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VHF COMMUNICATIONS

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ESPECIALLY COVERING VHF, UHF AND MICROWAVES

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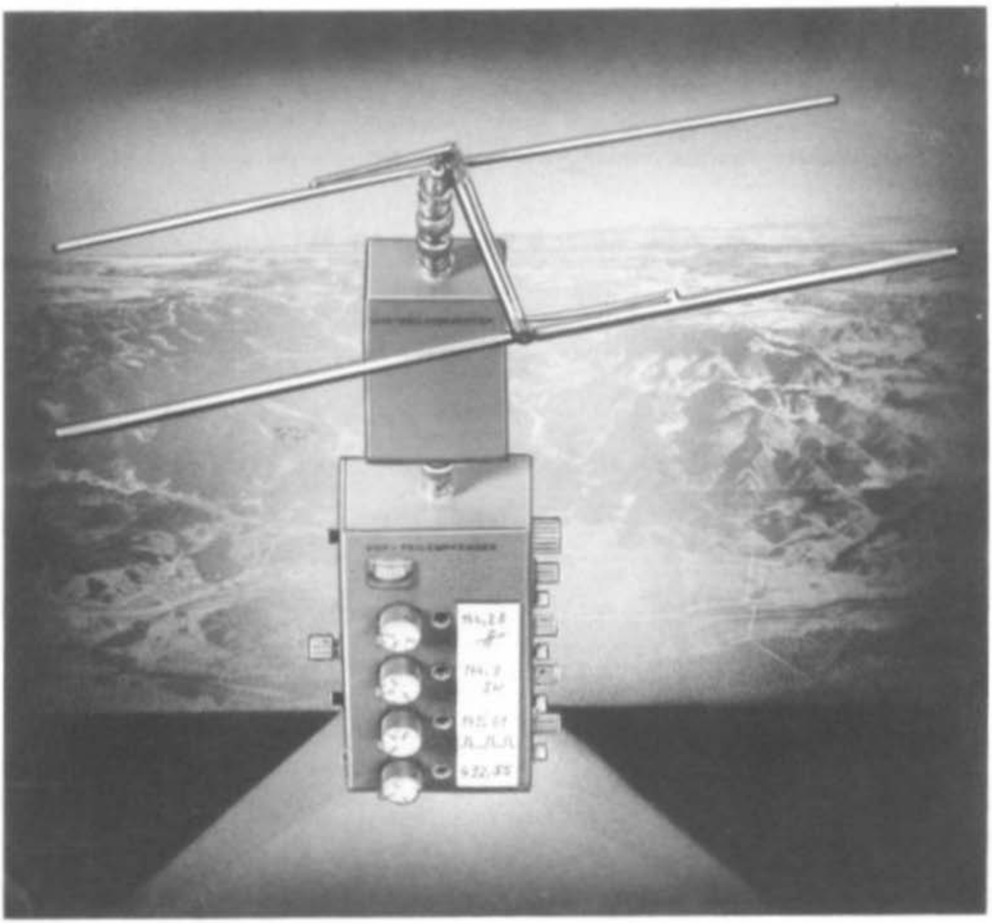

VHF COMMUNICATIONS

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The extremely large price increases of components throughout the world are having a great effect on the sales price of equipment and kits. Unfortunately, this is also true of our kits. The recent price increases of over 20% with crystals and crystal filters have forced us to also increase the sales prices of these items, which also increases the price of all kits that contain a crystal or crystal filter. Normally, we have a large stock of components so that we can "live from stock" for a short period. However, the deliveries of crystals from all over the world are so terrible that we are still waiting for crystals and crystal filters promised us for October 1973. This means that we are virtually out-of-stock and must immediately purchase for the increased prices. An insert is included in the magazine showing how the new kit prices can be calculated until we can send you a new price list together with edition 3/74 of VHF COMMUNICATIONS. Please remember the difficulties in obtaining the more specialized components should deliveries be taking longer than we would like.

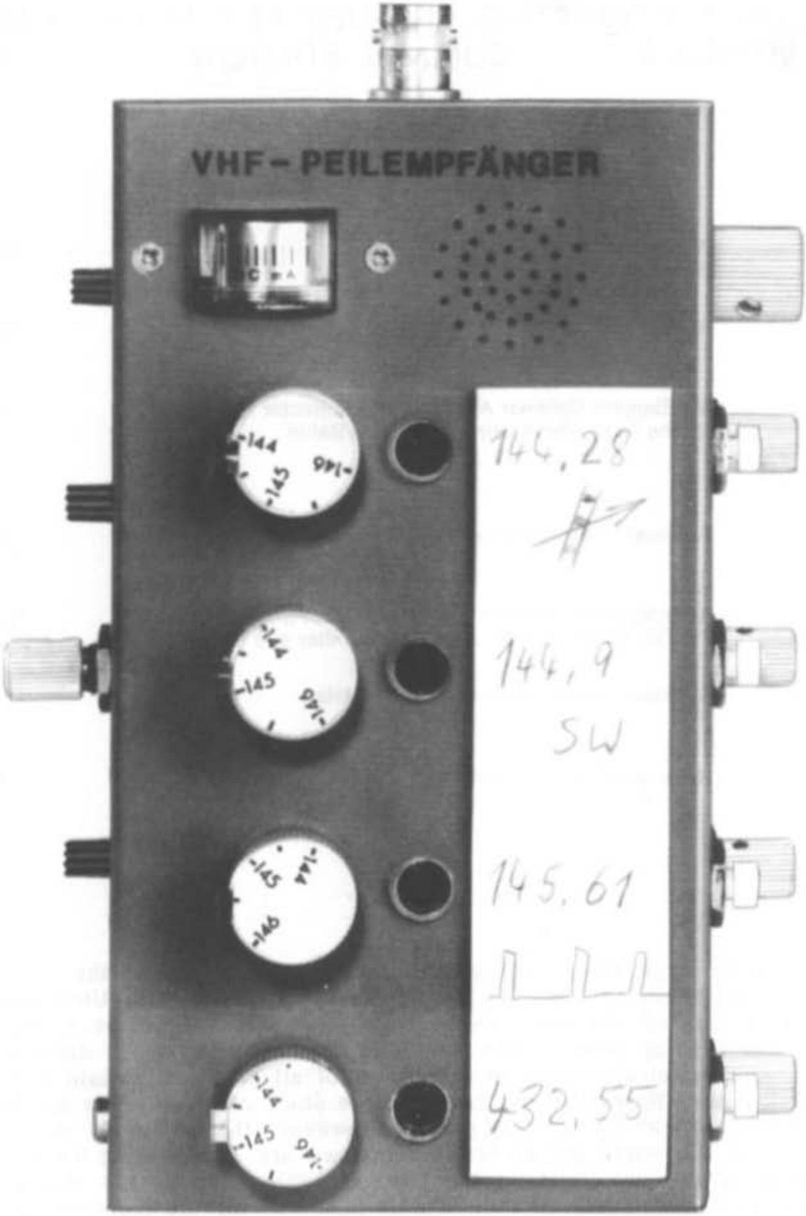


Fig. 1: Prototype of the programmable DF-receiver

A PROGRAMMABLE FOX-HUNT (DF) RECEIVER FOR 2 METRES

by G. Hoffschild, DL 9 FX

Fox hunts (DF-meetings) have become very popular in recent years. However, due to the non-availability of special equipment for this mode of operation, the person taking part has usually been hindered by equipment not really suitable for the purpose e.g. too large or too heavy or both. Furthermore, there are often more than one "fox" to be found which are operating on different frequencies. For this reason, it is necessary to keep changing frequencies and to tune the band to find the next frequency.

The following article is to describe a special DF-hunt receiver that has been designed specifically for this application. It is the result of a discussion between DJ 1 NB, DL 3 WR, DK 1 XZ and the author. It possesses the following special features: Programmable receiver tuning and separate gain controls for each frequency. The preprogrammed frequencies are set up in a similar manner to the station buttons of broadcast and television receivers. They are tuned in at the beginning of the fox hunt according to the sequence in which the foxes are to be found and it is possible to change over to each frequency at the touch of a button.Since each preprogrammed frequency possesses its own gain control, it is immediately possible to determine whether this transmitter has increased or decreased in strength. These characteristics are also available for fox-hunts in the 70 cm band when the matching 70 cm/2 m converter is used that was described in (1).

Of course, the receiver is not only suitable for fox-hunts, but can be used for any other receive applications, e.g. for monitoring certain fixed frequencies, or for monitoring the fieldstrength of various 2 m or 70 cm beacons.

A further special feature of this receiver is the "acoustic" S-meter which allows the field strength to be indicated as a change in frequency which saves having to keep looking at the built-in S-meter. This could also have applications for blind amateurs.

1. CHARACTERISTICS

Figure 1 shows the author's prototype, it is built into a case having the dimensions 75 mm x 140 mm x 40 mm. With built-in battery, the completed receiver weighs only 420 g. A 9 V battery is used.

Each of the four tuning elements can be tuned within the range of 144 MHz to 146 MHz. A common fine tuning (vernier) has been provided to ease tuning.

The receiver possesses a very good selectivity due to the use of a crystal filter comprising two crystals in the 9 MHz IF. A S-meter is provided to indicate the field strength. The receiver can be switched from automatic to manual gain control. Four potentiometers are provided for this that are switched electronically together with the four tuning potentiometers. It is advisable to program the frequencies of the "fox" transmitters at the beginning of the fox-hunt with the automatic gain control selected and to switch over to manual control during the fox-hunt in order to obtain a larger S-meter sensitivity.

A tone generator can be switched on during the reception of unmodulated carriers whose frequency is dependent on the field strength. This "acoustic" S-meter allows variations of the field strength to be determined without having to keep looking at the small S-meter on the receiver. Finally, it should be mentioned that the receiver possesses a small loudspeaker which will be switched off on placing the plug of an earphone into the appropriate connector.

2. CIRCUIT DETAILS

The receiver is split into two main circuits: The receive and switching modules. For this reason, the circuit is built up on two separate PC-boards. In the type of construction used, the case comprising the 10 potentiometers, 7 switches, loudspeaker, S-meter, and connectors forms a third group. Figure 2 shows a block diagram of the receiver and control elements. It should be mentioned that the positive pole of the 9 V battery is grounded to the case.

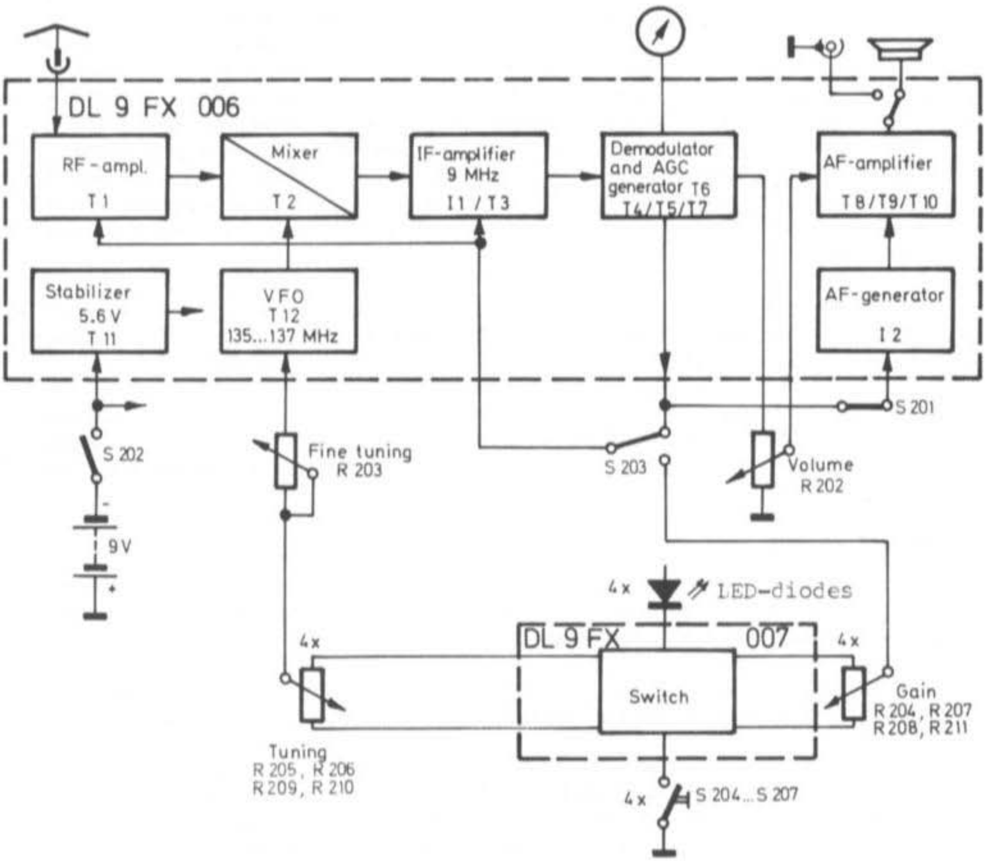


Fig. 2: Block diagram of the DF-receiver

2.1. RECEIVER MODULE

Figure 3 shows the circuit diagram of the complete programmable receiver. The upper half and the left part of the lower half of the circuit diagram are of the actual receiver module. The components of the receiver module are accommodated on PC-board DL 9 FX 006 and are designated by one or two digit numbers.

All components with numbers between 100 and 200 belong to the switching module and are accommodated on PC-board DL 9 FX 007. The components with numbers greater than 200 are mounted on the case.

2.1.1. PREAMPLIFIER, MIXER AND OSCILLATOR STAGES

The input signal is fed via the input circuit to the base of the preamplifier transistor T 1. A low-noise transistor with a high transit frequency is used in this stage. The AGC is also active on the preamplifier stage. Since this stage determines the noise figure of the receiver, it is possible to use a cheap IF transistor (BF 224) in the subsequent mixer stage (T 2). The oscillator equipped with the switching transistor 2 N 709 (T 12) oscillates 9 MHz below the required frequency. It is tuned with the aid of a varactor diode whose operating voltage is stabilized. The oscillator voltage is inductively coupled to the mixer stage due to the adjacent location of the two stages.

2.1.2. IF AMPLIFIER

The output signal from the mixer is passed via a bandpass filter tuned to the intermediate frequency of 9 MHz and fed to the first crystal bridge of the IF amplifier which uses the integrated circuit CA 3028. The second crystal filter is to be found in the collector circuit of this stage. After both crystals have been aligned correctly (R 7, R 8) the total bandwidth will be approximately 4 kHz with good ultimate selectivity. The relatively inexpensive carrier crystals used together with the KVG crystal filters are used in our application. The subsequent amplifier stage comprising transistor T 3 is also equipped with a transistor BF 224 and the amplified signal is then filtered and passed to the subsequent demodulator stage. The gain of both IF stages is controlled.

2.1.3. DEMODULATOR AND AF STAGES

In the demodulator circuit, transistor T 4 amplifies the injected voltage aperiodically and capacitively couples it to transistor T 14 that operates as a diode. The signal is then demodulated and the resulting AF voltage is passed back to transistor T 4 for amplification. The collector voltage of this stage is kept constant via transistor T 6 whose base is connected to the stabilized operating voltage. This guarantees that the S-meter reading is independent of the battery voltage and that the electrical zero of the meter is not altered.

The subsequent stage equipped with transistor T 5 amplifies the AF voltage. Since its collector current follows the mean value of the rectified voltage due to the galvanic coupling to the demodulator stage, it is possible for the indication of the field strength to be made at this point. The AGC voltage is also obtained here. It is fed via a voltage divider and a filter capacitor (C 18) to an impedance converter equipped with transistor T 7. This also controls the tone generator. The audio-frequency signal is taken from potentiometer R 202 and is fed to the driver stage (T 8) and then to the complementary output stage comprising T 9 and T 10. With a loading of 200 Ω, the output power is small and capacitance of 1 μF is sufficient for the coupling capacitor C 22.

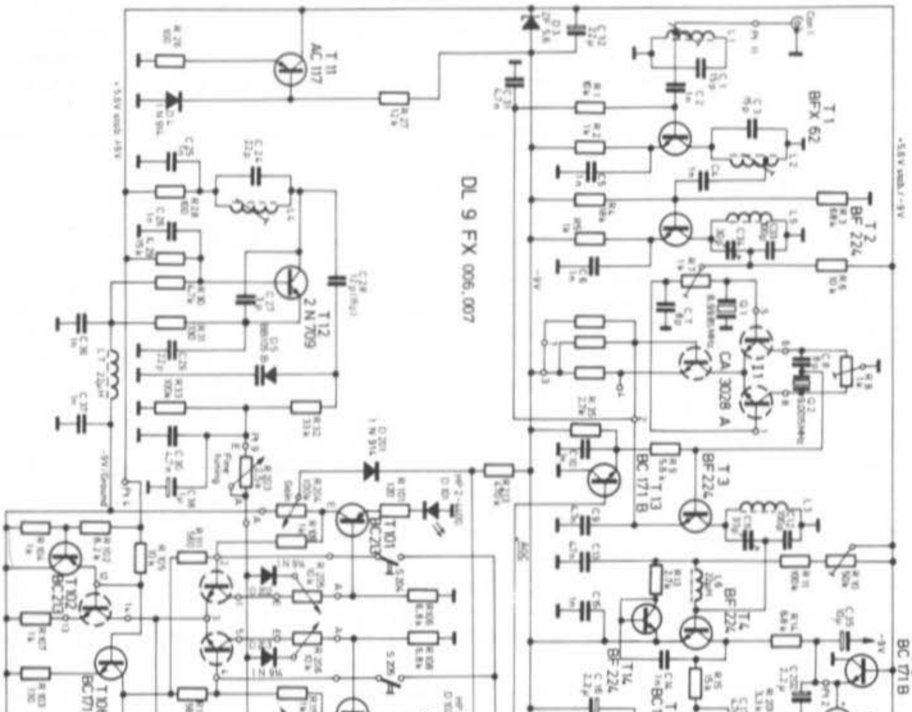


Fig. 3: Circuit diagram of the complete DF-receiver

2.1.4. TONE GENERATOR

A multivibrator is used as tone generator. Figure 4 shows the principle of the circuit and the tuning curve. It can easily be constructed using an integrated circuit. The frequency is determined by the time constant of the coupling capacitor C 19 and resistor R 21, as well as by the collector current. This is impressed into the actual multivibrator with the aid of transistor T 3. The value of this current can be controlled by the control voltage fed to the base. In the case of the given circuit, the tone frequency varies from approximately 600 Hz to approximately 1200 Hz at FSD of the S-meter. Diode D 2 in the collector circuit of T 2 ensures a current-dependent output voltage that is directly fed to the AF output stage.

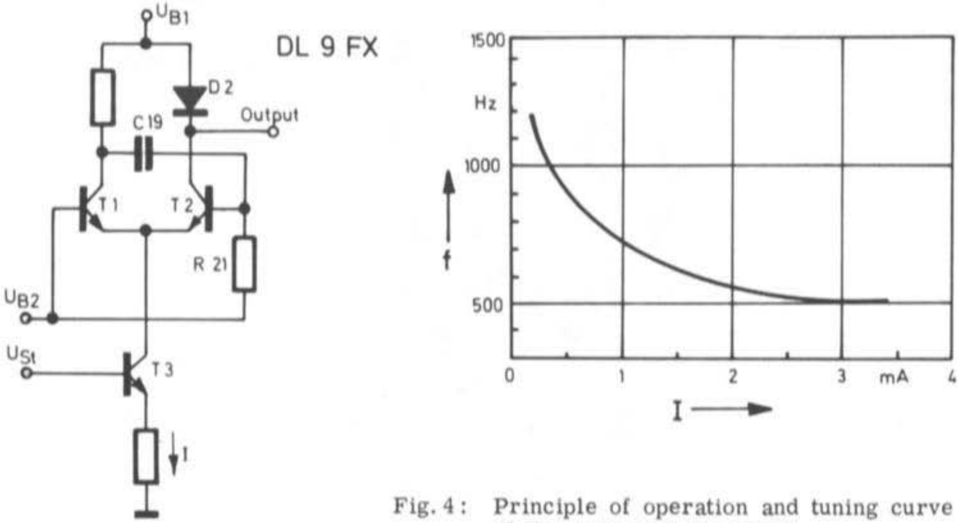
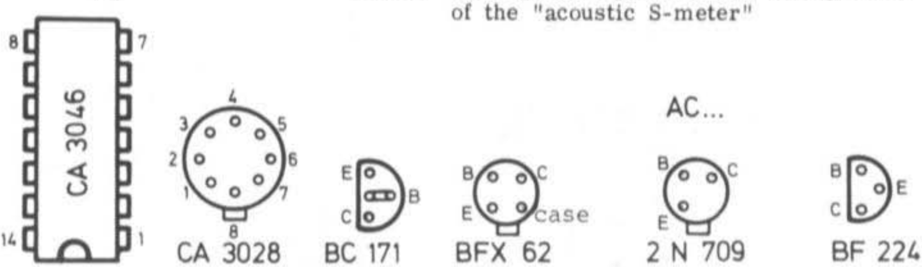


Fig. 4: Principle of operation and tuning curve of the "acoustic S-meter"



2.1.5. VOLTAGE STABILIZATION

Since the stability of a varactor-tuned oscillator is directly dependent on the stability of the tuning voltage, it is necessary for this voltage to be carefully stabilized (2). A zener diode is used for stabilization and its current is virtually impressed with the aid of a transistor (T 11) so that the zener voltage possesses a very low dependence on the battery voltage. Due to the constant load, the consumer also ensures that the stabilized voltage remains constant. The stabilized voltage amounts to + 5.6 V when referred to the minus pole of the battery.

2.2. SWITCHING MODULE

The switching of the tuning elements is made electronically. The use of a conventional TV or FM switching unit was avoided since the mechanical contacts used often provide uncertain contact resistances after a certain time which would have an adverse effect on the reproduceability of the frequency selection. Figure 5 gives the circuit diagram of two tuning units and includes measured values. The given voltage values (switched on, switched off) are referred to the minus pole of the battery and are measured with a meter having an impedance of 10 kΩ/V.

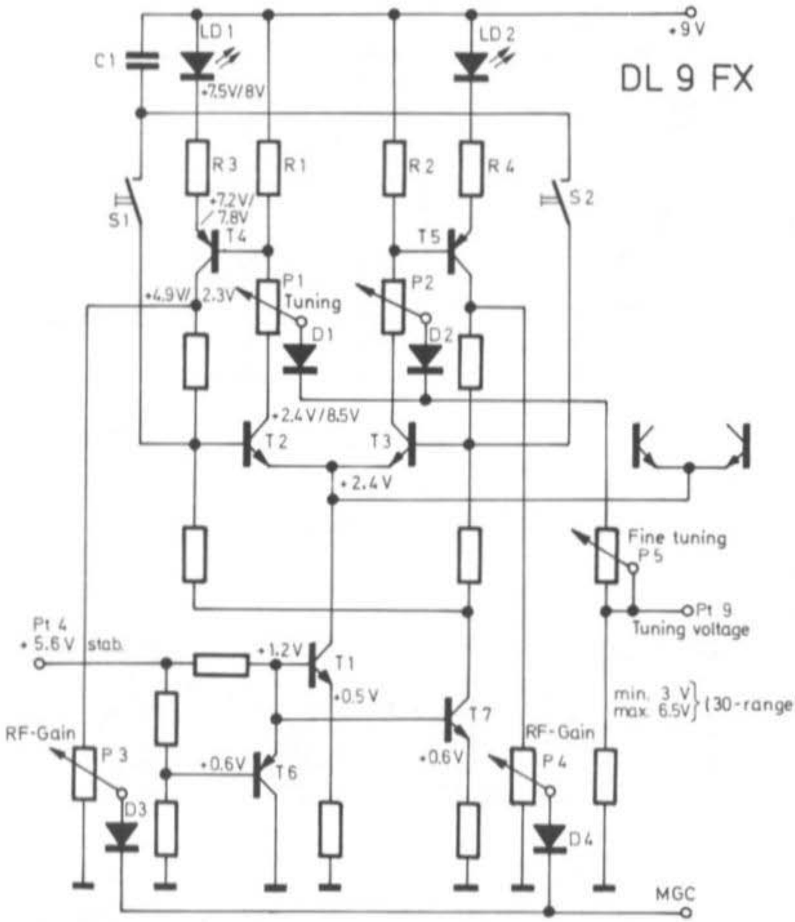


Fig. 5 : Principle of operation of two tuning controls with electronic switching

2.2.1. OPERATION

In the depressed position of switch S 1, transistor T 2 will possess a higher base current until capacitor C 1 is charged. Transistor T 2 will then take over the current impressed by T 1. This current then flows via the tuning potentiometer P 1 and resistor R 1. The voltage drop across R 1 will open transistor T 4 and cause the indicator diode LD 1 to light. The forward voltage of the indicator diode together with resistor R 3 determines the current flowing via transistor T 4. In addition to this, the switching condition of T 2 remains constant since the base current for T 2 is supplied by T 4 after releasing pushbutton S 1. The main portion of the current supplied by transistor T 4 flows, however, via T 7 and a smaller component is fed via P 3 in order to generate the voltage for the manual gain control. The current via transistor T 7 is also impressed and ensures that the base current of T 2 remains constant even when the battery voltage begins to fall. Otherwise, differing collector currents of transistor T 2 would be exhibited which would have an effect on the tuning voltage and thus cause a deterioration of the frequency stability.

Transistor T 6 compensates the temperature coefficient of the base-emitter voltage of T 1. The decoupling diodes D 1, D 2, D 3 and D 4 ensure that the active tuning controls are not affected by the inactive ones. Furthermore, the temperature coefficient of the varactor diode is compensated for with the aid of D 1 and D 2. The tuning voltage can be varied with potentiometer P 5 in order to provide a fine tuning of the receiver.

If pushbutton switch S 2 is depressed, T 3 will receive a base-bias current. The current supplied from T 1 will then flow via T 3 and the switching circuit will be actuated via T 5. This means that the circuit behaves similar to a bistable multivibrator circuit, however, the number of switchable stages can be extended as required. The current requirements amount to only approximately 2.5 to 3 mA due to the measures taken during the design of the circuit.

The circuit is equipped with the integrated circuit CA 3046 (RCA). It contains transistors T 1, T 2, T 3 as well as the transistors corresponding to T 2 and T 3 of the other tuning elements. The inexpensive type CA 3086 can also be used whose pins are directly compatible. This means that they can be directly exchanged.

3. COMPONENT DETAILS

Con 201: BNC-connector (UG 1094)
Con 202: Sub-miniature connector with change-over contact

Ceramic tubular capacitor, if possible NPO (black), spacing 5 mm:

C 7, C 8: 8 pF	C 1, C 3: 15 pF	C 24: 22 pF, between N 075
C 28: 12 pF	C 29: 22 pF	and N 220 (yellow)

Ceramic disc capacitors, spacing 5 mm:

C 2, C 4, C 5, C 6, C 9, C 10, C 14, C 15, C 25, C 26, C 36, C 37: 1 nF;
C 13, C 17, C 30, C 31: 4.7 nF
C 27: 3 pF N030 (brown) to N075 (red)

Styroflex capacitors, spacing 10 mm (Siemens B 31 310): C 12,
C 33: 100 pF;

Plastic-foil capacitors, spacing 10 mm: C 19: 0.1 μF

Tantalium electrolytic, drop-type, spacing 2.5 mm
C 18, C 20, C 22, C 38: 1 μF
C 16, C 21, C 202: 2.2 μF

Tantalium or aluminium electrolytics:
C 203: 10 μF C 201: 47 μF C 23: 100 μF C 204: 220 μF

Plastic-foil trimmers 7 mm dia.:
C 11, C 34: 2 - 22 pF (green) or 6 - 65 pF

D 2, D 4, D 201 - D 208: 1 N 4148 or 1 N 914 or similar silicon diode
D 3: BZY 85/C5V6, BZX 55/C5V6, ZF 5.6 or similar 5.6 V zener diode
D 5: BB 105 B (Siemens) 11.5 pF/5 V
D 101 - D 104: HP 5082 - 4400 (Hewlett-Packard) or similar LED's

I 1, I 2: CA 3028 A (RCA)
I 101: CA 3046 or CA 3086 (RCA)

Meter: 1 mA, approx. 100 Ω

Loudspeaker: Miniature dynamic loudspeaker, 200 Ω

L 1: 3.3 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire, coil tap 0.5 turns from the cold end, cold end facing upwards, coilformer 4.3 mm dia., shortened to 17 mm, pink core, coil length 6 mm
L 2: 3.5 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire, coil tap 0.25 turns from the cold end, cold end above, coilformer and core as for L 1, coil length 10 mm
L 3, L 5, L 6, L 7: Miniature ferrite choke 18 μH to 22 μH
L 4: 2.75 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire, coil length 6 mm, coilformer as for L 1, brown core

Q 1: Crystal XF 901 (KVG)
Q 2: Crystal XF 902 (KVG)

Standard carbon resistors for 10 mm spacing

56 Ω: 1 piece	3.3 kΩ: 1 piece	18 kΩ: 2 pieces
100 Ω: 2 pieces	4.7 kΩ: 1 piece	22 kΩ: 1 piece
120 Ω: 4 pieces	5.6 kΩ: 2 pieces	33 kΩ: 1 piece
330 Ω: 2 pieces	6.8 kΩ: 5 pieces	47 kΩ: 1 piece
560 Ω: 4 pieces	8.2 kΩ: 2 pieces	68 kΩ: 2 pieces
820 Ω: 1 piece	10 kΩ: 4 pieces	100 kΩ: 2 pieces
1 kΩ: 9 pieces	12 kΩ: 1 piece	470 kΩ: 1 piece
2.7 kΩ: 3 pieces	15 kΩ: 2 pieces	

Trimmer resistors for vertical, PC-board mounting, spacing 5 mm/2.5 mm:

1 kΩ: 2 pieces 50 kΩ: 1 piece

Potentiometers without switch, 16 mm dia. 4 mm shaft:

5 kΩ log. (R 202): 1 piece 2.5 kΩ lin (R 203): 1 piece
10 kΩ lin: 4 pieces 100 kΩ lin: 4 pieces

S 201 - S 203: Miniature sliding switches, one change-over contact
S 204 - S 207: Miniature push-button switches with one make contact

Case: 140 mm x 75 mm x 40 mm
Battery connector
10 knobs for 4 mm shafts

T 1: BFX 62 (Siemens) or BF 200 (Philips)
T 2 - T 4, T 14: BF 224 (TI) or BF 173, BF 311
T 5 - T 8: BC 108 B, BC 171 B or similar NPN-AF transistor with Bmin. 100, e.g. BC 415
T 9, T 11: AC 117 or similar germanium PNP transistor (AC 122, AC 131, AC 150, AC 170, AC 188)
T 10: AC 175 or AC 187 or similar germanium NPN transistor
T 12: 2 N 709 (TI) or similar fast switching transistor
T 13: BC 108 B, BC 171 B or similar silicon AF transistor with B min. 100, e.g. BC 415 (Siemens)
T 101 - T 105: BC 213 (TI) or similar silicon PNP transistor, e.g. BC 413 (Siemens)
T 106: BC 108 B, BC 171 B or similar, e.g. BC 415 (Siemens).

4. CONSTRUCTION

4.1. ELECTRICAL CONSTRUCTION

As previously mentioned, the main part of the circuit is accommodated on two PC-boards: The receiver module DL 9 FX 006 and the switching module DL 9 FX 007.

4.1.1. RECEIVER MODULE DL 9 FX 006.

The PC-board and component locations are given in Figure 6. The component side of the board possesses a continuous ground surface. The designations for the connections on the edge of the board are given to ease the interconnection of the PC-boards to another and to the case. They are also given in the corresponding positions of the circuit diagram.

Nearly all holes on the PC-board are made with the aid of a 0.7 mm diameter drill; the holes for the two crystals, trimmers, trimmer potentiometer's and wire connections are made with the aid of a 1.0 mm drill. The ground surface around all holes that are not directly grounded (indicated as rings on the component plan and not as black dots), should be removed with the aid of a 2.5 mm drill so that no short circuits occur. The three holes for the mounting screws on the centre line are 3.0 mm in diameter.

4.1.2. SWITCHING MODULE DL 9 FX 007

The PC-board and component locations of this module are given in Figure 7. The PC-board has the dimensions 100 mm x 27 mm. This board is single-coated. The four light emitting diodes (LED's) are soldered into place approximately 3 mm from the component side of the board. All holes for the components should be drilled with a 0.7 mm drill, whereas the holes for the connection wires should have a diameter of 1.0 mm.

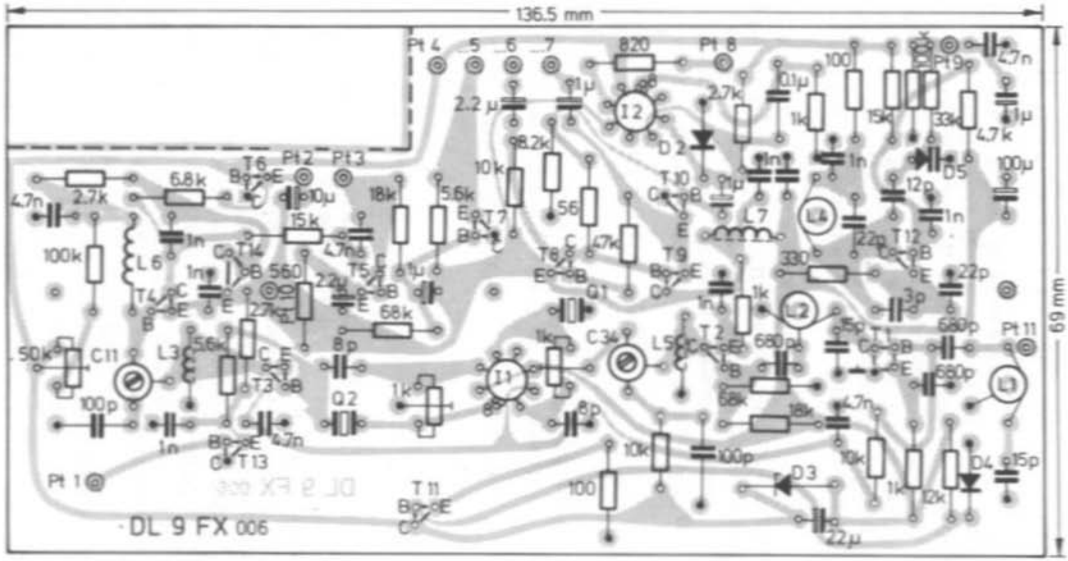


Fig. 6: Receiver board DL 9 FX 006
(Component side: continuous ground surface)

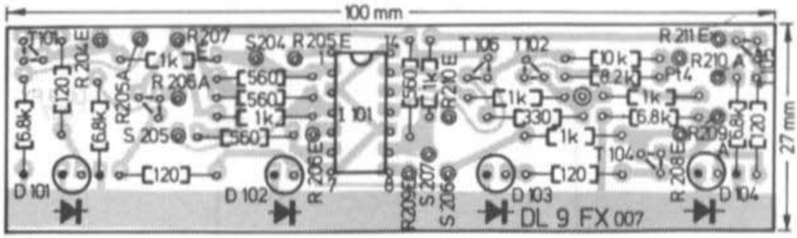


Fig. 7: Switching board DL 9 FX 007

4.2. MECHANICAL CONSTRUCTION

The receiver is accommodated in a case having the dimensions 140 mm x 75 mm x 40 mm. The position and dimensions of the holes and cut-out in the U-shaped case as well as the hole for the BNC connector Con 1 can be seen in Figure 8. The holes for the mounting of board DL 9 FX 006 possess a diameter of 2.2 mm. It is advisable for the drilled board to be placed into the bottom of the case before it is equipped. The mounting holes can then easily be marked. Self-taping screws are normally used for joining the base and cover. However, the author cut a 2 mm thread into the appropriate holes and used normal screws.

4.3. INSTALLATION AND INTERCONNECTION

The photographs of the author's prototype given in Figure 9 and Figure 10 allow further details regarding the construction to be seen.

4.3.1. COVER OF THE CASE

The installation is commenced with the mounting of the components accommodated on the cover of the case. The four push-button switches (S 204 - S 207), the RF gain potentiometers (R 204, R 207, R 208, R 211) and the fine-tuning potentiometer R 203 can be provided with washers or counter-nuts so that their threads are level with the mounting nuts. The connection tags of the manual gain potentiometers point in an upward direction towards the BNC connector, and the fine tuning potentiometer tags towards the front panel. The AF potentiometer R 202 on the opposite side can only be set back slightly, which is also valid for the four tuning potentiometers R 205, R 206, R 209 and R 210. The connection tags of AF potentiometer point towards the base of the case and those of the tuning potentiometers all face downwards with the exception of the lower potentiometer whose connections face upwards. This is followed by installing the switching board. The ground connection of the board is now connected to the three ground tags of the tuning potentiometers and soldered into place after the board has been positioned so that the indicator diodes are directly under their appropriate holes.

The loudspeaker is glued into place with the aid of a spacing ring (22 mm outer diameter, approximately 1.5 mm thick) after all components have been mounted, it is possible for the interconnection of the case to be commenced. It has been found that direct connection is more favourable than to bring all the wires into a common cable harness.

All resistors, electrolytic capacitors and diodes (with numbers above 200) that are not mounted on the PC-boards are soldered to the appropriate positions where they are self-supporting.

DL 9 FX

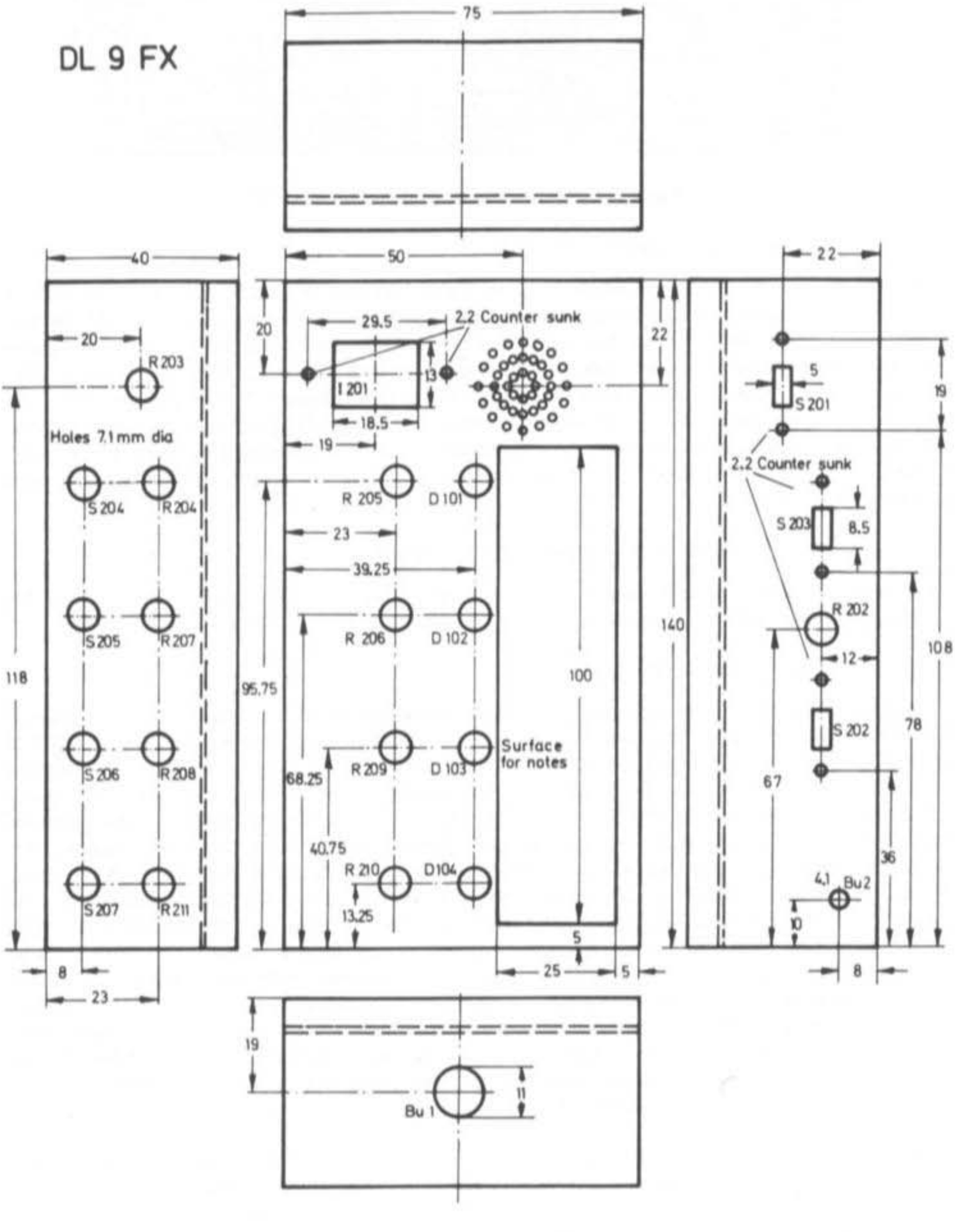


Fig. 8 : Case with all required holes

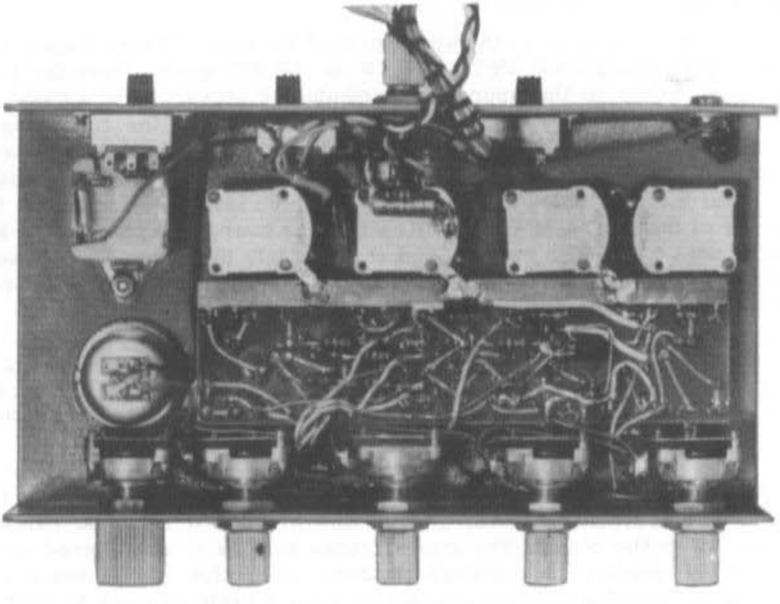


Fig. 9:
Original prototype of the DF-receiver. Cover part of the case with potentiometers, switches, loudspeaker, S-meter, and switching board DL 9 FX 007 (no components are located on the conductor lane side with the published boards)

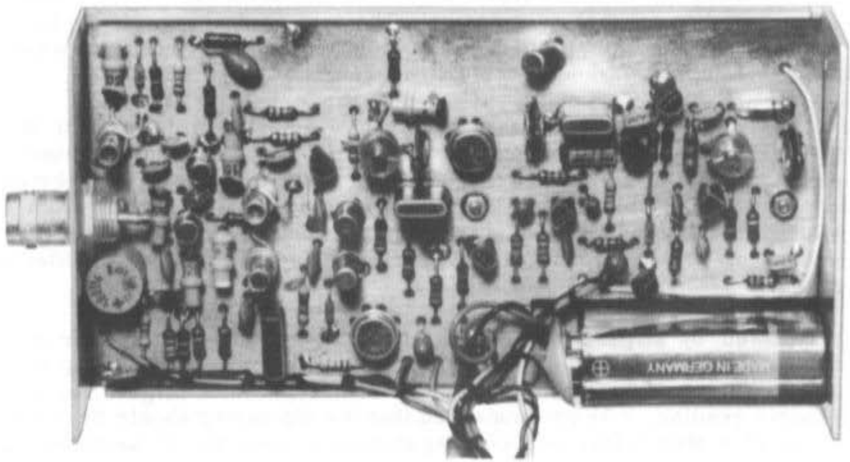


Fig. 10:
Receiver board DL 9 FX 006 in the base of the case. The three germanium transistors can be installed without their heatsinks

4.3.2. BASE OF THE CASE

Mount the BNC connector to the upper part of the base. Wires should be placed into the holes designated Pt 1 to Pt 10 on the PC-board from the component side and soldered to the appropriate conductor lanes. It is advisable to use insulated wire since the cable connection to the cover of the case must remain flexible. The interconnection should not be longer than absolutely necessary so that the wires do not lay around on the PC-board. An uninsulated piece of wire is inserted into the hole designated Pt 11 which is then soldered to the inner conductor of the BNC socket after installing the board into place. The PC-board is placed onto 4 mm high spacers and screwed into the base of the case. Spring washers should be used under the nuts to ensure that it is held firmly into place even under the most rugged conditions.

After mounting the board, all the wires except those from ground, Pt 3 and Pt 6 are formed into a cable harness and fed to the cover of the case just before the battery cut-out. The wires then branch off from here to the appropriate connection points.

In order to avoid a feedback between the AF lines, the connections from Pt 3 and Pt 6 are combined with the ground line to form a separate cable harness and are fed to the cover. The ground connection is then soldered to the case of the potentiometer. The positive battery connection is connected to the receiver board, and the negative connection via the cable harness to switch S 202.

5. ALIGNMENT

After carefully checking the boards and the wiring, the battery is connected via a mA-meter. The RF and AF potentiometers should be in their fully anti-clockwise position, the tone-generator should be switched off (S 201) and the receiver switched to manual gain-control. The current drain should amount to 8 - 10 mA at a battery voltage of 9 V. If one of the push-buttons is depressed, the appropriate indicator diode should light up.

The electrical zero point of the S-meter is now adjusted with resistor R 10. The adjustment is correct when the meter needle just starts to lift from the mechanical zero. If it is not possible to align R 10 to achieve this position, this will mean that the current gain of transistor T 4 deviates greatly from the mean value. If the pointer remains at the fully left-hand position, it is necessary for the value of R 11 to be increased, and decreased if full-scale deflection is indicated.

This is followed by aligning the IF amplifier. The gain-potentiometer of the selected range is brought approximately to a central position and a dip-meter is loosely coupled to L 5. Capacitors C 34 and C 11 are now aligned for maximum S-meter reading. It is recommended that the dip-meter should be allowed to warm up at 9 MHz before commencing alignment since the IF amplifier only possesses a bandwidth of approximately 4 kHz. The RF gain should be reduced in steps during the alignment process, or the coupling of the dip-meter to L 5 can be reduced. If the dip-meter can be modulated, it is possible for the alignment to be made by ear.

The two IF bridge circuits are then aligned in any order. The dip-meter is detuned so that it is 100 kHz too low and trimmer potentiometers R 7 and R 8 should be aligned for minimum reading on the S-meter. It will be necessary to increase the coupling to L 5 during alignment, or to increase the RF gain.

The alignment of the oscillator is made with L 4 at the lowest frequency of the tuning potentiometer (135 MHz). If the full tuning range of 2 MHz cannot be achieved, it will be necessary for C 28 to be increased from 12 pF to 15 pF.

A weak 144 MHz signal is now fed to the input and L 2 and subsequently L 1 aligned for maximum reading.

After completing this alignment, the other three tuning elements are checked as well as the fine tuning, automatic gain control and the tone generator. The frequency of the latter should increase on increasing the field strength. It is then possible for the tuning knobs to be calibrated. This can be made with the aid of a calibration spectrum generator such as (3).

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ANTENNA NOTEBOOK

by T. Bittan, DJ Ø BQ/G 3 JVQ

It is wellknown that a good antenna is by far the best RF-amplifier, especially since it is effective both in the transmit and receive mode. The gain of an antenna or antenna array is dependent on the aperture or capture area. It is possible to calculate the gain of the antenna from its horizontal and vertical beamwidth. This is shown in the nomograph (1) shown in Figure 1. However, this nomograph has been calculated assuming no losses in matching etc. so that the actual gain figure provided by an antenna will be slightly less.

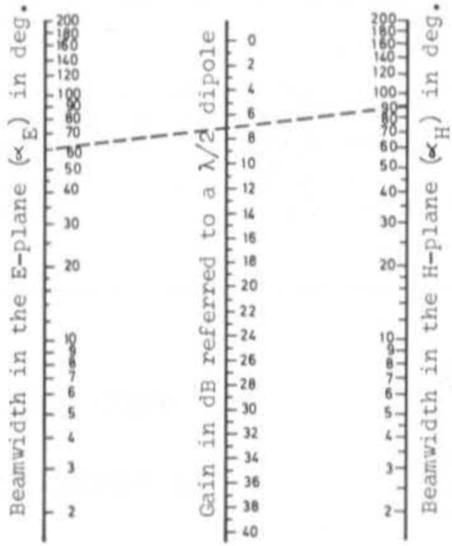

Fig. 1:
Nomogram for determining the gain of an antenna from its vertical and horizontal beamwidths

In the case of a Yagi antenna, the gain is mainly dependent on its length. However, this is not as true as it has been in the past since it is now possible for optimum dimensions and element spacings to be calculated using computers. This has allowed considerable reductions of the overall dimensions of Yagi antennas without loss of gain.

Another way of increasing the gain of antenna is to stack several identical antennas in either the vertical (stacking) or horizontal (baying) plane, thus reducing the beamwidth in that plane. In theory, it is possible to increase the gain of a single antenna by the full 3 dB when the correct spacing is provided. Unfortunately, a certain amount of sidelobes are generated when two antennas are stacked so that it is necessary to make a compromise either with respect to gain in order to obtain minimum sidelobes or to accept a certain amount of sidelobes in order to achieve maximum gain. According to (2) the maximum gain is achieved with sidelobes of -10 dB, i.e. when the sidelobes are only 10 dB down on the main lobes. Since the gain of an antenna is of greater importance in amateur applications than the directivity of the array and since virtually no gain will be exhibited when two antennas are stacked for minimum sidelobes, information is now to be given which allows the stacking distance to be read off from a graph for any known beamwidth in that plane. Figure 2 gives the stacking distance for two identical arrays and Figure 3 gives a similar graph for application when four identical antennas are stacked in one plane.

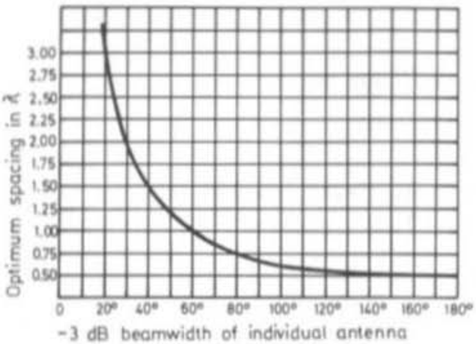

Fig. 2:
Stacking distance as a function of the beamwidth in the stacking plane for two identical antennas

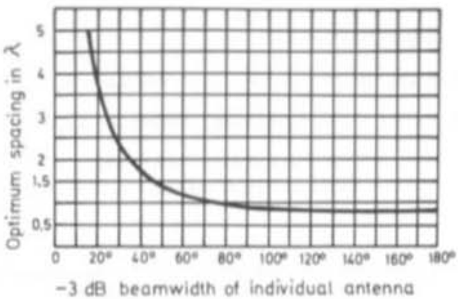

Fig. 3:
Stacking distance as a function of the beamwidth in the stacking plane for four identical antennas

1. MATCHING

Assuming that the two individual antennas are designed for low-impedance feeders, the easiest manner of interconnecting two stacked antennas to a common feedline is for each antenna to be fed with an equal length of coaxial feeder (multiple of an electrical half-wave). These two feeders are then connected in parallel and the resulting impedance of Z x 0.5 increased to the required value in a quarter-wave transformer. The impedance of this quarter-wave transformer can be calculated as given in (3) using the formula:

$$Z_1 = \sqrt{Z_a \times Z_f}$$

Where Z_1 is the impedance of the $\lambda/4$ line; Z_a is the characteristic impedance of the parallel-connected antennas, and Z_f is the characteristic impedance of the feeder.

In the case of four antennas and more, it has been found that it is more favourable to interconnect these in groups of two antennas and to interconnect each of these groups in a similar manner. The main difficulty is in obtaining the correct impedance for the $\lambda/4$ transformer. However, most manufacturers also supply the required stacking cables. It is, of course, possible to construct any required impedance using metal tubing.

It is also possible to increase the impedance in a $\lambda/4$ transformer before connecting the two feeders in parallel. In the case of two 50 Ω feeders, the impedance of each individual feeder can be transformed to 100 Ω in a $\lambda/4$ transformer section of 70 Ω impedance; after parallel connection, the resulting characteristic impedance is once again in the order of 50 Ω. The advantage of this method is that 70 Ω or 75 Ω coaxial cable is readily available.

This article has described how antennas can be stacked and bayed in order to achieve maximum gain. A later article will describe one of the easiest ways of stacking two antennas in the form of a Skeleton slot array. Further information will also be given on how antennas can be stacked and bayed in order to achieve maximum directivity as well as an infinite front-to-back-ratio.

2. REFERENCES

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(3) T. Bittan: Circular Polarization on 2 Metres
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J-BEAM SLOT-FED GROUP ANTENNAS

Colinear antennas are very popular as DX-antennas due to their wide horizontal and narrow vertical beamwidths which concentrate the transmitted energy at the horizon and allow a larger area to be covered in the horizontal plane. However, such colinear groups possess several disadvantages: The mechanical stability is limited by the long, fullwave (λ) elements; the most favourable stacking distance cannot be achieved with the type of feed used; the height and mast height requirements are often too great for many locations. It should also be noted that the max. gain of a 20-element colinear group antenna is in the order of 11.5 to 12 dB referred to a dipole and not 16 to 17 dB as often given.

The J-BEAM slot-fed groups represent a real alternative to colinear groups since they possess all the advantages of the latter as well as the following:

- Better mechanical stability due to the shorter ($\lambda/2$) elements. Seawater-proof, low corrosive aluminium is used throughout with exception of the strong, zinc-plated mast clamps.
- Lower height and mast requirements for a given gain: Max. 116 cm for individual group.
- Various types available from 10 to 16 elements for each individual group, which may be extended to form larger groups.

Technical data of individual groups:

Type	H. Beamwidth (-3 dB)	Dimensions	Gain ref. dipcle
D5/2 m	52°	103x161x116	10.8 dB
D8/2 m	45°	103x279x116	12.6 dB
D8/70 cm	45°	34x106x 43	12.6 dB

All slot-fed group antennas possess a very wide bandwidth that by far exceeds the amateur bands.

SIX-ELEMENT COLINEAR WITH REFLECTOR SHEET FOR THE 24 cm BAND USING A STRIPLINE BALUN

by M. Münich, DJ 1 CR and B. Lübbe, DJ 5 XA

The authors demonstrated the above antenna at the German VHF Convention in Weinheim. The described antenna was provided with a new-type of balun transformer using the printed stripline technology. The construction of this antenna is now to be described.

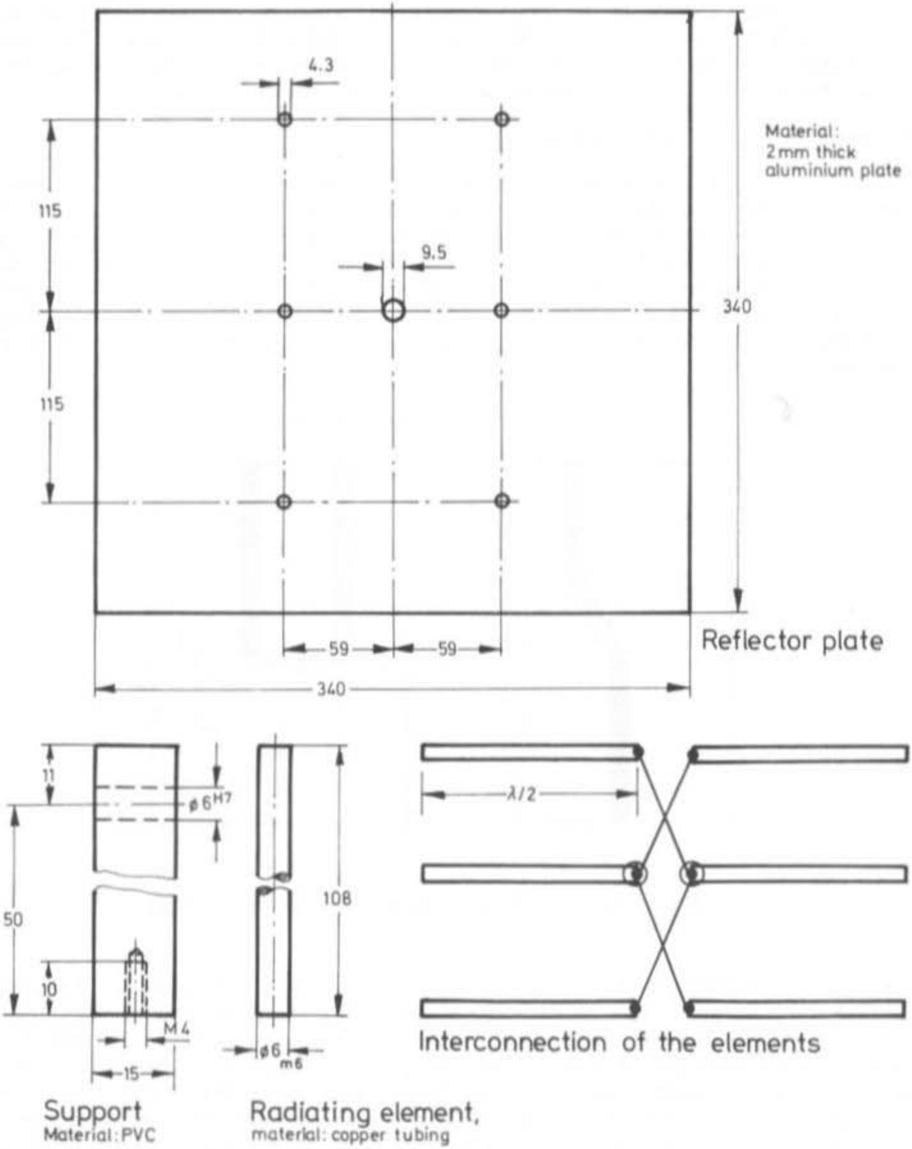

Fig. 1: Construction of a 6-element colinear antenna with reflector panel for the 24 cm band

1. SIX-ELEMENT COLINEAR ANTENNA WITH REFLECTOR SHEET

The antenna comprises six $\lambda/2$ dipoles mounted before a common reflector plate. The construction, dimensions and material used for the antenna are given in Figure 1. The three pairs of dipoles are end-fed at high impedance and are crosswise connected with the aid of 2 mm diameter (12 AWG) silver-plated copper wire as shown in Figure 1. Since the impedance of each individual λ-dipole amounts to approximately 600 - 700 Ω with this ratio of element thickness to length, an impedance of approximately 200 to 240 Ω will be present after interconnecting all three pairs of dipoles. The balanced 200 - 240 Ω is then transformed to unbalanced 50 - 60 Ω in a 4 : 1 balun transformer. The balun transformer is connected to the centre pair of dipoles and can be built up from thin, coaxial cable or in stripline technology as described here.

2. STRIPLINE BALUN TRANSFORMER

Normally, a balun transformer consists of an electrical $\lambda/2$ length of coaxial cable. Since such a cable for 24 cm would only be 8 - 9 cm in length according to the dielectric used, difficulties can be encountered with conventional cables since it is not possible to bend them in a loop. This means that very thin cables should be used, and only cables with teflon (PTFE) dielectric provide sufficient stability and low-attenuation characteristics. In order to avoid these difficulties, DJ 5 XA developed the described stripline balun for use with the antenna.

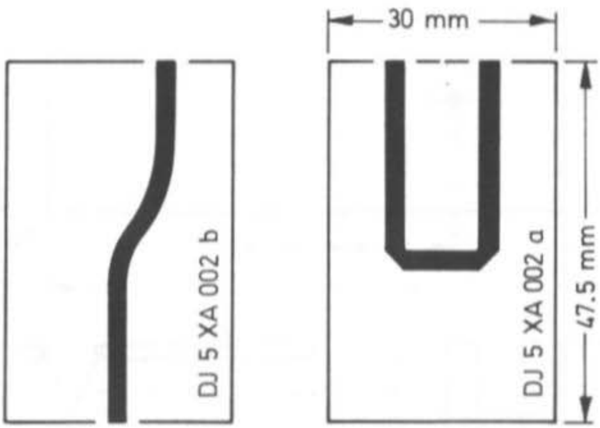

Fig. 2: PC-boards of a stripline balun for 24 cm

The two PC-boards of the stripline balun are shown in Figure 2. The first board possesses a U-shaped $\lambda/2$ 50 Ω stripline and the other board consists in principle of a 50 Ω line of uncritical length. The latter is bent so that it is directly adjacent to one arm of the U. The ground surfaces at the back of both boards are glued together and the connection to the centre dipoles is made with short pieces of 2.5 mm wide copper foil strip. The other end of the balun is directly soldered do a BNC-connector mounted in the reflector plate of the antenna.

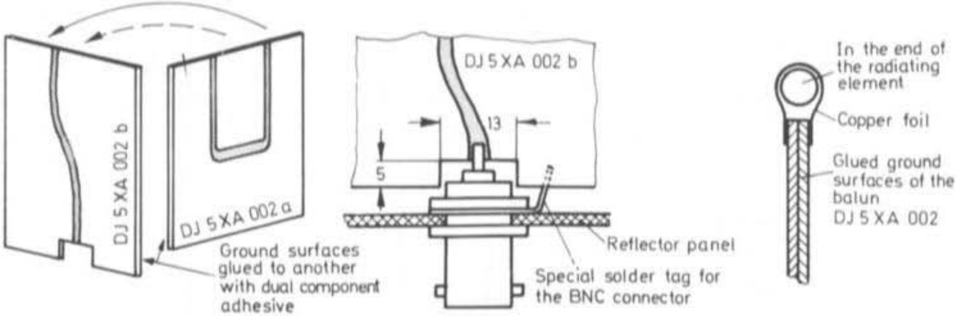

Fig. 3 : Construction of the stripline balun and connections

The striplines shown in Figure 2 with their width of 2.5 mm are designed for use with an epoxy glass fibre material with $\varepsilon_r = 5$ and a thickness of 1.5 mm. If another material, e.g. PTFE (teflon) is to be used, it is necessary for the striplines to be altered to suit the different dielectric constants (1), (2). The described stripline balun can, of course, be used for other 24 cm antennas. It is also possible for the U-shaped loop to be made narrower, for instance, when used together with yagi antennas. Figure 4 shows how a coaxial cable should be connected low-reflectively when it is not possible for a connector to be mounted suitably at the correct spacing from the radiating element.

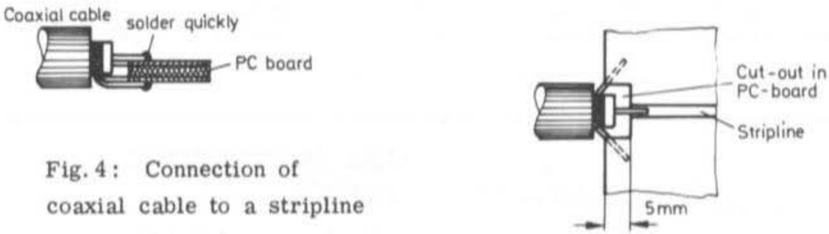

Fig. 4 : Connection of coaxial cable to a stripline

3. MEASURED VALUES AND POSSIBLE EXTENSIONS

According to measurements made by DJ 1 CR, the described antenna possesses a gain of approximately 10 dB over a reference dipole and a horizontal beamwidth (-3 dB) of 60°. These values coincide to those of similar antennas described in (3). The reflection factor of the antenna when using the stripline balun was measured by DJ 9 XN. Figure 5 shows the measured result. Since the bandwidth of the antenna is greater than that of the 24 cm amateur band, it is extremely suitable for ATV transmissions.

Two or more such antennas could be combined to increase the total gain. In such cases, the reflector sheets should form a common surface. The dimensions of the reflector sheets have been designed so that suitable spacings between the radiating elements result. The interconnection of two individual antennas should be made with the aid of $\lambda/4$ transformer sections with an impedance of $Z = 75\ \Omega$. Since $\lambda/4$ is too short mechanically, they may be extended by any desired multiple of $\lambda/2$. Of course, it is necessary for the velocity factor of the cable to be taken into consideration, e.g. RG-11 A ($75\ \Omega$) velocity factor 0.66.

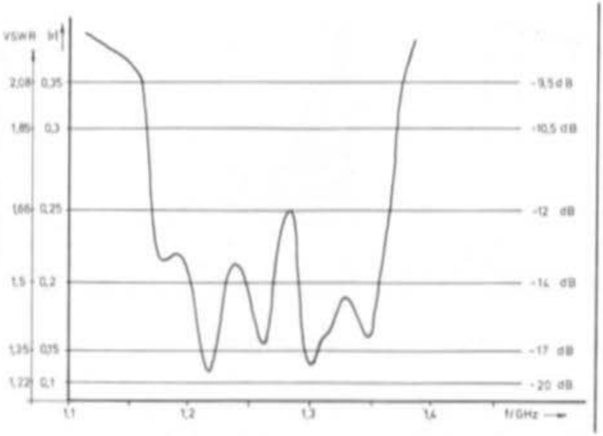

Fig. 5 : Frequency dependence of the reflection factor of the 24 cm antenna with stripline balun

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A LINEAR TRANSVERTER FOR 2 m / 70 cm WITH DOUBLE CONVERSION

by W. Rabe, DC 8 NR

1. INTRODUCTION

Two main methods are used at present in order to provide linear (SSB) operation on the 70 cm band. The 70 cm signal is either obtained by conversion of a 10 m signal as described in (1), or from the 2 m band as was used in (2). Whereas the relatively low frequency spacing of the local oscillator frequency from that of the required signal has disadvantages, which are even worse for the other shortwave bands, the second principle has the disadvantage that a tripling of the 2 m signal cannot be completely avoided. The degree of suppression of this unwanted multiplication will be even reduced on increasing the power level of the frequency conversion in order to save several stages in the amplifier chain. Even the selection of a shifted local oscillator frequency as was used in (2) is not of great assistance since the unwanted harmonics will only have a spacing of 500 kHz from the required signal under the most unfavourable conditions. This means that the circuits of the subsequent linear amplifier will not be able to suppress it. This shows that such a frequency plan is not completely suitable for a home station when high demands are to be placed on the freedom of spurious transmissions. Unfortunately, the method of elimination and restoration of the envelope as described in (3) has not gained any real popularity in amateur circles. Furthermore, the construction of a separate station for 70 cm with a suitable intermediate frequency of approximately 50 to 70 MHz would be extensive technically and would be extremely costly. It is true that crystal filters are available for such high frequencies and are not too expensive. However, they exhibit bandwidths that are still far too wide for the SSB mode which means that 9 MHz or 10.7 MHz crystal filters must still be used in the generation of the SSB signal. The use of a single-conversion principle for 70 cm is therefore hardly to be realized at present at a reasonable cost and technical complement.

2. CONCEPT

The concept used in the linear transverter to be described virtually provides a compromise between the two methods. It also uses the 2 m band for provision of the 70 cm signal, which is an advantage for those VHF-UHF amateurs that do not possess a shortwave transceiver.

In order to achieve a sufficient suppression of the tripled 2 m signal, a double conversion principle is used. The frequency plan and the various stages of the transverter are shown in Figure 1. Since all auxiliary frequencies are obtained from a master oscillator, complete tracking is guaranteed between transmitter and receiver. Due to the principle used, no inversion of the sidebands of the 2 m signal takes place.

Fundamentally speaking, various crystal frequencies could be used for the oscillator: Approximately 32 MHz, 57.6 MHz or 96 MHz. These frequencies allow the required three auxiliary frequencies to be obtained by frequency multiplication. Since the number of stages is far less at a frequency of 96 MHz and because the spurious rejection when using a frequency of 57.6 MHz is hardly better than at 96 MHz, the latter frequency was selected for this application. Furthermore the use of a lower frequency is far more liable to produce unwanted

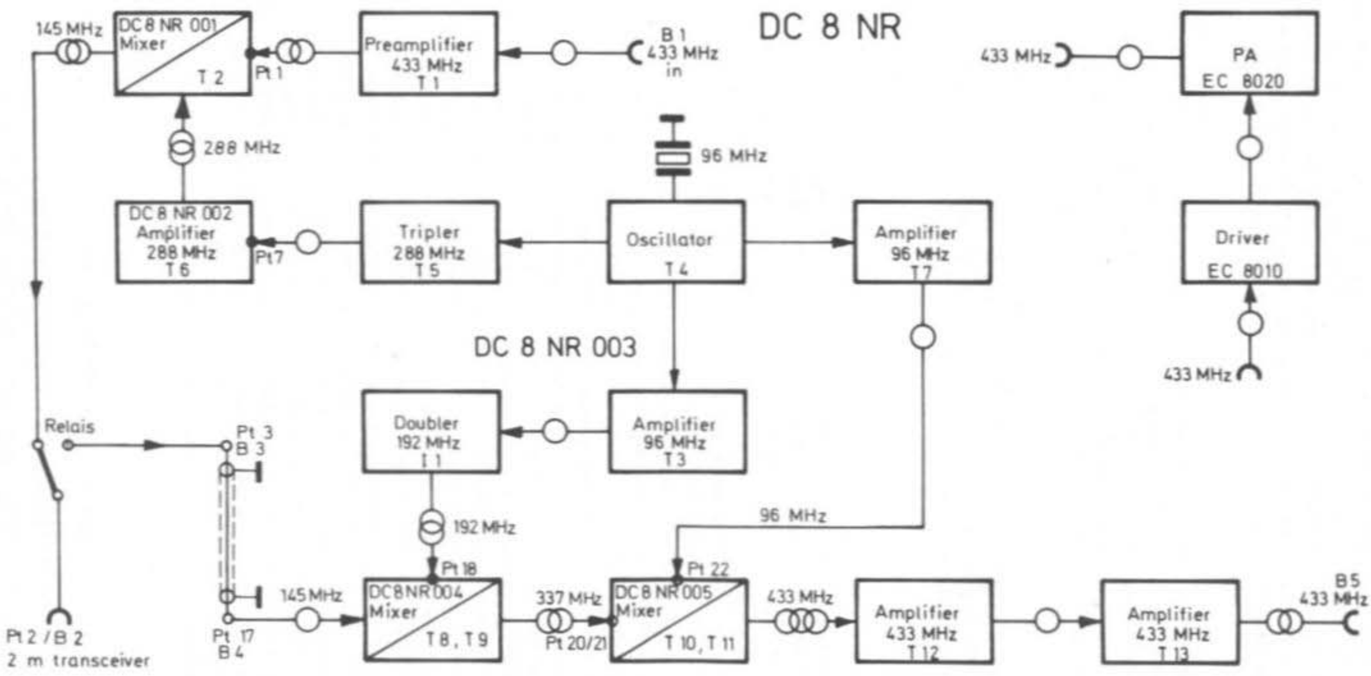

Fig. 1: Block diagram of the linear 144 MHz - 432 MHz transverter with double conversion

harmonics due to the multiple conversion than a higher frequency. Due to the use of push-pull mixers equipped with field-effect transistors or MOSFETS, local oscillator frequencies are obtained that possess very low harmonic contents which together with the low-level conversion and extensive selectivity ensures a very low spurious content of the output signal. The suppression of the third harmonic of the 2 m input signal amounts to more than 80 dB. The actual power amplification of the 433 MHz signal is carried out in a two-stage grounded grid amplifier which provides an extremely linear amplification and high rejection of intermodulation products. In order to obtain the same gain with a sufficient rejection to intermodulation, it would be necessary for at least a five-stage amplifier to be used which would be more expensive.

Virtually all 433 MHz circuits are constructed as coaxial lines of high Q. This is especially valid for the receive converter and the linear amplifier of the transmit converter.This is the reason why these stages use completely screened metal chambers which provide excellent screening between the individual stages. All other stages are accommodated on PC-boards which are enclosed in sheet-metal chambers and mounted on a frame chassis.

Since the gain and selectivity at a predetermined Q is dependent on the unloaded Q, printed striplines were avoided, especially since they would not be able to reduce the mechanical dimensions.

The circuit diagram of the transmit and receive converters are given in Figures 2, 3 and 4.

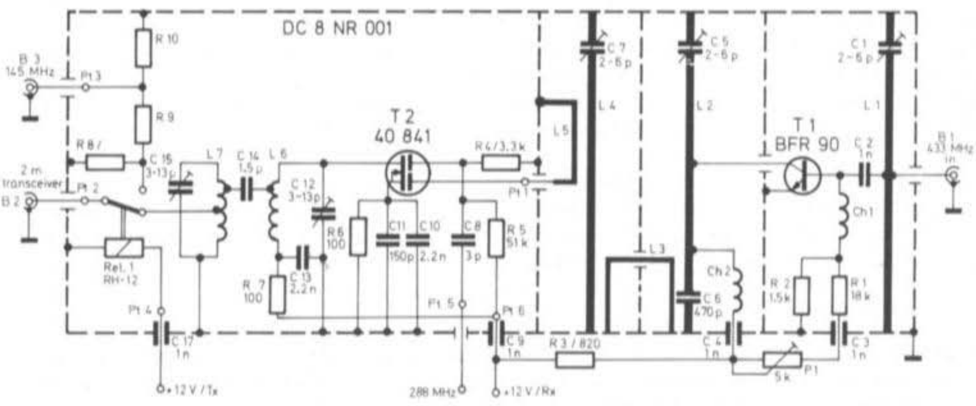

Fig. 2: Circuit diagram of the receive converter using resonant lines in the preamplifier stage

3.1. RECEIVE CONVERTER

3.1.1. PREAMPLIFIER

Due to the use of a high-gain stripline transistor type BFR 90, one preamplifier stage is sufficient in this converter. This transistor exhibits a noise figure of approximately 2.5 dB at 450 MHz which guarantees excellent characteristics for the converter. Unfortunately, the noise figure of the converter was not measured but it is most certainly better than 4 dB. Comparison to the well-known converters described in (2) and (4) and single-stage converters equipped with a transistor AF 279 or 3 N 200 in the preamplifier stage showed that the sensitivity of the described converter was somewhat better than the first mentioned converters, whereas it was considerably better than the latter. Experiments made using different FET and MOSFET transistors in converters with a single preamplifier stage were not satisfactory. Transistor types E 300 were used in a common-gate circuit, type 2 N 5397 in a common-source circuit and the MOSFET 3 N 200 was also tried. In all cases, the attainable degree of preamplification of approximately 8 dB to 10 dB appears to be too low.

The emitter is directly grounded in order to avoid an oscillation of the BFR 90 at lower frequencies. A ferrite choke between the base connection and the voltage divider comprising R 1/R 2 ensures that none of the required signal is lost over the base-voltage divider. Inductance L 2 is isolated from ground by capacitor C 6, so that the collector can be directly connected to a tap on the inner conductor. It is possible in this manner to avoid the use of a critical collector choke when using parallel feeding. The coaxial circuit comprising L 4 provides additional selectivity in front of the mixer stage.

An inductive coupling of the transistor to the input stage L 1 would be an alternative to the capacitive coupling of the preamplifier transistor as shown in the circuit diagram. This would mean that this circuit would not be dampened to such a degree and the input selectivity would increase considerably. However, the noise figure will increase and will then be in the order of the converters described in (2) or (4). This type of coupling is achieved by connecting the BFR 90 to an approximately 25 mm long line of 1 mm dia., silverplated copper wire via C 2. This wire is then grounded and spaced approximately 2 - 3 mm parallel to L 1. All other values remain as given. If any tendencies to oscillation are observed, these can be neutralized by adding a resistor of approximately 4.7 Ω between the collector and the tap on L 2.

Furthermore, it is recommended that the stages comprising transistors T 3, T 5, and T 7 remain connected to the operating voltage as well as the oscillator in order to ensure that no frequency spacing exists between the transmit and receive due to variations of the load.

3.1.2. RECEIVE MIXER

A MOSFET-type 40841 was used in the mixer stage due to its large dynamic range and the isolation between the input and local oscillator signal. It was also very suitable due to its mechanical construction. The more expensive transistor type 3 N 200 did not offer any better results in this application; the well-known type 40673 was even less suitable. Gate 1 of the mixer transistor is grounded via the coupling link L 5. The IF bandpass filter is slightly over-critically coupled. The passband width amounts to 2 to 3 MHz. The overall gain of the receive converter amounts to approximately 15 - 18 dB.

A relay is accommodated on the PC-board of the receive mixer for transmit-receive switching. This is switched with the aid of a control line from the 2 m transceiver. It is also possible for an attenuator suitable for the drive power (R 8 to R 10) to be accommodated if the drive power does not exceed approx. 5 W, otherwise it will be necessary for the attenuator to be mounted externally.

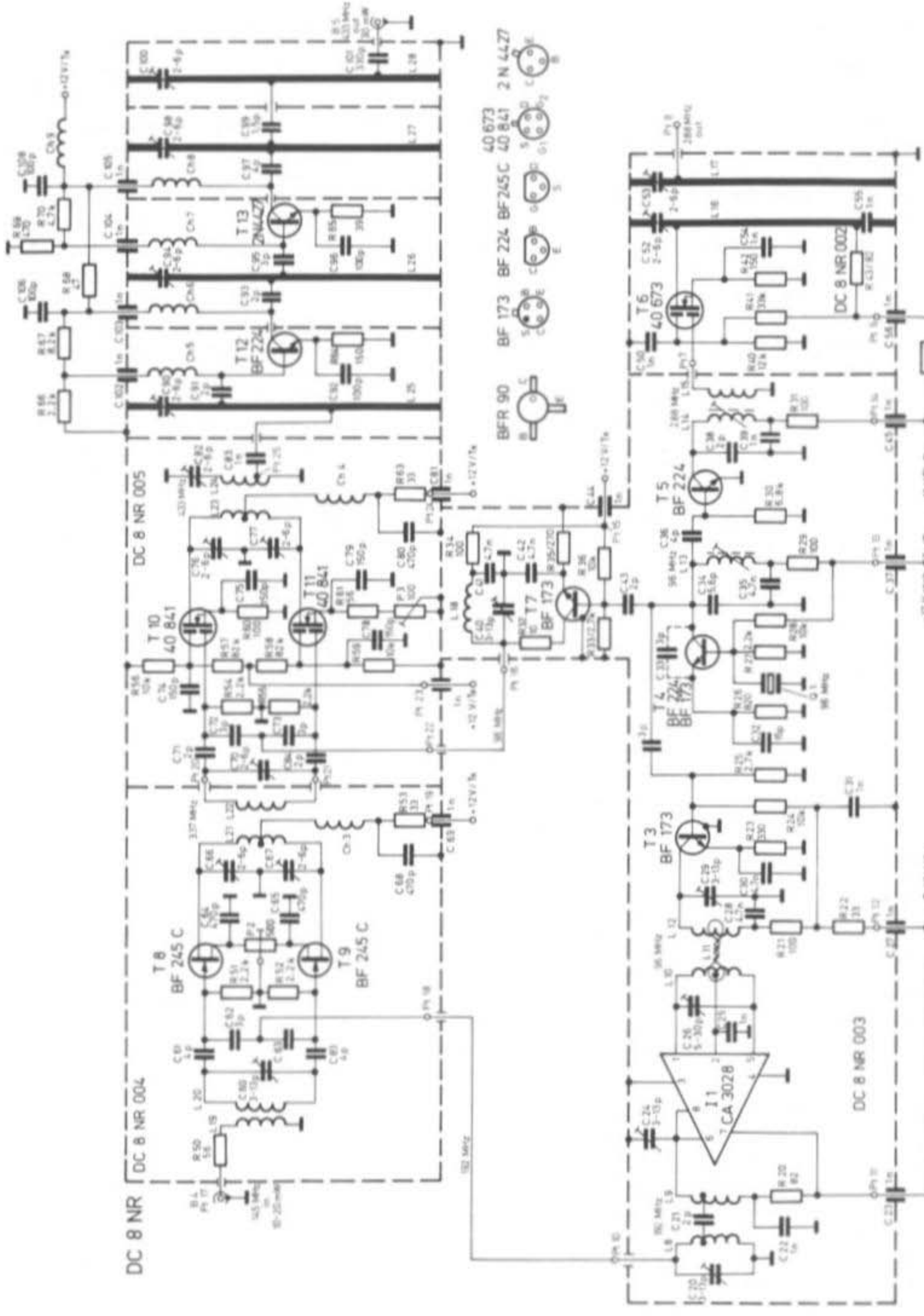

Fig. 3 : Circuit diagram of all other, transistorized stages of the transverter

3.2. GENERATION OF THE LOCAL OSCILLATOR FREQUENCY (Fig. 3)

The circuit for generation of the local oscillator frequency does not offer any special features. The oscillator transistor T 4 operates in a common-base circuit, in which the crystal oscillates at the 5th harmonic in series-resonance and the base of the transistor is grounded with respect to the generated frequency. The feed-back is made partially via the internal collector-emitter capacitance, and partially with the aid of capacitor C 33. If other transistor types are used, it is possible that the value of C 33 will have to be varied. The two 96 MHz buffer and amplifier stages equipped with transistors T 3 and T 7 are loosely coupled to the resonant circuit of the oscillator by 2 or 3 pF. Both stages operate approximately in class A in order to avoid any generation of unwanted harmonics. Only low-reactive transistor types can be used here in order to avoid a frequency shift of the oscillator due to variations of load. A part of the oscillator voltage is passed via capacitor C 36 to the tripler stage equipped with transistor T 5, which operates in Class C. This is followed by an amplifier stage (T 6) which is followed by a coaxial line filter and the required 288 MHz local oscillator frequency for the receive mixer is taken from the secondary circuit.

The 96 MHz signal is coupled with the push-pull input circuit (L 10) of the push-pull/push-push frequency doubler equipped with the integrated circuit CA 3028 (5) from inductance L 12 and via a coupling link. Two of the internal transistor systems of this integrated circuit are therefore biased to class B by the third system and their collectors are connected in parallel. Twice the input frequency (192 MHz) is then available at the output circuit. The circuit operates very reliably and suppresses the fundamental wave and odd harmonics in a very efficient manner.

3.3. TRANSMIT CONVERTER (Fig. 3)

3.3.1. FIRST AND SECOND TRANSMIT MIXER

Both transmit mixers (T 8/T 9 and T 10/T 11) are built up in a similar manner and operate in push-pull. This ensures a good suppression of the fundamental and even harmonics of the input and local oscillator signals. Since the degree of suppression is mainly dependent on the balance of the circuit, potentiometers P 2 and P 3 have been provided for DC-balancing; the RF-balancing is made with the aid of trimmer capacitors C 66 and C 67 or C 76/77.

In contrast to normal practice, additive conversion is used in the second transmit mixer by feeding the signal and local oscillator voltage to gate 1. This allows a higher conversion gain to be obtained.

The good experience the author had with the MOSFET's in the second transmit mixer led to him re-dimensioning the first transmit mixer (DC 6 NR 004) so that it could be equipped with either junction FETs or MOSFETs without any considerable changes to the circuit. Due to the higher conversion gain, this will probably lead to a higher overall gain of the transmit converter. With the complement given in the circuit diagram, a drive power of approximately 10 to 20 mW is required at 144 MHz for full output. The drive-power should be limited to the given value in order to keep spurious signals and intermodulation at a minimum, especially since the output power will not increase to any extent due to limiting.

3.3.2. LINEAR AMPLIFIER

The 70 cm signal is then passed through a three-link bandpass filter comprising inductances L 23, L 24, L 25 and fed to a two-stage transistor amplifier which amplifies the signal up to approximately 30 mW. The coupling capacitances should not be changed to any extent. They have been selected for optimum matching to the given transistor types. The two emitter bypass capacitors C 92 and C 96 are ceramic disc types without connection leads. They are directly soldered to the screening panel.

A special UHF transistor type 2 N 5913 was tried instead of transistor 2 N 4427, however, it did not provide any increase of the output power. Furthermore, it is only specified for class C operation. Due to this and the fact that it was more expensive, it was decided not to use this transistor in the described amplifier. If a higher output power is required, e.g. for portable operation, it is possible for the last chamber to be equipped with a similar stage and the secondary circuit of the bandpass filter used as collector circuit. In this case, the two feedthrough capacitors required should be soldered into place beforehand since a subsequent removal of the coaxial, inner conductor would entail difficulties.

3.4. THE TUBE AMPLIFIER (Fig. 4)

A two-stage grounded-grid amplifier equipped with the tubes EC 8010 and EC 8020 is used for the actual power amplifier of the 70 cm signal. The overall gain amounts to at least 25 dB at the given plate voltages. According to the individual tubes, approximately 8 - 11 W is available at the output.

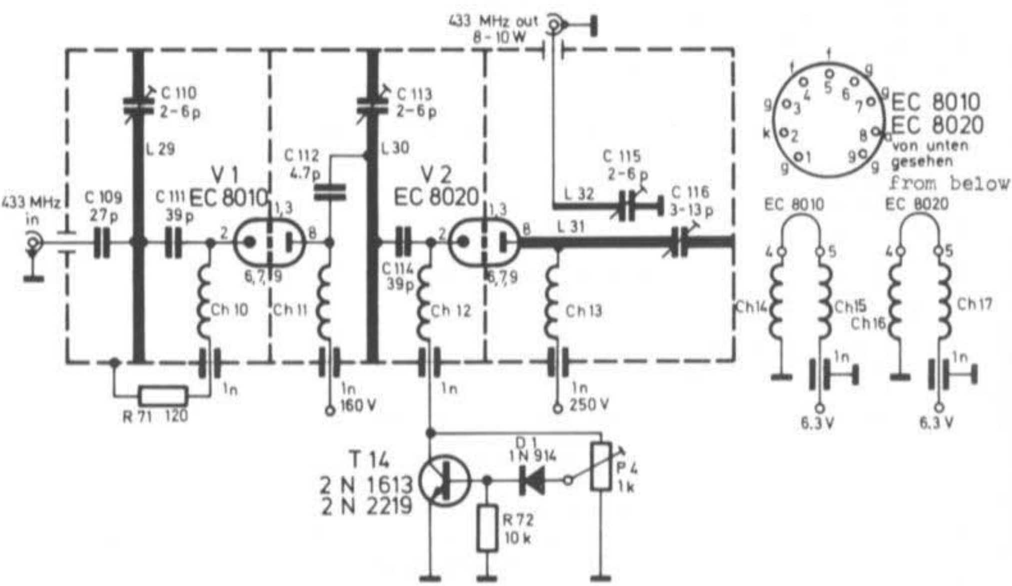

Fig. 4: Circuit diagram of the linear power amplifier

The grid bias voltage for the driver tube is generated across a resistor, and a constant voltage source is provided in the cathode line in the case of the EC 8020. The input and output circuits of the EC 8010 are capacitively shortened $\lambda/4$ coaxial circuits. The output circuit of the tube EC 8020 is $\lambda/2$ in length and is constructed in the form of a plate-metal strip in order to increase the impedance. The antenna is inductively coupled to this line with the aid of a loop which is provided with a variable capacitor at the ground end.

The capacitive coupling to the cathode of each tube (C 111, C 114) was found to be superior to a π-circuit or inductive coupling both with respect to the ease and correctness of the alignment. The most favourable matching can easily be found in this manner.

In order to ground the anode circuit of the EC 8010, the line has been isolated with respect to DC-voltages with the aid of capacitor C 112. The plate voltage is fed with the aid of an insulated wire which is fed in through a small hole at the cold end of the inner conductor and passed through it up to the anode. Since the voltage distribution within the tube of the circuit accurately coincides to that on the outer surface, it is possible in this simple manner to use the blocking characteristics of the $\lambda/4$ line. The choke Ch 11 is only provided as an additional means in order to ensure that any residual RF-voltage is by-passed. Since this form of construction cannot easily be shown in the circuit diagram, Ch 11 is drawn as if it were directly connected to the anode.

4. COMPONENT DETAILS

T 1: BFR 90 (Valvo, Philips), BFR 34 (A) (Siemens)
T 2: 40841 (RCA)
T 3: BF 173 (AEG-Tfk, Siemens, Valvo)
T 4: BF 173 or BF 224
T 5: BF 224 (TI) or BF 173
T 6: 40 673 (RCA) or 40 841
T 7: BF 173
T 8, T 9: BF 245 C (TI), W 245 C (Siliconix) or similar FET
T 10, T 11: 40841
T 12: BF 224
T 13: 2 N 4427
T 14: 2 N 1613, 2 N 2219 or similar

D 1: 1 N 914, 1 N 4148 or similar silicon diode

I 1: CA 3028 (RCA)

V 1: EC 8010 (AEG-Tfk, Valvo, Siemens)
V 2: EC 8020 (AEG-Tfk)

Rel 1: RH-12 (National) encapsuled 12 V miniature relay.

L 1: Brass tube of 6 mm outer diameter, 91 mm long, tap 18 mm from the cold end.
L 2: As L 1, tap 37 mm from the chamber end.
L 3: Silver-plated copper wire of 1 mm dia. (18 AWG) fed for a length of 20 mm parallel to L 2/L 4, spacing approximately 2 mm
L 4: As L 1
L 5: As L 3, placed with a length of 30 mm parallel to L 4 with approx. 1 mm spacing.

L 6: 5 turns of 1 mm dia. (18 AWG) enamelled copper wire wound on a coil-former of 5 mm dia. with the cold end of the coil facing the board

L 7: 7 turns, otherwise as L 6, coil tap 1.5 turns from the cold end

L 8, L 9: 4 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 6 mm former, self-supporting, coil tap for C 21: 0.5 turns from the hot end, spacing between inductances 1 - 2 mm

L 10: 8 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 5 mm former, self-supporting.

L 11: Coupling link from 0.4 mm dia. (26 AWG) enamelled copper wire wound as a loop in the centre of L 10 and the other end towards the cold end of L 12

L 12: 8.5 turns of 0.8 mm dia. (20 AWG) enamelled copper wire wound on a 5 mm coilformer, close wound. Collector end facing the board

L 13: 5 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 5 mm coilformer with core (violet), collector end facing the board

L 14: 2.5 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 5 mm coilformer with core (violet), coil length approx. 8 mm, cold end facing the board

L 15: 1 turn of 0.4 mm (26 AWG) enamelled copper wire placed towards the cold end of L 14

L 16: 2 mm dia. (12 AWG) silver-plated copper wire, length corresponding to the holes in the PC-board DC 8 NR 002, spacing approx. 4 mm from the board.

L 17: As L 16

L 18: 6 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 5 mm coilformer, collector side facing the board

L 19: 1 turn of 0.4 mm dia. (26 AWG) enamelled copper wire wound in the centre of L 20

L 20: 6 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 8 mm former, self-supporting

L 21: 4 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 6 mm former, self-supporting, coil length according to the holes on the PC-board, centre tap

L 22: 3 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 6 mm former, self-supporting, coil length approx. 12 mm, spaced 3 mm from the board

L 23: U-shaped bracket from 1.5 mm dia. (15 AWG) silver-plated copper wire, length 26 mm, width corresponding to the holes on the PC-board, spacing approx. 4 mm, centre tap

L 24: U-shaped bracket from 1 mm dia. (18 AWG) silver-plated copper wire, length 17 mm, width corresponding to the holes on the PC-board, spacing approx. 6 mm, tap approx. 15 mm from the cold end

L 25: Brass tubing of 4 mm outer diameter, 47 mm long, tap 14 mm from the cold end

L 26: As L 25, but with tap 30 mm from the cold end

L 27: As L 25, but without tap

L 28: As L 25, but with tap 10 mm from the cold end

L 29: Brass tubing, 6 mm outer diameter, 50 mm long, tap 14 mm from the cold end

L 30: As L 29, but with tap 16 mm from the cold end

L 31: Brass strip of 6 mm x 0.7 mm x 90 mm long, bent at 15 mm. Cut the bent end somewhat so that it is suitable for soldering to the tube socket

L 32: 1 mm dia. (18 AWG) silver-plated copper wire placed approx. 4 - 5 mm from L 31 and bent by 90°. Length is given by the spacing between the trimmer and connector

Ch 1: Ferrite bead placed on the line.

The following chokes consist of 0.4 mm dia. (26 AWG) enamelled copper wire, close wound on a 3 mm former, self-supporting (exceptions: Ch 5, 7, 17).

Ch 2: 12 turns
Ch 3: 22 cm of wire ($\lambda/4$ choke)
Ch 4: 17 cm of wire ($\lambda/4$ choke)
Ch 5: 3 turns placed through a ferrite bead
Ch 6: 6 turns
Ch 7: As Ch 5
Ch 8: 7 turns
Ch 9 - 16: 17 cm of wire ($\lambda/4$ choke)
Ch 17: Ferrite choke with 6-hole core

C 1, C 5, C 7: Ceramic, tubular trimmer 0.8 - 6.8 pF or similar
C 12, C 15: Ceramic, disc trimmer 3 - 13 pF, 7 mm dia. or plastic-foil trimmer
C 20, C 24, C 29, C 40: Air-spaced trimmer 3 - 13 pF with 4 or 2 pins for PC-board mounting
C 26: Air-spaced trimmer of 5 - 30 pF with 2 or 4 pins
C 52, C 53: Air-spaced trimmer 2 - 6 pF with 4 or 2 pins
C 60: 3 - 13 pF as C 20
C 66, C 67, C 70, C 76, C 77, C 82: 2 - 6 pF as C 52
C 90, C 94, C 98, C 100, C 110, C 113, C 115: 0.8 - 6.8 pF as C 1
C 116: Ceramic tubular trimmer or air-spaced trimmer with 2 connections, 3 - 13 pF
C 6: Approx. 470 pF (value not critical) ceramic disc capacitor without connection leads, approx. 12 mm diameter
C 92, C 96: Approx. 100 pF (value not critical) ceramic disc capacitor without connection leads

All feed-through capacitors approx. 1 nF (value not critical)

All other capacitors: Ceramic miniature disc capacitors

All resistors for 10 mm spacing

Crystal: 96.000 MHz, HC 25/U directly soldered onto the board.

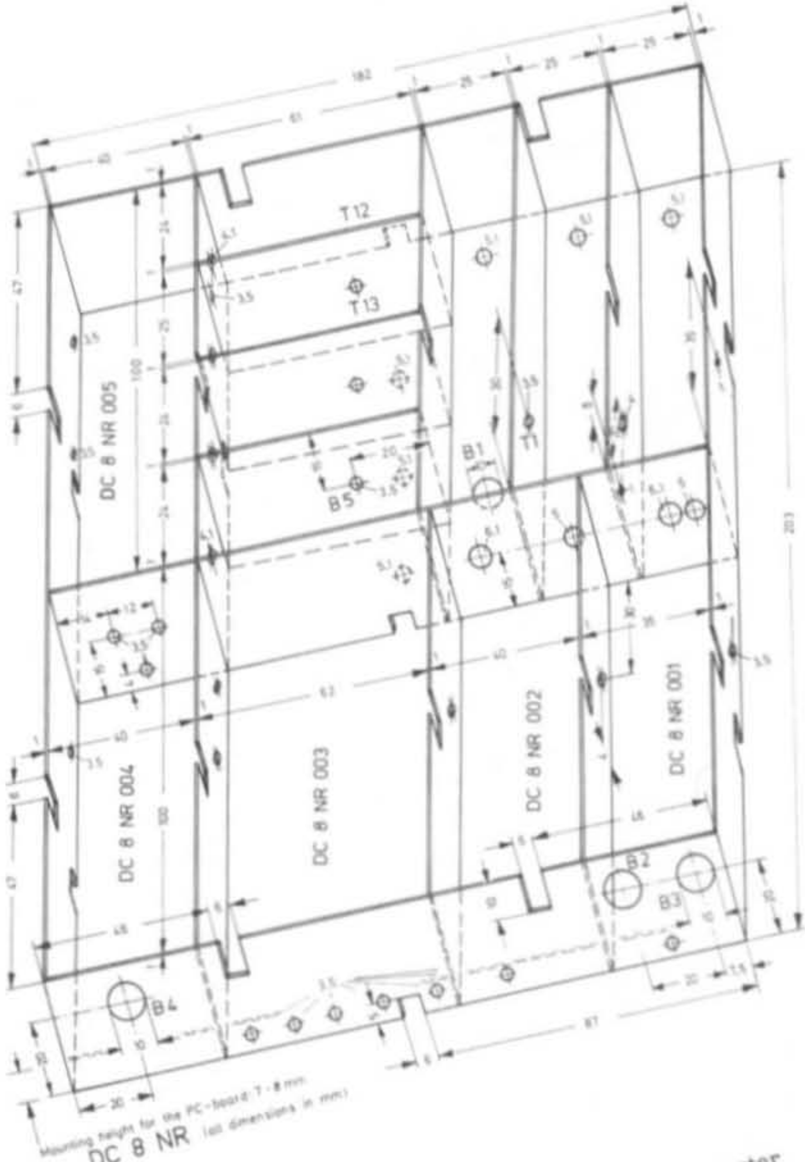

Fig. 5: Metal work of the transverter

5. MECHANICAL CONSTRUCTION

The actual converter and tube power amplifier are accommodated in two separate chassis. The mechanical construction is given in Figure 5 and Figure 6. Further details are also given in the photos shown in Figure 7 to Figure 9. A 0.7 mm thick brass plate is used for all panels. The side and intermediate panels are made from 30 mm wide strips of the correct length that are subsequently soldered into place. This method was found to be more satisfactory than bending the framework. The construction used is extremely stable which means that even thinner plates can be used. However, these are not very suitable for machine cutting.

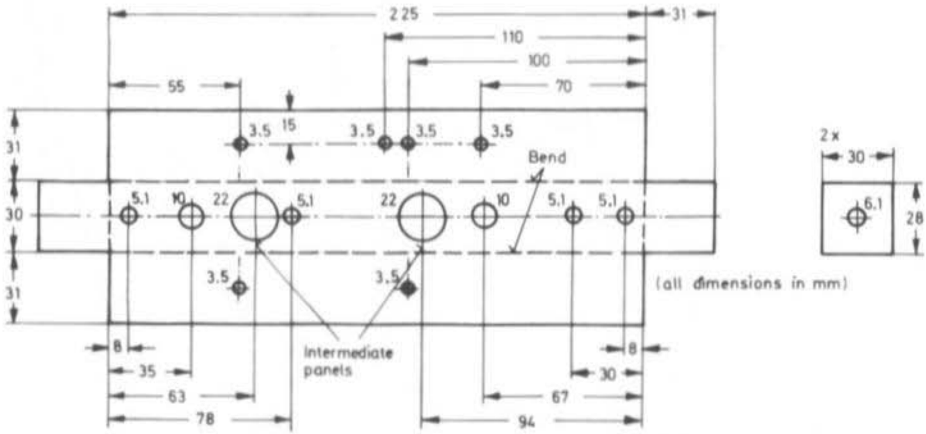

Fig. 6 : Metal chassis of the power amplifier

The following order of construction was found to be advisable: Firstly, drill the holes for the BNC connectors designated B 2, B 3, and B 4 after which the framework of the converter comprising the four strips are soldered together. The 3.5 mm holes for the feedthrough capacitors can be drilled afterwards to ensure greater accuracy and in order not to bend the metal plate. This is followed by soldering in the intermediate panels in the order given in Figure 5. All dimensions are inner values. It is necessary for these panels to be drilled before soldering into place. It is very important that all cut-outs, slots and holes are made before soldering the individual plates into place.

The next process is to solder the nuts for holding the covering plates into place and installing the base plate for the stages shown in Figure 7. This part is soldered into the required cut-out in the frame so that a height of 30 mm is maintained. This allows the bottom of the cabinet to be used as base for the whole converter. After completing the whole framework, it is possible for any other holes to be made whose positions are mainly dependent on the connection points on the PC-board. After drilling the boards but before mounting the components, each of the boards is placed into the framework and the lower side of the PC-board soldered approximately every 7 to 8 mm to the frame.

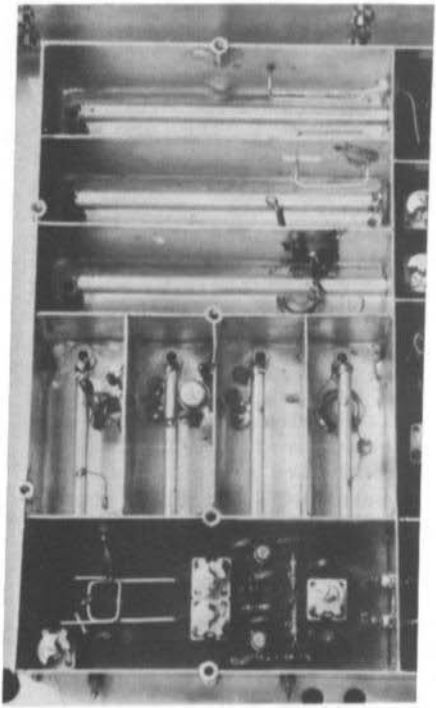

Fig. 7 :
Author's prototype: Preamplifier stage of the converter (above), linear amplifier (centre) and DC 8 NR 005

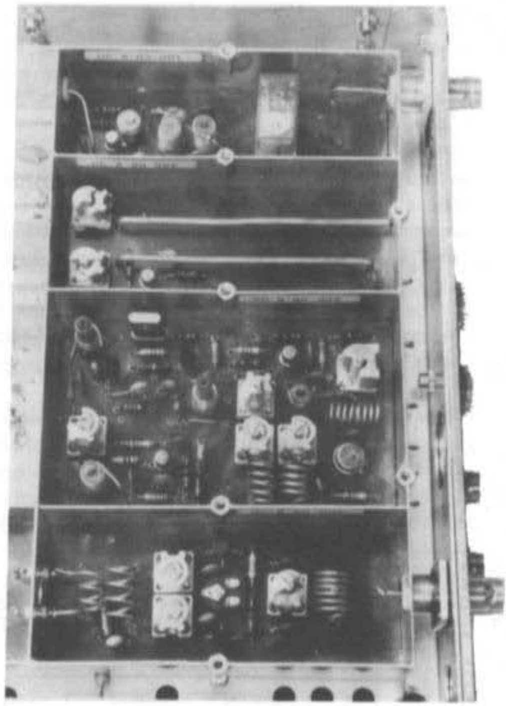

Fig. 8 :
Author's prototype: DC 8 NR 001 to 004

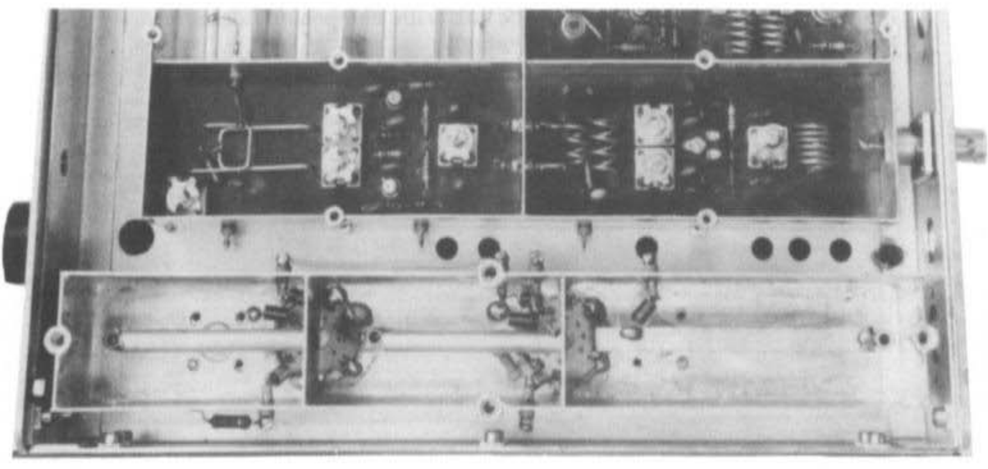

Fig. 9 : Author's prototype: The power amplifier

This should be made very quickly and with a relatively small soldering iron to ensure that the conductor lanes of the PC-board are not damaged. After removing the protective surface, the framework should be silver-plated to a thickness of 10 μ. The brass tubing used for the resonant lines is silver-plated before installation and then cut to the required dimensions. They are approximately 2 mm longer than the given inner dimensions so that they can be soldered to the outside of the chamber. Inner conductor L 2 is soldered to a disc capacitor of sufficient diameter (vertical) which is then soldered to the appropriate panel that is then provided with a small hole. The soldering process should be carried out as quickly as possible so that the coating of the capacitor is not damaged. If necessary, the disc capacitor can be replaced by the lower part of ceramic tubular trimmer without spindle and the inner conductor placed onto the ceramic portion. In this case, a capacitor with very short connection is used for bypassing.

Finally, all feedthrough capacitors and the BNC connectors are mounted into place (the flange of the BNC connectors should be soldered to the inside of the chamber), after which the trimmers and inner conductors are mounted. This is followed by installing the components of PC-boards DC 6 NR 001 to 005. It is important that all transistors should be soldered into place with the shortest possible connections.

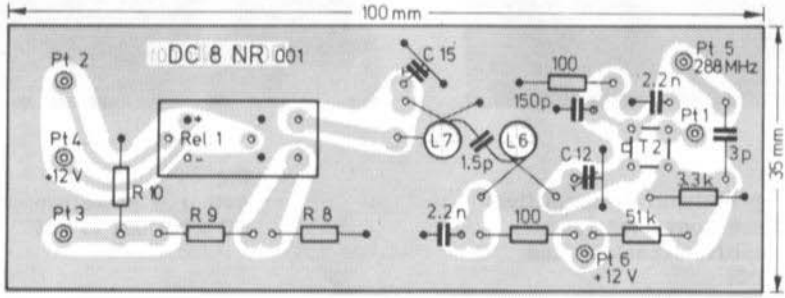

Fig. 10: PC-board DC 8 NR 001 with component locations (receive mixer, relay)

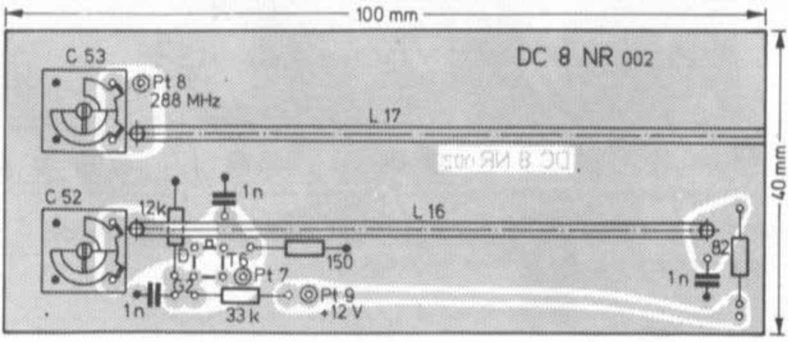

Fig. 11: PC-board DC 8 NR 002 with component locations (288 MHz amplifier)

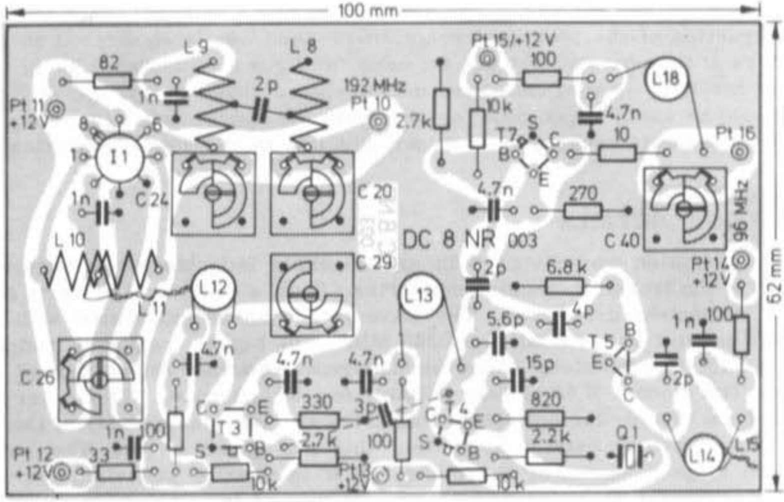

Fig. 12: PC-board DC 8 NR 003 with component locations (local oscillator module)

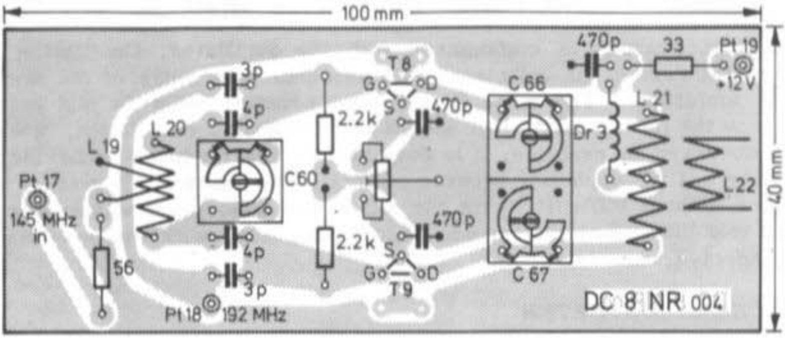

Fig. 13: PC-board DC 8 NR 004 with component locations (first transmit mixer)

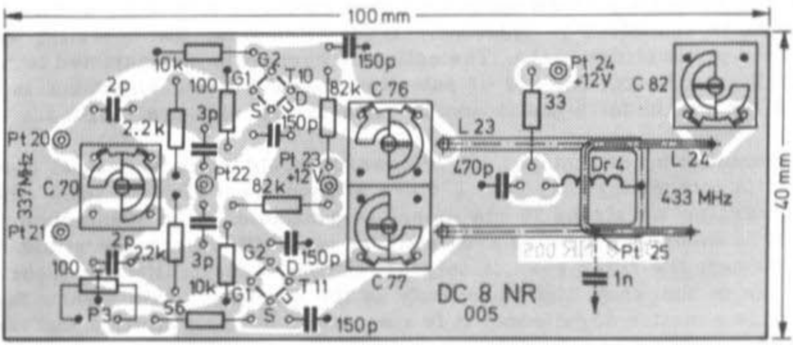

Fig. 14: PC-board DC 8 NR 005 with component locations (second transmit mixer)

The construction of the tube power amplifier need not be described in detail. A good idea of the construction can be seen in Figure 9, as well as in (2) and (6). Only the UHF-type of tube base manufactured from epoxy, glassfibre material should be used, if possible together with the appropriate screening collar. Pertinax or ceramic sockets are not suitable due to the long connection leads.

6. ALIGNMENT

6.1. GENERAL DETAILS

A simple absorption wavemeter with exchangeable inductances is required for alignment of the transmit-receive converter. Such a unit is also very suitable for other alignment processes. The wavemeter can be calibrated with a conventional dipmeter at approximately 280 MHz. The higher frequency ranges can then be easily interpolated, since an accurate frequency measurement is not required. The diode 1 N 82 is very suitable for this purpose. With other diodes that have been tested, the indication was not very clear, especially at the higher frequencies. Further details regarding this are given in (8). The individual stages are aligned one after the other and it is usually sufficient for the absorption wavemeter to be placed in the vicinity of the circuit in question. If this is not sufficient, it is coupled to the circuit to be aligned with the aid of an approximately 1 pF capacitor.

6.2. ALIGNMENT OF THE LOCAL OSCILLATOR MODULE

The alignment process is commenced with the oscillator. Oscillation should cease when the core of the inductance is rotated to both sides of the maximum oscillator amplitude. The most favourable operating point is just below the maximum on the flatter skirt. If it should not break into oscillation, which has not been observed in practice, it is possible for the feedback to be increased by inserting a 1 pF capacitor between collector and emitter. The other stages are aligned without difficulty with the aid of the absorption wavemeter. It is important that the alignment is made under load, e.g. with the coupling capacitors connected.

6.3. RECEIVE CONVERTER

The mixer stage of the converter is now connected to the operating voltage. The IF bandpass filter is now aligned for maximum noise indication on the S-meter of the connected 2 m receiver. The 288 MHz secondary circuit can now also be carefully corrected for maximum noise. If the alignment is correct, a signal of approximately 30 dB above the noise should be perfectly readable when the antenna is connected to inductance L 5. After this, the operating voltage is fed to the preamplifier stage. The collector current is now adjusted to approximately 2 - 4 mA with the aid of potentiometer P 1 (this alignment is made later for the maximum signal-to-noise ratio at low signal strength).

Before commencing alignment, all trimmers should be in a central position. The circuits comprising L 1, L 2, L 4 are aligned from back to front when no signal generator or strong 70 cm signal is available, e.g. commencing with L 4. This is necessary because even a signal of 40 dB over noise would not be passed through the filter even if only one of the two circuits are incorrectly aligned due to the very high selectivity of the circuits. This means that the alignment is a matter of patience. It is also rather difficult to align the circuits for maximum noise since the noise only increases very slightly even when the circuits are correctly aligned.The antenna is therefore connected to the coupling link L 3, and L 4 tuned for maximum indication. After this, the antenna should be connected to the input connector. Inductance L 2 should then be aligned firstly which is then followed by L 1. Due to the large degree of damping, the adjustment of L 1 is quite wide.

6.4. TRANSMIT CONVERTER

The alignment of the transmit converter is made in steps. A 2 m transmitter of the required power (attenuator) is now connected. The alignment is commenced with the 145 MHz input circuit which is followed by alignment of the 337 MHz output circuit. If an indication is present, it is possible for the 192 MHz bandpass filter comprising C 20, C 24 to be corrected. This is followed by aligning P 2 and C 66/67 for maximum suppression of the local oscillator frequency. The electrical centre point is found by touching the individual windings with a screwdriver and noting the position where the reduction of power is at a minimum. This point is shifted to the connection point of the choke with the aid of the two trimmers. The same is valid for the alignment of the second mixer. The RF bandwidth of the converter is mainly determined by the output circuit of this stage. The linear amplifier is also aligned in the same manner.

6.5. TUBE AMPLIFIER

The alignment of the tube amplifier is made by connecting a 50 Ω resistor (dummy-load) or antenna with SWR meter to the output. After allowing a sufficientwarm-up time,the various plate voltages are connected one after the other to the tubes. The quiescent current of the tube EC 8010 should amount to approximately 15 mA, and approximately 25 - 30 mA is adjusted for the tube EC 8020 with the aid of potentiometer P 4.

The alignment of the input and output circuits of the driver is also made with the aid of the absorption wavemeter. If both stages are correctly aligned, a small power indication is usually present at the output even without connecting the plate voltage to the output stage. The trimmers are aligned alternately for maximum output power. The alignment is completely by carefully correcting all stages for minimum ripple in the passband range.

7. NOTES

The 96 MHz crystals manufactured by KVG that are provided with the kit seem to be considerably more active than the crystal used in the prototypes, and that are available from other sources. This caused the three subsequent stages to be overdriven so that the collector currents sometimes exceeded several mA (presuming that the correct harmonic has been selected, which requires a dipmeter that is able to tune at least upto 200 MHz). In order to avoid this and oscillation of the crystal at 57.6 MHz (3rd overtone) which has also been observed, the following modifications are recommended:

Increase value of C 32 to 27 pF
Increase value of C 34 to 12 pF
Reduce value of C 43 to 1 pF

The coupling capacitor to the base of T 3 should be reduced from 3 pF to 1 pF. Connection Pt 16 is no longer connected to the hot end of inductance L 18 but to a coil tap 3 turns from the cold end. It is then possible to use coaxial cable for the interconnection to Pt 22.

The alignment of the local oscillator module is now made by monitoring the collector current of the subsequent stage and decreasing the coupling to the previous stage until the collector current just increases when the oscillator commences operation.

Since it is necessary for the attenuator for the 2 m transmitter to be located after the relay, it is necessary to check the operation of the relay before transmitting at high power for the first time because the transmit power could otherwise damage transistor T 2.

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AN INTEGRATED RECEIVER SYSTEM FOR AM, FM, SSB and CW

PART VI: POWER SUPPLY, AF-LOWPASS FILTER AND S-METER STAGES

by H. J. Franke, DK 1 PN

The following article describes the power supply, AF-lowpass filter and S-meter stages of the 9 MHz receiver for the AM, FM, SSB/CW modes. This receiver is mainly equipped with modern communication ICs. The following modules have already been described:

SSB/CW IF module with Plessey ICs DK 1 PN 003 (1)
Carrier oscillator module DK 1 PN 002 (2)
AF-amplifier with active CW filter DK 1 PN 004 (3)
Input module with FM circuit (CA 3089 E) DK 1 PN 005 (4).

The following are still to be described:
A simple AM-IF-module DK 1 PN 006 and - as recommended by D. E. Schmitzer, DJ 4 BG - a printed circuit board with connectors for all modules. When all modules are plugged onto this main board, it is only necessary for an external converter, S-meter, loudspeaker, switches and potentiometers to be connected, as well as the power line cable.

The described module DK 1 PN 007 comprises a completely encapsuled power transformer and three voltage stabilizer circuits for the various voltages required by the receiver. Furthermore, it comprises an active low-pass filter which provides a considerable improvement of the signal-to-noise ratio especially in the AM and FM modes, as well as two S-meter stages. A block diagram of the whole 9 MHz receiver system including all external connections is given in Figure 1.

1. CIRCUIT DETAILS

The circuit diagram of the power supply, AF low-pass filter and S-meter stages is given in Figure 2.

1.1. POWER SUPPLY

The encapsuled power transformer is only 47.5 mm by 36 mm by 25 mm high. It provides an output voltage of 24 V with centre tap (5 VA). This voltage is rectified in push-pull so that approximately 14 V is available at the filter capacitor at full load. With the low loading of the receiver, the voltage at the filter capacitor amounts to approximately 19 V. Under non-load conditions, this will increase to approximately 24 V. This means that the filter capacitor C 701 must be at least a 25 V type.

Three voltage stabilizer circuits are provided. The output voltage of the integrated voltage stabilizer I 701 can be adjusted with the aid of the voltage divider comprising resistors R 703/R 704. This voltage is required for the oscillators of the receiver and will amount to 6 V if the component values given in the circuit diagram are used. This output can be short-circuited without damage. Current limiting ensures that only 60 mA can be taken from the output. The output voltage of the two, simple voltage stabilizer circuits comprising zener diode and pass transistor can be varied by selection of the appropriate zener

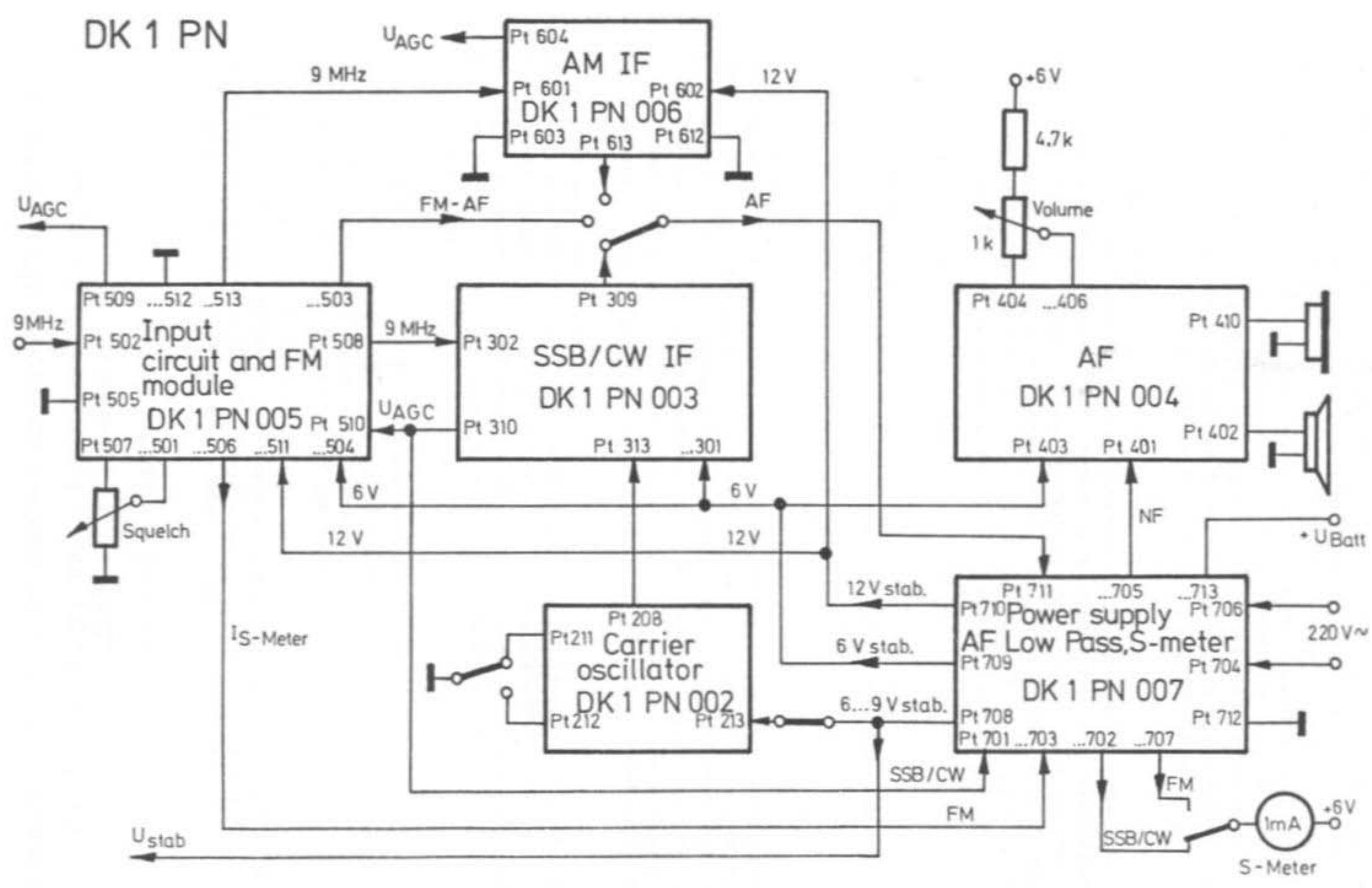

Fig. 1: Block diagram of the complete 9 MHz receiver

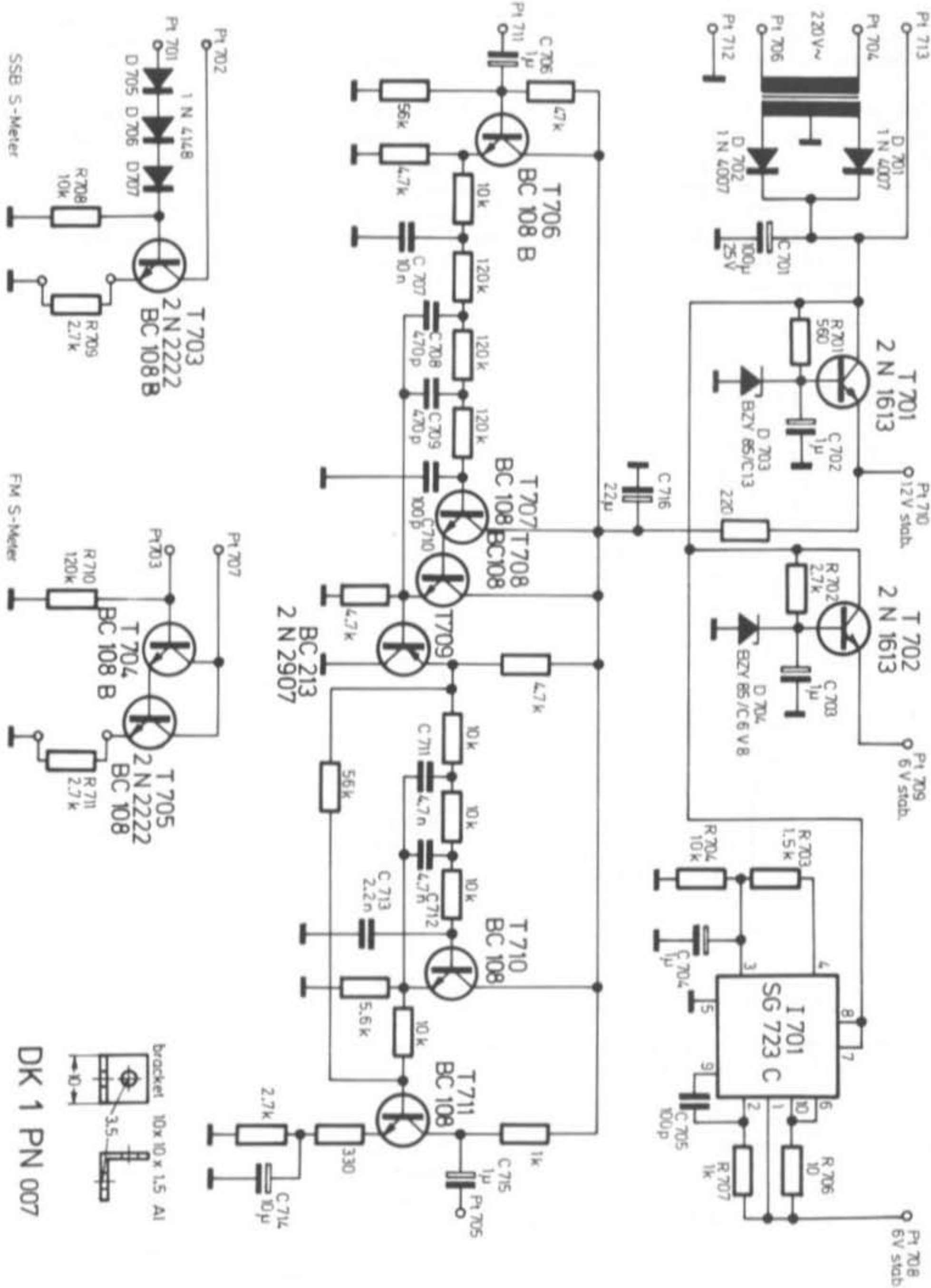

Fig. 2: Circuit diagram of power supply, AF low-pass filter and S-meter stages

diode. The voltage of the diode should be selected so that it is 0.7 V higher than the required output voltage. Both of these circuits can be loaded to a maximum of 100 mA if the pass transistors are provided with cooling fins.

The stabilizer circuit comprising T 702/D 704 is adjusted to 6 V and is used to feed all Plessey integrated circuits. The stabilizer circuit comprising T 701/D 703 provides 12 V for all other stages. During battery operation of the receiver, it is necessary for the last voltage stabilizer circuit to be bridged, since the battery voltage could fall below 14 V and means that this stabilizer circuit will no longer operate. The battery should be connected to connection point Pt 713. During power-line operation, a small battery can be charged via a dropper resistor from this connection.

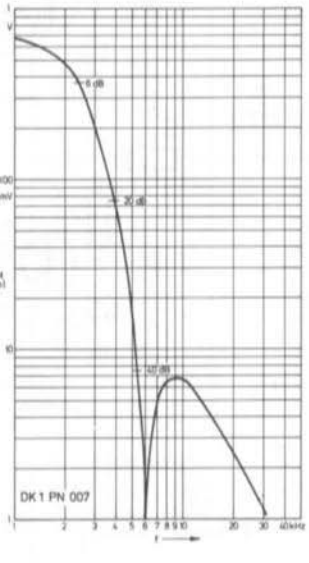

Fig. 3:
Frequency response of the AF low-pass filter

1.2. AF-LOW-PASS FILTER

The steep, active low-pass filter comprises two filters of 3 links each and is dimensioned as described in (5). The cut-off frequency amounts to 2 kHz and the attenuation pole is at 6 kHz. Figure 3 gives the frequency response of this filter. As can be seen in this circuit diagram, the active filter is directly connected to 12 V on the PC-board. At this operating voltage, the maximum AF input voltage is 7 V peak-to-peak, or approximately 2.5 V RMS. In the case of the 9 MHz receiver, the AF-voltage only amounts to approximately 10 to 20 mV. In order not to limit the suitable AF output amplifier by the low sensitivity and in order to avoid an emitter follower at the output, an amplifier stage has been provided at the output of the filter. The emitter resistor of 330 Ω can be increased or decreased according to the required output voltage.

As can be seen in the block diagram, the AF low-pass filter is effective in all modes. It is used to suppress all noise voltages whose frequencies are in excess of approximately 2.5 kHz. In the SSB mode, such noise is generated in the wideband integrated amplifiers subsequent to the crystal filter. A tailoring of the frequency response is extremely necessary in the FM mode since the noise voltages appearing after demodulation are not distributed homogeneously but increase towards higher frequencies. In the AM mode, it is possible that the AF low-pass filter will make a special AM crystal filter unnecessary at least with respect to the sensitivity threshold. However, if an optimum of adjacent selectivity is required, it will be necessary for a crystal filter to be used.

1.3. S-METER STAGES

Several voltages are available in the receiver that are dependent on the field strength. Two of these are used and processed so that they are suitable for driving an S-meter.

The integrated circuit CA 3089 E in module DK 1 PN 005 provides a current for an S-meter as shown in Figure 2 of (4). This current can be read off in the range of 80 dB, and this range can be extended when the converter is also controlled from connection Pt 509. A voltage is generated across resistor R 710 from this current which is fed to a Darlington-circuit comprising transistors T 704 and T 705. The meter is connected between the operating voltage of 6 V and connection Pt 707. The quiescent current of module DK 1 PN 005 will not be indicated since the voltage across R 710 must firstly exceed the threshold voltage of the two-base emitter paths. The indication will increase from this threshold with increasing signal up to a maximum value determined by resistor R 711. The given value is only valid for a meter having a full scale deflection of 1 mA. However, any meter can be used as long as resistor R 711 is correspondingly selected so that full-scale deflection is obtained at the maximum signal to be expected.

Virtually the same is valid for the SSB S-meter circuit. The Plessey-integrated circuits provide a control voltage between 2.0 V and approximately 4.4 V with a control range of approximately 130 dB. The bias voltage of 2 V is suppressed by four diode paths (D 705 to D 707 and T 703). Resistor R 709 is selected so that full scale deflection is achieved at an input voltage of 4.4 V. This corresponds to an RF voltage of at least 10 mV at the input (Pt 502) of module DK 1 PN 005. This information was given in an application sheet published by Plessey, but it was found that four diode paths were necessary instead of the given three since the threshold voltage was approximately 0.5 V at the very low base current.

In contrast to several publications regarding the Plessey integrated circuits, it is not possible for the scale of the S-meter to be divided linearly in dB. The lower 60 dB are further apart than the upper 60 dB. This means that a calibrated attenuator is necessary for calibration. The "Plessey-S-meter" has the disadvantage that the indication does not follow fluctuations of the field strength continuously due to the special time constants of the control voltage IC SL 621. This means that the FM S-meter should be used for tuning and for adjusting the antenna direction. The Plessey S-meter can also be used for FM and AM if the carrier oscillator is switched on when it is desired to read off

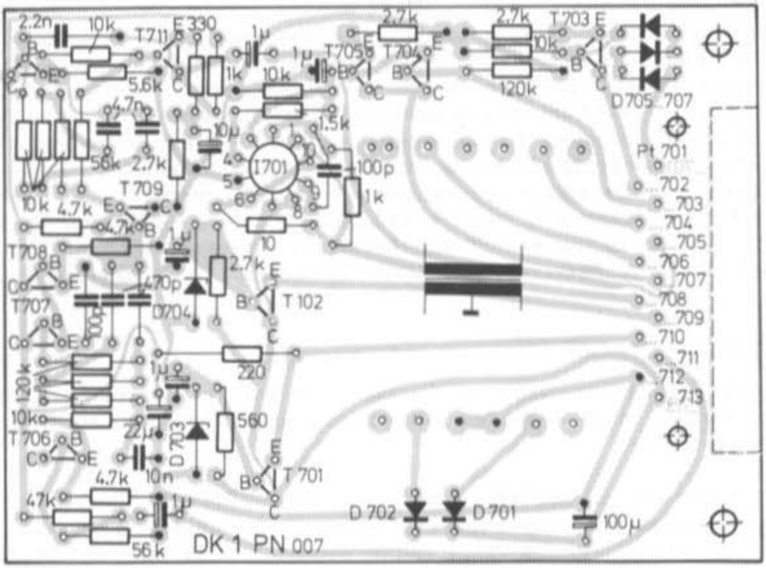

Fig. 4: Component locations on PC-board DK 1 PN 007

the signal strength. For this purpose it is possible for the S-meter change-over switch to be provided with a further contact. However, it is very useful for two S-meters to be installed; the FM S-meter can then be a cheap meter without calibration.

2. COMPONENT DETAILS

T 701, T 702: 2 N 1613, 2 N 2219 or similar silicon transistors in TO 5 case.
T 703 to T 708; T 710, T 711: BC 108 B or similar silicon AF-transistors with B min. 100
T 709: BC 213, 2 N 2907 or similar silicon PNP transistor

I 701: SG 723 C (Silicon General) or equivalent

D 701, D 702: 1 N 4001 to 1 N 4007 or similar rectifier diode
D 703: BZY 85/C 13 or similar 13 V zener diode
D 704: BZY 85/C6V8 or similar 6.8 V zener diode
D 705 to D 707: 1 N 914, 1 N 4148 or any silicon diode

Transformer: TFB 468/012 - 01 - N2 (available from the publishers)

C 701: 100 μF or 220 μF/25 V

All other electrolytics: Tantalium drop-types, 16 V

C 705, C 707 - C 713: Styroflex capacitors, or if not available ceramic disc types.

All resistors: For 10 mm spacing.

1 13-pin connector
3 cooling fins for TO 5
2 brackets 10 mm x 10 mm x 10 mm.

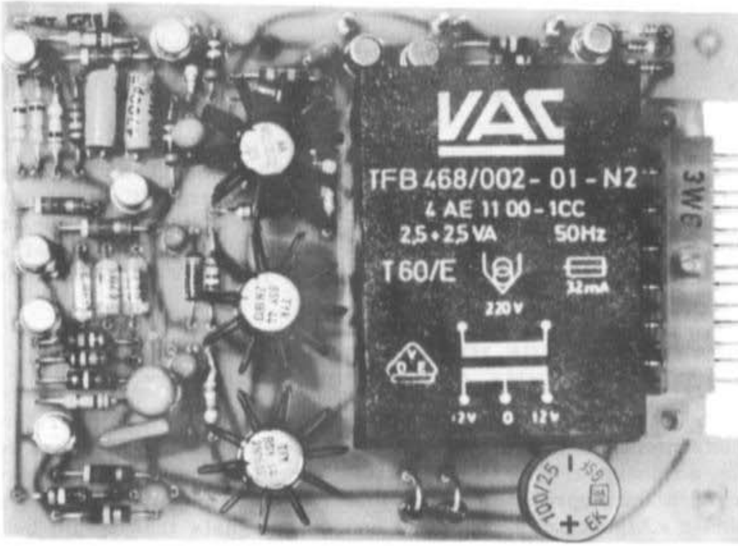

Fig. 5 : Photograph of a prototype module DK 1 PN 007

3. CONSTRUCTION

The circuit given in Figure 2 is accommodated on a single-coated PC-board whose dimensions are 98 mm x 72 mm. This means that this module has the same dimensions as a TEKO box size 3. Although this module is not enclosed in a TEKO box, it is of the same dimensions as the other modules of the 9 MHz receiver. Figure 4 gives the component locations of the PC-board which has been designated DK 1 PN 007. The area enclosed by the dashed lines should be cut out so that the 13 pin connector can be mounted. The connector end of the PC-board is then mounted on the chassis where the other half of the connector has been mounted, and can be fixed with two brackets. Figure 5 shows a photograph of the author's prototype.

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PHASE-LOCKED OSCILLATOR FOR 144 MHz

by J. Kestler, DK 1 OF

A phase-locked oscillator is to be described which has been designed as local oscillator module for 2 m transmitters and receivers. It is possible, for instance, for the phase-locked oscillator to be used in a multi-mode transceiver where it is fed from a multi-channel frequency synthesizer with either 10 kHz or 25 kHz channel-spacing in the FM mode, or from a VFO in all other modes.

1. COMPARISON OF THE METHODS FOR OBTAINING THE REQUIRED LOCAL OSCILLATOR FREQUENCY

It has been found that it is far superior to use a single-conversion principle in the construction of amateur SSB equipment for 2 m. This has led to considerable advantages over the earlier method of using a shortwave station together with the appropriate transverters. It is possible in this manner to easily obtain a higher suppression of spurious signals than would be case with double or triple conversion since only one single oscillator is used. For this reason, far less unwanted conversion products can be formed from the fundamental wave and harmonics of the signal, and local oscillator frequencies. The large-signal capabilities of the receiver are also improved since the main selectivity of the unit (crystal filter) is placed directly after the single mixer used.

Since the intermediate frequency of a single-conversion superhet is fixed (crystal filter), but the operating frequency is to be variable within the 2 m band, it is necessary for the local oscillator frequency to be also variable. The required local oscillator signal can be obtained using one of the three principles described below:

The variable frequency oscillator (VFO) oscillates at a half, quarter or a sixth of the required local oscillator frequency and is subsequently multiplied (as in the Braun SE 600).

The VFO frequency is mixed with a crystal-controlled oscillator signal (e.g. DJ 5 HD, DC 6 HL).

The required local oscillator frequency is obtained using a phase-locked system where the oscillator operates at the required frequency.

It is also possible for combinations of the above methods to be used, for instance, as in the 80-channel synthesizer described by DK 1 OF.

The first of the above principles provides an excellent suppression of spurious signals but places great demands on the frequency stability of the VFO, whose drift will be multiplied together with the signal. With the second version, there is a danger of producing unwanted conversion products. This means that a considerable amount of filtering is required in order to ensure that only the required signal is obtained from the spectrum produced during each conversion process.

The above problems are avoided if the phase-locked oscillator principle is used. The frequency stability of a phase-locked oscillator is only determined by that of the VFO which can operate at a relatively low frequency. A high rejection of spurious signals can be achieved when the oscillator is carefully constructed.

2. OPERATION OF A PHASE-LOCKED OSCILLATOR

Figure 1 gives the block diagram of a phase-locked oscillator. This circuit uses a phase-locked loop (PLL). A voltage-controlled oscillator (VCO) operates at the required frequency f (required local oscillator frequency), which means that a high suppression of spurious waves is guaranteed. The frequency stability of this oscillator is not of importance since it will be synchronized continuously with the aid of a varactor diode. It is only necessary for the drift to be kept low enough so that the VCO does not leave the hold range of the phase-control circuit. The signal generated by the VCO is fed to a mixer stage and converted with the aid of a crystal-controlled signal f_q to a considerably lower frequency. Due to the relatively low difference in frequency between the VCO and crystal frequency, this will cause no problems and a simple mixer can be used.

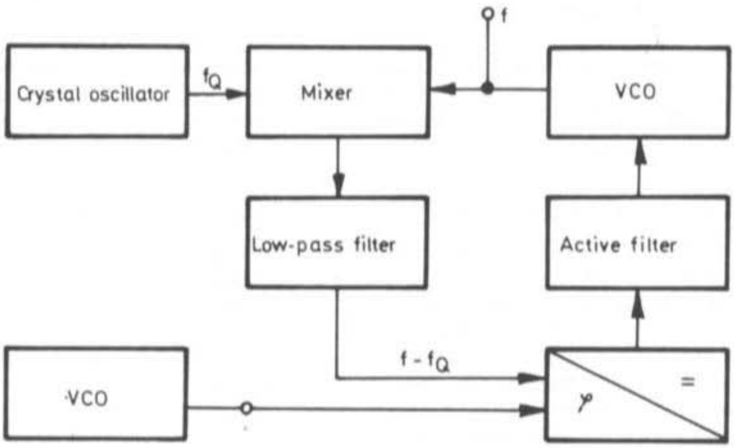

Fig. 1: Block diagram of a phase-locked oscillator

The frequency difference is then fed via a simple low-pass filter to the phase comparator where it is compared to the VFO signal. If both frequencies coincide, a DC-voltage will appear at the output of the comparator that is proportional to the phase difference between both input signals. This DC-voltage is fed via an active filter, and will control the VCO with the aid of a varactor diode so that the phase angle between the VFO and frequency difference remains constant. This results in a complete frequency control circuit. In order to ensure that the control circuit remains stable (e.g. that no oscillations of the control voltage occur) certain demands are placed on the frequency response of the whole system. These demands are satisfied by the active filter on which the dynamic behaviour of the phase-locked oscillator is mainly dependent. The theory and mathematic relationships of PLL - circuits are not to be explained here. However, further details were given in references (1) to (4).

As has already been mentioned, the phase angle between the VFO and the difference frequency remains constant in the phase-locked mode. However, since the frequency difference of two oscillations is generally equal to the time derivative of the phase-difference, it is necessary for the two frequencies fed to the phase comparator to be exactly equal. This means that the system will not exhibit any residual control difference. This is one of the great advantages of the PLL-technology when compared to frequency control circuits using frequency discriminators.

There is usually a very large deviation between the required and actual frequency of the VCO directly after switching on that is not within the hold range of the phase-control circuit. This means that it is necessary for some means to be provided so that the VCO is brought into the vicinity of the required frequency. Up till now, oscillators were used that were swept over the frequency range of interest until they locked in. Phase-discriminators are now available that operate as frequency discriminators when large deviations of the frequency are present. They are automatically switched to the phase-comparator mode as soon as the frequency difference is sufficiently small. In this manner, it is possible for the locking range of the control circuit to be dimensioned so that it is equal to the hold range.

Such phase-detectors are now available as integrated circuits. The most well-known types are the Motorola type MC 4044 (for TTL-signals) and RCA type CD 4046 (MOS technology). The appropriate application sheets give durther details as to the operation and applications of these circuits.

3. DESIGN OF A PHASE-LOCKED OSCILLATOR FOR THE 2 m BAND

It has been found that it is favourable for practical operation when the 2 m band from 144 MHz to 148 MHz is divided into eight bands of approximately 500 kHz each. It is true that it is easier to tune over the band in one range, but since simplex operation is usually made in relatively narrow portions of the band, it is felt that the frequency accuracy and bandspread is better when several ranges are used. Furthermore, it would be possible for the lower part of the band to be covered with the aid of the VFO, whereas the upper portion could be covered with a multi-channel synthesizer.

3.1. SELECTION OF THE FREQUENCIES FOR VFO AND CRYSTAL-CONTROLLED OSCILLATOR

In order to obtain a sufficiently stable VFO without any extreme measures being required, it is advisable for it to oscillate in the order of 5 MHz. This frequency range has been used for previously published concepts such as DC 6 HL. Any harmonics (28th harmonic) of the VFO frequency that fall into the frequency range of interest will be so weak that they will be below the noise level if the receiver is constructed carefully. Attention should be paid that the harmonics of the crystal-controlled oscillators do not fall into the receive or image-frequency range and that no harmonics of the VFO fall in the IF-range. The following frequency plan is recommended for an intermediate frequency of 9 MHz (all details in MHz):

Operating Frequency	Oscillator Range	Crystal Frequency	Crystal	VFO
144.0 - 144.5	135.0 - 135.5	130.0	65.00	5.0 - 5.5
144.5 - 145.0	135.5 - 136.0	130.5	65.25	5.0 - 5.5
145.0 - 145.5	136.0 - 136.5	131.0	65.50	5.0 - 5.5
145.5 - 146.0	136.5 - 137.0	131.5	65.75	5.0 - 5.5
146.0 - 146.5	137.0 - 137.5	132.0	66.00	5.0 - 5.5
146.5 - 147.0	137.5 - 138.0	132.5	66.25	5.0 - 5.5
147.0 - 147.5	138.0 - 138.5	133.0	66.50	5.0 - 5.5
147.5 - 148.0	138.5 - 139.0	133.5	66.75	5.0 - 5.5

If an intermediate frequency of 10.7 MHz is to be used, it is not possible for the VFO to be operated in the range of 5.0 to 5.5 MHz since the second harmonic of 10 MHz to 11 MHz will fall into this IF-range. In this case, a VFO frequency of 4.0 to 4.5 MHz should be used. This results in the following frequency plan:

Operating Frequency	Oscillator Range	Crystal Frequency	Crystal
144.0 - 144.5	133.3 - 133.8	129.3	64.65
144.5 - 145.0	133.8 - 134.3	129.8	64.90
145.0 - 145.5	134.3 - 134.8	130.3	65.15
145.5 - 146.0	134.8 - 135.3	130.8	65.40
146.0 - 146.5	135.3 - 135.8	131.3	65.65
146.5 - 147.0	135.8 - 136.3	132.3	66.15
147.0 - 147.5	136.3 - 136.8	132.3	66.15
147.5 - 148.0	136.8 - 137.3	132.8	66.40

Further details regarding the VFO are not to be given here since a sufficient number of detailed descriptions have already been published, e.g. in (4). Any variable oscillator can be used that is able to provide 100 mV into 50 Ω. It can be tuned with a varactor diode or variable capacitor.

A multi-channel synthesizer is to be described in one of the next editions of VHF COMMUNICATIONS that allows the range of 4.0 to 4.5 MHz to be covered in steps of either 10 kHz or 25 kHz. Such an oscillator can be used together with the phase-locked oscillator instead of the VFO. This will allow an 80-channel oscillator to be obtained with 25 kHz spacing, e.g. for Europe or a 400-channel oscillator in the case of 10 kHz spacing at 144 MHz to 148 MHz in the case of North America. The intermediate frequency is 10.7 MHz in both cases.

Of course, it would also be possible for a synthesizer to be provided that covers the range of 5.0 to 5.5 MHZ in 5 kHz steps. However, this would only be needed for non-FM operation and since no channel spacings have been agreed for these modes, it is still too early to commence such operation.

3.2. CIRCUIT DESCRIPTION

The circuit diagram of the phase-locked oscillator is given in Figure 2. The circuit is designed for use with four crystal-controlled oscillators and the required oscillator is selected by grounding points Pt 101 to Pt 104 with the aid of a switch. The common drain circuit can be tuned with the varactor diode

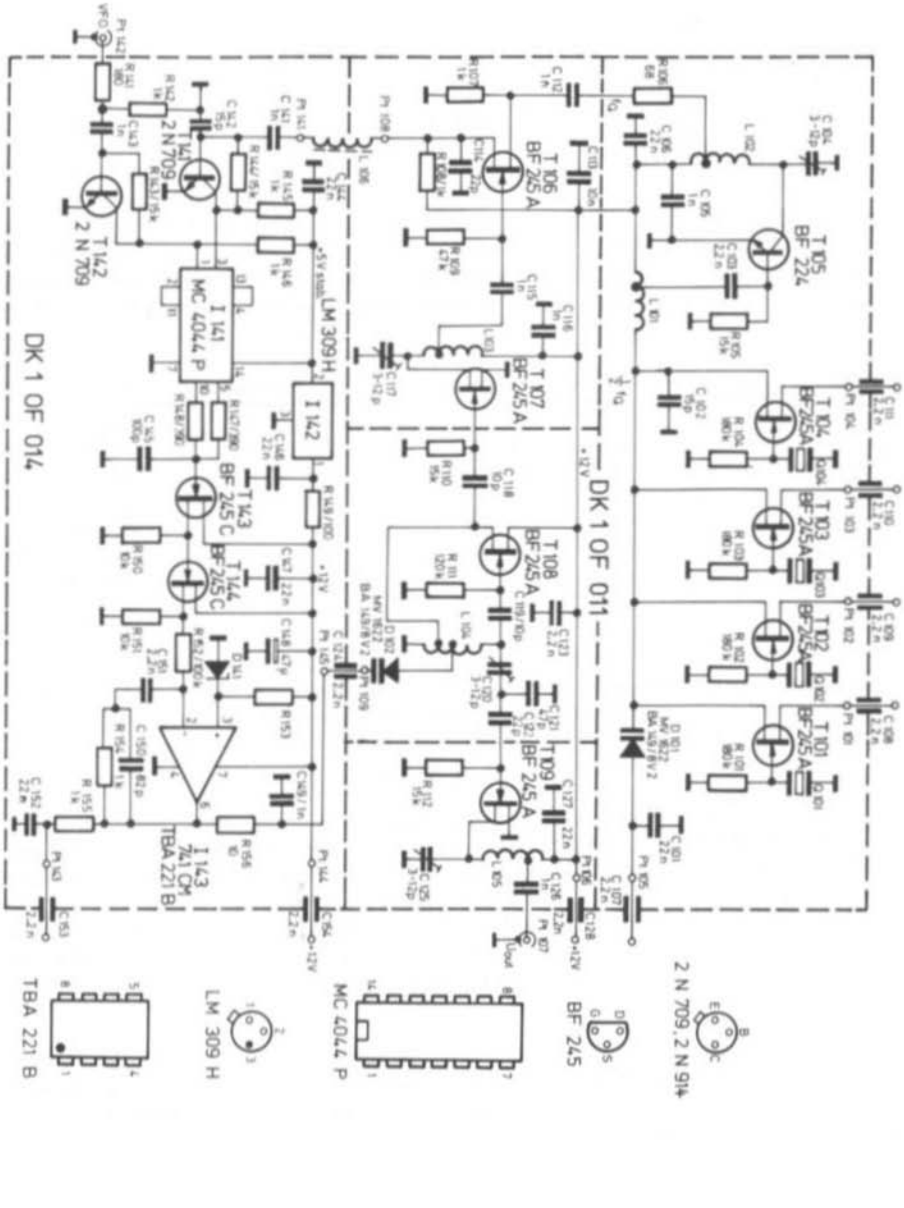

Fig. 2: Circuit diagram of a phase-locked oscillator for 2 m

D 101 and the required tuning voltage (Pt 105) is also switched with the same switch as before. The subsequent frequency doubler stage equipped with transistor T 105 is connected to the drain circuit via an inductive voltage divider.

The difference between the crystal-controlled signal and the VCO frequency is formed in the mixer T 106. The VCO comprises transistor T 108 and the resonant circuit L 104 which is tuned with the varactor diode D 102. The VCO frequency is fed via the buffer T 107 to the mixer and via the buffer T 109 to the output of the phase-locked oscillator.

The difference frequency signal generated in mixer T 106 is fed via a π-type lowpass filter to the pulse-shaper equipped with transistor T 141. The output capacitance of the π-link is partially formed by the input capacitance of transistor T 141, which then feeds the phase detector I 141. The reference frequency from the VFO is fed via connection Pt 142 and an identical pulse-shaper stage (T 142) to the reference input of the phase discriminator. The required operating voltage of 5 V is provided by an integrated voltage stabilizer I 142 (TO 5 case with three connections).

The output signal of the phase-comparator is passed via a simple RC-lowpass filter to the two-stage source-follower comprising transistors T 143 and T 144. These two stages increase the DC-voltage level by approximately 5 V as the output voltage of I 141 is only in the range of 0 to 2 V. This allows the subsequent operational amplifier I 143 to operate in its most optimum working range: Input voltage approx. equal to half the operating voltage. The non-inverting input of I 143 is connected to a constant voltage provided with the aid of zener diode D 141. The required zener voltage depends on the pinch-off voltage of transistors T 143 and T 144, but is most certainly within the range between 3.9 V and 8.2 V. Further details regarding this are given in Section 6.

The RC-network comprising R 155, C 150, and C 151 provided in the feedback link of the analog amplifier I 143 determines the frequency response of the active filter; the given values were found experimentally. The short-time phase stability of the phase-locked oscillator is mainly dependent on the ratio between the phase-comparator frequency (VFO frequency) and cut-off frequency of the active low-pass filter. This ratio amounts to approximately 10 000 in the given circuit which means that only a very low sideband noise is to be expected. The cut-off frequency of the lowpass filter also determines the lock-in time (until synchronization takes place) with a given frequency jump (e.g. when switching the crystals). In the case of a frequency jump of 1 MHz, the time taken until synchronization occurs amounts to approximately 10 ms.

The output voltage of the active filter controls the frequency of the VCO with the aid of the varactor diode D 102. This voltage is available externally at connection Pt 143 for alignment and monitoring purposes.

4. COMPONENT DETAILS

T 101 - T 104:	BF 245 A (TI), W 245 A (Siliconix) or similar FET
T 105:	BF 224
T 106 - T 109:	BF 245 A or similar
T 141, T 142:	2 N 709, 2 N 914, BSY 18 or similar fast switching transistors
T 143, T 144:	BF 245 C or similar

D 101, D 102: MV 1622 (Motorola), BA 149/8V2 (AEG-Telefunken)
BA 110 or similar (approx. 9 pF/2 V)

D 141: Zener diode of the BZY 85 or BZX 55 series or similar, value see text.

I 141: MC 4044 P (Motorola)

I 142: LM 309 H (National Semiconductor) or SG 309 H (Silicon General)

I 143: 741 CN (various manufacturers) or TBA 221 B (Siemens)

Q 101 - Q 104: HC-6/U crystals soldered directly to the board, frequency: see text.

C 104, C 117, C 120, C 125: Ceramic disc trimmers 3 - 12 pF, 10 mm dia., or 3 - 13 pF, 7 mm dia., or plastic foil trimmers 2 - 13 pF (7 mm dia., yellow)

9 feed-through capacitors of 2.2 nF or more.

Capacitors below 100 pF: Ceramic tubular types, ceramic disc types, spacing 7.5 mm.

Spacing 5 mm for all other capacitors.

All coils are wound with a right-hand thread on a 6 mm former, and are self-supporting. They are wound from 1 mm dia. (18 AWG) silver-plated copper wire, turns spaced approximately 1 mm.

L 101: 9 turns, coil tap at 2.75 turns from the cold end
L 102: 7 turns, coil tap at 0.75 turns from the cold end
L 103: 7 turns, coil tap at 6.25 turns from the cold end
L 104: 8 turns, 1st tap at 1.75 turns, 2nd tap at 3.25 turns from the cold end
L 105: 8 turns, coil tap at 3.25 turns from the cold end
L 106: Ferrite wideband choke with 6 hole core (Philips)

A spacing of 15 mm is available for all resistors.

5. CONSTRUCTION

The single-coated PC-boards DK 1 OF 011 and 014 have been developed for accommodating the phase-locked oscillator. Board DK 1 OF 011 accommodates the four crystal-controlled oscillators and frequency doubler, the mixer VCO and both buffers (Fig. 3). The two pulse shapers, the phase detector and the active filter are accommodated on PC-board DK 1 OF 014. The component locations are given in Figure 4. The photograph given in Figure 5 shows the author's prototype which uses slightly different PC-boards. The PC-boards are directly soldered to the screening panels. The height of the screening panels around the edges of the PC-boards is approximately 30 mm. Brass plates of 1 mm thickness were used for the prototype. This ensures a sufficiently stable construction. After completing the alignment, it is possible to solder double-coated PC-board material into place as upper and lower cover of the module. All DC-voltage connections should be provided with feedthrough capacitors of 2 nF or greater. The VFO input (Pt 142) and output of the phase-locked oscillator (Pt 107) are provided with miniature coaxial sockets or low-capacitive feedthroughs. The ferrite choke L 106 is mounted half-way up the intermediate panel between Pt 108 and Pt 141. The same is also valid for feedthrough capacitor C 124.

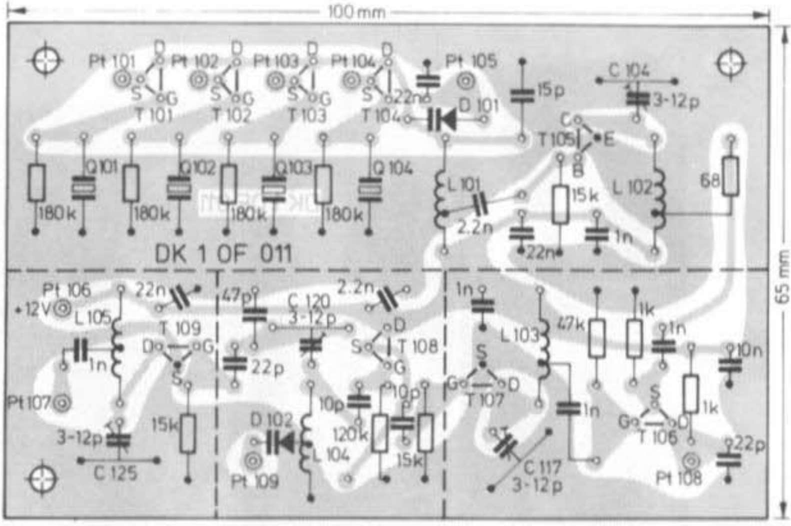

Fig. 3 : Component locations and PC-board DK 1 OF 011

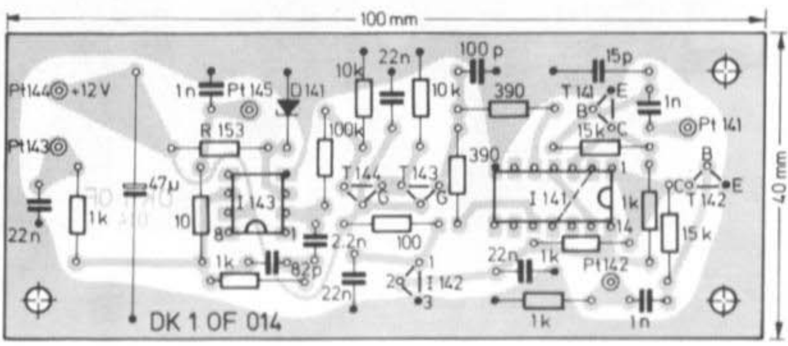

Fig. 4 : Component locations and PC-board DK 1 OF 014

6. ALIGNMENT OF THE PHASE-LOCKED OSCILLATOR

The zener diode D 141 and its voltage dropper resistor R 153 should still not be installed at this time. The interconnection between R 156 and the output of I 143 is broken. Varactor diode D 102 is provided with a variable bias voltage of 0 - 10 V (positive ground) via R 156. This is followed by connecting the operating voltage (+ 12 V) to Pt 106 and measuring the frequency of the VCO at the output of the module (Pt 107) with the aid of a VHF frequency counter or calibrated receiver. The required oscillator frequency range (See Sect. 3.1.) can be adjusted with the aid of trimmer capacitor C 120. The tuning voltage should then be in the range of 5 - 8 V. Finally, the output circuit comprising C 125 is aligned at the centre of the band for max. output signal. The resonant circuit comprising L 103/C 117 is aligned by measuring the voltage drop across the impedance of the mixer transistor (R 108). Trimmer capacitor C 117 should be adjusted so that an increase of the drain current of T 106 by approx. 1 - 2% can be observed in the centre of the VCO range.

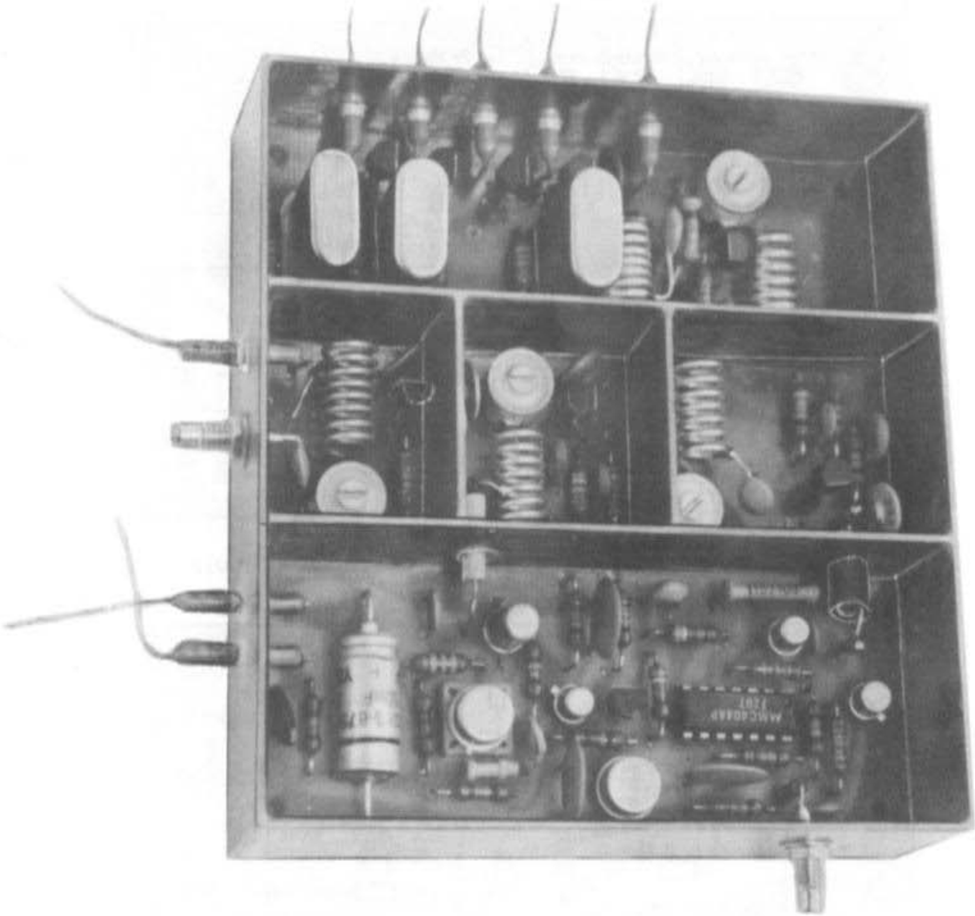

Fig. 5 : Author's prototype of the phase-locked oscillator comprising modules DK 1 OF 011 and 014

This is followed by aligning the frequencies of the crystal oscillators. A variable bias voltage of 0 to +12 V is fed via a dropper resistor of 1 kΩ to diode D 101, and one of the points Pt 101 to Pt 104 is grounded via a mA-meter. The bias voltage at Pt 105 is now adjusted for maximum current reading; this is carried out one after the other for all four oscillators. The individual voltage values at connection Pt 105 are noted; Four switchable voltage dividers or adjustment potentiometers provide these voltages in the completed unit. The values should be in the order of +5 V and +10 V. If necessary, L 101 or C 102 can be slightly altered until the given range is obtained.

The output circuit of the frequency doubler is aligned with the aid of trimmer C 104. The criterion is the RF voltage at the coil tap of L 102. It can be measured with a simple test circuit comprising a germanium diode, disc capacitor and μA-meter.

The VFO is now connected and tuned approximately to the centre of its range. Pt 144 is then connected to +12 V and one of the four crystal-controlled oscillators brought into operation. The voltage at the source of T 144 should be measured with a voltmeter having a range of +10 V. The characteristic given in Figure 6 should be present on varying the bias voltage of diode D 102. The voltage value at which the jump takes place can be shifted by switching the crystal oscillators and by varying the VFO frequency. The two voltages U_1 and U_2 are determined and the arithmetical mean value U_M is formed:

$$U_M = \frac{U_1 + U_2}{2}$$

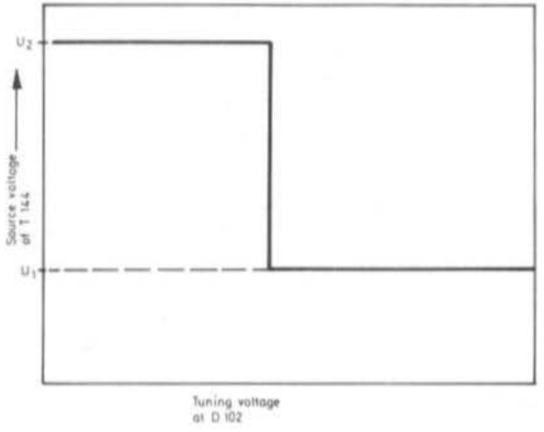

Fig. 6 :
Voltage at the source of T 144

The difference between these two voltages (U_2 - U_1) should amount to approx. 2 V. The zener diode D 141 should stabilize a voltage that is approximately equal to U_M ($\pm$ 0.5 V are permissible). It is therefore necessary for a suitable diode to be selected; the fine alignment of this voltage can be carried out with dropper resistor R 154 which should be in the order of 0.5 to 10 kΩ.

Finally, R 157 is reconnected to the output of I 143 and the tuning voltage present at point Pt 143 should amount to approximately 5 V at the lower end of the band (crystal 1 active, VFO to lowest frequency) and approximately 8 V at the upper limit (crystal 4, highest VFO frequency).

It is advisable for a meter to be provided on the front panel that indicates the voltage present at Pt 143 so that it is possible to monitor the synchronization. This is to avoid transmitting out-of-band should one of the critical components fail. Professional transceivers possess a logic circuit for this purpose that does not allow the transmitter to be switched on if the synchronization is not present.

7. PRACTICAL EXPERIENCE

Unfortunately, the author did not possess any precision measuring equipment in order to exactly measure the sideband noise and shortterm drift (jitter). However, experiments were made with simple measuring equipment which is now to be described:

The output signal of the phase-locked oscillator was mixed with a crystal-controlled signal so that the frequency difference was audible when fed to an AF amplifier and loudspeaker. A perfectly clean heterodyne was audible down to very low frequencies of approximately 100 Hz. This shows that the residual frequency deviation is well below 10 Hz, and means that the short-term stability of the described oscillator is more than sufficient even for SSB and CW operation.

The tuning time-constant is most certainly small enough for all practical operation. It was not possible to leave synchronization even with very rapid jumps of the VFO tuning.

A further experiment allowed the speed of the control circuit to be seen more clearly. The VFO was frequency-modulated with the aid of a varactor diode and the output signal of the phase-locked oscillator was monitored on a FM receiver. The control circuit was able to follow this modulation which was reproduced perfectly with the exception of some limiting of the higher frequencies. This effect can, however, compensated for by preemphasis in the modulator. This process could therefore also be used for practical operation.

The described phase-locked oscillator forms the heart of the author's 2 m station. No spurious signals are heard within the band and even the strongest signals in the order of 100 dB above noise can only be heard at one position in the band.

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(2) SIGNETICS: Linear Phase-Locked Loops Applications Book
Signetics Corp. 1972

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RSGB VHF-UHF MANUAL

Unfortunately, the RSGB has informed us that this popular manual is now out of print. A reprint is not planned and this means that this VHF-UHF Manual will be no longer available until the new manual has been published.

NOTES and MODIFICATIONS

1. FREQUENCY COUNTER BOARD DL 8 TM 002

Many of our readers will have equipped this board with IC-sockets to facilitate exchange or selection of the IC's. However, there are some sockets whose contacts must be bent back on the lower side of the board and can cause the following type of short-circuit: Such sockets possess wider metal strips before the pins enter the socket. When the IC-socket is depressed onto the PC-board during the bending back or soldering process, these wider pieces could cause short circuits between adjacent conductor lanes. Since this short-circuit will not always be present, it will be difficult to find and different types of errors could be indicated. This fault can be avoided either by using other types of IC-sockets or by leaving a spacing of 1 mm when soldering in the IC-sockets.

If such a fault is present, it is necessary for the IC-sockets to be removed by placing a screwdriver between board and plastic part of the IC-socket and levering it away from the board.The contacts can then be unsoldered individually until either the whole IC-socket is free or the require spacing is present. The holes can then be drilled through with a 0.75 mm drill, after which all filings should be removed completely. If the through contacts have been damaged, the component in question should be soldered on the upper and lower sides of the boards.

2. COUNTER PREAMPLIFIER DL 8 TM 003

It was mentioned in the description of this preamplifier in Edition 2/73 of VHF COMMUNICATIONS that a switching transistor must be used for the last transistor (T 308) of the AF-amplifier in order to ensure that the pulse rise and fall times were short enough. If this is still not sufficient, a resistor of approximately 3.3 kΩ can be connected between the base of T 308 and ground. This increases the upper frequency limit considerably.

3. MODIFICATION POSSIBILITIES FOR FREQUENCY COUNTER AND 250 MHz PRESCALER

A block diagram of a complete frequency counter system was given on page 159 of VHF COMMUNICATIONS, Edition 3/73 in the description of the DJ 6 PI 001 250 MHz prescaler. It should be noted that the relay RH-12 is a polarized type whose connection 3 must go to the cathode of the diode and connection 5 to the switch.

The counter sometimes indicates a "1" in the unit position when no input signal is present. This is caused by flip-flop I 302 (SN 74 S112) which can have either a high or low level at the output after receiving an input pulse. If the counting process is ceased with a high level at this position, a "1" indication will remain in the unit position. This minor error can be avoided by one of the two following methods:

a) Only the AF-portion of the preamplifier DL 8 TM 003 is used without I 301 and I 302. All frequencies over 1 MHz will now be measured via the 250 MHz prescaler DJ 6 PI 001. In this case the resolution will be 10 Hz instead of 1 Hz. This is made by connecting the output of the AF-amplifier (pin 1 of I 301) and the output of the prescaler (Pt 2) via the relay to Pt 208.

b) Only I 302 (SN 74 S 112) is removed and the input directly connected to the output of this IC. In this manner all three inputs are still available. The HF-input, however, only operates in the frequency range upto 50 or 60 MHz. The full resolution is available down to 1 Hz. If this method is preferred, another suitable preamplifier with high-impedance input will be described in one of the next editions of VHF COMMUNICATIONS.

4. TEKO VFO MODULE DJ 4BG 012

In the component location plan (Fig. 5, page 245, Edition 4/73) of this module, the resistors R 6 and R 9 have been exchanged. The 10 kΩ resistor should be placed below T 3 and the 1.5 kΩ resistor to the left directly over the PC-board's designation DJ 4 BG 012. The output voltage is distorted when these two resistors are inserted in the incorrect positions. The tuning diodes D 3 and D 4 must be connected in their blocked direction. They are shown in Figure 3 for a positive, and in Figure 6 for a negative tuning voltage.

MATERIAL PRICE LIST OF EQUIPMENT
described in Edition 2/74 of VHF COMMUNICATIONS

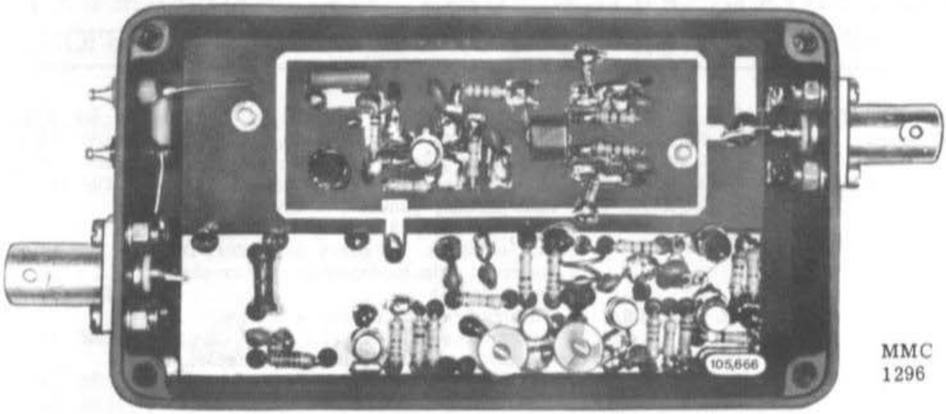
DL 9 FX 006/007	2-m-DF-HUNT RECEIVER		Ed. 2/74
PC-board	DL 9 FX 006	(double coated)	DM 16.--
PC-board	DL 9 FX 007	(with printed plan)	DM 8,--
Semiconductors	DL 9 FX 006/7	(3 IC's, 20 transistors, 16 diodes)	DM 132.--
Minikit 1	DL 9 FX 006/7	(1 case already punched, 1 BNC-connector, 1 earphone jack, 3 sliding switches, 4 push-button switches, 1 battery connector, 10 knobs, 1 meter, 1 min. loudspeaker, silver-plated wire, 2 drills)	DM 98.--
Minikit 2	DL 9 FX 006/7	(3 coilformers with cores, 4 chokes, 2 trimmer caps.)	DM 19.--
Minikit 3	DL 9 FX 006/7	(24 ceramic caps., 3 plastic foil caps., 11 tantalium or aluminium caps.)	DM 28.--
Minikit 4	DL 9 FX 006/7	(54 resistors, 3 trimmers, 10 potentiometers)	DM 64.--
Crystals	XF 901, XF 902 compl.		DM 40.--
Complete kit	DL 9 FX 006/7	with above components	DM 405.--

DJ 5 XA 002	STRIPLINE BALUN FOR 24 cm ANTENNAS		
PC-boards	DJ 5 XA 002 a+b	(double coated), pair	DM 16.--

DC 8 NR	2-m/70-cm LINEAR TRANSVERTER		Ed. 2/74
PC-board	DC 8 NR 001	Receive mixer (with plan)	DM 9.--
PC-board	DC 8 NR 002	288 MHz amplifier (with plan)	DM 9.--
PC-board	DC 8 NR 003	Local oscillator module (with plan)	DM 10.50
PC-board	DC 8 NR 004	First transmit mixer (with plan)	DM 9.--
PC-board	DC 8 NR 005	Second transmit mixer (with plan)	DM 9.--
Semiconductors	DC 8 NR	(1 IC, 15 transistors)	DM 94.--
Tube set	DC 8 NR	(1 EC 8020, 1 EC 8010)	DM 95.--
Minikit 1	DC 8 NR	(1 miniature relay, 6 coilformers, 2 cores, 3 ferrite beads, 1 ferrite choke, 2 UHF-tube sockets)	DM 19.50
Minikit 2	DC 8 NR	(10 tubular trimmers, 17 trimmers, 3 ceramic disc. caps.)	DM 75.--
Crystal	96.000 MHz	HC-25/U	DM 34.--
Kit	DC 8 NR	with above parts	DM 358.--
PC-board set	DC 8 NR 001 to 005		DM 45.--

DK 1 PN 007	INTEGRATED RECEIVER SYSTEM POWER SUPPLY, AF-LOWPASS FILTER, S-METER STAGES		Ed. 2/74
PC-board	DK 1 PN 007	(with printed plan)	DM 11.50
Semiconductors	DK 1 PN 007	(1 IC, 11 transistors, 7 diodes, 1 cooling fins)	DM 48.50
Minikit 1	DK 1 PN 007	(1 aluminium and 7 tantalium electrolytics, 7 styroflex and 1 ceramic capacitors)	DM 14.--
Minikit 2	DK 1 PN 007	(30 carbon resistors)	DM 10.--
Transformer	DK 1 PN 007		DM 23.--
13-pole connectors			DM 9.50
Kit	DK 1 PN 007	(with above parts)	DM 115.--

DK 1 OF 011/014	PHASE-LOCKED OSCILLATOR		Ed. 2/74
PC-board	DK 1 OF 011	(with printed plan)	DM 10.--
Semiconductors	DK 1 OF 011	(9 transistors, 2 diodes)	DM 42.--
Minikit	DK 1 OF 011	(4 trimmer caps., 1 ferrite choke, 11 ceramic caps., 7 feedthrough caps.)	DM 20.--
Crystals	HC-6/U	(65.0 MHz, 65.25 MHz, 65.5 MHz, 65.75 MHz)	DM 80.--
Kit	DK 1 OF 011	(with above parts)	DM 150.--
PC-board	DK 1 OF 014	(with printed plan)	DM 9.--
Semiconductors	DK 1 OF 014	(2 IC's, 5 transistors)	DM 60.--
Minikit	DK 1 OF 014	(2 feedthrough caps., 7 ceramic caps., 1 electrolytic)	DM 8.--
Kit	DK 1 OF 014	(with above parts)	DM 75.--



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Overall gain: typ. 30 dB
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Type TA   Type TA-S   Type TA 4

Type U 3   Type U 4   Type U 5

Model	Type	Frequency	Gain	Weight	Features
TA	5/8 λ	144-175 MHz	3 dB	275 g	Glass-fibre whip
TA-S	5/8 λ	144-175 MHz	3 dB	275 g	Glass-fibre with 5 m cable
TA 4	1/4 λ	144-175 MHz	0 dB	130 g	Stainless steel (PH 17-7)
U 3	5/8 λ	400-470 MHz	3 dB	100 g	Silver-plated, epoxy coated
U 4	Colinear	420-470 MHz	4 dB	150 g	Stacked λ/4 and 5/8 λ
U 5	Colinear	420-470 MHz	5 dB	175 g	Stacked 5/8 λ and 5/8 λ

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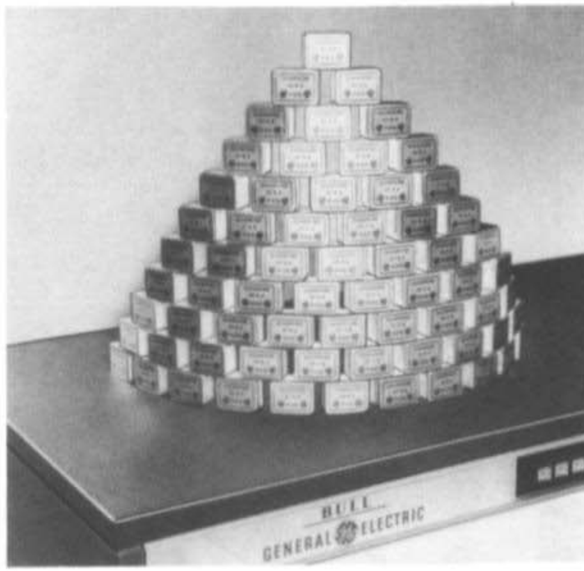

Filter Type		XF-9A	XF-9B	XF-9C	XF-9D	XF-9E	XF-9M
Application		SSB-Transmit.	SSB	AM	AM	FM	CW
Number of Filter Crystals		5	8	8	8	8	4
Bandwidth (6dB down)		2.5 kHz	2.4 kHz	3.75 kHz	5.0 kHz	12.0 kHz	0.5 kHz
Passband Ripple		< 1 dB	< 2 dB	< 2 dB	< 2 dB	< 2 dB	< 1 dB
Insertion Loss		< 3 dB	< 3.5 dB	< 3.5 dB	< 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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