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VHF COMMUNICATIONS

A PUBLICATION FOR THE RADIO AMATEUR
ESPECIALLY COVERING VHF, UHF AND MICROWAVES

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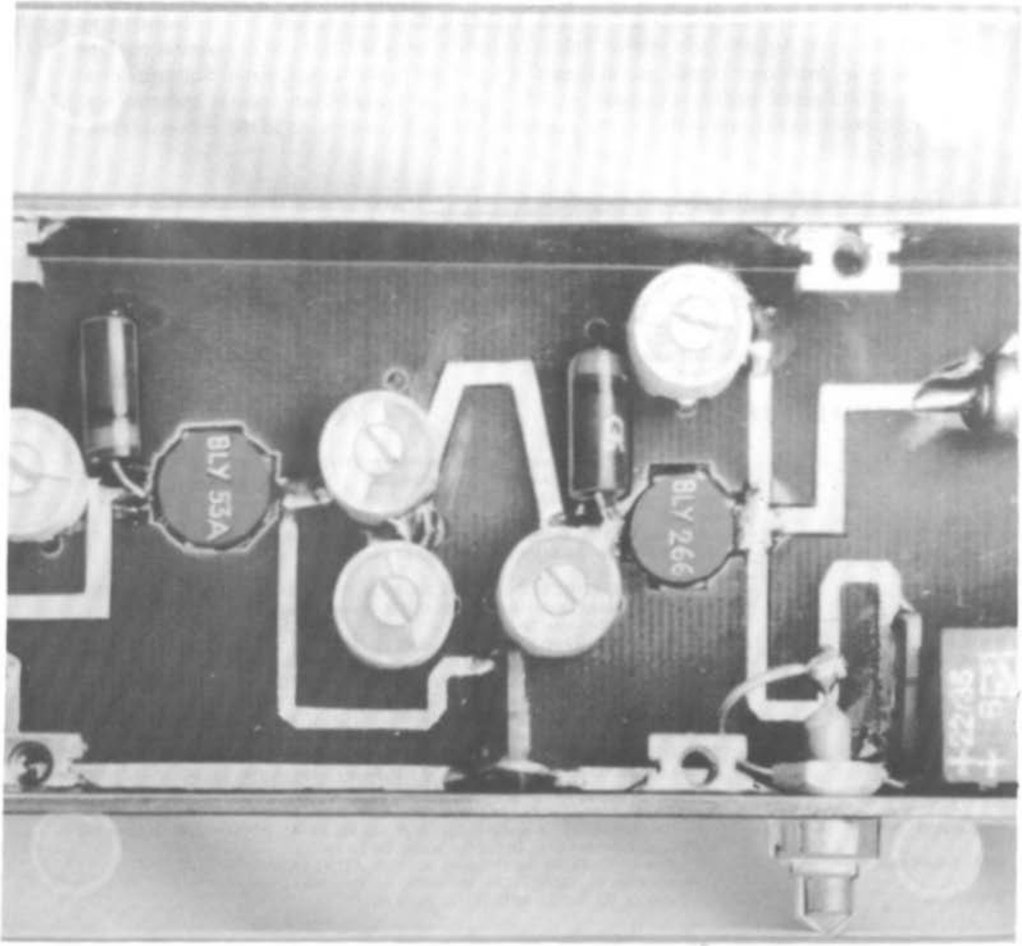

VHF
COMMUNICATIONS

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LATEST AMSAT NEWS

Due to delays with the various systems of OSCAR B (A-O-B) which were not expected to be completed in time for the scheduled launch in July, it has been decided to prepare a second, simplified spacecraft from the available modules to meet the launch date.

A-O-C will carry the two-to-ten metre linear translator and 24 channels morse telemetry encoders. It is also planned to include the WIA-Australis 35-function command decoder.

If completed in time, the following modules may also be launched: CODESTORE a shift-register memory for storage of up to 15 words of morse and retransmission on the 29.450 MHz beacon frequency. A 435.10 MHz beacon with an output of 350-450 milliwatts.

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

PART II: CONSTRUCTION AND ALIGNMENT

by G. Otto, DC 6 HL

The characteristics and circuits of this portable or fixed transceiver were described in Part I of this article. This is now to be followed by details of the components, construction and alignment. The FM-attachment will be described in Part III.

3. CONSTRUCTION

Since the layout of the transceiver is dependent on many preferences and requirements, this description is to be limited to the construction of the individual modules. However, the photograph in Part I could be of some assistance for the total layout. The construction and component details are given module for module.

3.1. THE MAIN BOARD DC 6 HL 001

Figure 9 which is given on the centre pages of this magazine, shows the printed circuit board DC 6 HL 001. The dimensions of this board are 245 mm x 65 mm. It accommodates the main part of the receiver and transmitter circuit shown in Figure 3 of Part I. A photograph of the completed module is given in Figure 10. A number of thin screening plates are provided to screen the various stages from another. The dimensions and positions of the screening panels are given on the component location plan in Figure 9. The height amounts to 18 mm. They are grounded and mounted with the aid of short pieces of wire that are placed through the board at the positions shown in Figure 9 where they are soldered to the ground surface. This arrangement represents a slight disadvantage for the receiver: The excellent ultimate selectivity of the crystal filter XF-9 B cannot be utilized to the full with a PC-board construction without use of considerable screening measures. Of course, this disadvantage is present with all equipment with an open PC-board construction. A certain improvement can be obtained with the screened 9 MHz coupling circuits in the IF-amplifier. However, this disadvantage can only be avoided by complete screening. It should be mentioned, that the selectivity of this transceiver is by no means inferior to any previously published equipment and it is considered that the improvement that can be obtained is not worth the extensive screening measures required.

The power output transistor T 118 and its collector circuit are mounted on a 1 mm thick copper cooling plate (heat sink) that is flatly mounted onto the PC-board. The longest sides are bent up to a height of 18 mm. The shape of the heat sink is given in the component location plan and room is provided on the PC-board for this. Trimmer capacitor C 193 is mounted in a cut-out in the heat sink since both connections are at RF voltage. The thermistor R 192 is mounted on the heat sink in the vicinity of the output transistor. Since this thermistor is grounded at one end, it is not necessary for it to be insulated. It is important that the connection leads of the driver transistor T 117 are kept as short as possible. This, as well as a sufficiently high local oscillator voltage at the ring mixer (and an optimum alignment), are decisive for the output power value.

3.1.1. SPECIAL COMPONENTS FOR DC 6 HL 001

The smallest available components should be used for construction.

T 101...T 103: 40673, 40820, 3N187 (protected gate, dual-gate MOSFET (RCA)
T 104 - T 106: 40602, 40604 (RCA) or gate-protected types
T 107 - T 108: BF 245 A (Texas Instruments Germany, Siliconix), W 245 A (Siliconix)
T 109 - T 111: BC 167 - BC 169, BC 107 - BC 109, 2 N 2926
T 112 - T 113: BSX 39 (Fairchild), BSX 20 (Philips), BF 224 (Texas Instruments Germany), 2 N 3304
T 114 - T 115: BFY 90 (Philips), BF 224 (Texas Instruments Germany)
T 116 : 40673 (RCA)
T 117 : 2 N 4427 (RCA, Motorola) with cooling fins
T 118 : 2 N 5641 (Motorola)

D 101 - D 102: OA 90 (Philips), 1 N 914, 1 N 4148 or similar
D 103 - D 107: BAY 83 (Fairchild), 1 N 914, 1 N 4148
D 108 : ZF 6, 2 (ITT-Intermetall), BZY 85/C6V2, 1 N 753
D 109 - D 119: BAY 83, 1 N 914, 1 N 4148 or similar
D 120 - D 123: 1 N 4148

L 101 - L 104: 6 turns of 0.8 mm diameter (20 AWG) silver-plated copper wire wound on a 5 mm former, self-supporting.
L 101: coil tap 2 turns from the cold end.
L 102 - L 104: without coil tap.
L 105 : 6 turns as L 101 except coil tap 4.5 turns from the cold end
L 106 : 7 turns of 0.35 mm diameter (27 AWG) enamelled copper wire wound on a 5 mm coilformer with core (red), coil tap 4 turns from the cold end.
L 107 - L 110: 10.7 MHz IF-filters for transistor receivers (Type FM-FB) with built-in capacitors and approx. 22 pF external capacitance
L 111 : 15 turns of 0.35 mm diameter (27 AWG) enamelled copper wire wound on a 5 mm coilformer with core (red).
L 112 : 4 turns of 0.35 mm diameter (27 AWG) enamelled copper wire wound onto L 111.
L 113 : 30 turns of 0.2 mm diameter (32 AWG) enamelled copper wire wound on a 5 mm diameter coilformer with core (red), honeycomb wound.
L 114 : 5 turns of 0.2 mm diameter (32 AWG) enamelled copper wire wound onto L 113, with centre tap
L 115 : 5 turns of 0.8 mm diameter (20 AWG) silver-plated copper wire wound on a 5 mm coilformer with core (red), centre tap
L 116 : As L 115 but without tap
L 117 : As L 116 but self-supporting
L 119 : 3.5 turns of 0.8 mm diameter (20 AWG) silver-plated copper wire wound on a 5 mm former, self-supporting
L 120 : 4 turns of 1.5 mm diameter (15 AWG) silver-plated copper wire wound on a 9 mm former, self-supporting
L 121, L 122 : 7 turns of 0.3 mm diameter (29 AWG) enamelled copper wire wound on a 5 mm coilformer with core (red).

Ch 101	: 10 turns of 0.3 mm diameter (29 AWG) enamelled copper wire wound on a ferrite core (Philips 4312 020 30161) or on a core for the IF frequency
Ch 102 - Ch 104	: Ferrite bead with three turns of 0.3 mm diameter (29 AWG) enamelled copper wire, pulled through the bead.
C 101, C 103, C 106, C 109, C 111, C 182, C 194	: 3.5 - 13 pF ceramic disc trimmers of 7 mm dia.
C 120, C 192, C 193	: 2 - 22 pF foil trimmer capacitors of 7 mm diameter.
C 121	: 10 - 40 pF ceramic disc trimmer, 10 mm diameter.

17 ceramic capacitors for bypassing of VHF frequencies (approx. 1 nF)

22 ceramic capacitors for bypassing lower frequencies (4.7 nF to 22 nF)

1	Tantalium electrolytic capacitor (drop type)	1 μF/16 V
6	Tantalium electrolytic capacitors (drop type)	2.2 μF/16 V
1	Tantalium electrolytic capacitor (drop type)	10 μF/16 V
3	Tantalium electrolytic capacitors (drop type)	22 μF/16 V
3	Tantalium or aluminium electrolytic capacitors	47 μF/16 V

R 192: 100 Ω Thermistor

3.2. CARRIER OSCILLATOR DC 6 HL 002

This module is accommodated on printed circuit board DC 6 HL 002 which is shown in Figure 11. The dimensions of this double-coated PC-board are 55 mm x 30 mm. A photograph of the completed module is given in Figure 12. The printed circuit board is not drilled; the components are soldered directly to the conductor lanes. After mounting the components and checking the operation of this module, the printed circuit board DC 6 HL 002 should be provided with side panels of 18 mm in height. These side panels can be manufactured from metal plate or PC-board material. Figure 13 shows the construction in the form a drawing. The cover can be screw-fitted if a nut is soldered to each corner. The operating voltage and the switching voltage for the diodes are fed in via feedthrough capacitors (C 202, C 204, C 213) that have been soldered in the side panels.

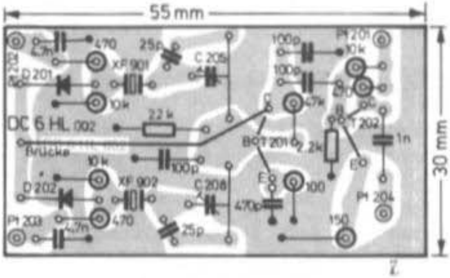


Fig. 11: PC-board DC 6 HL 002

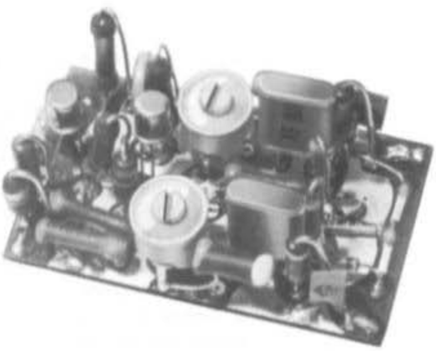


Fig. 12: Photograph of module DC 6 HL 002

3.2.1. SPECIAL COMPONENTS FOR DC 6 HL 002

T 201: BC 109 (low noise AF)
2 N 708 or similar.
T 202: BC 107 - BC 109,
2 N 708 or similar.

D 201;D 202: BAY 83, 1N914, 1N4148

C 205, C 208: 10-40 pF ceramic disc trimmers, 10 mm dia.
3 ceramic feedthrough capacitors for solder mounting, approx. 2 nF

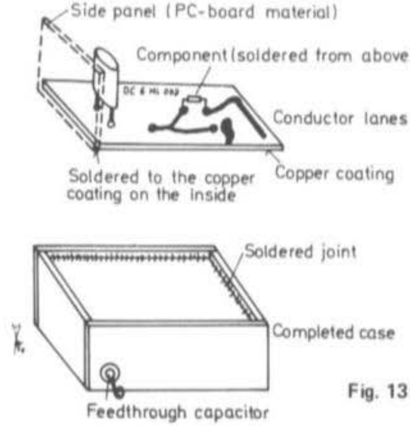


Fig. 13: Construction principle for modules DC 6 HL 002 + 003

3.3. LOCAL OSCILLATOR MODULE DC 6 HL 003

This printed circuit board contains the circuit shown in Figure 6 of Part I. It is accommodated on a double-coated PC-board in a similar manner to that of module DC 6 HL 002. Figure 14 gives the component locations on the 115 mm by 40 mm PC-board DC 6 HL 003. The components are mounted on the printed side of the board in the same manner as for module 002 and the components directly soldered to the conductor lanes. This means that the PC-board should not be drilled. It is only the large circles marked with a cross, that are drilled with a 5 mm diameter drill for mounting the coilformers. With the exception of this, the copper surface on the lower side of the PC-board remains as screening surface.

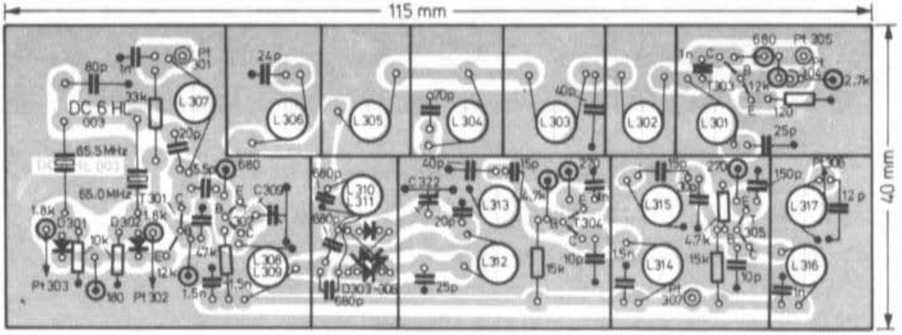


Fig. 14: PC-board DC 6 HL 003

This module is also provided with a number of thin, 18 mm high brass screening panels. The shape and size of these panels can be taken from the component location plan. The panels are provided with small cut-outs with the aid of a round file where no ground surface is present. At other positions, it can be directly soldered to the ground surface. The module is also provided with screening panels which are soldered to the edges of the printed circuit board as shown in Figure 13. The feedthrough capacitors C 301, C 307, C 308, C 311 and C 334 are soldered to the required position. Figure 15 shows a photograph of the completed module. The construction should be carried out in the following order:

a) It is advisable for the resonant frequency of the inductances to be checked before installation (dipmeter). Firstly mount the crystal oscillator and doubler. The crystal connections should not be shortened so that room is available for smaller components underneath the crystals.

b) Solder the centre screening panel into place and mount the first lowpass filter chamber.

c) Solder the screening panel to the next lowpass filter chamber and so on until all lowpass filters have been mounted.

d) The chambers for the ring modulator and selective amplifier are now mounted in a similar manner.

e) Align according to the instruction.

f) Place the feedthroughs and feedthrough capacitors into the side panels.

g) Correct the alignment.

Note: If the crystal oscillator oscillates wildly, it will be necessary for the value of C 304 to be increased (up to 150 pF).

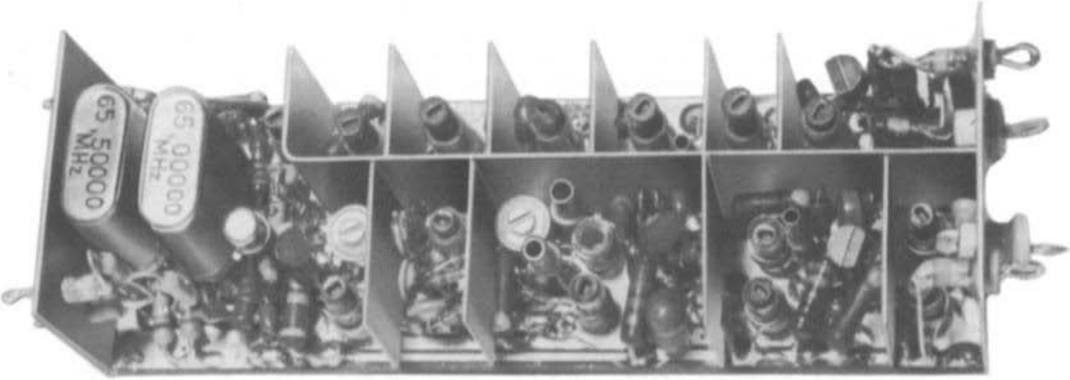


Fig. 15: Photograph of the completed module DC 6 HL 003

3.3.1. SPECIAL COMPONENTS FOR DC 6 HL 003

The smallest components available should be used.

T 301: BC 169, BC 109 (low noise), 2 N 708 or similar
T 302, T 303: BF 224, BF 173, BF 167, BF 115
T 304, T 305: BF 167 (C_{12e} = 0.15 pF !)

D 301, D 302: BAY 83, 1 N 914, 1 N 4148 or similar
D 303 - D 306: 1 N 4148, 1 N 914 (should be selected for the same forward resistance)

Inductances L 310 and L 311 are wound at random, e.g. honeycomb-wound.

L 301, L 306: 13.5 μH, 55 turns of 0.15 mm dia. (35 AWG) enamelled copper wire close wound on a 5 mm coilformer with core (red).
L 302: 20.3 μH, 70 turns, otherwise as L 301
L 303: 23.7 μH, 75 turns, otherwise as L 301
L 304: 1.85 μH, 21 turns, otherwise as L 301

L 305: 18.6 μH, 66 turns, otherwise as L 301
L 307: 6 turns of 0.45 mm diameter (25 AWG) enamelled copper wire wound on a 4 mm coilformer with core (red)
L 308: 3.5 turns, otherwise as L 307
L 309: 2 turns on L 308
L 310: 17 turns of 0.15 mm diameter (35 AWG) enamelled copper wire wound at random on a 4 mm coilformer with core (red). Wound onto a small piece of adhesive tape (sticky side towards the wire) to form a movable roll on the top of the coilformer.
L 311: 33 turns wound on the lower end of the coilformer of L 310
L 312: 4 turns of 0.45 mm dia. (25 AWG) enamelled copper wire wound on a 4 mm coilformer with core (red).
L 313 - L 317: As L 312
L 317: Coil tap at 2 turns from the cold end.

1 crystal of 65.0 MHz and 65.5 MHz directly soldered without holder.

C 317 - C 319: 680 pF miniature styroflex capacitors.
C 309, C 322: 3.5 - 13 pF ceramic disc capacitor of 7 mm dia.

7 ceramic bypass capacitors of approx. 1 nF
1 ceramic bypass capacitor of approx. 10 nF
5 ceramic feedthrough capacitors of approx. 2.2 nF.

3.4. CONVENTIONAL AF AMPLIFIER DC 6 HL 004

This PC-board contains the circuit given in Figure 7 of Part I. The dimensions of printed circuit board DC 6 HL 004, which is shown in Figure 16, are 93 mm x 40 mm. The complimentary transistor pair is provided with an aluminium heat sink as shown in Figure 17. Diode D 401 must be glued with a two-component adhesive (UHU-Plus, Araldite) to the heat sink. Further details can be taken from the photograph given in Figure 18. This module is very uncritical.

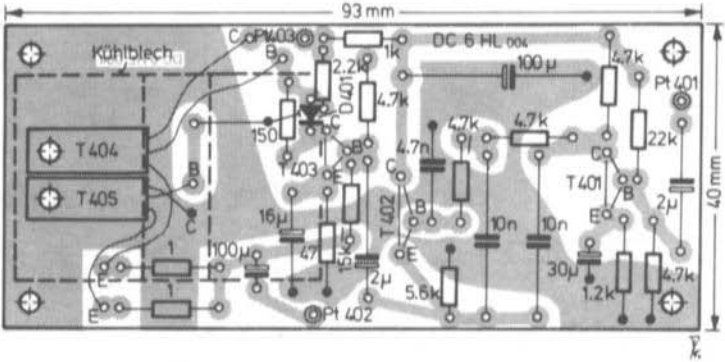


Fig. 16: PC-board DC 6 HL 004

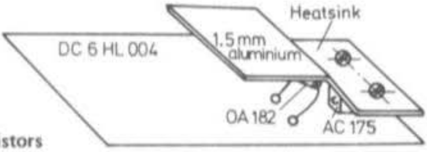


Fig. 17: Drawing of the heatsink for the output transistors

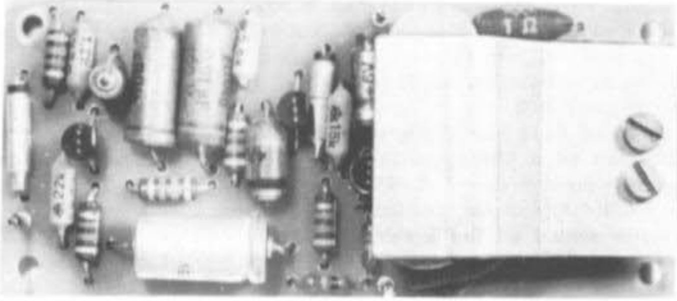


Fig. 18: Photograph of module DC 6 HL 004

3.4.1. SPECIAL COMPONENTS FOR DC 6 HL 004

T 401 - T 403: BC 167 - BC 169, BC 107 - BC 109, 2 N 2926 or similar

T 404 : AC 175, AC 187 K

T 405 : AC 117, AC 188 K

D 401 : OA 182 (AEG-Telefunken), AA 144 (ITT-Intermetall), 1 N 277

R 411 : 150 Ω trimmer potentiometer for PC-board mounting (spacing 12.5/10 mm).

3.5. INTEGRATED AF AMPLIFIER DC 6 HL 005

Due to the non-availability of the original round IC-type, the circuit shown in Figure 8 of Part I was modified for a split-Dip case. Please note the new connections on the circuit diagram shown in Figure 19. The printed circuit board DC 6 HL 005 and component locations are shown in Figure 20, and a photograph of the module is given in Figure 21.

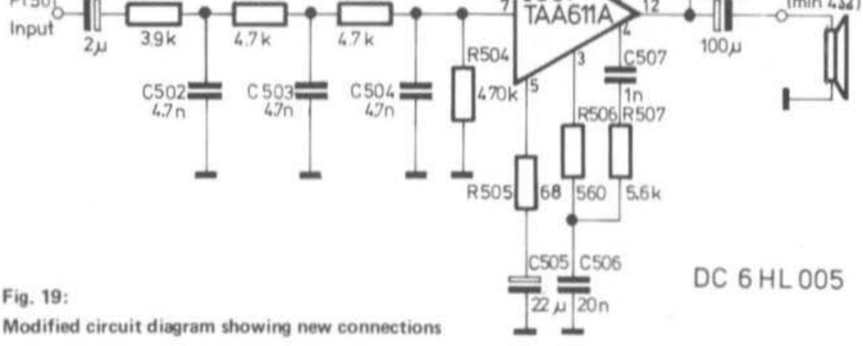
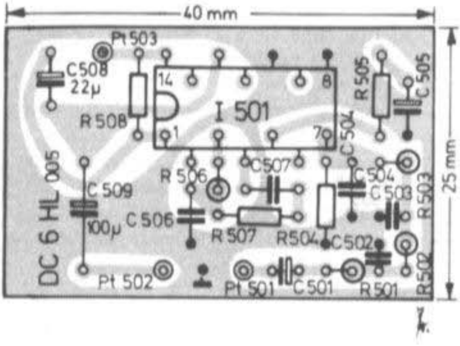


Fig. 19:
Modified circuit diagram showing new connections



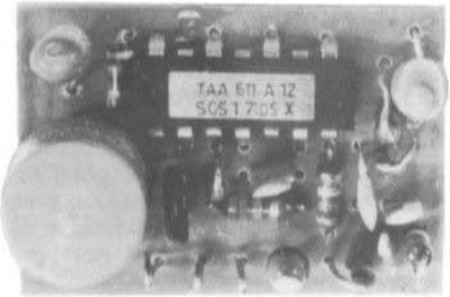


Fig. 20: PC-board DC 6 HL 005

Fig. 21: Photograph of module DC 6 HL 005

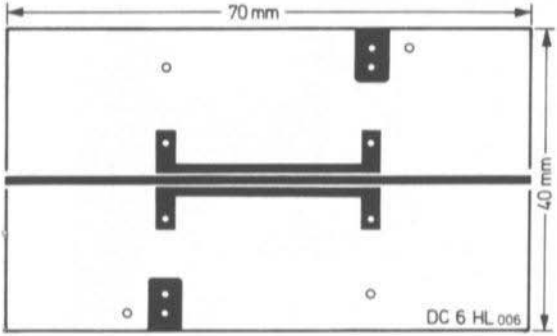
3.5.1. SPECIAL COMPONENTS FOR DC 6 HL 005

I 501: TAA 611 A 12 (12 V type, split-DIP, Fairchild Germany)

C 501, C 505, C 508 and C 509: Tantalium electrolytic capacitors
C 502 - C 504: Ceramic disc capacitors, spacing 5 mm

3.6. REFLECTOMETER DC 6 HL 006

The reflectometer is accommodated on printed circuit board DC 6 HL 006 which possesses the dimensions 70 mm x 40 mm (Figure 22). This PC-board is double-coated and the small number of components are soldered to the printed side of the board as shown in the photograph given in Figure 23. Only 4 holes must be drilled, through which one end of the terminating resistors and the bypass capacitors are connected to the ground surface. The coaxial cables are directly soldered to the centre conductor and the ground surface.



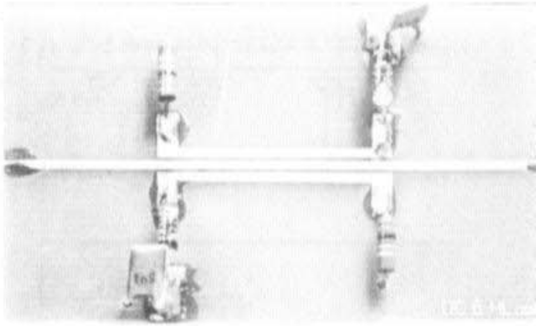


Fig. 22: PC-board DC 6 HL 006

Fig. 23: Photograph of the reflectometer module

3.6.1. COMPONENTS FOR DC 6 HL 006

D 601, D 602: AA 112, 1 N 87 A or similar germanium demodulator diodes

2 resistors 60 Ω (56 Ω or 62 Ω) carbon resistors

2 ceramic bypass capacitors of 1 nF, spacing: 5 mm

3.7. VARIABLE FREQUENCY OSCILLATOR

The variable frequency oscillator is the only module that is not mounted on a PC-board. In order to increase the stability, the VFO is built up on a 5 mm thick aluminium plate between ceramic supports. Figure 24 and Figure 25 show further details of the construction. The given components are non-expensive and provide a good temperature compensation so that an excellent frequency stability is obtained with the aid of the thick casing. Since it is difficult, if not impossible, to give a foolproof constructional description for a highly stable VFO, it is recommended that the general rules should be observed (high quality, new components, stable wiring, well soldered, etc.) and not to rely on any circuit. The VFO should not provide more than 100 mV_{rms} when connected to point Pt 304. Otherwise harmonics will be present at Pt 306.

3.7.1. SPECIAL COMPONENTS FOR THE VFO

T 50: BF 194 (AEG-Telefunken, Philips). Do not use an equivalent type !
T 51: BF 194, BF 224

L 50: 45 turns of 0.45 mm diameter (25 AWG) enamelled copper wire random wound on a 5 mm diameter trolitul coilformer glued with a trolitul adhesive (UHU-PLAST) (Do not use a two-component adhesive) with core (red). Approx. 3.5 μH. Use core padding.

C 50: 100 pF variable capacitor
C 52: 27 pF

C 51, C 55: 100 pF ceramic tubular cap.
positive TC: P 100

C 53: 800 pF, C 54: 1000 pF styroflex capacitors

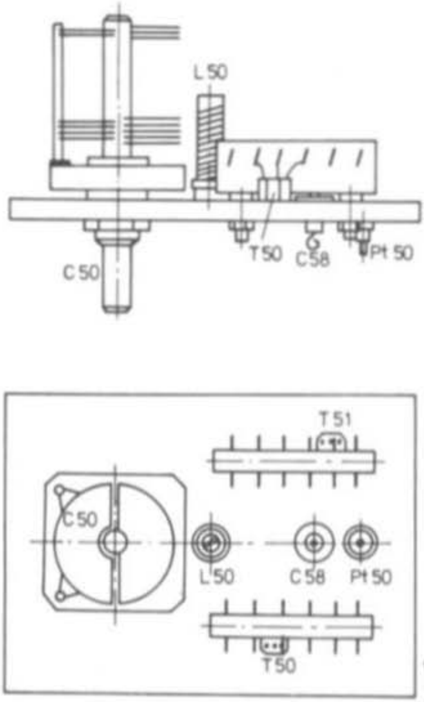


Fig. 24: Construction of the VFO

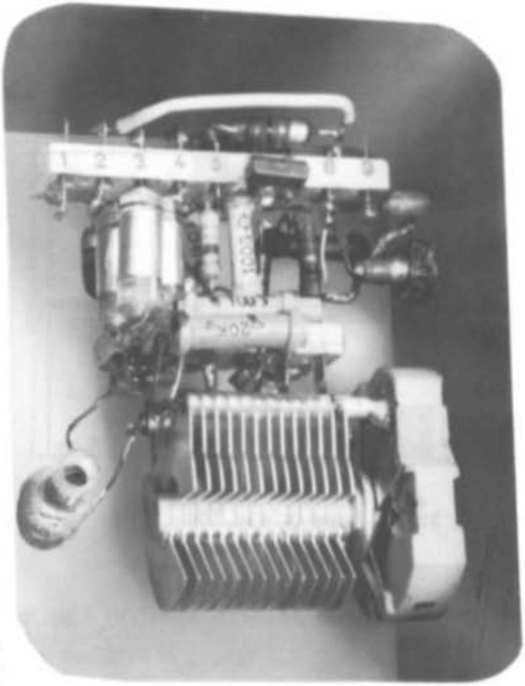


Fig. 25: Photograph of the VFO

3.8. OTHER CIRCUITS

The three transistor stages shown in the block diagram Figure 2 of Part I for the voltage stabilisation, S-meter and for driving the impulse relay are not critical and can be constructed in any suitable position.

3.8.1. SPECIAL COMPONENTS

T 1: BC 167 - BC 169, BC 107 - BC 109, 2 N 2926
T 2: BF 245 C (TI, Siliconix) or similar FETs
T 3: BC 178, BC 213, 2 N 3905 (PNP)
T 4: BC 167 - BC 169, BC 107 - BC 109, 2 N 2926

D 1: ZF 9,1 (ITT-Intermetall), BZY 85/C9V1 (AEG-Telefunken), 1 N 4103
D 2, D 3: BAY 83, 1 N 914, 1 N 4148 or similar.

4. ALIGNMENT

Before commencing alignment, the individual modules should be checked to see whether any components have been incorrectly mounted. This should be followed by interconnecting the modules. Decoupling resistors of approximately 100 Ω should be placed between the individual modules, e. g. between the AF amplifier and the rest of the receiver, as well as eventually further bypass capacitors.

4.1. AF-AMPLIFIERS DC 6 HL 004 and 005

Disconnect the collector lead of transistor T 404 (AC 175) and connect a mA-meter in series.

Adjust the quiescent current to approximately 5 mA with the aid of trimmer resistor R 411. The amplifier must be capable of providing 2.23 V (6.3 V peak-to-peak) into a 5 Ω load at low distortion when driven with the appropriate input voltage.

With the integrated AF amplifier, no adjustment of the quiescent current is required and it is only necessary to check operation of the amplifier.

4.2. CARRIER OSCILLATOR DC 6 HL 002

The trimmers C 205 and C 208 are firstly placed in their centre position. After connecting the stabilized operating voltage of 8.5 V to connection Pt 201 and Pt 202 (or Pt 203), a voltage of at least 1.5 V should be measured at the output Pt 204. If the oscillator does not commence oscillation, transistor T 201 should be exchanged for one with a higher transit frequency. If a frequency counter is available, trimmer C 205 can be aligned to 8.9985 MHz and C 208 to 9.0015 MHz.

Since the carrier oscillator frequency is introduced into the IF amplifier in spite of the screening, the carrier oscillator is switched off during AM reception. This injection is of no importance during SSB reception since the AGC voltage is taken from the audio frequency signal.

4.3. VFO

It is only necessary to align the frequency range of this module. The core of inductance L 50 is aligned until the VFO is able to tune over the frequency range of 5.0 to 6.0 MHz. Attention must be paid to ensure that the VFO does not provide more than 100 $\mathrm{mV_{rms}}$ output at Pt 304 so that no harmonics are present at Pt 306. It should be mentioned that it is possible to deviate slightly from the frequency concept used in this transceiver without having to alter the circuits. For instance, the author uses crystals of 64.9 and 65.4 MHz in module DC 6 HL 003 and a VFO frequency range of 5.2 to 6.2 MHz.

4.4. LOCAL OSCILLATOR MODULE DC 6 HL 003

The alignment of this extensive module is carried out in steps:

4.4.1. LOW-PASS FILTER

a) Insert the cores of L 301 to L 306 by approximately half.

b) Solder a 1 nF coupling capacitor to the input connection Pt 304 and inject a signal (e.g. from a dipmeter).

c) Disconnect the output of the filter (L 306) from the bandpass filter (C 317) and terminate with a 560 Ω resistor.

d) Connect the RF-probe of a VTVM or the arrangement shown in Figure 26 and feed the operating voltage to Pt 305.

e) Align L 302, L 303 and L 305 so that the output signal falls as much as possible above 6.5 MHz.

f) Align the signal generator (dipmeter) to 11 MHz (harmonic of the VFO centre frequency) and increase the drive to the filter until the meter indicates a clear reading.
Align L 301 and L 306 for minimum reading.

g) Tune the signal generator from 7 MHz towards higher frequencies. The output signal should firstly fall and then increase at frequencies over about 12 MHz. Inductance L 304 should be aligned for minimum reading at the frequency of maximum signal.

h) A low, constant insertion loss between 5 MHz and 6 MHz is obtained by carefully correcting L 302, L 303 and L 305. If the attenuation commences lower than 6 MHz, it will be necessary to correct the alignment of L 301 and L 306.

i) Remove the terminating resistor and reconnect L 306 to C 317/L 310.

4.4.2. CRYSTAL OSCILLATOR AND DOUBLER

a) Provide connections Pt 301 and Pt 302 (or Pt 303) with a stabilized operating voltage.

b) Align the core of inductance L 307 until the coupled dipmeter indicates RF-energy.

c) Correct alignment of L 307 so that the oscillator commences oscillation and offers the same output with each of the crystals.

d) Align L 308 and C 309 for maximum output voltage at twice the oscillator frequency. Check for equal output at both frequencies, using the circuit shown in Figure 26.

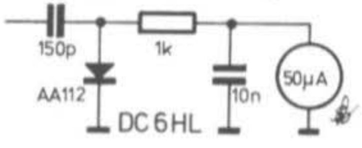


Fig. 26: Alignment aid if no VTVM available

4.4.3. RING MODULATOR

a) Adjust the connected VFO to 5.5 MHz and place trimmer capacitor C 322 to its centre position.

b) Align L 310 and L 311 for max. reading at the output of the ring modulator.

c) Disconnect L 306 from C 317/L 310 and measure the RF-voltage on L 306.

d) Reconnect the filter and remeasure the output voltage. It should have been reduced by half (power matching). If this is not the case, the spacing between L 310 and L 311 must be changed, after which the measurement should be repeated.

e) Switch the crystal oscillator to the lower frequency and adjust the VFO to 6 MHz. A frequency of 136 MHz should be present at L 312 (measure with dipmeter). Align L 312 and L 313 for maximum reading.

4.4.4. SELECTIVE AMPLIFIER

a) Align L 314, L 315, L 316 and L 317 for maximum output at 136 MHz. If the output voltages differ greatly from that at 135 MHz or 137 MHz, L 314 and L 315 should be corrected.

b) If a measuring receiver is available, it is possible for trimmer capacitor C 322 of the ring modulator to be aligned for minimum output of the fixed frequency (connect the receiver input to the base of T 304).

4.5. MAIN BOARD DC 6 HL 001

4.5.1. TRANSMITTER

Connect the carrier oscillator and VFO/mixer module to PC-board DC 6 Hl 001 and terminate the output of the transmitter.

a) The quiescent current of output transistor T 118 is adjusted to 20 - 30 mA with the aid of the base trimmer potentiometer R 191. After this, R 190 and R 191 can be replaced by a fixed resistor of the same value.

b) Feed an AF signal to input Pt 110 from a microphone or audio generator. A signal of 4 V_{pp} should be present at the emitter of the second AF transistor T 111 (at low distortion).

c) Place trimmer potentiometer R 159 of the ring modulator to one of the stops. The carrier signal should be present at L 111/L 112. The core of these inductances should be aligned for maximum carrier output.

d) The core of L 113/L 114 should also be aligned for maximum.

e) The signal should now be received on a two metre receiver. Adjust the carrier to 145 MHz by tuning the VFO.

f) Align L 115, L 116 and trimmer capacitor C 182 for maximum carrier output.

g) Screw out the core of L 121 until the signal at L 119 begins to fall.

h) It is now possible for the driver and PA to be aligned for maximum power gain. L 115, L 116 and C 182 should be aligned to the correct frequency from a higher frequency so that they are not incorrectly aligned to 137 MHz.

i) It is necessary to spread the resonance points of the circuits in the linear amplifier slightly so that there is no fall of at the band limits.

j) Align trimmer potentiometer of the ring modulator for minimum carrier output (max. carrier suppression).

k) The transmitter should now be driven at low level by the AF generator. Now align trimmers C 120 and C 121 of the crystal filter for minimum ripple in the pass band.

l) Align the frequency of the XF 901 crystal so that the output power of the transmitter is 3 dB lower at 3 kHz than in the voice frequency range.

4.5.2. RECEIVER

a) Connect earphones or an AF-amplifier to the SSB-output Pt 107.

b) Connect the S-meter circuit to the control voltage line. The meter should be adjusted with R 2 so that a very slight reading results.

c) Connect a signal generator via a 10 pF capacitor to the "hot end" of L 110. A heterodyne should be heard when the signal generator is tuned to 9 MHz. L 110 should now be aligned for maximum.

d) This procedure should be repeated for L 109, L 108 and L 107.

e) Inject a 145 MHz signal to L 105 and align C 111 for maximum.

f) Inject the signal to L 103 and align trimmers C 106, C 109 and C 111 for maximum reading.

g) Inject the signal to the antenna input and align L 101, L 102 and L 103 for maximum.

h) Align L 106 for the highest conversion gain.

i) Align L 122 for best image rejection.

5. NOTES AND IMPROVEMENTS

5.1. INTERMODULATION MEASUREMENT ON THE TRANSMITTER

The author carried out intermodulation measurements on the transmitter of his SSB-transceiver.

Measuring equipment:	two AF-generators SBF (Rohde u. Schwarz) and attenuator, Spectrum Analyzer 8554 L with probe (HP), terminating resistor 50 Ω.
Measurement:	Two-tone modulation 350 Hz and 2350 Hz. RF-output (two-tone): 0.5 W
Result:	Greatest intermodulation product: -33 dB.

5.2. AUTOMATIC GAIN CONTROL FOR SSB RECEPTION

The SSB control voltage output Pt 108 remains connected to the control line during the alignment of the receiver. If the gain of the IF-amplifier is too great (tendency to self-oscillation), the value of the 4.7 nF source capacitors should be reduced or one of them or more should be removed. During normal operation, the control voltage will vary between approximately + 2.2 V without input signal and approximately + 0.5 V with local signals. If strong signals appear distorted, this will indicate that too little control voltage is being generated. This can be avoided by increasing the values of R 138 and R 139 from 680 Ω to 1.2 kΩ.

Since the AGC is distributed over two RF and three IF stages, it is possible for the first mixer to be overdriven in spite of full gain control. This will be observed as intermodulation and cross modulation in conjunction with strong signals. In order to avoid this, the last IF stage should remain uncontrolled and operate at full gain. The total control range is practically not affected by this. However, since only four stages are influenced by the AGC, the control range of each of these stages is higher, and thus the attenuation of the signal before the first mixer will be better. The modification is carried out easily by disconnecting resistor R 125 (390 Ω) from the control line and connect it to the basic bias voltage of 2.2 V (cathode of diodes D 101, D 103, D 108).

If the SSB-demodulator is not able to handle higher IF-voltages (distortion), it will be necessary to increase the oscillator voltage at connection Pt 106. This can be made using the intermediate transformation circuit shown in Figure 27.

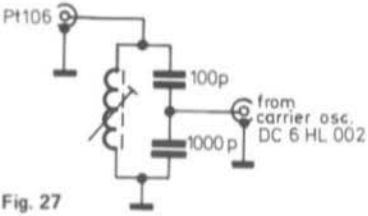


Fig. 27

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PHASE-LOCKED CIRCUITS

by T. Schad, DJ 8 ES

In phase-locked loops (abbreviated PLL), the phase angles of two signals are synchronized in a closed loop. Phase-locked loops allow construction of selective circuits without use of conventional components, e. g. inductances. The following applications are of special interest to radio amateurs: FM-demodulator, FSK-demodulator, AM-demodulator, SSB-demodulator, analog or digital variable phase-locked oscillators, frequency multiplier, frequency divider, tracking narrow-band filters.

The fundamentals and advantages of phase-locked circuits have been known since approximately 1930. However, the large number of components required limited their use to applications where cost was not of primary concern. Even the construction of phase-locked circuits using individual integrated circuits would be rather extensive. Now that whole phase-locked systems are available as monolithic integrated circuits, this technology is rapidly gaining ground, and the prices of such integrated phase-locked loops will most certainly fall.

The principle of phase-locked loops as well as several interesting applications using partly integrated construction are to be discussed in this article.

1. FUNDAMENTALS

In PLL-circuits, the phase angle of two signals is compared in a phase discriminator (see Fig. 1). Since the output signal of the phase comparator usually possesses an AC component, a low-pass filter is provided to form the DC component "U" ($\Delta\varphi$). This error voltage U ($\Delta\varphi$) is dependent on the phase difference $\Delta\varphi$ between the two input signals and is used to control the frequency of a voltage-controlled oscillator (VCO) via a DC voltage amplifier. The output frequency f_o of the VCO, or another frequency controlled by it, represents one of the two signals fed to the phase discriminator. The other signal is the reference signal u_r with the frequency f_r. The control circuit synchronizes the frequency of the VCO to the reference frequency. With the exception of a small, constant phase difference, both frequencies will remain synchronized. Fundamentals of the control circuit were discussed in detail in (1), (2) and (3) and need not be repeated here.

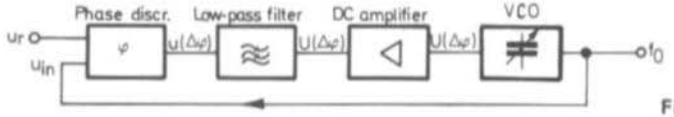

Fig. 1: Block diagram of a PLL-circuit

2. PHASE DISCRIMINATOR

The operation of several circuits for a phase discriminator that often appear in the integrated circuit technology, are now to be explained. For simplification, it is assumed that the frequency of both signals are equal so that only the phase angle, and not the variation of the phase angle as a function of time must be considered.

2.1. AND-GATE

The operation of such gates has been described in (4). If one assumes that the reference and input signal at the phase discriminator are in the form of squarewave pulses with a duty cycle of 1 : 1 (Fig. 2), a squarewave signal will also be present at the output of the AND-gate whose duty cycle is determined by the phase difference between the two signals. The mean value as a function of time is thus dependent on the phase difference as shown in Figure 2c. If the reference and input signal have the same phase angle, a logic 1-signal will be present at the output of the AND-gate for one half cycle. If a phase difference is present, the duty cycle at the output will decrease until no pulses occur at $\Delta\varphi = 180^0$. Such circuits can be constructed using all types of gates.

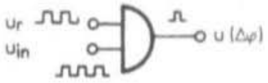

Fig. 2 a: An AND-gate as phase-comparitor

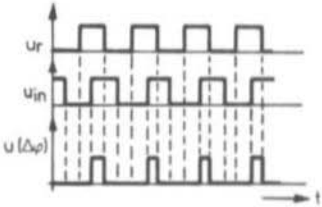

Fig. 2 b: Input and output signals

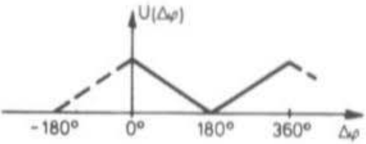

Fig. 2 c: Mean value of the output voltage as a function of the phase difference

2.2. R-S FLIP-FLOP

R-S flip-flops were also described in (4) and (6). They allow the following phase discriminator circuits to be constructed. Short pulses are obtained at a certain phase angle from the input and reference signals (e.g. on leaving the zero pass in a positive direction). These pulses are fed to the R or S inputs of the flip-flop (Fig. 3).

Figure 3b shows the position of the pulses to another, and Figure 3c the mean value of the output voltage as a function of the phase difference. A DC voltage is again obtained that is dependent on the phase difference. It should be noted that the duration of the input pulses must be as small as possible in comparison to the period of the signal since they would coincide at $\Delta\varphi = 0^0$ and $\Delta\varphi = 360^0$. By shifting the pulses towards each other, e.g. by 180^0, the zero pass of curve $U(\Delta\varphi)$ can be shifted to another phase angle. The exact working of an R-S flip-flop in a PLL- circuit is explained in great detail in (2) and (3).

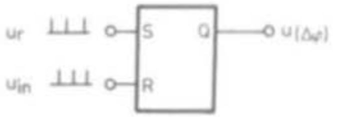

Fig. 3 a: R-S flip-flop as phase discriminator

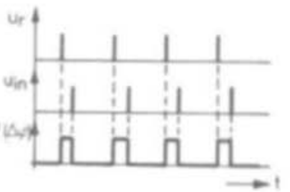

Fig. 3 b: Input and output signals

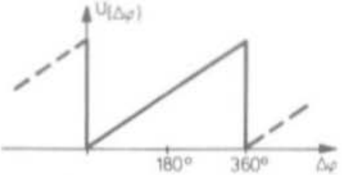

Fig. 3 c: Mean value of the output voltage as a function of the phase difference

2.3. RING MODULATOR

The most important considerations that must be taken into consideration during construction of a ring modulator were described in (5). The application of such a circuit as a phase discriminator is to be discussed here (see Fig. 4). If u_r is positive, diodes D 2 and D 3 will conduct whereas D 1 and D 4 will be blocked. This means that point 3 is connected to point 2 and $u(\Delta\varphi) = -u_{in}/2$ (if the transformation ratio is 1 : 1). During the other half cycle of u_r, diodes D 1 and D 4 will conduct and D 2 and D 3 will be blocked. In this case, $u(\Delta\varphi) = +u_{in}/2$. The principle of this electronic switching is shown in Figure 4b. The switch is actuated in the zero pass of u_r. Figure 4c shows the output voltage as a function of the phase difference. It is assumed here (as in (5)) that the diodes switch exactly at the zero pass of u_r, e.g. that the forward voltage does not amount to 0.6 V but to 0 V. The distortions that are caused by the finite forward voltage can be avoided by using fast switching diodes such as Schottky (hot carrier) diodes or planar diodes that are driven with squarewave pulses.

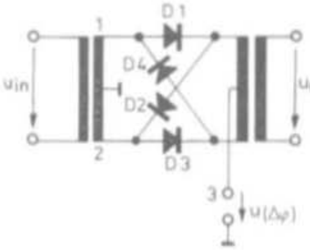

Fig. 4 a: Ring modulator as phase discriminator

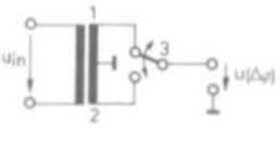

Fig. 4 b: Switching principle of the ring modulator

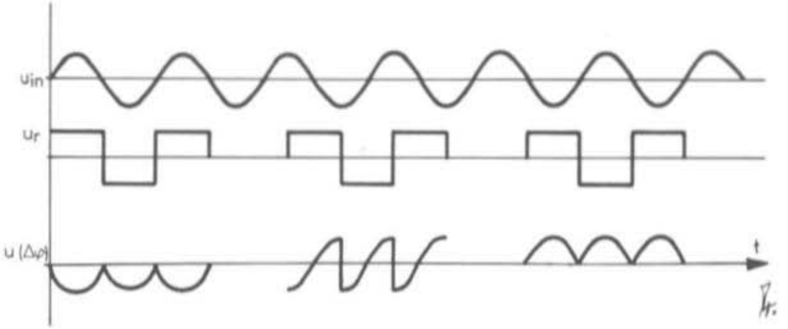

Fig. 4 c: Output signal as a function of the phase difference between the two input signals

2.4. SAMPLING DISCRIMINATOR

A sampling discriminator mainly comprises an electronic switch and a storage capacitor (see Fig. 5). The electronic switch is controlled by the reference signal so that the switch is closed for a short period after the zero pass of the reference signal. If the sampling period (switch closed) is short with respect to the period duration of the reference signal, the storage capacitor will be charged with the momentary voltage of the input signal during the short sampling period. As can be seen in Figure 6, the voltage across the storage capacitor will have the same waveform as the time-dependent input signal as a function of the phase difference. With this phase discriminator circuit, the input and reference signal are better suppressed than with the previously mentioned circuits. This can be very important for many applications.

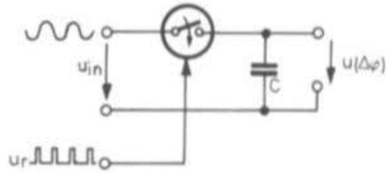

Fig. 5: Principle of a sampling discriminator

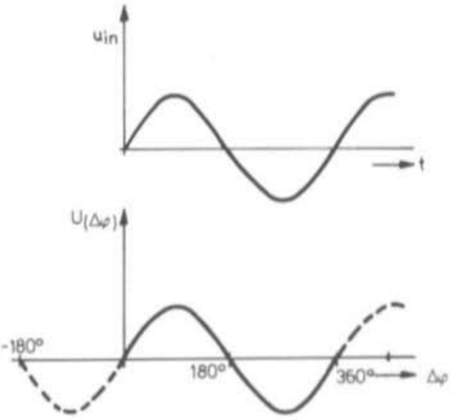

Fig. 6:
Voltage across the capacitor as a function of the phase difference

3. VOLTAGE CONTROLLED OSCILLATOR

Either an oscillator equipped with a varactor diode or a multivibrator is used as a voltage controlled oscillator (VCO). In the case of a multivibrator, a capacitor is charged at a constant current until a certain potential is reached. The circuit then switches to its other state where the capacitor is discharged. The charge and discharge times and thus the frequency are dependent on the value of the constant current. This current is controlled by the DC-voltage coming from the phase comparator via a DC amplifier.

4. FURTHER CIRCUIT ELEMENTS

4.1. LOW-PASS FILTER

The phase discriminators that have been mentioned previously do not generate a DC voltage at the output. This means that a low-pass filter must be placed between the discriminator and the VCO since the VCO would otherwise be modulated by the reference signal. Simple RC low-pass filters (Fig. 7a) or a so-called lag filter (Fig. 7b) are usually sufficient. Both filters (especially the lag filter) only produce a very small phase shift which can be very important with respect to the stability of the control circuit. The cut-off frequency of the low-pass filter must be far lower than the reference frequency in order to obtain a sufficient suppression of the signal. However, this has an adverse effect on the transient behaviour of the PLL-circuit.

Both single or multiple section configurations of both filters can be used. The disadvantages can be reduced when an active low-pass filter is used. The bandwidth of this low-pass filter influences the hold range of the PLL circuit and thus the selectivity.

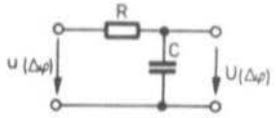

Fig. 7 a: Simple RC low-pass filter

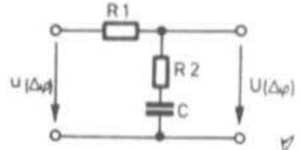

Fig. 7 b: Lag filter

4.2. DC VOLTAGE AMPLIFIER

The higher the amplification of the error voltage, the smaller will be the remaining phase difference between the reference and VCO frequency. The DC amplification in the control loop determines the range of synchronization and the input sensitivity. However, limits are placed on the degree of amplification by the frequency behaviour of the control circuit which is mainly dependent on the characteristics of the low-pass filter. Operational amplifiers are very suitable for use in PLL-circuits.

5. APPLICATIONS

5.1. FM DEMODULATOR

Figure 8 shows the block diagram of an FM demodulator constructed according to the PLL-technology. The input signal u_{in} is frequency modulated, which means that the input frequency deviates from the centre frequency in the rhythm of the modulation (7). The VCO is now controlled via the control circuit in rhythm with the input frequency. It is necessary for the cut-off frequency of the low-pass filter to be higher than the highest modulating frequency since, otherwise, it would not be possible to control the VCO above a certain frequency. The control voltage $U\ (\Delta\varphi)$ is then the required audio voltage. The PLL-circuit can be used in a similar manner to convert an FSK signal (e.g. RTTY) into a digital DC-signal and thus save the extensive AF-filters normally required.

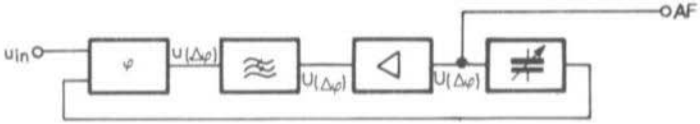

Fig. 8: Block diagram of a PLL FM demodulator

5.2. AM DEMODULATOR

The operation of the AM demodulator is somewhat more complicated, as can be seen in Figure 9. The input voltage U_{in} is an amplitude modulated signal which is fed to one input of a product detector (x) and as reference signal to a PLL-circuit. A phase shifter $(\Delta\phi)$ is provided in front of the phase-locked loop, and the reason for this will be given later. The output signal of the VCO is fed to the other input of the product detector (x). As is known from the operation of the PLL-circuit, the VCO frequency is locked onto the frequency of the AM carrier. Since the VCO signal is of constant amplitude (not amplitude modulated) the output signal of the product detector includes the AF signal which is filtered out in a low-pass filter in the normal manner. Since the amplitude of this AF signal is greatest when the two signals at the product detector have a phase angle of 0° or 180°, this value should be selected with the aid of the phase-shifter network.

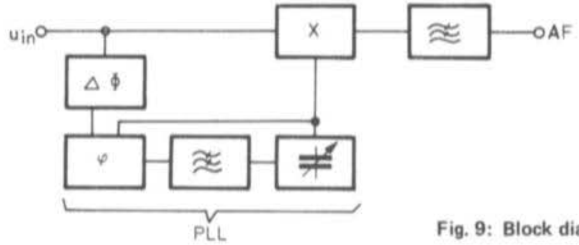

Fig. 9: Block diagram of a PLL AM demodulator

5.3. SSB DEMODULATOR

The PLL-circuit for AM demodulation can easily be extended for demodulating SSB transmissions. Normally, a signal originating from the carrier oscillator is fed to the product detector. However, in our case it is possible to lock the PLL-circuit onto the frequency of the carrier signal by feeding it to the input of the phase shifter instead of the AM signal. In this manner, the same relationships exist as with AM demodulation and it is not necessary to compensate the different gain of the two demodulators.

5.4. PHASE-LOCKED OSCILLATOR

It is not clear in engineering circles whether the described method should be called a phase-locked or frequency synthesis oscillator. The theory is that a free-running oscillator is locked to a multiple of a reference frequency with the aid of a phase-locked loop (Fig. 10). In this application, the output frequency of the VCO is digitally divided and the phase of the divided signal compared with the reference signal (crystal oscillator). Often, the output frequency is mixed with another fixed frequency and is divided. This means that the stability of the VCO is equal to the long term stability of the crystal-controlled signal. Since various frequency division ratios can be selected, the VCO can be adjusted to any frequency n x f_r. The reference frequency f_r therefore determines the spacing between the channels that can be selected. The transient behaviour on switching from one channel to another is described in (1). It should be mentioned that the hold range, e.g. the difference between n x f_r and the frequency of the VCO amounts to 2% - 10% of f_r x n. The short term stability of the output frequency is dependent on the characteristics of the control circuit.

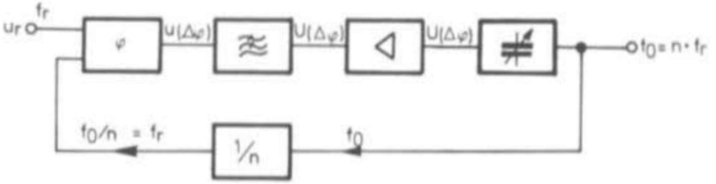

Fig. 10: Block diagram of a phase-locked oscillator using PLL-circuits

5.5. FREQUENCY DIVISION AND MULTIPLICATION

If the reference signal contains harmonics, the VCO can be locked to any required harmonic of the reference signal (e.g. frequency multiplication). Frequency dividers can be constructed in a similar manner.

The most important, but not all, of the applications for (integrated) PLL circuits have been described. The author hopes that this will lead to new developments based on these interesting components.

6. AVAILABILITY

Motorola offer economical TTL integrated circuits for phase-locked oscillators in the shortwave range: Phase-frequency detector MC 4044 P, Dual voltage controlled multivibrator MC 4024 P and frequency dividers up to 60 MHz (SN 74196). However, it is still necessary for VHF oscillators to be built up as frequency synthesis oscillators for price reasons. Both frequency and phase are compared in the MC 4044 P which means that a tuning range of several octaves is possible.

The Signetics Corporation offers PLL-circuits for FM and AM demodulation (NE 560 B, NE 561 B, NE 562 B). The prices are around DM 40,-- at the moment. Since the prices of semiconductors fall considerably when the market requirements for such components increase, reasonable prices can be expected in the future. When considering the price, however, one should not forget that three modulation modes can be detected and that these circuits provide additional selectivity.

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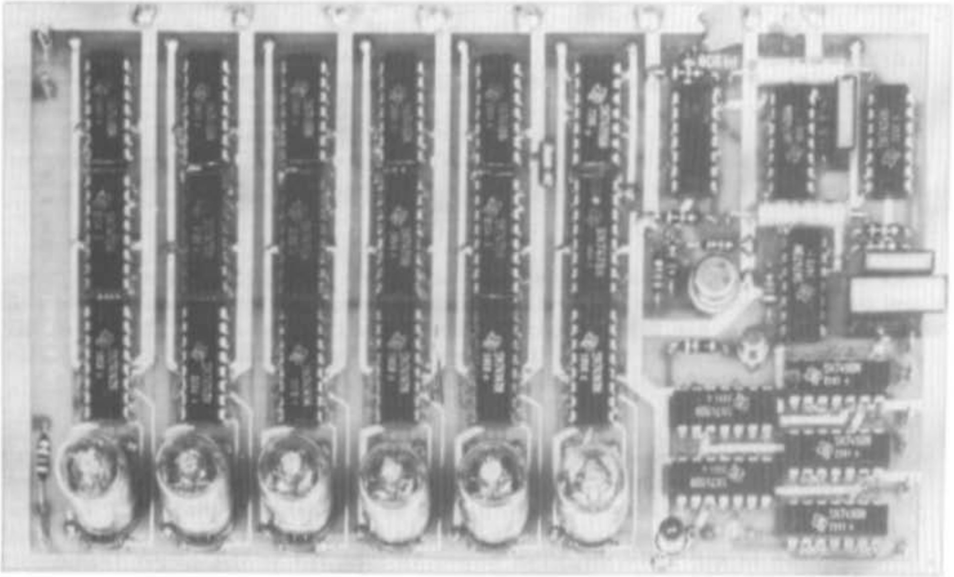
Introduction and Application Notes on PLL-circuits.



AN 18 W POWER AMPLIFIER FOR 432 MHz WITH PRINTED STRIPLINES

by K. Hupfer, DJ 1 EE

A new generation of equipment has been developed for mobile services of public utilities and business communications at UHF. The semiconductor manufacturers have developed new power transistors for this equipment that offer output power levels of up to 20 W at 12 V to 14 V. This makes such transistors very suitable for mobile operation on the 70 cm amateur band where they can be used to build up an efficient power amplifier. A three-stage power amplifier with an output power of 18 W is to be described. All inductances are in the form of printed striplines etched on a double-coated epoxy board. The reproducibility is thus extremely good and the dimensions are very compact.

1. CHARACTERISTICS

The following values were measured when equipped with the given Philips transistors:

Output power:	16 W min., 20 W max. (50 Ω)
Drive power:	300 mW min., 500 mW max.
Current drain at 13.5 V:	2.9 A to 3.0 A
Suppression of first harmonic (864 MHz):	approx. 30 dB

This amplifier has also been constructed using RCA transistors: for T 1 and T 2: 2 N 5914 and for T 3: 2 N 5915. With this version, an output power of approximately 8 W was obtained.

2. CIRCUIT DETAILS

The circuit diagram of the three-stage amplifier is given in Figure 1. Input and output impedance are 50 Ω. All inductances required for transformation and selectivity are in the form of printed striplines (L 1 to L 9). They were dimensioned as described in (1). It is planned to explain how the matching elements can be calculated from the transistor data in a later article.

The drive power of approximately 350 mW is matched to the low and complex base-emitter impedance of transistor T 1 with the aid of matching link C 1, C 2, L 1. A power of approximately 2 W is available at the collector, which is fed via the matching network of C 4, C 5, L 3 to the base of transistor T 2. This stage also amplifies by approximately 4 times which means that about 8 W can be measured at the collector. The power is then fed to the output transistor T 3 via a similar matching network. A somewhat more extensive network transforms the load impedance of 50 Ω to the required working impedance for transistor T 3.

Chokes Ch 1 to Ch 3 for DC grounding of the base connections are not critical. Miniature ferrite chokes with an inductivity of approximately 100 nH are used. The collector DC-voltages are fed to the transistors via the low-inductive inductances L 2, L 4, L 6, which are also in the form of striplines.

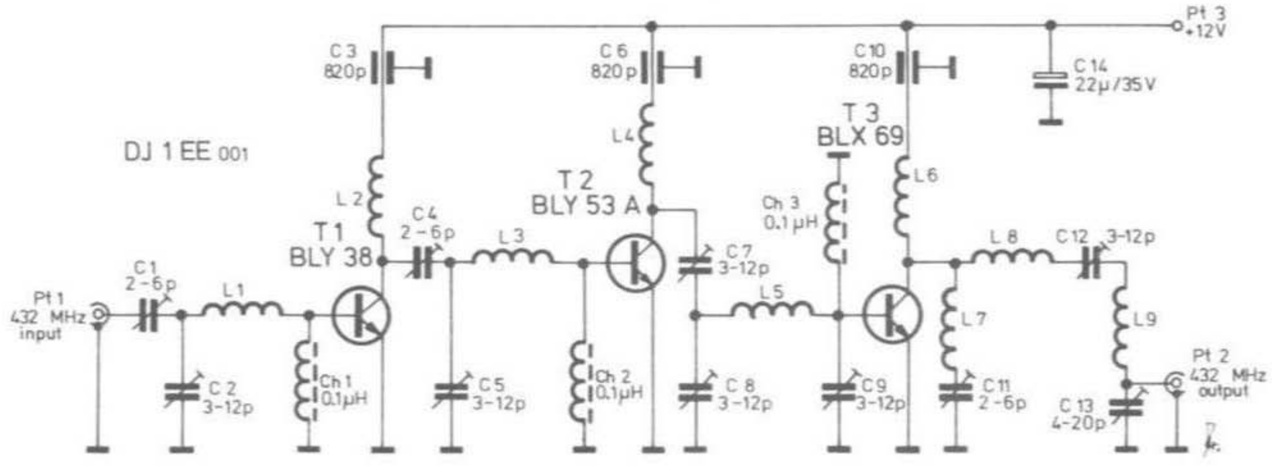


Fig. 1: Circuit diagram of the three-stage 432 MHz 18 W amplifier

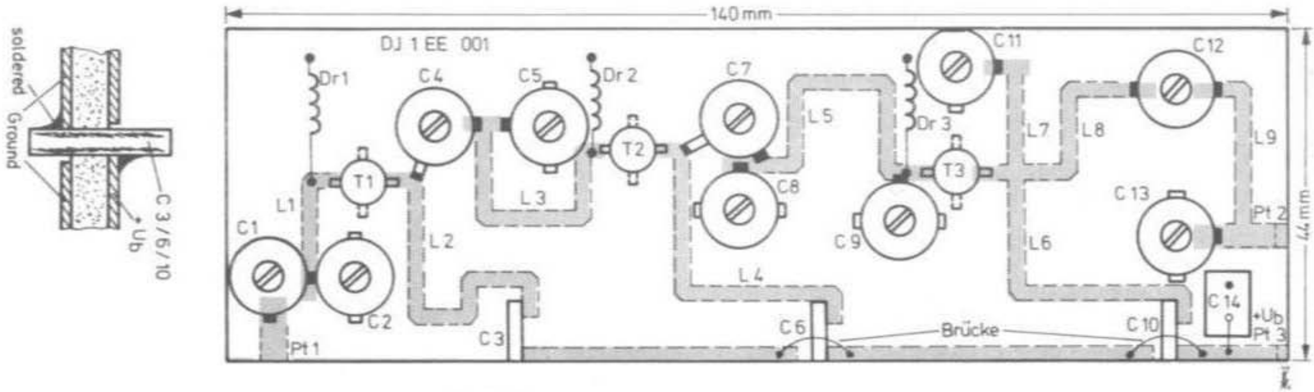


Fig. 2: Components and printed striplines of PC-board DJ 1 EE 001

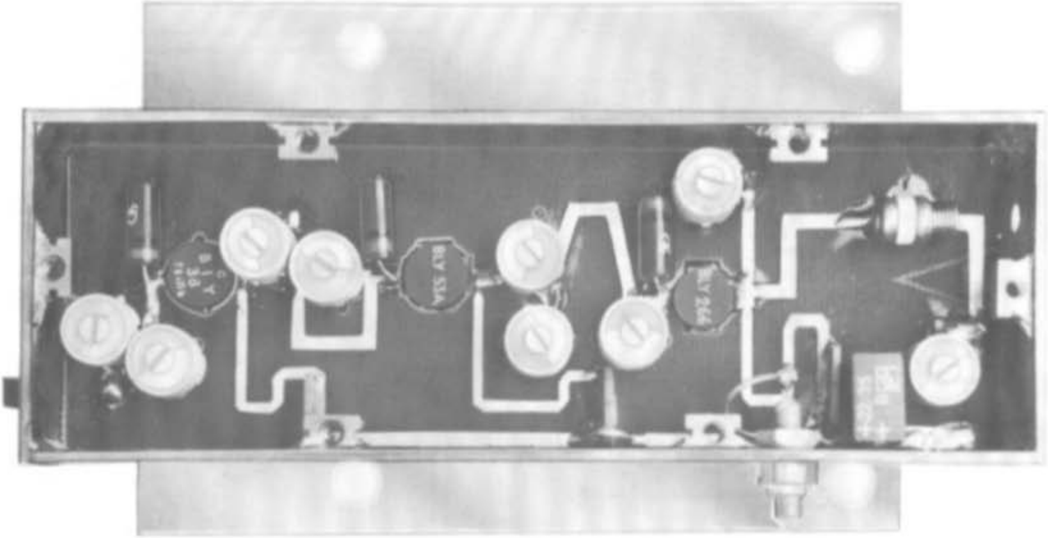


Fig. 3

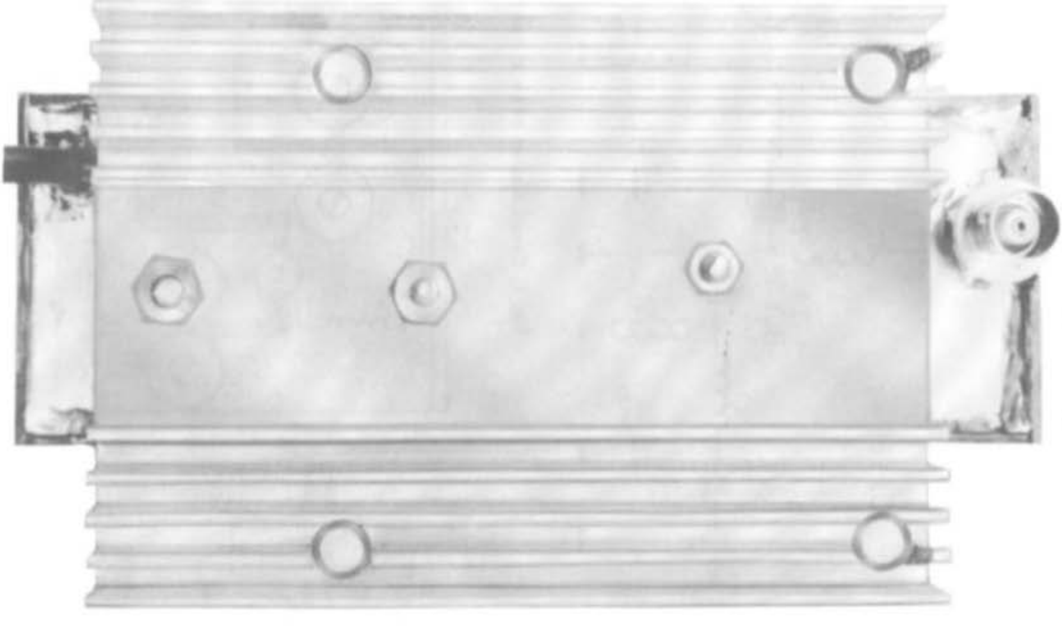


Fig. 4

The power transistor BLX 69 (formerly 266 BLY) has only been developed recently by Philips. The price of approximately DM 150,-- is of course very expensive for amateur equipment but will not doubt drop after production of professional communications equipment gets under way. However, the modification has shown that less expensive, lower power transistors such as the mentioned RCA types can be used. In the case in question, it was not necessary to carry out any modifications of the circuit or printed circuit board.

3. COMPONENTS

T 1: BLY 38 (Philips)
T 2: BLY 53 A (Philips)
T 3: BLX 69 (Philips)

C 1, C 4, C 11: 2-6 pF ceramic disc trimmer capacitors (10 mm diameter)
C 2, C 5, C 7, C 8, C 9: 3-12 ceramic disc trimmer capacitors (10 mm diameter)
C 3, C 6, C 10: 820 pF ceramic disc capacitors without wires
C 12: 3-12 pF ceramic tubular trimmer
C 13: 4-20 pF ceramic disc trimmer (10 mm diameter)
C 14: 22 μF tantalium electrolytic capacitor

Ch 1 to Ch 3: 100 nH ferrite choke (Delevan 1025-94)

4. CONSTRUCTION

All components are accommodated on the double-coated printed circuit board DJ 1 EE 001 whose dimensions are 140 mm x 44 mm. This PC-board is shown in Figure 2 together with the components and striplines. Holes must be drilled in the board for all ground connections and transistor supports. All "hot" connections are directly soldered to the conductor lanes. Slots should be sawn into the PC-board for the bypass capacitors which are mounted as shown in the drawing. The positive bar of the operating voltage is fed via bridges to the bypass capacitors so that a mA-meter can be connected instead of the bridge during the alignment procedure.

Further details regarding the construction can be seen in the photographs Figures 3 and 4. The printed circuit board is mounted on a somewhat shorter heatsink which cools all three transistors. Walls made from 15 mm high PC-board material are soldered to the edge of the PC-board, and a cover should also be provided.

Due to the non-linear relationship between input and output RF-voltage, the amplifier is best suited for FM-transmissions. At lower drive levels with correct carrier adjustment, it can also be used for amplification of AM signals. In this case, however, an AF distortion of approximately 10% will result.

5. AVAILABLE PARTS

See material price list.

6. REFERENCES

(1) K. Hupfer: Striplines for VHF and UHF
VHF COMMUNICATIONS 3 (1971), Edition 4, Pages 207-216

23 cm PREAMPLIFIER WITH PRINTED MICRO-STRIPLINES

by K. Hupfer, DJ 1 EE

Both Siemens and Philips brought out low-signal transistors in the second part of 1971 that exhibit a considerable gain in the lower GHz range with low noise figures. The most important specifications of these non-expensive semiconductors are given in the following table:

Type	Manufacturer	f_T (GHz)	G_p (dB)	and F (dB)	at f (GHz)
BFR 14	Siemens	3.4	9	5	2
BFR 34/35	Siemens	3.0	9	5.5	2
BFR 90/91	Valvo/Philips	5	13	3	0.8

The specifications indicate that these transistors are suitable for use as preamplifiers, mixers or frequency multipliers in 23 cm receivers. Using the above transistor types, the author firstly constructed a preamplifier for 1296 MHz.

1. PROTOTYPE

The circuit shown in Fig. 1 was used for the first experiments. The circuit was built up on single copper-coated PC-board material using air-spaced striplines. This construction is shown in Fig. 2 and 3. The inductances L 1 and L 2 are silver-plated metal strips of approximately 27 mm in length and 7 mm wide. The tuning capacitors should have a variable capacitance range of 0.8 pF to 8 pF. Very important for the operation is the shortest possible ground connection for the emitter. An example of a suitable construction is given in Fig. 5. The transistor type BFR 35 was tried in the described circuit. The result was a power gain of $G_p \approx 10$ dB and an additional noise figure $F_a \approx 5$ dB.

If this preamplifier is connected to a receiver having a noise figure of $F_r \approx$ 9 dB, the total noise figure (insert the values as a power ratio) will amount to:

$$F_{tot} = F_a + \frac{F_r - 1}{G_p} = 3.15 + \frac{7.9 - 1}{10} \approx 4 = 6 \text{ dB}$$

Since these transistors exhibit a very high gain at lower frequencies, it is necessary for them to be neutralized at low frequencies by suitable networks in the interconnections. The decoupling at the operating frequency is made with the aid of a $\lambda/4$ choke at the base (L 3) and collector (L 4) with bypass capacitors (C 4 and C 9) of approximately 15 pF. The damping at lower frequencies is made with the aid of the connected RC-low-pass links.

2. MICRO-STRIP CONSTRUCTION

Fig. 4 shows the circuit diagram and approximate stripline shapes for a 23 cm preamplifier built up on a double-coated glass fibre reinforced teflon (PTFE) board. No alignment elements are required. The striplines were dimensioned according to (1). However, different stripline configurations will be necessary to suit the various transistor types. These are indicated in Fig. 4 by the dashed lines. The basic design is valid for transistor type BFR 90. If the BFR 14 is to be used, it will be necessary for the angled part of L 2a to be 15 mm long instead of 6 mm and the output must be provided with a stub line of 20 mm by 3 mm. For transistor type BFR 35, an additional stub-line of 10 mm by 7 mm will be necessary at the collector.

DJ 1 EE

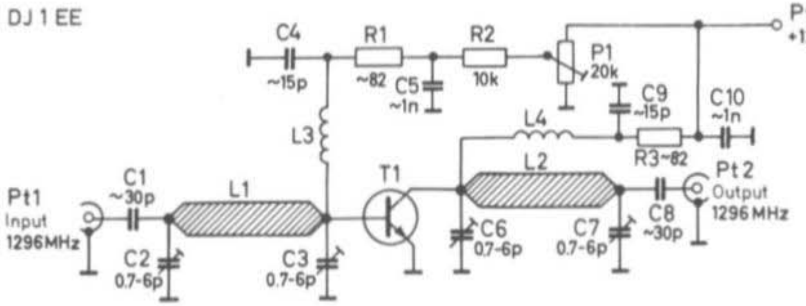


Fig. 1: Basic circuit diagram of a 1300 MHz preamplifier

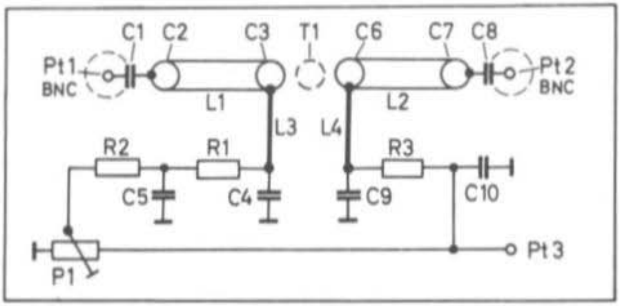


Fig. 2: Construction principle using air-spaced striplines

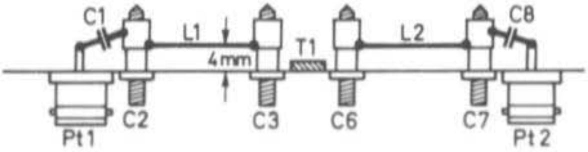


Fig. 3: Side view

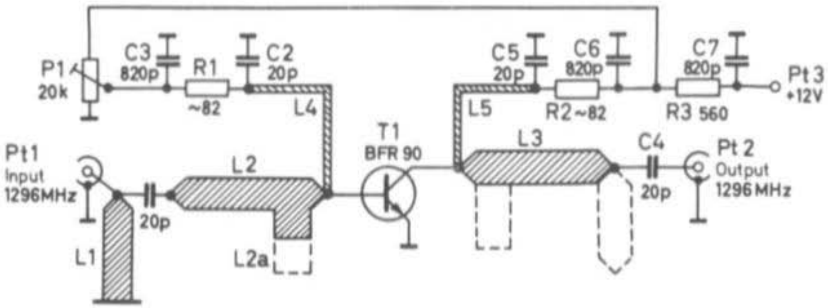


Fig. 4: Circuit diagram of a 23 cm preamplifier using transistor type BFR 90
(dashed lines: modifications required BFR 14 or BFR 35)

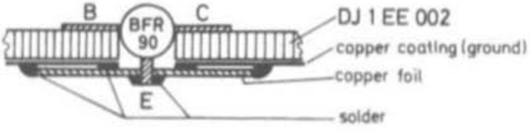


Fig. 5: Installation details for transistor type BFR 90

This circuit also obtains the previous values of $G_p \approx 10$ dB and $F_a \approx 5$ dB. Stripline L 1 shortcircuits all frequencies that are lower than the working frequency. This means that interference due to conversion effects will be virtually avoided. The bypassing of the operating voltage connections is made as described in Section 1. However, ceramic bypass capacitors without leads are placed into slots on the PC-board. Very decisive for the operation of the amplifier is that inputs and outputs are terminated with a resistive load of 50 Ω. The following converter should be aligned for power matching and not for minimum noise. Various lengths of cable can be placed between the antenna and preamplifier for checking the matching. If the field strength indication on the receiver does not vary greatly, it can be assumed that a good matching has been obtained.

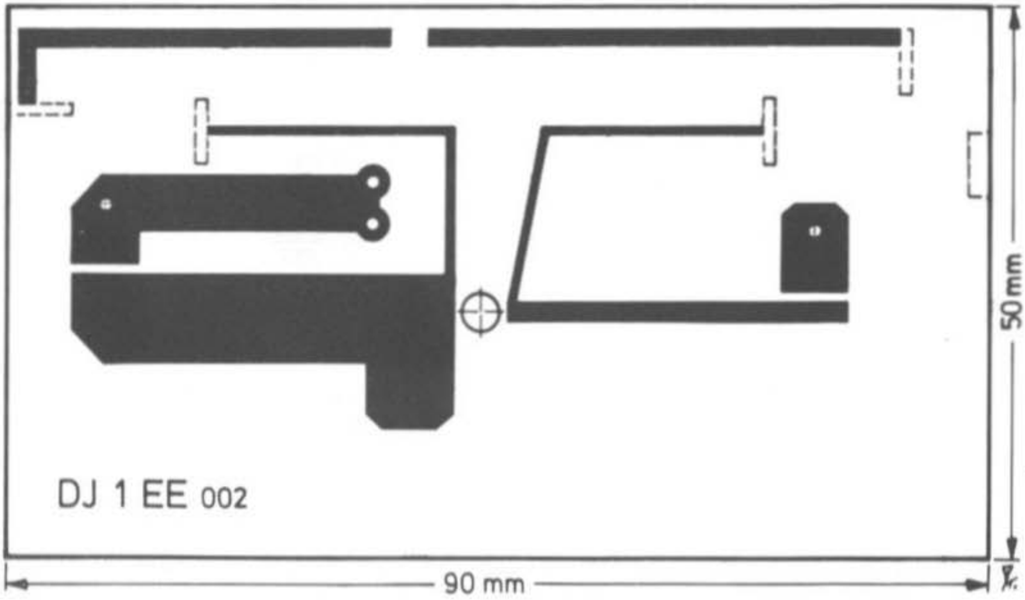


Fig. 6: Printed circuit board DJ 1 EE 002 for the 23 cm preamplifier
(Material: glass-fibre reinforced teflon (PTFE), doublecoated)

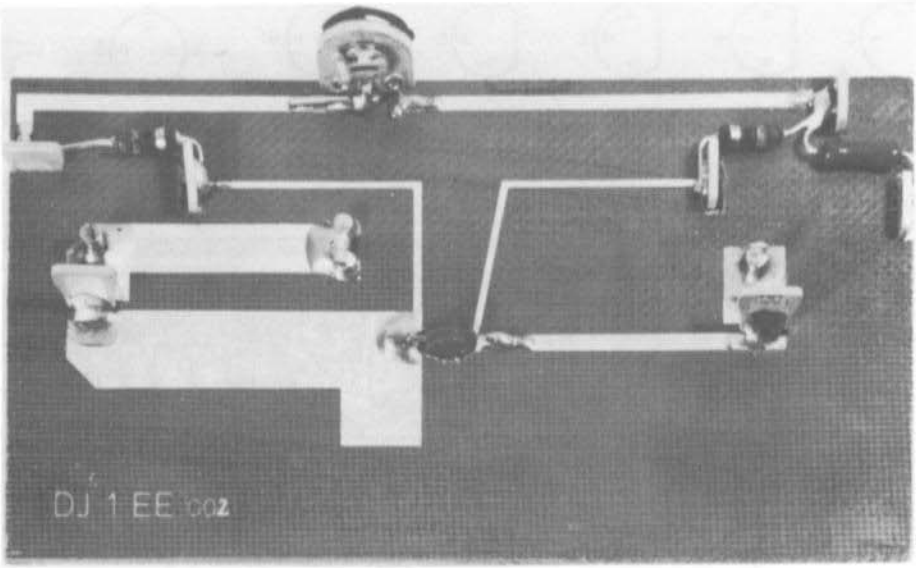


Fig. 7: Photograph of the author's prototype

3. CONSTRUCTION

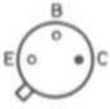
Fig. 6 shows the printed circuit board for the described preamplifier. This board is designated DJ 1 EE 002. The dimensions are 90 mm x 50 mm and it is made from a double-coated glass-fibre reinforced teflon (PTFE) board. The striplines are dimensioned for the transistor type BFR 90. It is necessary for holes to be drilled for the transistor, the two BNC sockets as well as for the short-circuit on L 1; slots must be sawn for the bypass capacitors. The soldering of these chip capacitors must be carried out very quickly since the very thin ceramic plates are damaged very easily when heated on one side (in contrast to a solder bath). Fig. 5 shows how the transistor is installed. A photograph of the prototype is given in Fig. 7 and shows further details of the construction. The potentiometer P 1 is aligned so that a current of approximately 2 mA flows through T 1. After the alignment P 1 may be replaced by two fixed resistors (approx. 12 kΩ / 1.2 kΩ).

3.1. COMPONENTS

T 1:	BFR 90 (Philips)
C 1, C 2, C 4, C 5:	approx. 20 pF ceramic bypass capacitor chips
C 3, C 6, C 7:	approx. 820 pF ceramic bypass capacitor chips
P 1:	20 kΩ miniature trimmer potentiometer
R 1, R 2, R 3:	82 Ω carbon resistors
2 BNC connectors	

4. REFERENCES

(1) K. Hupfer: Striplines for VHF and UHF
VHF COMMUNICATIONS 3 (1971), Edition 4, Pages 207-216



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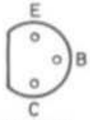

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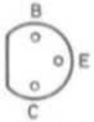

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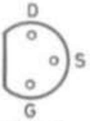

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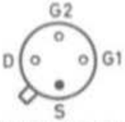

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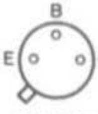

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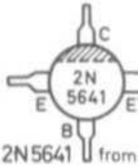

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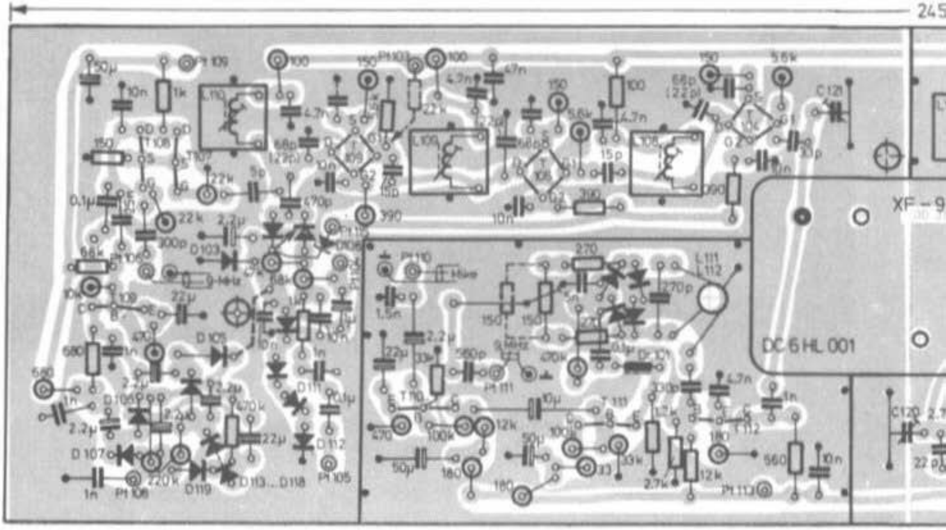

Fig. 9: Component locati

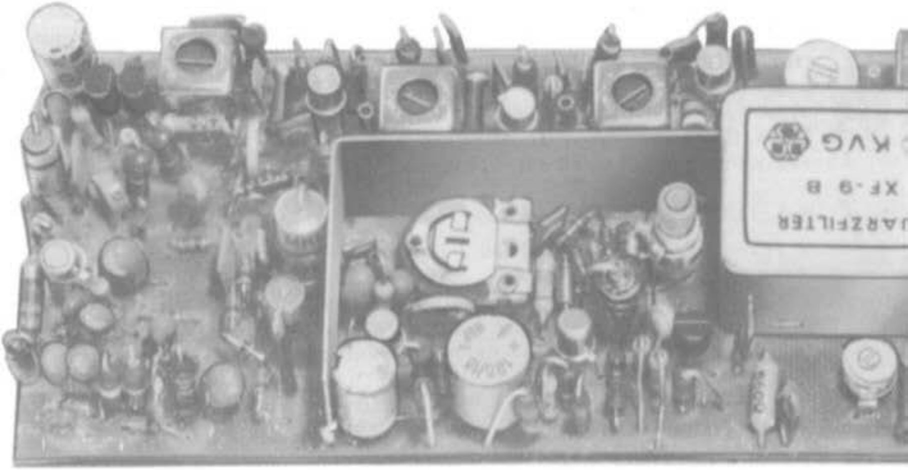

Fig. 10: Photograph

above

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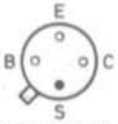

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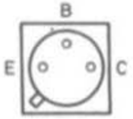

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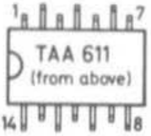

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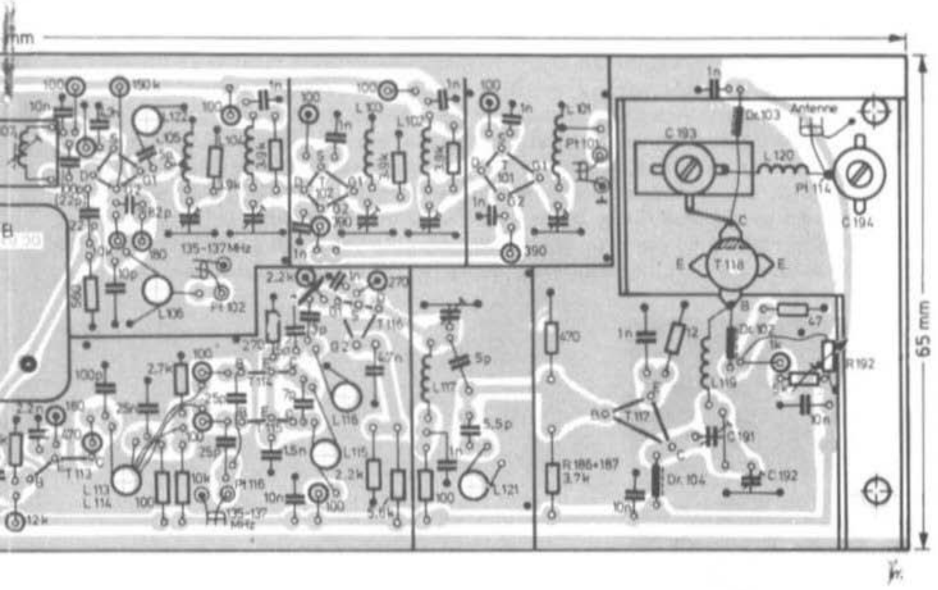

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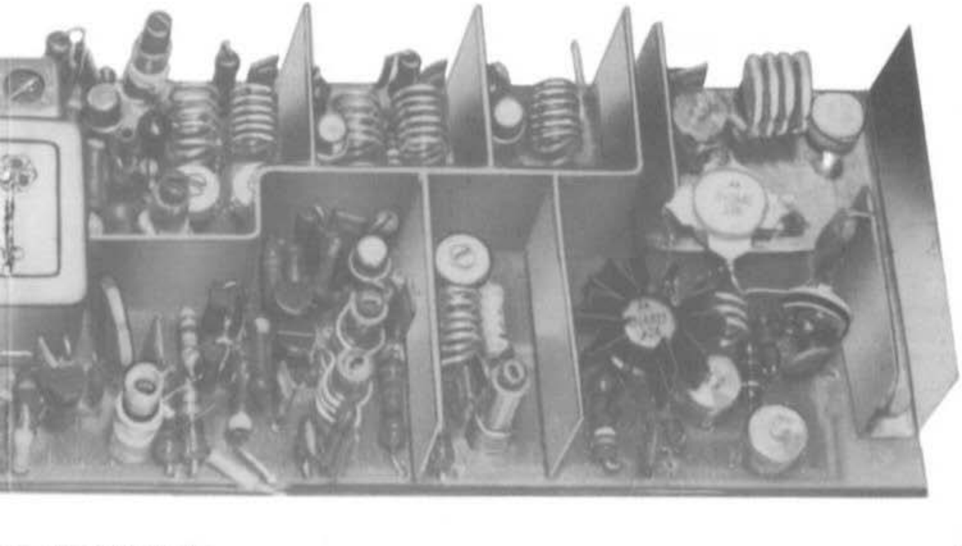

of module DC 6 HL 001

CIRCULATORS AND ISOLATORS
Part II

by R. Lentz, DL 3 WR

3.1.1. ISODUCTOR OR GYRATOR

One special type of circulator with lumped elements is a component which has been designated ISODUCTOR (Melabs) or GYRATOR (Microwave Ass.). This component does not have one important characteristic of a circulator: the matching at the connections. One can assume it to be the core of a circulator with lumped elements but without transformation links, connections and matching elements (Fig. 10). The isoductor or gyrator is therefore smaller and less expensive than a circulator. However, it will be able to be used for the same applications as a circulator if external matching links are provided. For this reason, this component is extremely suitable for use in stripline circuits. Another advantage is that the resonant frequency of a circulator built up in this manner is determined by the matching links and can thus be shifted in a relatively wide frequency range. Of course, it is also possible for such a circulator to be matched with the aid of the matching links to other impedance values than 50 Ω.

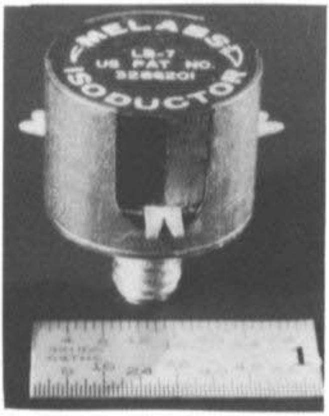

Fig. 10: An isoductor

The impedance at the three connections of a gyrator or isoductor consists of the ohmic resistance r_o and a parallel inductivity l_o. For this reason, it is necessary for matching links to be provided at connection 1 and 2 whereas connection 3 is terminated with a resistor (R 3) and a parallel capacitance (C 3) as shown in Fig. 11. The values of R 3 and C 3 can be determined with the following formulas:

$$R\,3 = r_o\,; \quad C\,3 = \frac{1}{\omega^2 \times l_o}$$

Whereby $\omega = 2\pi f$; with f = required operating frequency or centre frequency of the required frequency range. The isoductor or gyrator will operate as a circulator within the frequency range in which the inductivity l_o is compensated for by C 3. At resonance, the stopband attenuation is theoretically infinity; at other frequencies, the stopband attenuation will fall. The bandwidth is determined by the loaded Q_L :

$$Q_L = \frac{r_o}{\omega \times l_o}$$

The lower the value of Q_L the wider will be the bandwidth.

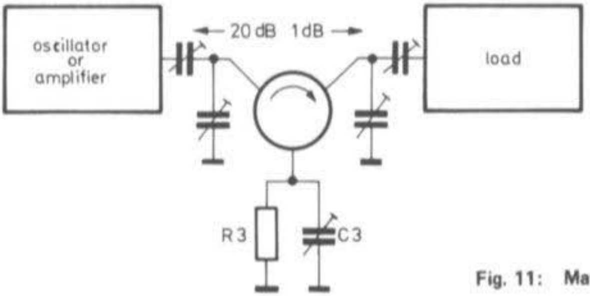

Fig. 11: Matching of an isoductor or gyrator

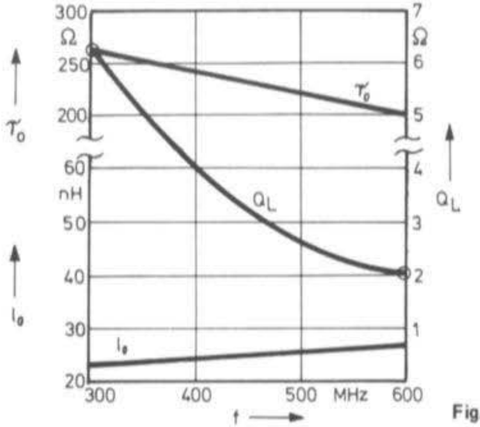

Fig. 12: Typical characteristics of an isoductor LB-3

Typical calculated values for an isoductor covering the 70 cm band are given in Fig. 12. This allows the values for R 3 and C 3 and for the transformation networks to be calculated. Fig. 13 gives the stopband and passband attenuation as a function of frequency. As an example, the frequency characteristics of an isoductor type LB-3 when tuned to 435 MHz have been drawn below this. It can be seen from this that the bandwidth of this component is more than sufficient for radio amateurs.

Table 1 gives the nominal specifications of the available isoductor types.

Frequency Range	100-200 MHz	200-400 MHz	300-600 MHz
Insertion loss	1 dB	< 1 dB	< 1 dB
Stopband attenuation	> 20 dB	> 20 dB	> 20 dB
r_o	225 Ω	250 Ω	230 Ω
l_o	32 nH	26.5 nH	24 nH
Q_L	7.5	5	2.7
Designation	LB-1	LB-1	LB-3
Price (Dec. 1969)	300. - DM	300. - DM	300. - DM

Further specifications and applications can be taken from (6), (7) and (8).

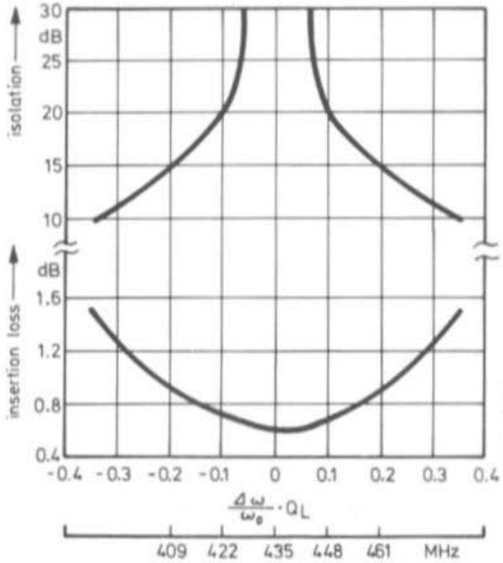

Fig. 13: Standardized attenuation curves of the isoductor series LB-1, LB-3, LB-7. Below: Frequency characteristics calculated for a type LB-3 tuned to 435 MHz

3.2. JUNCTION CIRCULATORS

Junction circulators are closely related to circulators with lumped elements. This relationship can be compared to that between cavity resonators and resonant circuits constructed from inductances and capacitors. Fig. 14 shows a junction circulator manufactured in stripline technology. The designation of this type of circulator originates from the symmetrical Y-shaped line junction.

The ferrite disc in the centre of the junction represents a dielectric resonator. The smaller the diameter, the higher the frequency. The relatively high dielectric constants of ferrite materials, which are 12 to 15 times as great as that of vacuum, allow the construction of circulators for microwaves that are very compact. In the VHF and UHF range, on the other hand, it is possible to obtain more compact dimensions when using concentrated elements since concentrated elements are very much smaller at lower frequencies than waveguide resonators or $\lambda/4$ transformation links.

Junction circulators (1), (5), (9) are constructed in stripline, waveguide and micro-stripline configurations. At microwave frequencies, bandwidths of up to an octave are covered; at VHF and UHF, the bandwidth is limited to approximately 40% of the operating frequency. However, if only a relatively low bandwidth is required, more favourable specifications, lower cost and more compact dimensions are possible. Of special interest in this context is a special circulator manufactured by Melabs for parametric amplifiers on the 70 cm, 24 cm and 12 cm bands. Table 2 gives the specifications and price (1965) of these three circulators.

Frequency	432 MHz	1296 MHz	2300 MHz
Bandwidth (min)	2 MHz	4 MHz	4 MHz
Insertion loss (max)	0.5 dB	0.3 dB	0.3 dB
Stopband attenuation (min)	20 dB	20 dB	20 dB
VSWR (max)	1.5	1.2	1.2
Impedance	50 Ω	50 Ω	50 Ω
Connectors	TNC	N	N
Designation	J-6969	J-6970	J-8017
Price in US $	97.50	95.50	88.00

Waveguide junction-circulators exhibit lower losses and a better standing wave ratio for the same stopband attenuation and bandwidth than coaxial types (e.g. an SWR of 1.05 and an insertion loss of 0.1 dB over a bandwidth of 15%). However, waveguide circulators are larger than the appropriate stripline types at VHF and UHF.

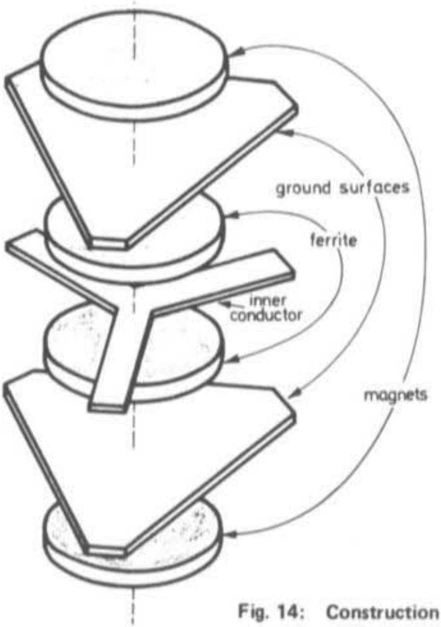

Fig. 14: Construction of a stripline junction circulator

3.3. SWITCHING CIRCULATORS

Since the direction of circulation is dependent on the polarity of the magnetic bias, it is possible for the circulator to be used as a simple switch (without moving parts) by controlling this magnetic field. The field is then generated with the aid of an electromagnet or by a permanent magnet whose polarity is reversed by a short, high-current surge through a magnetizing coil. The symbol of a switching circulator is given in Fig. 15.

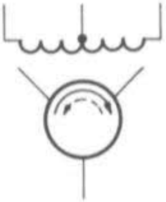

Fig. 15: Symbol of a switching circulator

The radio frequency characteristics of switching circulators correspond to those of equivalent conventional circulators. Further details are given, for instance, in (1), (5) and (9).

4. ISOLATORS

Isolators consist of a RF line (stripline, coaxial line or waveguide) into which a magnetically biased ferrite body is placed. The RF-energy is absorbed in one direction by the ferrite material and is not affected in the other direction. For optimum attenuation values the operating frequency must coincide with the gyromagnetic resonant frequency of the ferrite material. In this context, the operation of an isolator coincides to that of a circulator. In principle, a three-port circulator will work as an isolator as long as one port is terminated with the correct impedance.

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A 50 MHz TRANSVERTER BY MODIFICATION OF RECEIVE CONVERTER DL 6 HA 001 AND TRANSMIT CONVERTER DL 6 HA 005

by Robert R. Eide, W 0 ENC

1. RECEIVE CONVERTER DL 6 HA 001

The following modification allows the reception of the 50 MHz to 54 MHz (6 meter) band in conjunction with an intermediate frequency range of 14 MHz to 18 MHz. The original circuit diagram (1) is given in Figure 1.

With the exception of the resonant circuits and the crystal, it is not necessary for any component values to be changed. The modifications are as follows:

Crystal: 36.000 MHz HC-6/U (for 14 MHz - 18 MHz IF)
Capacitors C 4 and C 7: 5 pF

All inductances except L 1 are close-wound using enamelled copper wire. L 1 is spaced one wire diameter between turns.

L 1: 4 turns of AWG 24 (0.50 mm) on L 2 near coil tap.
L 2: 15 turns of AWG 24 (0.50 mm) on a 3/16" (4.3 mm) coilformer with VHF-core. Coil tap 3.5 turns from gate connection.
L 3: 13 turns of AWG 24 (0.50 mm). Coilformer 3/16" (4.3 mm) diameter, with VHF-core.
L 4: 11.5 turns wire and coilformer as for L 2. Coil tap 3.25 turns from ground end.
L 5: As L 3.
L 6: 11 turns wire and coilformer as for L 2.
L 7: 36 turns of 28 AWG (0.30 mm) on 1/4" (6 mm) coilformer with SW core.
L 8: 14.5 turns of 24 AWG (0.50 mm). Coilformer as for L 7, with VHF core.
L 9: 20.5 turns, wire as for L 8. Coilformer 3/16" (4.3 mm) diam., VHF core.
L 10: 22 turns, wire as for L 8. Coilformer and core as for L 9.

1.1. ALIGNMENT

The alignment should be made as described in (1). It should be remembered that resonant circuits of FET stages can only be checked for resonance with a dipmeter when the operating voltage is connected.

2. TRANSMIT CONVERTER DL 6 HA 005

This modification allows the use of the transmit converter in the six meter band of 50 MHz to 54 MHz. The original circuit diagram is given in Figure 2 (2). The required local oscillator frequency of 36 MHz is taken from the receive converter and fed to connection Pt 302 on the transmit converter board.

The following components must be changed:

Capacitors:
Increase value of the 5 pF capacitor between L 304 and base of T 303 to 8.2 pF.
Increase value of the 8 pF capacitor between L 305 and base of T 304 to 10 pF.
Increase value of 470 pF capacitor between L 306 and output Pt 303 to 560 pF.

All inductances are close wound on 9/32" (7 mm) diameter coilformers.
Wire: 20 AWG (0.80 mm) enamelled copper.

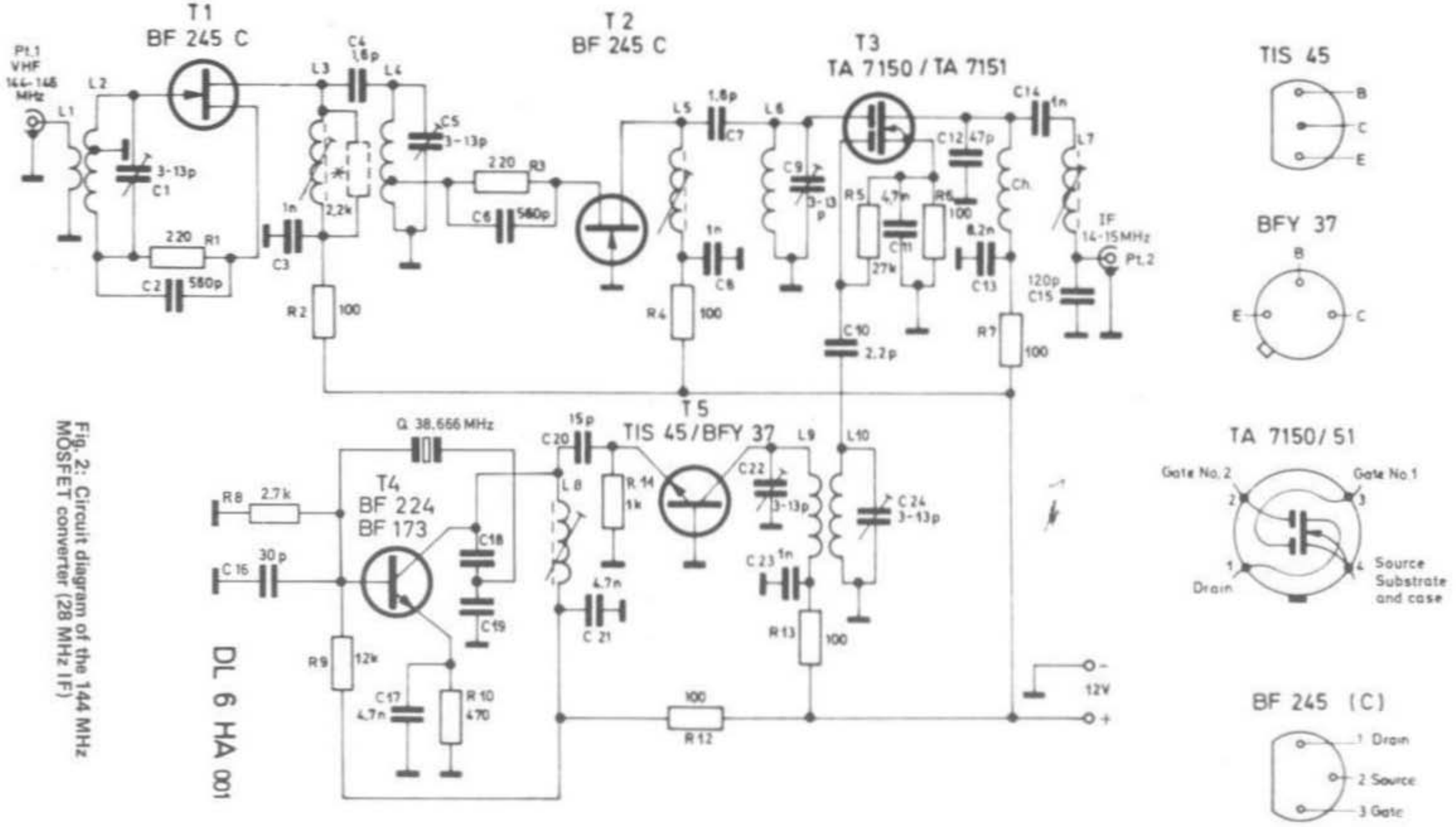
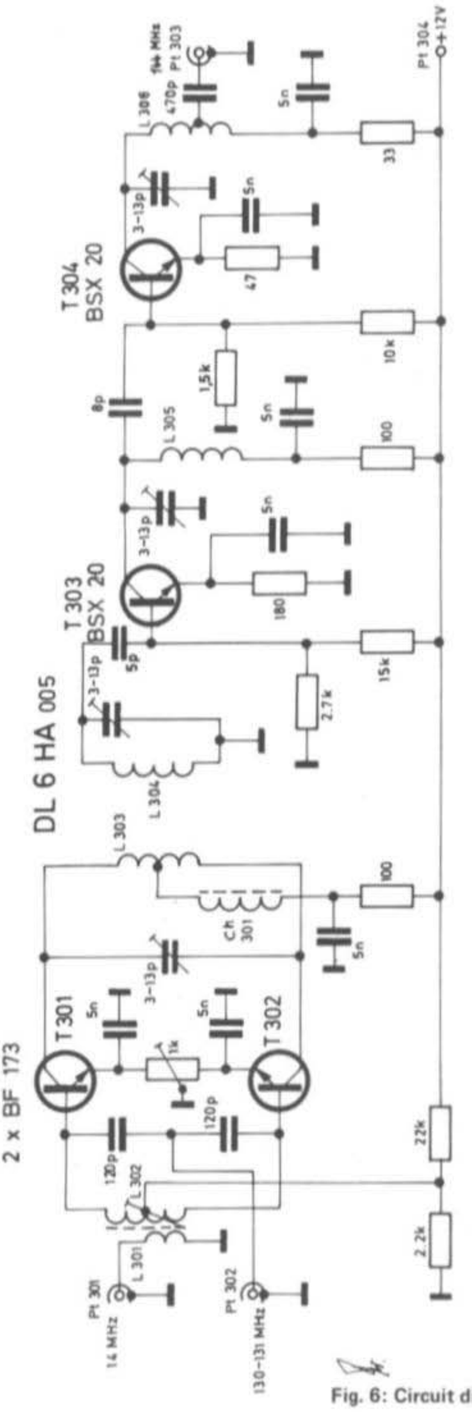


Fig. 2: Circuit diagram of the 144 MHz MOSFET converter (28 MHz IF)



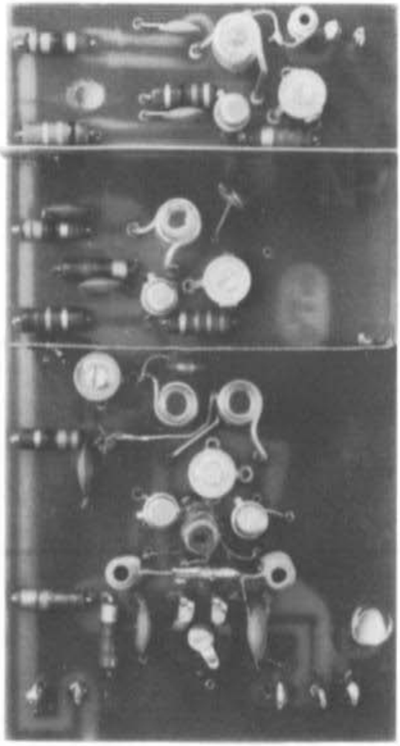


Fig. 9: Photograph of the 14 MHz – 144 MHz transmit converter

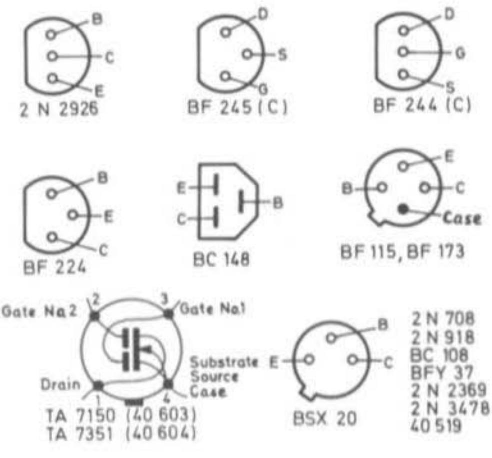


Fig. 6: Circuit diagram of the 14 MHz – 144 MHz transmit converter

L 303: 11.5 turns with center tap
L 304: 10.5 turns
L 305: 11.5 turns
L 306: 12.5 turns. Coiltap 5.25 turns from cold end.

The 36 MHz injection frequency is taken via a 100 pF capacitor connected to a coil tap on L 9 of the receive converter DL 6 HA 001. The coil tap is one turn from the cold end. If a coaxial cable is to be used to connect the 100 pF capacitor to Pt 302, it should be as short as possible.

For operation at the high end of the band, it may be necessary to spread the turns or to remove one or two turns from L 303, L 304, L 305 and L 306.

Note:
It is, of course, possible for other intermediate frequency ranges to be obtained by changing crystal and inductances L 7 to L 10 to suit the frequency.

3. AVAILABLE PARTS

See material list at the end of the magazine.

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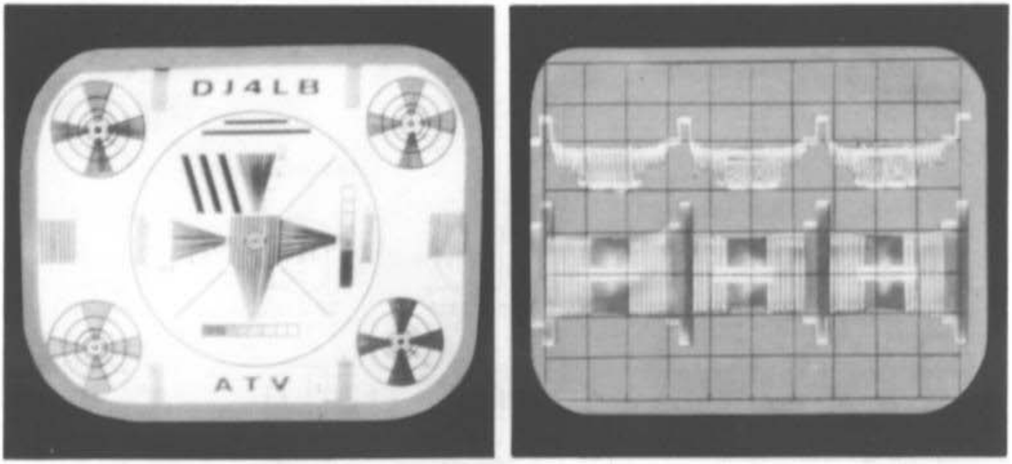

A 12 W DC-DC CONVERTER FOR 12 V/28 V

by H.-J. Franke, DK 1 PN

Although power transistors for 12 V operation on the 2 m and 70 cm bands have been available for some time now, many radio amateurs prefer to use the 28 V types since they are less expensive. Power transistors for frequencies above approximately 1 GHz usually require an operating voltage of 28 V. Since usually only 12 V are available during portable and mobile operation, a DC-DC converter is often required to convert the available 12 V to 28 V.

Such a DC-DC converter with a maximum output power of 12 W is to be described here. Since the converter operates at a frequency of 18 kHz, it has been possible to design a light-weight and compact unit. The efficiency is in excess of 80% due to the use of modern components. Fig. 1 shows the author's prototype.

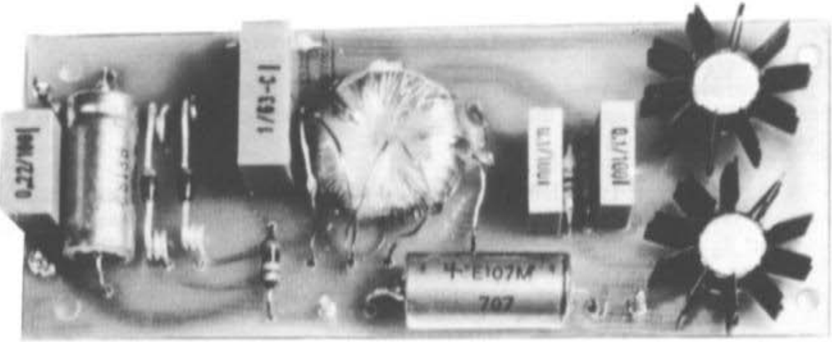

Fig. 1: 12 W DC-DC converter 12 V/28 V DK 1 PN 001

1. CHARACTERISTICS

U_{in} = 12 V

U_{out} = 28 V

$P_{out\ max}$ = 12 W

Operating frequency f = 18 kHz

Dimensions:
100 mm x 40 mm, height 15 mm

Weight: 50 g

Output power P_{out}	Efficiency
3.2 W	77%
6.3 W	83%
9.2 W	86%
11.5 W	84%

2. CIRCUIT DETAILS

Fig. 2 shows a circuit diagram of the DC-DC converter. One of the most common push-pull circuits has been used. In order to obtain the very small transformer dimensions and to design a very low-weight and compact unit, it was necessary for a relatively high operating frequency to be used. This frequency of 18 kHz is above the audible range of the human ear. A high efficiency can only be obtained at this frequency when using special transformer material (high saturation induction = less turns = less copper loss) and fast switching transistors, since the square wave voltages contain harmonics of up to ten times the operating frequency.

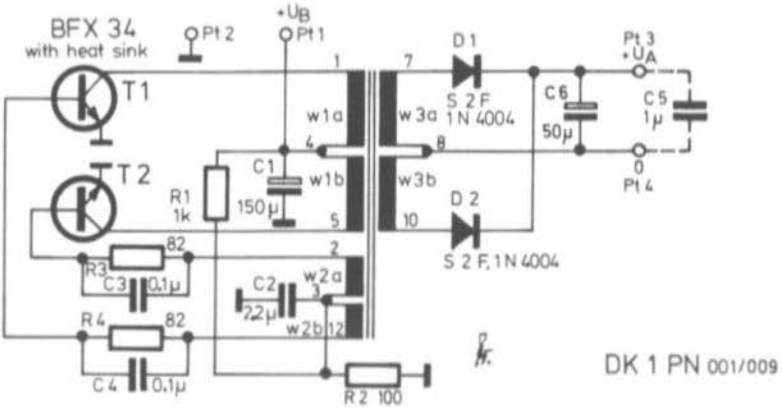

Fig. 2: Circuit diagram of the 12 W DC-DC converter

The transformer is wound on a toroid core manufactured from Ultraperm 10 material. The winding of this transformer is somewhat difficult, but only a few turns are required.

The TO 5 transistor type BFX 34 is used for the switching transistors. They possess a transit frequency of $f_T = 70$ MHz and a residual voltage of $U_{CE(S)} = 1.0$ V at $I_c = 5$ A. The power dissipation is actually so small that the cooling fins are not really necessary. They have only been provided for mobile operation during the summer months where high ambient temperatures can occur.

In order to rectify the fundamental and harmonics of the 18 kHz square-wave voltage at high efficiency, it is necessary for fast diodes, e.g. diodes with a short reverse recovery time to be used. Diodes type 1 N 4004 are suitable for this. In addition to the 50 µF capacitor, a capacitor of 1 µF is provided for filtering the higher frequency components. The push-pull rectifier circuit used in this DC-DC converter has the advantage over a bridge rectifier circuit that the voltage drop of only one diode is present per half-cycle. The output is not grounded, which means that the outputs of several DC-DC converters can be connected in series. Under non-load conditions, the output voltage will increase to more than 30 V due to the overshoot of the squarewave voltage. If capacitor C 6 is not able to withstand this voltage, a load resistor of approximately 1 kΩ should be soldered to the output.

Experience has shown that such DC-DC converters have difficulty in commencing oscillation under load, especially at low temperatures. The voltage divider comprising resistors R 1 and R 2 is responsible for starting oscillation. It is usually dimensioned so that the transistors are approximately biased into class B ($U_{BE} \approx 0.3$ V for germanium transistors; $U_{BE} \approx 0.7$ V for silicon transistors). In the case of a short circuit or exchanged windings, no current will flow. The described DC-DC converter is dimensioned so that it will commence oscillation under load even at temperatures of down to -40 °C. This is also valid when the battery voltage has dropped to 9 V. However, the transistors are biased to class AB, which means that in the case of a short circuit or exchanged windings, a quiescent current will flow and heat up the transistors.

During oscillation, the base current is gained from the feedback winding w2 via the combination of R 3/C 3 and R 4/C 4. The higher the current gain of the transistors, the higher the impedances may be (higher efficiency). However, it should be remembered that the current gain decreases with decreasing temperature.

3. CONSTRUCTION

The DC-DC converter is built-up on a printed circuit board whose dimensions are only 100 mm x 40 mm. The printed circuit board, which has been designated DK 1 PN 001, is shown in Fig. 3. The cooling fins of the two transistors should not be allowed to touch. The windings should be wound equally over the surface of the toroid core and fixed with two-component adhesive (Araldit, UHU-Plus). This adhesive is also used for mounting the transformer onto the PC-board after connection. If the DC-DC converter does not commence oscillation, this will mean that either the base or the collector connections to the transformer must be exchanged.

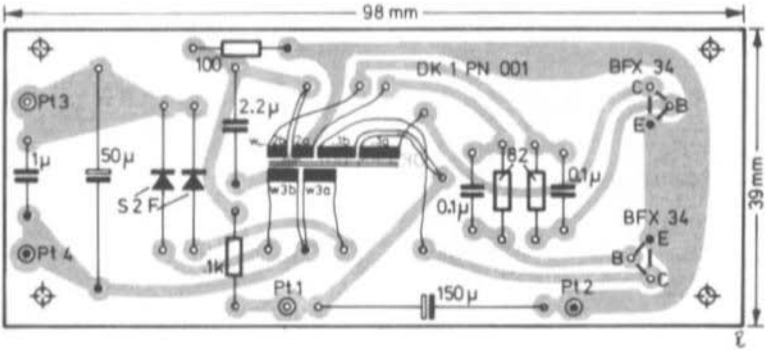

Fig. 3: Printed circuit board DK 1 PN 001 and component locations

3.1. COMPONENTS

T 1, T 2 : BFX 34 (Fairchild)
D 1, D 2 : S 2 F (Semtech), 1 N 4004

C 1 : approx. 150 µF tantalium electrolytic capacitor max. 25 mm long
C 2 : approx. 2.2 µF, plastic foil capacitor, spacing 15 mm
C 3, C 4 : 0.1 µF, plastic foil capacitor, spacing 10 mm
C 5 : approx. 1 µF, foil capacitor, spacing 10 mm
(do not use a tantalium capacitor)
C 6 : approx. 50 µF tantalium electrolytic capacitor, max. 25 mm long

All resistors are for 12.5 mm spacing.
Toroid core 16 x 8 x 5 ZKF Ultraperm 10 x 0.015 (VAC)

Windings:

w1a = w1b: 17 turns of 0.45 mm diameter (25 AWG) enamelled copper wire, both windings wound together (for U_{in} = 6 V: 9 turns of 0.65 mm diameter (22 AWG)
w2a = w2b: 4 turns of 0.1 - 0.3 mm diameter (38 - 29 AWG) enamelled copper wire, both windings wound together.
w3a = w3b: 41 turns of 0.45 mm diameter (25 AWG) enamelled copper wire, both windings wound together.

4. MODIFIED VERSION DK 1 PN 009

Due to the demand for this DC-DC converter in Germany another version has been developed using a ready-wound transformer. The printed circuit board of this model had to be changed to match this transformer and the new board has been designated DK 1 PN 009. The transformer is completely enclosed in plastic and has been especially developed and manufactured for this DC-DC converter. A photograph of the completed unit is given in Fig. 4. The dimensions of the PC-board were changed so that the unit could be accommodated in a Teko case type 2 A. The connection numbers given in Fig. 2 refer to the connections of the ready made transformer. The new PC-board whose dimensions are 65 mm x 50 mm is shown in Fig. 5, together with the component location plan.

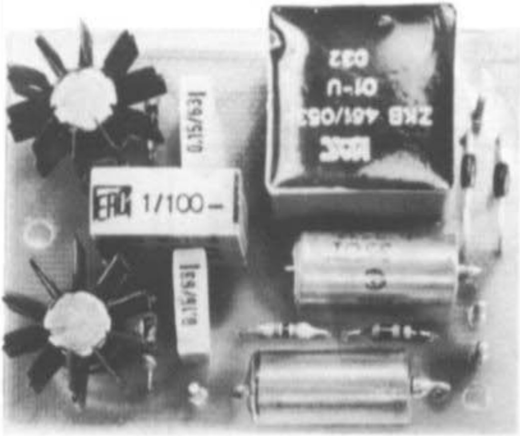

Fig. 4: Photograph of DK 1 PN 009

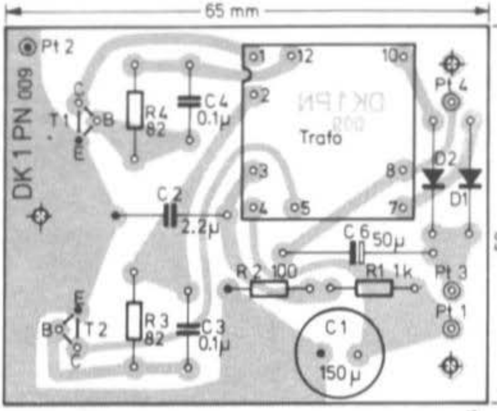

Fig. 5: PC-board DK 1 PN 009

The ready wound transformer also results in better characteristics:

Output voltage (V)	Output current (mA)	Output power (W)	Efficiency %
27.2	400	10.9	
27.0	445	12.0	93
26.9	500	13.5	90

The input voltage was 12.0 V. In practical mobile operation, the input voltage is always slightly higher which also increases the output voltage.

With the exception of the new transformer, the same components are used as in the first version DK 1 PN 001. Capacitor C 5 (1 µF) is no longer accommodated on the new board since it has been found that the filtering of the output voltage is better when connections Pt 3 and Pt 4 are fed through the casing using feedthrough capacitors and C 5 is soldered outside the case.

Of course, the ratio of the transformer is fixed. If other voltages than 28 V are required, the first version DK 1 PN 001 should be used and the transformation ratio changed to suit the required voltage.

5. AVAILABLE PARTS

Please see material price list.

A 200 kHz RECEIVER FOR SYNCHRONIZING 1 MHz OSCILLATORS TO THE DROITWICH LONGWAVE TRANSMITTER

by D. E. Schmitzer, DJ 4 BG

There are a number of transmitters whose frequencies are kept to within very narrow limits and can therefore be used as a standard for frequency measurements. The British longwave transmitter Droitwich transmits on 200.000 kHz and is extremely suitable for such applications, especially since the frequency tolerance of the transmitted carrier was increased from 10^{-8} to approximately 10^{-11} some time ago. It is not intended to give details regarding the short term and long term stability here.

The Droitwich transmitter can be received all over Central and Western Europe and it is only necessary to dimension the antenna to suit the prevailing field strength. In Eastern Europe, difficulties can be caused by Radio Moscow, which operates on the same frequency. Since the Droitwich transmitter is modulated by voice and music information, it is advisable that the actual frequency is not used directly but used to synchronize a crystal oscillator whereby the modulation is completely suppressed. This also has the advantage that the oscillator signal is also present during failure of the standard frequency (Droitwich is closed down between 01.15 to 03.00 GMT), even if at reduced accuracy.

1. CIRCUIT DETAILS

The circuit diagram of the receiver is shown in Figure 1. A gate-protected MOSFET is provided as preamplifier. A separate operating voltage for gate 2 can be fed via connection point Pt 4 of the PC-board. In this manner, it is possible for this voltage to be adjusted with the aid of a potentiometer during alignment so that the level of the input signal is reduced to a level at which the integrated amplifier I 1 will not limit and the various circuits can be aligned with the aid of a meter. After alignment, the potentiometer is removed and point Pt 4 connected to Pt 6.

The loosely-coupled bandpass filter (L 2, L 3, C 7/C 8/C 9) suppresses a good deal of the modulation which means that the carrier is emphasized. The subsequent integrated circuit (I 1) type CA 3076 exhibits a very high gain and possesses very good limiting characteristics. This means that any residual amplitude modulation is suppressed at input levels in excess of approximately 3 µV. A squarewave voltage is present at the output of the integrated circuit which possesses odd harmonics in addition to the fundamental wave. The 5th harmonic, e.g. 1 MHz, is filtered out with the aid of the resonant circuit comprising L 4 and C 13. Since a different frequency is now to be amplified, there is no danger of self-oscillation in spite of the high gain.

The 1 MHz signal is tapped-off via the emitter-follower T 2 and fed firstly to the rectifier circuit for indicating the level, and secondly to the phase discriminator. In the latter case, the field effect transistor T 3 is used as switch which is fed with the 1 MHz signal of the crystal oscillator to be synchronized (3). Transistor T 3 should not require too large a negative voltage at the gate in order to block since it would otherwise not be fully driven by the output voltages of the oscillators used in (1) and (2).

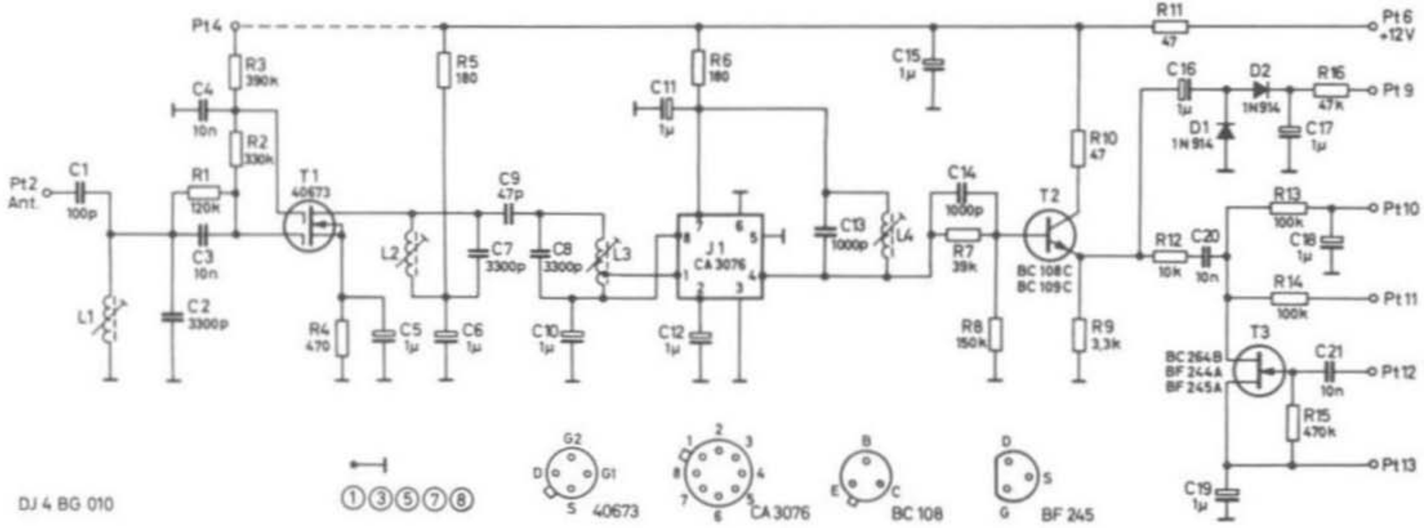

Fig. 1: Circuit diagram of the standard frequency receiver

The control voltage obtained in this manner is filtered by the RC-link comprising R 13/C 18. It can be tapped-off from connection Pt 10 and fed to the varactor diode of the crystal oscillator. The basic bias for the varactor diode can be injected at connection Pt 13. A sensitive meter with centre zero (5 - 0 - 5 μA) can be connected between connections Pt 11 and Pt 13 which can be used for indicating the control voltage. Since such sensitive meters are expensive and difficult to obtain, it is more advantageous to use a meter of 50 or 100 μA with centre zero and to provide a balanced DC-amplifier. A suggested circuit is shown in Figure 2.

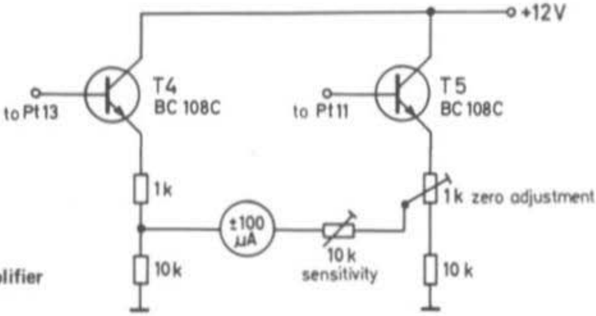

Fig. 2:
Recommended circuit for a DC-amplifier

2. SPECIAL COMPONENTS

I 1: CA 3076 (RCA)
T 1: 40673, 40820, 3 N 187 (RCA), or similar
T 2: BC 108 C, BC 109 C or similar NPN AF transistors with B≥100.
T 3: BF 244 A, BF 245 A, BC 264 B or similar field effect transistors that are blocked at low negative gate bias voltages.
T 4, T 5: BC 108 C, BC 109 C or similar NPN AF transistors with B≥100.

D 1, D 2: 1 N 914 or similar silicon diodes

C 2, C 7, C 8: 3300 pF styroflex capacitors
C 13, C 14: 1000 pF styroflex capacitors

All electrolytic capacitors: Tantalium electrolytics.

2.1. INDUCTANCES

All inductances are wound with stranded wire (10 x 0.05 mm). The thinnest possible silk or cotton-covered wire should be used and wound carefully with a slight opposite pressure since the available space will otherwise not be sufficient. This is especially valid for L 3 since the tapping requires additional space. It is therefore advisable for L 1 and L 2 to be wound firstly in order to obtain an idea of the available space. If these measures are followed, there should be no difficulties. Solid wires are not suitable for this case since the Q of such inductances is not even half that of inductances wound with stranded wire. This would mean a larger bandwidth and subsequently a considerable reduction of the gain and sensitivity.

L 1, L 2: 87 turns of 10 x 0.05 mm dia. silk-covered stranded copper wire.
L 3 : 87 turns, tapped at 12 turns, wire as for L 1
L 4 : 32 turns, wire as for L 1.

The connections of the inductances are given in the component location plan shown in Figure 3. The coil former set provided with the kit contains all required parts including screening cans and stranded wire. If an inductivity meter is available, inductances L 1 to L 3 can be prealigned to 190 μH and L 4 to 25 μH; this simplifies the alignment of the receiver considerably. During this, it is necessary for the two halves of the potted core to be depressed between the fingers. It is only after it has been established whether the required inductivity can be obtained that the two halves of the potted core are glued together with a thin film of two-component glue (UHU-Plus, or similar). After glueing, the two halves should be pressed together and the excessive adhesive carefully removed. The potted cores should then be allowed to stand overnight to harden.

3. CONSTRUCTION

The described circuit is accommodated on the printed circuit board DJ 4 BG 010 whose dimensions are 65 mm x 90 mm. These dimensions are suitable for the modular system described in (4), which means that the receiver can be accommodated in a screened Teko-case. The connections can be in the form of a 13-pole connector or in the form of connection pins.

The printed circuit board and the component location is shown in Figure 3. A photograph of the prototype is given in Figure 4. With the exception of the four inductances, which have been described in detail, the construction will not present any difficulties.

4. ALIGNMENT

The standard frequency receiver can be aligned extremely easily if a calibration generator is available that is able to provide frequency markers at 1 MHz and 100 kHz and harmonics thereof. Firstly, a 1 MHz signal is fed via a low-value capacitor (e.g. 5 pF) to the hot end of L 4 and its core is aligned for maximum reading on the level meter (Pt 9). By injecting a weak 200 kHz signal (harmonics of the calibration generator in the 100 kHz position) to L 3, and afterwards L 2, it is possible for these inductances to be aligned in the same manner. If the level meter indicates a value of virtually full scale deflection, that does not alter, this will indicate that the integrated amplifier I 1 is limiting. The input level should then be reduced, e.g. the injection to L 3 or L 2 should be reduced. The 200 kHz signal is injected to the input at even lower level and inductance L 1 is aligned in the same manner.

If it was possible for the inductances to be prealigned on an inductivity meter it is possible for the 200 kHz signal to be connected directly to the input and for L 3, L 2 and L 1 to be aligned in that order. If the integrated amplifier I 1 limits (the output level does not alter on rotating the cores), it is possible for the voltage at Pt 4 (gate 2 of T 1) to be reduced with the aid of a potentiometer. The gain will be reduced in this manner and it is once again possible for the alignment to be observed on the meter. The nearer one comes to the final alignment of all circuits, the lower will be the voltage at gate 2 in order to keep I 1 from limiting. If the adjustment range of the potentiometer is not sufficient, it will be necessary for the injection from the calibration generator supplying the 200 kHz signal to be reduced. The manual adjustment range can be extended if the base connection of the potentiometer is not grounded but fed with a negative voltage, e.g. with -4.5 V.

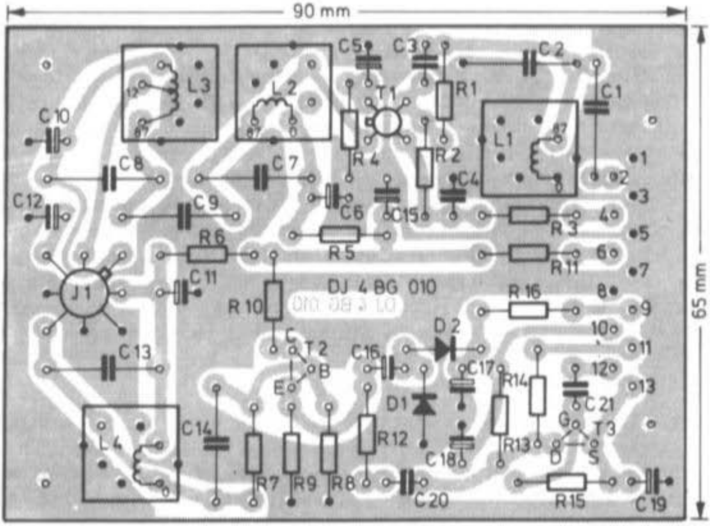

Fig. 3: Printed circuit board DJ 4 BG 010 and components

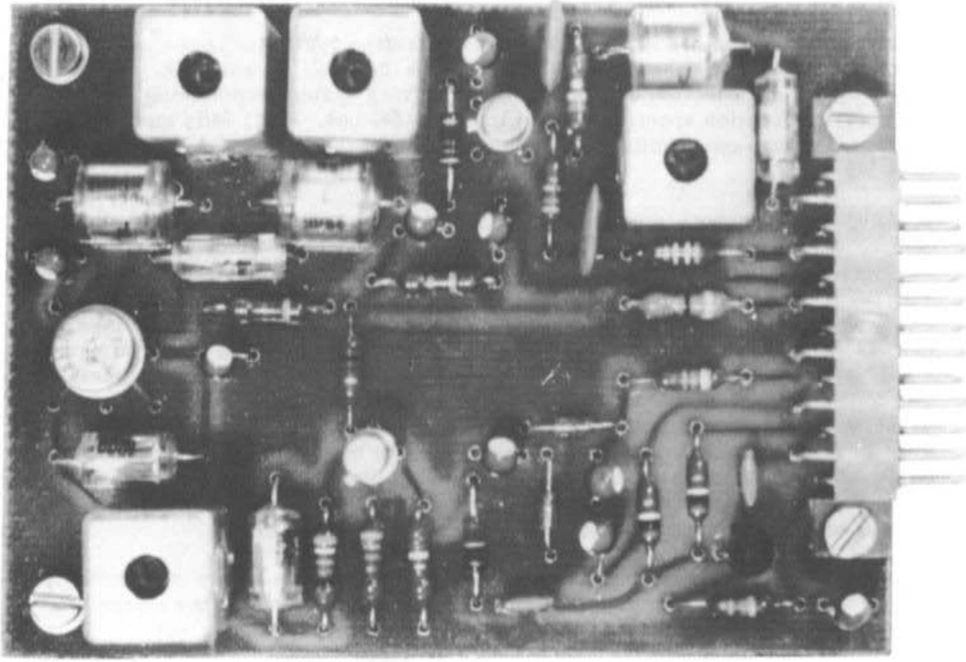

Fig. 4: Prototype of the standard frequency receiver

After all circuits have been aligned in this manner, the antenna should be connected to the input. With a long-wire antenna of at least 10 m in length, it will be possible for the Droitwich transmitter to be received anywhere in Central and Western Europe and to obtain full scale deflection on the level meter at full gain. The gain is now reduced until the level indicator falls distinctly after which all circuits are carefully corrected. It will be inductance L 1 that will be detuned by the antenna and will require the most correction. After this, the potentiometer can be removed from connection Pt 4 and the latter connected to Pt 6 ($+U_b$). The alignment is thus completed. It should be mentioned that the previous comments with respect to the 10 m long-wire antenna are valid for Central Europe. Due to the higher signal strength in such countries as Great Britain, Holland, Belgium, France, etc. it will not be necessary for such large antennas to be used. It is therefore only necessary for the antenna to be dimensioned to suit the field strength at the actual location.

5. LIGHTNING PROTECTION

Since the standard frequency receiver will often be used on an outdoor antenna, it will be necessary for lightning protection measures to be taken. These should be located as near to the antenna socket as possible. In most cases, two silicon diodes connected parallel to the antenna socket with opposite polarity will provide sufficient protection. It is important that the receiver is well grounded.

6. CONNECTION TO THE OSCILLATOR CIRCUITS

The synchronization of the 1 MHz crystal-controlled oscillator in the calibration spectrum generator described in (2) is to be used as an example. Figure 5 shows the interconnection of a complete measuring system comprising the three modules: calibration spectrum generator DJ 4 BG 004, 1.001 MHz auxiliary oscillator and voltage stabilizer DJ 4 BG 005 and standard frequency receiver DJ 4 BG 010.

Firstly, a shortcircuit is made between connections Pt 10 and Pt 13 so that the control voltage does not have any effect on the varactor diode of the crystal oscillator. The pointer of the control voltage meter should vibrate in rhythm with the frequency difference between standard and crystal frequency, of course, only when this frequency difference is not too great. The basic bias voltage for the varactor diode is adjusted to approximately 8 to 10 V and the trimmer of the crystal oscillator is aligned until the pointer oscillates as slowly as possible. The alignment tool should be made from plastic or similar material since the capacitance of a normal screwdriver would detune the oscillator considerably, even if it is insulated. The potentiometer for the bias voltage of the diode can be corrected. If the shortcircuit between Pt 10 and Pt 13 of the standard frequency receiver is now removed, the crystal-controlled oscillator will be synchronized to the standard frequency.

If the bias voltage of the diode is carefully altered, the phase discriminator will generate a control voltage (of approximately ± 0.5 V) and the control voltage meter will indicate this change. If the variation of the bias voltage is too great, the hold range of the control circuit will be exceeded and the pointer of the meter will start swinging back and forth. The synchronized condition can be reobtained by reducing the bias voltage. The lock-in and hold range of the synchronizing circuit can be increased somewhat when the oscillator is not pro-

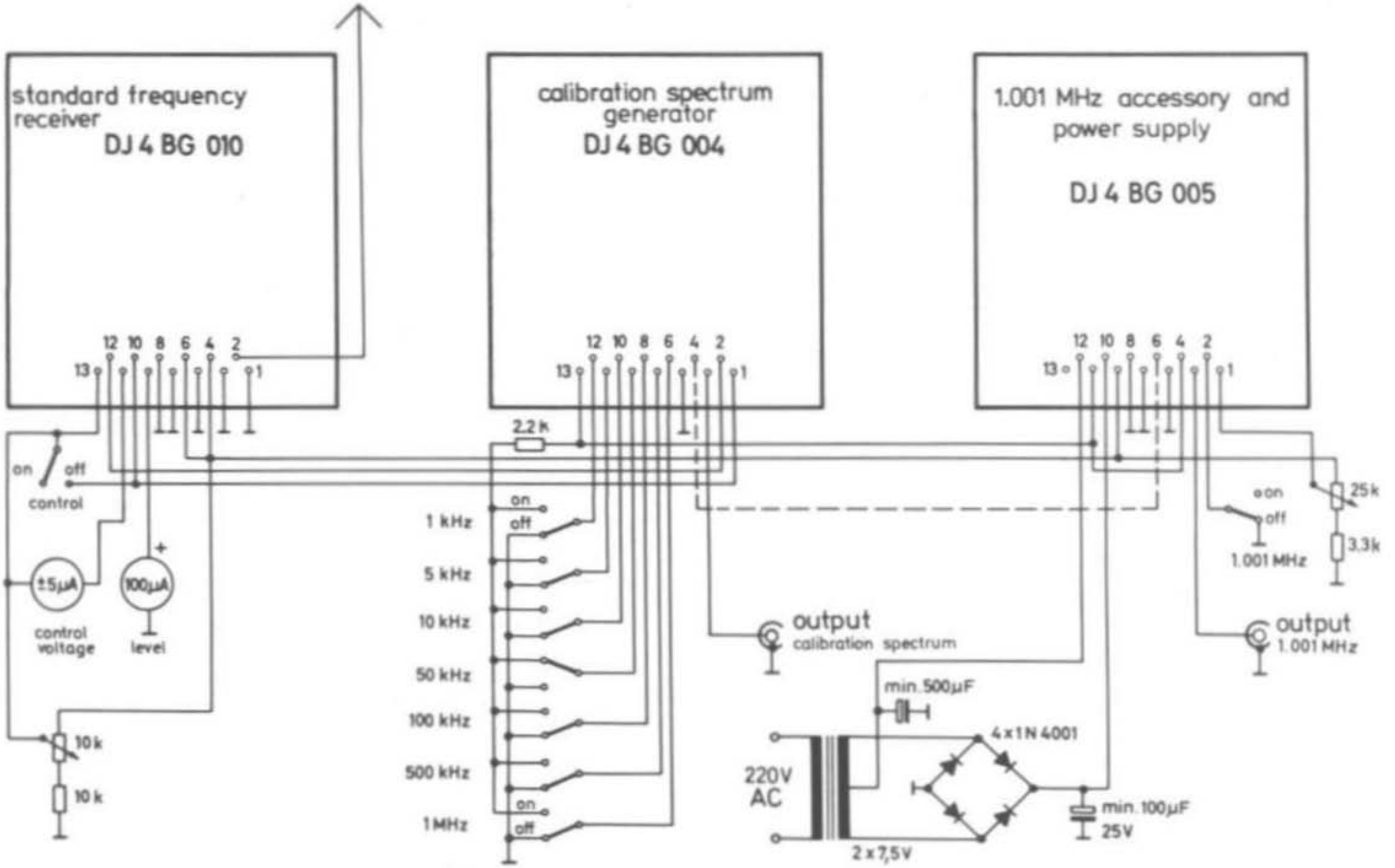

Fig. 5: Interconnection of the standard frequency receiver and the calibration spectrum generator

vided with one varactor diode (BA 110), but with two in parallel. The second diode can then replace C 2 (see Fig. 3 in (2)). In this case, capacitor C 3 should be decreased accordingly.

When once aligned, the synchronization maintains itself over a considerable period. Although the Droitwich transmitter interrupts transmission for a period of 2.5 hours each night, the circuit re-synchronizes the crystal oscillator immediately on reappearance of the signal.

The signal of the 1.001 MHz accessory can be separately taken from connection Pt 3 of module DJ 4 BG 005 and fed to a separate output socket. The 1 kHz higher signal can also be tapped off from connection Pt 6 of module DJ 4 BG 005 and fed to the last vacant input of the 8-gate in module DJ 4 BG 004. In this case, the signal is fed via the pulse shaper of this module and fed to the same output as the calibration spectra of the calibration spectrum generator. This is shown in Figure 5 in the form of a dashed line. If this is the case, the pulse shaper of module DJ 4 BG 005 can be deleted and components I 2, R 13, R 14 and C 8 will not be required.

The synchronized calibration spectrum generator shown in Figure 5 can also be used for other applications. If, for instance, a 1 MHz signal of high accuracy is required as control oscillator for a frequency counter or clock, it is possible for the module DJ 4 BG 005 to be equipped with a 1 MHz crystal which can be synchronized to the frequency of the Droitwich transmitter instead of the calibration spectrum generator DJ 4 BG 004.

If the resistor network at the output of the pulse shaper is deleted and the signal is directly fed to the output, the pulse present at this position will be sufficient to drive further frequency divider stages. In this manner, it is possible to divide, for instance, the frequency down to 50 Hz in order to drive the synchronous motor of a clock via a power amplifier. The same is valid for module DJ 4 BG 004 where the frequency can be divided down to 1 kHz. In this case, only a 10 to 1 divider (SN 7490 N) and a flip-flop (e. g. SN 7472 N) are required to divide the signal to 50 Hz. Of course, it is again necessary for the output resistors of the module to be deleted. Naturally, the pulse shaper and the gates can be deleted and the extra dividers to be connected directly to the output of the last divider (connection 11 of I 7 of DJ 4 BG 004).

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FM REPEATERS IN GERMANY

by T. Bittan, G 3 JVQ/DJ 0 BQ

A large number of repeater stations are now operating in Germany. Since radio amateurs in other countries are thinking of establishing such networks and since the repeaters can also be used as beacons, it is thought that a short report regarding the German network may be of interest.

Repeater stations have been operating in Germany for several years now mainly as a result of the inexpensive FM-mobile stations that were available on the surplus market. This equipment was designed for operation from about 146 MHz to 158 MHz with a channel spacing of 50 kHz and can easily be modified for the amateur band. Most repeaters are located at hill top locations usually together with police or other radio services. Two outstanding locations are the repeater on the Zugspitze mountain (the highest mountain in Germany) and the Stuttgart repeater on their famous television tower. The Zugspitze-repeater has a coverage of about 300 km in diameter, of course, from a mobile station. Fixed stations with directional antennas can work over the repeater from far greater distances. However, the normal coverage of most repeaters is in the order of 60 - 70 km in most directions.

1. CHANNELS

The first repeater frequency plan used a channel spacing of 50 kHz which was sufficient at first due to the limited number of stations and suited the surplus transceivers which used a greater frequency deviation. However, with growing popularity of this operating mode and the large number of repeaters operating, it has been necessary for a new frequency plan to be established with a 25 kHz spacing between channels.

All repeaters have a common spacing of 1.6 MHz between transmit and receive channels. The repeater receives low-band in the range of 144.150 - 144.300 MHz and retransmits this signal in the range of 145.700 - 145.850 MHz. The exceptions to the 1.6 MHz spacing are channels R 7 and R 8 which are later editions and were placed in between the existing channels R 2 to R 6. The channels are as follows:

STATIONS:

Channel R 2

DB 0 WF	Berlin	GM 47 a
DB 0 YC	Cham	GJ 74 c
DB 0 UC	Coburg	FK 55 c
DB 0 WW	Duisburg	DL 44 c
	Feldberg (Rhein-Main)	EK 63 h
DB 0 XH	Hamburg	EN 40 d
DB 0 WH	Hannover	EM 49 d
DB 0 ZF	Kaiserstuhl (Freiburg)	DI 79
DB 0 XE	Kassel	EL 57 e
DB 0 WK	Konstanz	EH 26 d
DB 0 WL	Lahr	DI 60 a
DB 0 ZM	Muenchen/Munich	FI 78 a
DB 0 UN	Nuernberg/Nuremberg	FJ 47 a
DB 0 UO	Oldenburg	EN 62 f
DB 0 ZO	Osnabrueck	EM 61
DB 0 WR	Stuttgart	EI 17 d
DB 0 WB	Winterberg (Au/Inn)	GI 62 j

Channel R 3

DB 0 VB	Bad Koenig (F 35)	EJ 15 d
DB 0 XY	Bocksberg (Harz)	FL 12 b
DB 0 WG	Goeppingen	EI 30 g
DB 0 VK	Koeln/Cologne	DK 05 j

Channel R 4

DB 0 WA	Aachen	DK 11
DB 0 UA	Augsburg	FI 55 b
DB 0 XB	Baltic coast	FO 74 b
DB 0 UB	Bamberg	FJ 05 a
DB 0 UG	Bentheim-Lingen	DM 56 c
DB 0 WC	Bremerhaven	EN 33 c
	Darmstadt	EJ 24 a
DB 0 XD	Deggendorf	GI 15
DB 0 WD	Deister	DM 58
DB 0 ZR	Dortmund	DL 47 c
DB 0 XR	Border of DL/HB/F	DH 30 a
DB 0 ZZ	Grab	EJ 78 c
DB 0 XG	Greding	FJ 77 c
	Hersfeld	EK 19 a
DB 0 YK	Homberg-Kaiserslautern	DJ 47 e
	Koblenz	DK 49 j
DB 0 WO	Leer (Ostfriesland)	DN 68 a
DB 0 YN	Lindau-Northeim (Hann.)	FL 21 g
DB 0 ZL	Luechow (Elbe)	FN 65 j
DB 0 YS	Siegen	EK 01 f
DB 0 WX	Triberg	EI 72 a
DB 0 ZW	Weiden	GJ 22 c

Channel R 5

	Berlin-Neukoelln	GM 48 j
DB 0 WE	Essen (at present still Ch 6)	DL 45 d
DB 0 VF	Frankfurt	EK 64 e
DB 0 XM	Hoher Meissner	EL 7 0 a
DB 0 WN	Ochsenwang	EI 38 j
DB 0 VP	Pirmasens	DJ 69 g

Channel R 6

	Andernach-Mayen	DK 47 b
DB 0 ZA	Aschberg (Rendsburg)	EO 49 g
DB 0 XO	Bergheim	DK 0 4 a
DB 0 WU	Bremen	EN 75 g
DB 0 WT	Detmold	EL 0 5 g
DB 0 WE	Essen	DL 45 d
DB 0 WS	Goslar-Steinberg	DL 0 3 f
DB 0 ZH	Heidelberg	EJ 44 e
DB 0 YH	Hoechenschwand (Black Forest)	EH 21 b
DB 0 WV	Hoechsten (Oberschwaben)	EH 17 c
	Koeln-Bergheim	DK 0 4 a
DB 0 XU	Knuell	EK 0 8 f
DB 0 XS	Merzig (Saar)	DJ 43 c
DB 0 WM	Muenster (Westf.)	DL 0 9 h
DB 0 ZB	Ochsenkopf	FK 8 0 f
DB 0 WZ	Wuerzburg	

Channel R 7

DB 0 ZU	Zugspitze (at present still Ch 6)	FH 46 g

Channel R 8

DB 0 XA	Altenwalde	EN 14 f
	Kalmit	DJ 51 j
DB 0 YY	Ludwigsburg	EJ 76 f

A number of callsigns have still to be issued or will be changed to the DB-series.

Most repeaters have an output power of 5 to 20 watts and use vertically polarized, omnidirectional antennas. They are opened by a frequency modulated sinusoidal calling tone of 1750 Hz. In addition to the automatic callsign generator (F 3), the repeaters are equipped with numerous electronic gadgets such as circuits for limiting the transmission time of each station, automatic switch-off when the frequency deviation of one of the stations is too great, automatic alarm signal etc. In contrast to some repeater stations in the USA, no attempt has been made to limit the accessibility by use of coded calling tones or tone bursts.

2. CONSEQUENCES OF REPEATER OPERATION

At first one might assume that the main result would be that the two metre band would be full of repeater stations and that there would be no room left for other modes. However, the opposite is the case. Activity is now virtually limited to the input and output frequencies of the FM-repeaters and to a few tens of kHz each side of the SSB centre frequency of 145.4 MHz. AM-activity has virtually ceased to exist in Germany.

Repeaters are, of course, extremely favourable for mobile operation where the FM-mode offers considerable advantages over AM, and the mobile amateur does not have to keep tuning over the band whilst driving. Repeaters have also proved extremely useful for emergencies and road accidents where the amateur on the spot can request assistance of police and ambulance via the repeater.

Experience has shown that the repeater network must be organized on a national basis to avoid interference between neighboring repeaters. The less channels are used for the network the more important is nationwide planning of sites, frequencies and coverage.

Since reciprocal licence agreements exist between Germany and a large number of other countries, this frequency list may be of advantage to you in equipping your mobile station for the holidays.

A WIDEBAND RING MIXER WITH SCHOTTKY DIODES

by R. Lentz, DL 3 WR

Wideband ring mixers have been used in the carrier frequency technology for decades. Even the first, copper oxide diodes that were used in ring mixers exhibited a high carrier suppression. Unfortunately, the high noise level and conversion loss of these diodes limited their application to audio and intermediate frequencies. Modern Schottky or hot-carrier diodes do not exhibit these disadvantages. In contrast to professional electronics, such ring mixers have not been used to any extent in amateur radio equipment. However, they offer considerable advantages over transistor push-pull or even single stage mixers:

1. They suppress both the signal and local oscillator frequency. This means that only the required frequency and its image are available at the output. Due to this characteristic, far less filtering is required, or the spectral purity will be greater if the same amount of filtering is used.

2. In contrast to ring mixers with resonant circuits, wideband ring mixers have the advantage that they maintain their high balance, and thus the characteristics mentioned in 1., even during temperature fluctuations, and are not affected by aging or mechanical shock.

3. They possess excellent large-signal characteristics, in other words, they are able to process high signal levels before noticeable intermodulation or cross modulation effects are observed.

4. The conversion is made with a very low harmonic distortion factor which means that very few harmonics are generated.

5. Mixers equipped with Schottky diodes exhibit a very low noise figure which means that they can be used even in a receiver without RF-amplifier stage in order to obtain optimum large signal characteristics. However, the subsequent intermediate frequency amplifier has an essential effect on the total noise figure due to the conversion loss.

All transmit and receive mixers will profit from one or more of the above mentioned characteristics. The extra work and expenditure required to construct a ring mixer is very small.

1. CIRCUIT DETAILS

Figure 1 gives the circuit diagram of a practical wideband ring mixer which is additionally equipped with a pulse-shaper circuit for the local oscillator or VFO frequency. This results in very steep pulses at the local oscillator frequency which switch the mixer diodes very rapidly during the very short rise and fall times. The fast switching is very important with respect to obtaining a low distortion factor and low conversion loss. In order to obtain a rise time and fall time in the order of 5 ns, fast switching transistors having a low collector-base capacitance are required.

The low-pass filter comprising choke Ch 1 and capacitor C 1 ensure that no residual pulses can be fed via the power supply connections to other modules.

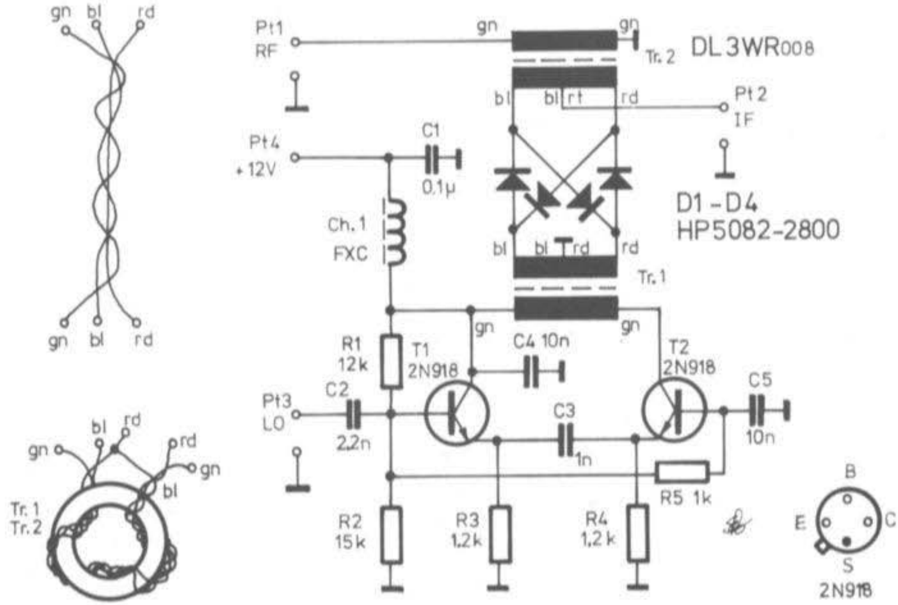

Fig. 1: Circuit of a wideband ring mixer with Schottky diodes and pulse-shaper

The balance of the mixer and thus the characteristics given in 1. are dependent on the matching of the four diodes and on transformers Tr 1 and Tr 2. Both transformers are identical and comprise toroid cores provided with three windings of three wires. This special winding technique - which has been taken from professional electronic applications - is also shown in Figure 1. Since all windings are equally long and closely coupled, both good balance and wideband characteristics are guaranteed. The colours are given only as an example to show the interconnection. Of course, windings of the same colour wire can be used and each winding checked with an Ohmmeter; however, any error will make the mixer completely useless.

2. CHARACTERISTICS

The mixer module was examined in a professional laboratory and was found to have the following characteristics:

Parameter: VFO (local oscillator) signal: 37-39 MHz, 300 mV
Signal frequency: 28-30 MHz
Intermediate frequency: 9 MHz

Measured values:
Rise time of the local oscillator frequency pulses: approx. 4 ns
Fall time of the local oscillator frequency pulses: approx. 4 ns
Can be driven to a signal level of 100 mV without distortion
3 dB desensitization at a signal level of 450 mV
Conversion loss: 7.4 dB; Carrier suppression: 36 dB

Unmatched diodes of type HP 5082-2800 were used. The carrier suppression could be increased to 46 dB by placing a grounded metal tab near one of the diodes (must be found by experiment).

The level of the local oscillator (VFO) frequency may be in the order of 100 mV to 700 mV without noticeably affecting the characteristics of the mixer. The conversion loss increases rapidly on reducing the level below 100 mV.

With the dimensioning given for the described mixer, the lowest permissible local oscillator frequency will be in the order of 3 MHz. However, if the mixer is to be used for lower frequencies, the number of turns on transformers Tr 1 and Tr 2 can be easily increased to ten which means that the lower frequency limit is in the order of 500 kHz. The wire used for the transformers should not be thicker than 0. 2 mm diameter (32 AWG).

The upper frequency limit of the mixer is dependent on the rise time of the local oscillator frequency pulses; e.g. the switching time should be short in comparison to one cycle of the signal frequency. With a rise time of 4 ns, this will be valid for signal frequencies of up to approximately 100 MHz. At higher frequencies, the pulse shaper will no longer be useful; it is then deleted and the frequency of the local oscillator or VFO directly fed to the primary of transformer Tr 1. The other end of the primary winding is then grounded. A local oscillator level of at least 300 mV is then required.

Two very important characteristics, namely the large signal capacity and the relatively low noise factor of less than 10 dB, will be maintained up to approximately 400 MHz (1).

Finally, it should be mentioned that the impedance connected to the RF-input will also appear at the IF-output and vice versa. This means that an impedance ratio of 1 : 1 exists between Pt 1 and Pt 2. The upper limit of the impedance is given by the intrinsic impedance of the transformer winding which is dependent on the A_L-value of the ferrite toroid, the number of turns and the frequency. It is calculated as follows:

$$X_L = t^2 \times A_L \times 2\pi f$$

where t = number of turns; A_L = inductivity constant in 10^{-9} H/turn^2;
f = frequency in Hz.

With the described toroids and number of turns, a value of approximately 1200 Ω results at 9 MHz. This allows, for instance an FM crystal filter XF-9E to be connected to the output of the mixer and the required load impedance of 1200 Ω to be realized at the RF-input.

3. SPECIAL COMPONENTS

T 1, T 2:	2 N 918, BSX 27 (SGS), BFY 90 (Philips)
D 1 - D 4:	HP 5082-2800, HP 5082-2811 (Hewlett-Packard), FD 700, FD 777 (SGS), BAX 25 (AEG-Telefunken)
Tr 1, Tr 2:	Ferrite toroid R 6 x 2 x 2; material 500 M 25; A_L = 890 (e.g. Siemens R 6,3 B 64290-A 0037-X 030) Wind with five turns of 0.2 - 0.3 mm dia. (32 AWG) silk covered or enamelled copper wire
Ch 1:	Wideband ferrite choke (Philips 4312 020 36701)

The given capacitance values are not too critical. However, low-inductive types should be used.

4. CONSTRUCTION

The mixer module is accommodated on a printed circuit board with the dimensions 65 mm x 50 mm. It is designated DL 3 WR 008. Figure 2 illustrates this board and shows the location of the individual components. It is important that the connections in the area where the fast local oscillator pulses exist are made at low capacitance (short connections). The printed circuit board should be located approximately 10 mm from the casing for the same reason.

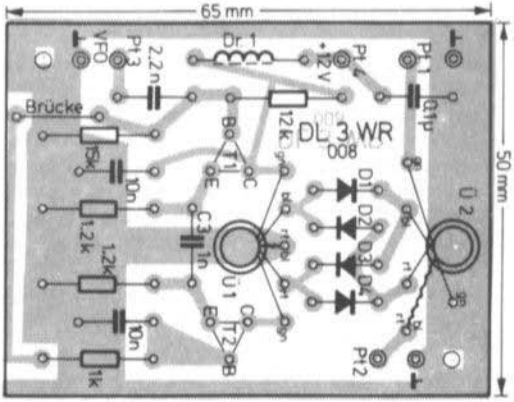

Fig. 2: Printed circuit board DL 3 WR 008

The mixer module can be installed in a screened casing of its own or, for instance, in a VFO casing (The author used the latter as can be seen in Figure 3). The casing is only provided with a feedthrough capacitor for the operating voltage and two low-capacitance feedthroughs for the RF or IF connection. Since the toroids exhibit a very low field, this construction ensures that the local oscillator (or VFO) frequency, and the harmonics contained in the pulses, are not radiated from the connections and circuit lanes.

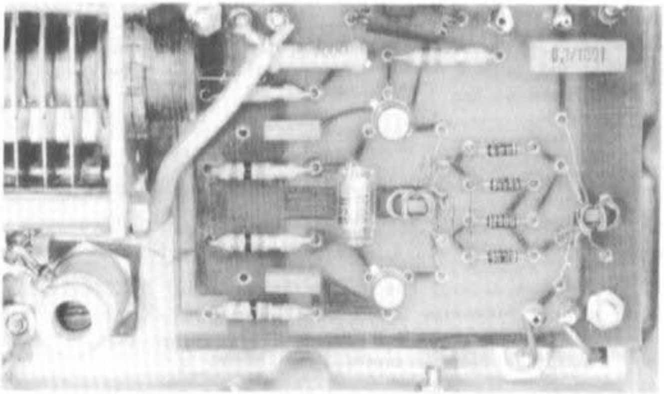

Fig. 3: The wideband ring mixer installed in a VFO casing

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Ham radio magazine, vol. 3 (1970), Edition 3, Pages 8-17

MATERIAL PRICE LIST OF EQUIPMENT
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Semiconductors	DC 6 HL 009	(B 12-12, B 25-12, CTC)	DM 108.--
Minikit 1	DC 6 HL 009	(1 transistor, 4 diodes, 1 IC, 5 ferrite beads, 1 ferrite choke, 6 trimmer cap., 3 feedthrough cap., 1 heatsink, 2 miniature relays, 1 TEKO box 3 A)	DM 68.75
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U 3	5/8 λ	400-470 MHz	3 dB	100 g	Silver-plated, epoxy coated
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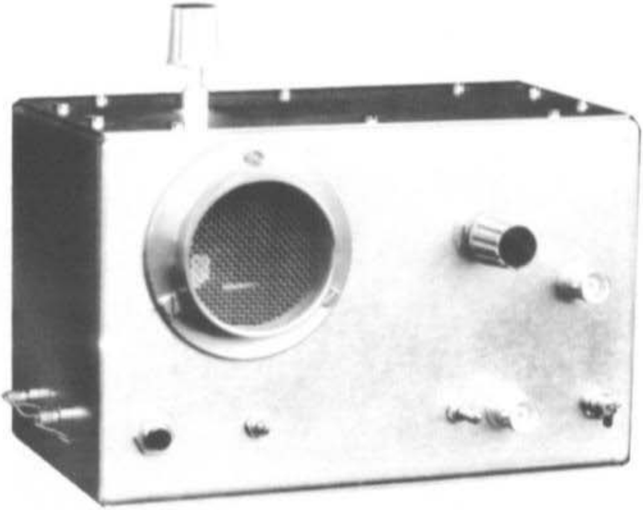
LINEAR AMPLIFIER FOR 70 cm USING A 2 C 39 TUBE

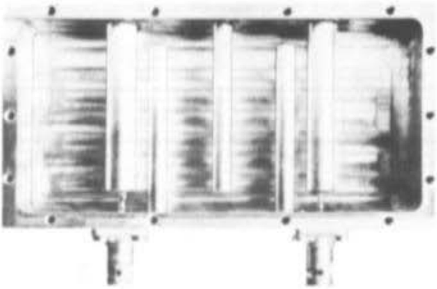
Further development of the amplifier described in edition 3/72 VHF COMMUNICATIONS

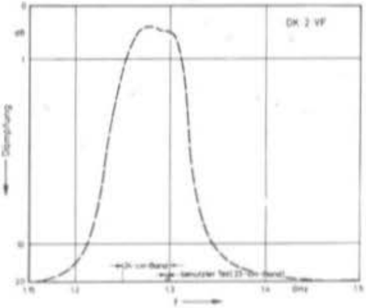


- Silver-plated
- Ready-to-operate
- Supplied without tube, power supply and blower
- Price DM 438,—

INTERDIGITAL BANDPASS FILTER FOR 23 cm

Similar to that described in edition 3/71 VHF COMMUNICATIONS





The advantages of interdigital filters is well known: High stopband attenuation and low insertion loss. The curve and insertion loss values are virtually independent of the matching (the filter can be shorted or have an open circuit when an isolation of at least 6 dB exists between this point and the filter).

Price with BNC sockets	DM 92.00
Price without sockets	DM 86.50

Verlag UKW-BERICHTE, H. Dohlus oHG

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CRYSTAL FILTERS - FILTER CRYSTALS - OSCILLATOR CRYSTALS

SYNONYMOUS for QUALITY and ADVANCED TECHNOLOGY

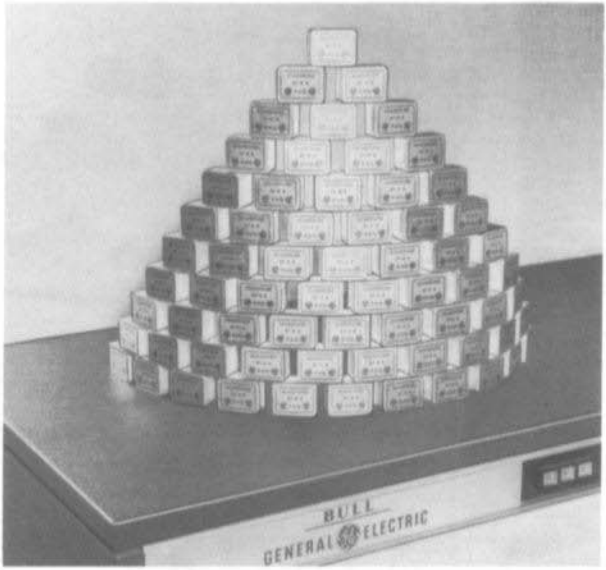
PRECISION QUARTZ CRYSTALS. ULTRASONIC CRYSTALS.

PIEZO-ELECTRIC PRESSURE TRANSDUCERS

Listed is our well-known series of

9 MHz crystal filters for SSB, AM, FM and CW applications.

In order to simplify matching, the input and output of the filters comprise tuned differential transformers with galvanic connection to the casing.



Filter Type		XF-9A	XF-9B	XF-9C	XF-9D	XF-9E	XF-9M
Application		SSB-Transmit.	SSB	AM	AM	FM	CW
Number of Filter Crystals		5	8	8	8	8	4
Bandwidth (6dB down)		2.5 kHz	2.4 kHz	3.75 kHz	5.0 kHz	12.0 kHz	0.5 kHz
Passband Ripple		< 1 dB	< 2 dB	< 2 dB	< 2 dB	< 2 dB	< 1 dB
Insertion Loss		< 3 dB	< 3.5 dB	< 3.5 dB	< 3.5 dB	< 3 dB	< 5 dB
Input-Output	Z_t	500 Ω	500 Ω	500 Ω	500 Ω	1200 Ω	500 Ω
Termination	C_t	30 pF	30 pF	30 pF	30 pF	30 pF	30 pF
Shape Factor		(6:50 dB) 1.7	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:40 dB) 2.5
			(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 2.2	(6:60 dB) 4.4
Ultimate Attenuation		> 45 dB	> 100 dB	> 100 dB	> 100 dB	>90 dB	> 90 dB

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