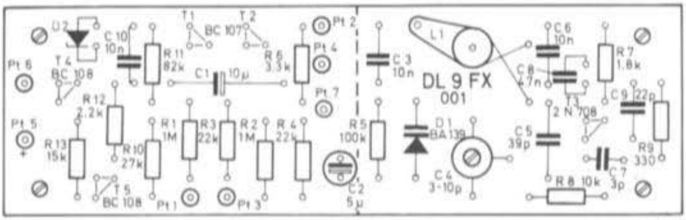


Fig. 4:
Component location plan
to DL 9 FX 001

2.3. PRELIMINARY OPERATION OF THE SIMPLE AUTOMATIC BAND SCANNER

After checking the voltage stabilizer by varying the operating voltage between 11 V and 20 V, whereby the stabilized voltage should not alter noticeably, the LC oscillator is coarsely tuned to the required frequency. With the given values for L 1 and C 5, the oscillator will operate at approx. 116 MHz so that an intermediate frequency range of 28 - 30 MHz will result at a tuning range of 144 - 146 MHz. If other frequencies are required, inductance L 1 and / or C 5 should be correspondingly dimensioned.

The fine alignment is made in the "manual tuning" mode. Potentiometer P 1 is rotated fully left or fully right towards the connection to potentiometer P 2. This is followed by aligning inductance L 1 to the upper frequency limit (P 2 in a central position). Potentiometer P 1 is then brought to the opposite stop and the lower limit frequency adjusted using trimmer capacitor C 4. The alignment should be corrected several times until the required frequency range is obtained.

If manual tuning is not required, a resistor of 10 kΩ should be connected between connection points Pt 1 and Pt 2. The alignment is made in a similar manner observing signals at the band limits.

- to be continued -

AVAILABLE PARTS

The printed circuit boards, semiconductors, coil formers, trimmer capacitors and a kit are available from the publishers or their national representatives. See advertising page.

REFERENCES

(1) H. Wilhelm: An Automatic Search Oscillator For 2 Metre Converters
VHF COMMUNICATIONS 1 (1970), Edition 4, Pages 215-217

(2) H. J. Franke: Stable Reference Voltages
VHF COMMUNICATIONS 2 (1970), Edition 2, Pages 76-86

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Crystal filter	XF-9A	(for SSB) with both sideband crystals	DM 106.--
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Crystal filter	XF-9D	(for AM; 5.00 kHz)	DM 137.--
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Crystal filter	XF-9M	(for CW; 0.5 kHz) with carrier crystal ...	DM 106.--
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Crystal	96.0000 MHz (HC- 6/U) for 70 cm converters	(DL 9 GU, DL 9 JU)	DM	21.50
Crystal	96.0000 MHz (HC-25/U) for 70 cm converters	(DL 9 GU, DL 9 JU)	DM	28.--
Crystal	95.8333 MHz (HC-18/U) for 70 cm converters	(DC 6 HY)	DM	28.--
Crystal	84.5333 MHz (HC- 6/U) for 24 cm converters	(DL 3 WR)	DM	21.50
Crystal	65.5000 MHz (HC- 6/U) for 2 m converters	(DL 6 HA)	DM	16.50
Crystal	65.0000 MHz (HC- 6/U) for 2 m converters	(DL 6 HA)	DM	16.50
Crystal	46.3333 MHz (HC-18/U) for phase-locked oscillator	) set	DM	49.--
Crystal	46.0000 MHz (HC-18/U) (DJ 7 ZV / DJ 9 ZR)	)		
Crystal	45.4780 MHz (HC-18/U) for VXO	(DJ 9 ZR)	DM	24.50
Crystal	42.0000 MHz (HC- 6/U) for 70 MHz converters	(G 3 JHM)	DM	15.60
Crystal	38.6667 MHz (HC- 6/U) for 2 m converters	(DL 6 SW, DL 6 HA)	DM	13.70
Crystal	27.8000 MHz (HC-18/U) for 24 MHz synthesis VFOs	(DL 3 WR 007)	DM	25.--

Standard frequency crystals

5.0000 MHz (HC- 6/U) for calibration spectrum generators (DC 6 HY)	DM	25.--
1.0000 MHz (HC- 6/U) for calibration spectrum generators (DJ 4 BG, DC 6 HY)	DM	20.50
100 kHz (HC-13/U) for calibration spectrum generators (DC 6 HY)	DM	28.--

Crystals 72.... MHz (HC-6/U) for 2 metre transmitters (DJ 1 NB, DL 3 WR)
Please state required frequency on ordering
(Delivery 4 to 6 weeks) .... DM 21.50

Crystals also available ex stock please see list with DJ 1 NB kit.

Crystals other frequencies available (Please state frequency and type)
(Delivery 6 to 8 weeks)

DL 6 HA 001/28	2 m MOSFET Converter (IF = 28-30 MHz)	Ed. 1/70
PC-board	DL 6 HA 001 (with printed plan)	DM 6.--
Minikit	DL 6 HA 001 (coil formers and trimmer set)	DM 5.20
Semiconductors	DL 6 HA 001 (5 transistors)	DM 20.50
Crystal	38.66667 MHz (HC-6/U)	DM 13.70
Kit	DL 6 HA 001 with above listed components	DM 45.40
Ready-to-operate	2 m MOSFET converter DL 6 HA 001/28	DM 134.80
DL 6 HA 001/14	2 m MOSFET converter (IF = 14-15 MHz)	Ed. 1/70
PC-board	DL 6 HA 001 (with printed plan)	DM 6.--
PC-board	DL 3 YK 001 (with printed plan) (Ed. 4/70)	DM 4.--
Minikit	DL 6 HA 001/14 (coil formers and trimmer set, 2 crystal holders)	DM 11.30
Crystal	65.00000 MHz (HC-6/U)	DM 16.50
Semiconductors	DL 6 HA 001/14 (7 transistors)	DM 24.50
Crystal	65.50000 MHz (HC-6/U)	DM 16.50
Kit	DL 6 HA 001/14 with above listed components	DM 78.80
DL 6 HA 002	9 MHz SSB Transceiver	Ed. 2/70
PC-board	DL 6 HA 002 (with printed plan)	DM 16.--
Minikit	DL 6 HA 002 (coil formers and trimmer set)	DM 6.--
Semiconductors	DL 6 HA 002 (12 transistors, 7 diodes)	DM 46.80
Kit	DL 6 HA 002 with above listed components	DM 68.80
Crystal filter	XF-9A with both sideband crystals	DM 106.--
Crystal filter	XF-9B with both sideband crystals	DM 137.--
DL 6 HA 003	9 MHz Carrier Oscillator	Ed. 2/70
PC-board	DL 6 HA 003 (with printed plan)	DM 3.--
Minikit	DL 6 HA 003 (coil formers and trimmer set, 2 transistors, 2 diodes)	DM 7.50
Kit	DL 6 HA 003 with above listed parts	DM 10.50
DL 6 HA 004	9/14 MHz Transmit-Receive Converter	Ed. 3/70
PC-board	DL 6 HA 004 (with printed plan)	DM 16.--
Minikit	DL 6 HA 004 (coil formers, trimmer and choke set)	DM 8.70
Semiconductors	DL 6 HA 004 (8 transistors, 5 diodes)	DM 49.10
Kit	DL 6 HA 004 with above listed components	DM 73.80
DL 6 HA 005	14/145 MHz Transmit Converter	Ed. 3/70
PC-board	DL 6 HA 005 (with printed plan)	DM 8.--
Minikit	DL 6 HA 005 (coil formers, trimmer and choke set)	DM 4.80
Semiconductors	DL 6 HA 005 (4 transistors)	DM 18.--
Kit	DL 6 HA 005 with above listed components	DM 30.80
DL 6 HA 006	V F O (5-6 MHz)	Ed. 3/70
PC-board	DL 6 HA 006 (with printed plan)	DM 8.--
Minikit	DL 6 HA 006 (2 transistors, 1 ceramic coilformer) ..	DM 6.50
Variable capacitor	100 pF (48.5 x 48.5 x 49.5 mm, two bearings)	DM 13.50 +
Kit	DL 6 HA 006 with above listed components	DM 28.--
DL 6 HA 007	Low-pass Filter	Ed. 3/70
PC-board	DL 6 HA 007 (with printed plan)	DM 3.--
Minikit	DL 6 HA 007 (coil former set, 1 transistor)	DM 3.20
Kit	DL 6 HA 007 with above listed components	DM 6.20
DL 6 HA 008	Stabilized Power Supply	Ed. 4/70
PC-board	DL 6 HA 008 (with printed plan)	DM 8.--
Minikit	DL 6 HA 008 (choke set, 5 transistors, 6 diodes) ...	DM 39.70
Kit	DL 6 HA 008 with above listed components	DM 47.70

DL 6 HA 009	Transformerless AF Amplifier (approx. 2.5 W)	Ed. 4/70
PC-board	DL 6 HA 009 (with printed plan)	DM 8.--
Semiconductors	DL 6 HA 009 (4 transistors, 1 diode)	DM 14.50
Kit	DL 6 HA 009 with above listed components	DM 22.50
All Kits	DL 6 HA 001/14 to 009 (without crystal filter).......	DM 350.--
All PC-boards	DL 6 HA 001 to 009 and DL 3 YK 001	DM 70.--

DJ 9 ZR KITS — TWO METRE SSB TRANSCEIVER

DJ 9 ZR 001	5 W SSB Transmitter for 145 MHz	Ed. 2/69
PC-board	DJ 9 ZR 001 (with printed plan)	DM 15.--
Minikit	DJ 9 ZR 001 (coil formers, trimmer and choke set)..	DM 7.50
Semiconductors	DJ 9 ZR 001 (9 transistors, 3 diodes)	DM 104.30
Kit	DJ 9 ZR 001 with above listed components	DM 126.80
Potted core kit	DJ 9 ZR 001 (for DC-DC converter)	DM 15.--
Crystal filter	XF-9A with both sideband crystals	DM 106.--
Crystal filter	XF-9B with both sideband crystals	DM 137.--

DJ 9 ZR 002	V X O (136.446 MHz + 20 kHz)	Ed. 2/69
PC-board	DJ 9 ZR 002 (with printed plan)	DM 3.--
Minikit	DJ 9 ZR 002 (coil former set, 3 transistors, 2 diodes)	DM 13.40
Crystal	45,478 MHz (HC-18/U)	DM 24.50
Kit	DJ 9 ZR 002 with above listed components	DM 40.90

DJ 9 ZR 005	IF-AF Portion (9 MHz)	Ed. 3/69
PC-board	DJ 9 ZR 005 (double coated with through contacts)	DM 42.--
Minikit	DJ 9 ZR 005 (coil formers, transformer and choke set)	DM 36.90
Semiconductors	DJ 9 ZR 005 (5 ICs, 2 transistors, 8 diodes)	DM 74.10
Kit	DJ 9 ZR 005 with above listed components	DM 153.--
Crystal filter	XF-9B with both sideband crystals	DM 137.--
Precision spindle	potentiometer 500 Ohm	DM 9.--

DJ 9 ZR 006	VHF Portion (145/9 MHz)	Ed. 3/69
PC-board	DJ 9 ZR 006 (teflon (PTFE) with printed inductances).	DM 43.50
Minikit	DJ 9 ZR 006 (coil formers and trimmer set)	DM 10.80
Semiconductors	DJ 9 ZR 006 (3 transistors)	DM 31.50
Kit	DJ 9 ZR 006 with above listed components	DM 85.80

DJ 9 ZR 008	Electronically Stabilized Power Supply	Ed. 3/70
PC-board	DJ 9 ZR 008 (with printed plan)	DM 14.--
Semiconductors	DJ 9 ZR 008 (5 transistors, 14 diodes, heatsink).....	DM 58.30 +
Potted core kit	DJ 9 ZR 008 (for DC-DC converter)..............	DM 12.50 +
Kit +	DJ 9 ZR 008 with above listed components	DM 84.80

DJ 9 ZV 001+002	Phase-locked Oscillator (135-137 MHz)	Ed. 1/69
PC-boards	DJ 7 ZV 001 + 002 (with printed plan)	DM 11.--
Minikit	DJ 7 ZV 001 + 002 (coil former set)	DM 3.20
Semiconductors	DJ 7 ZV 001 + 002 (10 transistors, 6 diodes)	DM 33.50
Crystals	46.0000 + 46.3333 MHz (HC-18/U)	DM 49.--
Kit	DJ 7 ZV 001 + 002 with above listed components	DM 96.70

CONVERTERS and TRANSVERTERS (See also DL 6 HA and DJ 9 ZR kits)

DL 6 SW 004/145	Two Metre FET Converter	Ed. 1/69
PC-board	DL 6 SW 004 (with printed plan)	DM 6.--
Minikit	DL 6 SW 004 (coil formers and trimmer set)	DM 5.20
Semiconductors	DL 6 SW 004 (5 transistors)	DM 19.--
Crystal	38.66667 (HC-6/U)	DM 13.70
Kit	DL 6 SW 004 with above listed components	DM 43.90

DL 6 SW 004/70	FET Converter for 70 MHz	Ed. 2/69
PC-board	DL 6 SW 004 (with printed plan)	DM 6.--
Minikit	DL 6 SW 004 (coil formers and trimmer set)	DM 5.20
Semiconductors	DL 6 SW 004 (5 transistors)	DM 19.--
Crystal	42.00000 MHz (HC-6/U)	DM 15.60
Kit	DL 6 SW 004/70 with above listed components	DM 45.80

DJ 6 ZZ 001	28 MHz - 145 MHz Transverter with FET Mixers	Ed. 4/69
PC-board	DJ 6 ZZ 001 (with printed plan)	DM 15.--
Minikit	DJ 6 ZZ 001 (coil formers and trimmer set)	DM 8.40
Semiconductors	DJ 6 ZZ 001 (12 transistors)	DM 39.--
Crystal	38.66667 (HC-6/U)	DM 13.70
Kit	DJ 6 ZZ 001 with above listed components	DM 76.10

DL 9 GU 001	70 cm Converter (IF = 145 MHz)	Ed. 2/69
PC-board	DL 9 GU 001 (with printed plan)	DM 6.--
Minikit	DL 9 GU 001 (coil formers and trimmer set, 2 disc and 3 tubular trimmers)	DM 8.10
Semiconductors	DL 9 GU (5 transistors)	DM 35.--
Crystal	96.000 MHz (HC-6/U)	DM 21.50
Casing	DL 9 GU 001 silver-plated with inner conductors and trimmers (delivery max. 6-8 weeks)	DM 65.--
Kit	DL 9 GU 001 with above listed components (but without tubular trimmers)	DM 129.60
Ready-to-operate	70 cm converter DL 9 GU 001 (delivery max. 6-8 weeks)	DM 196.50

DL 3 WR 001	Receive Converter 1296/28 MHz	Ed. 1/69
PC-board	DL 3 WR 001 (with printed plan)	DM 2.--
Minikit	DL 3 WR 001 (2 screening cans with coil formers, 6 trimmers)	DM 15.--
Casing	DL 3 WR 001 silver-plated with inner conductors (delivery max. approx. 3 months)	DM 136.--
Crystal	84.5333 MHz (HC-6/U)	DM 21.50
Kit	DL 3 WR 001 with above listed components	DM 175.--

DC 6 HY 001	Receive Converter 432/144 MHz)	Ed. 4/70
PC-board	DC 6 HY 001 (with printed plan)	DM 10.--
Minikit	DC 6 HY 001 (coil formers and trimmer set)	DM 13.20
Semiconductors	DC 6 HY 001 (5 transistors)	DM 20.30
Crystal	95.833 MHz (HC-18/U)	DM 28.--
Kit	DC 6 HY 001 with above listed components	DM 71.50
Ready-to-operate	receive converter DC 6 HY 001 (slight delay)	DM 179.60

DC 6 HY 002	Transmit Converter 144/432 MHz	Ed. 4/70
PC-board	DC 6 HY 002 (with printed plan)	DM 10.--
Minikit	DC 6 HY 002 (trimmer set and ferrite beads)	DM 21.50
Semiconductors	DC 6 HY 002 (3 transistors)	DM 18.50
Kit	DC 6 HY 002 with above listed components	DM 50.--
Ready-to-operate	transmit converter DC 6 HY 002 (slight delay)	DM 149.50

DC 6 HY	432 MHz Linear Amplifier	Ed. 4/70
Minikit	DC 6 HY/Lin (trimmer set and tube socket)	DM 10.90
Semiconductors	DC 6 HY/Lin (1 transistor, 1 diode)	DM 4.20
Tube	EC 8020	DM 27.--
Kit	DC 6 HY/Lin with above listed components	DM 42.10
Ready-to-operate	linear amplifier DC 6 HY (slight delay)	DM 142.40

TRANSMITTERS, VFOs and OSCILLATORS
(see also DL 6 HA- and DJ 9 ZR Kits as well as DJ 7 ZV, DC 6 HY

DJ 1 NB 004	2 W AM Transmitter for 145 MHz	Ed. 3/69
PC-board	DJ 1 NB 004 (with printed plan)	DM 10.--
Minikit	DJ 1 NB 004 (trimmer and choke set)	DM 12.60
Semiconductors	DJ 1 NB 004 (5 transistors, 2 diodes)	DM 56.50
Kit	DJ 1 NB 004 with above listed components	DM 79.10
Crystal	72,.... MHz (HC-6/U) on request, please state required frequency (delivery approx. 6 weeks)	DM 21.50
Crystals available ex. stock	72.0500 MHz / 72.0750 MHz / 72.1000 MHz / 72.1250 MHz 72.1500 MHz / 72.2000 MHz / 72.2500 MHz / 72.3000 MHz 72.3500 MHz / 72.4000 MHz / 72.4500 MHz / 72.5000 MHz when ordering please indicate frequency	DM 21.50

Modulation transformer kit: see DL 3 WR 003 transmitter

DL 3 WR 003	2 W Transmitter for 145 MHz for VFO Operation	Ed. 2/70
PC-board	DL 3 WR 003 (with printed plan)	DM 17.50
Minikit	DL 3 WR 003 (trimmer and choke set)	DM 14.40
Semiconductors	DL 3 WR 003 (11 transistors, 5 diodes)	DM 93.10
Modulation transformer kit	DL 3 WR 003 L for PC-board mounting	DM 7.50
Kit	DL 3 WR 003 with above listed components	DM 132.50
Crystals	72,.... MHz see DJ 1 NB 004	DM 21.50

DL 3 WR 007	24 MHz Synthesis VFO with FM Attachment +	Ed. 3+4/70
PC-board	DL 3 WR 007 (with printed plan)	DM 5.--
Minikit	DL 3 WR 007 (4 trimmers and 1 variable ceramic coil 1.8 - 2.5 µH)	DM 25.--
Semiconductors	DL 3 WR 007 (5 transistors, 2 diodes)	DM 30.50
Variable capacitor	100 pF (48,5 x 48,5 x 49,5 mm) two bearings	DM 13.50
Potted core kits	DL 3 WR 007 (4 potted cores)	DM 16.40
Crystal	27.800 MHz (HC-18/U)	DM 25.--
Kit	DL 3 WR 007 with above listed components	DM 115.40

DJ 8 PG 001	72 MHz VFO for FM Transmitters	Ed. 4/70
PC-board	DJ 8 PG 001	DM 2.50
Minikit	DJ 8 PG 001 (coil formers, trimmer and choke set)..	DM 3.85
Semiconductors	DJ 8 PG 001 (2 transistors, 1 diode)	DM 13.--
Kit	DJ 8 PG 001 with above listed components	DM 19.35

DL 3 YK 001	Oscillator Board for DL 6 HA 001	Ed. 4/70
PC-board	DL 3 YK 001 (with printed plan)	DM 4.--
Minikit	DL 3 YK 001 (coil former set, 2 crystal holders) ...	DM 6.10
Semiconductors	DL 3 YK 001 (2 transistors)	DM 4.--
Crystal	65.000 MHz (HC-6/U)	DM 16.50
Crystal	65.500 MHz (HC-6/U)	DM 16.50
Kit	DL 3 YK 001 with above listed components	DM 47.10

X Kit included in DL 6 HA 001/14. Serves to modify the 28-30 MHz IF to 14-15 MHz.

DK 1 PN	10 W Transmitter for 435 MHz ^{x})	Ed. 4/69
Minikit	DK 1 PN/a (trimmer set, diode BA 110, tube socket)	DM 29.90
Tube	EC 8020	DM 27.--
Kit	DK 1 PN/a with above listed components	DM 56.90

^{x}) see also DC 6 HY linear

DK 1 PN	Varactor Tripler 70 cm - 24 cm	Ed. 3/70
Minikit	DK 1 PN/b (trimmer set)	DM 10.--

DL 3 WR 004	RF-Voltage Indicator	Ed. 2/70
PC-board	DL 3 WR 004 (with printed plan)	DM 2.50
Kit	DL 3 WR 004 (PC-board, trimmer and diode)	DM 4.25
DL 9 FX 001	Simple Band Scanner	Ed. 4/70
PC-board	DL 9 FX 001 (with printed plan)	DM 4.--
Minikit	DL 9 FX 001 (coil formers and trimmer set, 5 transistors, 2 diodes)	DM 18.80
Kit	DL 9 FX 001 with above listed components	DM 22.80
DJ 4 BG 003	Calibration Spectrum Generator	Ed. 1/70
PC-board	DJ 4 BG 003 (with printed plan)	DM 9.--
Minikit	DJ 4 BG 003 (coil formers and trimmer set, Poti crystal holder)	DM 6.20
Semiconductors	DJ 4 BG 003 (6 transistors, 1 diode)	DM 15.70
Standard frequency	crystal 1.000 MHz (HC-6/U)	DM 20.50
Kit	DJ 4 PG 003 with above listed components	DM 51.40
DC 6 HY 003	Calibration Spectrum Generatur	Ed. 4/70
PC-board	DC 6 HY 003 (with printed plan)	DM 4.--
Minikit	DC 6 HY 003 (1 trimmer, 5 transistors)	DM 16.50
(Standard frequency	crystal 1.000 MHz (HC-6/U)	DM 20.50
(Kit	DC 6 HY 003/1 MHz with above listed components ...	DM 41.20
(Standard frequency	crystal 100 kHz (HC-13/U)	DM 28.--
(Kit	DC 6HY 003/100 kHz as above	DM 48.70
(Standard frequency	crystal 5 MHz (HC-6/U)	DM 25.--
(Kit	DL 6 HY 003/ 5 MHz as above	DM 45.70
DJ 4 BG 002	Digital Discriminator	Ed. 2/70
PC-board	DJ 4 BG 002 (with printed plan)	DM 5.--
Semiconductors	DJ 4 BG 002 (1 IC, 4 transistors, 2 diodes)	DM 17.50
Kit	DJ 4 BG 002 (with above listed components)	DM 22.50
DJ 4 BG 006	Speech Processor	Ed. 1/71
PC-board	DJ 4 BG 006 (with printed plan)	DM 7.--
Semiconductors	DJ 4 BG 006 (1 IC, 2 transistors)	DM 17.80
Kit	DJ 4 BG 006 with above listed components	DM 24.80
Connectors	Siemens (13-pole) for PC-board DJ 4 BG 006, set ...	DM 7.40
DL 3 WR 002	Electronic Fuse 24 V	Ed. 3/69
PC-board	DL 3 WR 002 (with printed plan)	DM 1.50
Semiconductors	DL 3 WR 002 (3 transistors, 1 diode)	DM 11.20
Kit	DL 3 WR 002 with above listed components	DM 12.70

OTHER EQUIPMENT and COMPONENTS

HB 9 CV antenna for 145 MHz, chrome-plated, detachable including post and packing (surface mail) .... DM 40.--

Interdigital bandpass filter for the 23 cm band, similar to the description in QST, March 1968; ready to operate with BNC connectors ...... DM 82.30
without connectors (delivery 6 to 8 weeks) .... DM 76.20

Completely ready-to-operate equipment

2 m converter	DL 6 HA 001 (IF: 28-30 MHz)	Ed. 1/70	DM 134.80
70 cm converter	DL 6 GU 001 (IF: 144-146 MHz)	Ed. 2/69	DM 196.50
70 cm receive converter	DC 6 HY 001	Ed. 4/70	DM 179.60
70 cm transmit converter	DC 6 HY 002	Ed. 4/70	DM 149.50
70 cm linear amplifier	DC 6 HY with EC 8020	Ed. 4/70	DM 142.40

tube EC 8020	for 70 cm transmitters DK 1 PN, DC 6 HY	DM 27.--
UHF tube socket	made of silicon glass fibre for EC 8020	DM 2.90
Dual-Gate MOSFET	with integrated protective diodes 40673 (RCA)	DM 12.50

Connectors for PC-boards

SIEMENS	13-pole connectors .... set	DM 7.40
SIEMENS	21-pole connectors .... set	DM 11.20
Potted core kit	for DC-DC converter (SSB transmitter DJ 9 ZR 001)	DM 15.--
Potted core kit	for DC-DC converter (power supply DJ 9 ZR 008)	DM 12.50
Potted core kit	für VFO (DL 3 WR 007) .... each	DM 4.10
Transformer kit	DL 3 WR 003/L horizontal PC-board mounting	DM 7.50
Transformer kit	DL 3 WR 003/S vertical PC-board mounting	DM 7.50

Various epoxy glass-fibre PC-boards with silver-plated conductorlanes.
Available for:

2 m bandpass filter	Ed. 4/69	DJ 4 KH 001	DM 1.--
AF universal filter	Ed. 4/69	DJ 4 BG 001	DM 3.--

Perforated experimental boards 2.5 mm spacing .... DM 6.--
Special 1.3 mm drills for epoxy PC-boards will be supplied free with orders above DM 40.-- .... DM 3.--

HANDBOOKS FOR ENGINEERS AND RADIO AMATEURS

RADIO COMMUNICATION HANDBOOK (RSGB)	incl. postage	DM 35.--
VHF - UHF MANUAL (RSGB)	incl. postage	DM 13.40
Empfangstechnik im UHF-Bereich, F. Möhring	(German)	DM 14.80
Fernseh-Servicetechnik, F. Möhring	(German)	DM 28.90
PAL-Farbfernsehtechnik, F. Möhring	(German)	DM 24.80
PAL-Farbfernseh-Servicetechnik, F. Möhring	(German)	DM 36.--
Schaltungstechnik von Schwarz-Weiß-Fernseh-Empfängern	(German)	DM 16.80
Taschenbuch für Funkamateure, Kortz/Lentz	(German)	DM 19.80
Antennenbuch, Rothammel	(German)	DM 24.--

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capacity 12 editions (3 volumes) incl. surface post and packing .... DM 5.--

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Some delays may be caused at the moment due to delivery difficulties at the manufacturers. Equivalent semiconductor types will be supplied if original types are not available.

Semiconductors, quartz crystals and crystal filters can not be exchanged.

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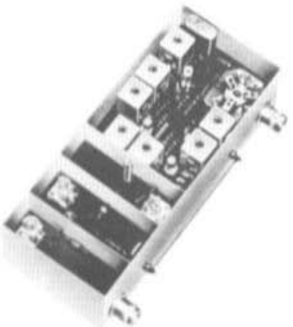
High performance equipment from Braun

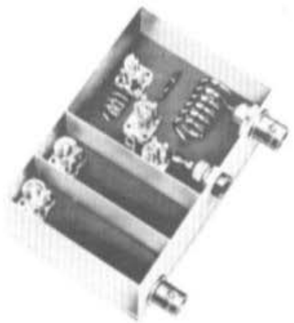
Two Metre Transceiver SE 600

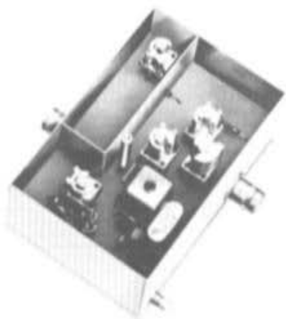
A selective two metre transceiver for all operating modes having a very low noise figure and extremely high cross-modulation rejection.
True transceiver operation or separate operation of transmitter and receiver are possible. Transmitter and receiver can be individually switched to the following modes: CW, LSB, USB, AM and FM. The separate operation and the possibility of selecting either LSB or USB make the transceiver suitable for operation with balloon carried translators or satellites.
Separate crystal filter for transmitter and receiver. True AM using plate/screen grid modulation of the PA tube. Bulit-in clipper. Crystal discriminator for FM demodulation, with IC limiting. Product detector for SSB. VOX and anti-trip. RF output and S meter. Built-in antenna relay. Power supply for 115 and 220 volt as well as a DC-DC converter for 12 volt are built in.
Price: DM 2725.—

Two Metre Converter DGTC 22

High performance dual-gate MOSFET converter. Very high sensitivity and cross-modulation rejection. Highest possible spurious signal rejection by using a 116 MHz crystal.
Price: DM 122,—







70-cm-Converter DGTC 1702

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

144 MHz/432 MHz Tripler LVV 270

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

144 MHz/432 MHz Transverter TTV 1270

This unit represents — in conjunction with a two metre station — the quickest and simplest means of becoming active on 70 cm. It has been especially developed for portable operation.
No antenna switching is required! The transverter is simply connected between the two metre station and the 70 cm antenna. Completely screened, silver-plated brass casing.
Input and output: 60 ohm BNC connectors. **Price: DM 142,—**

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