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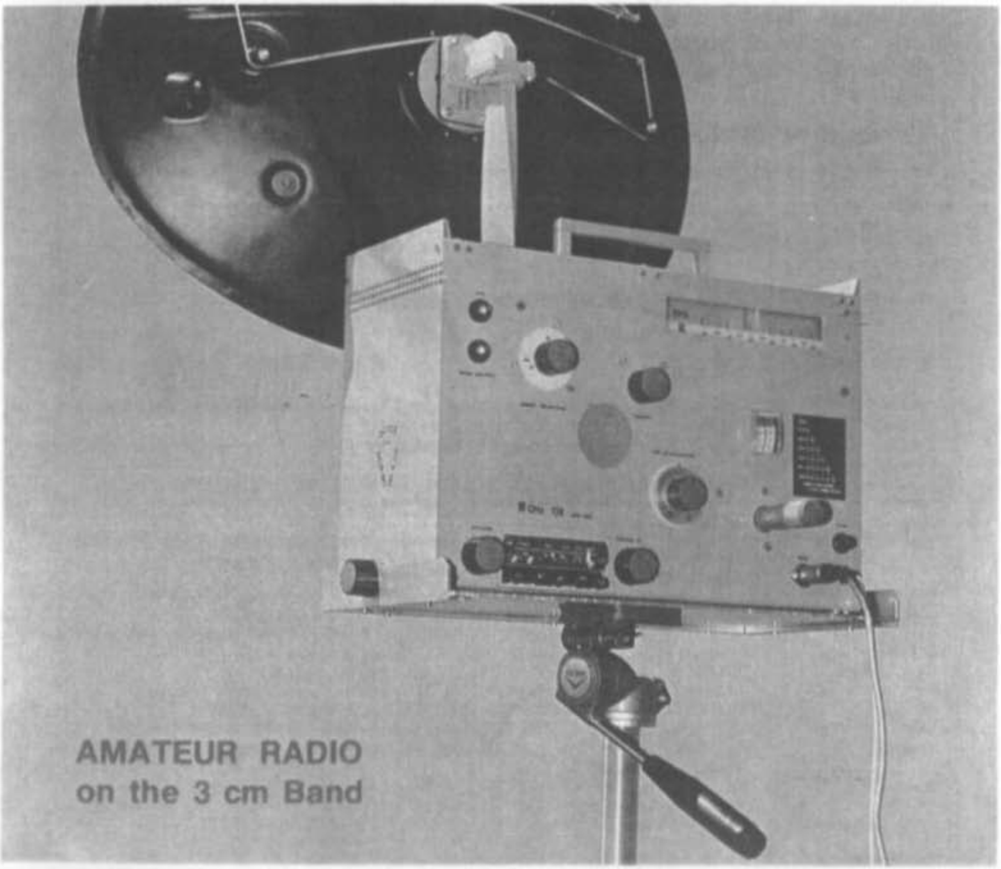
VHF COMMUNICATIONS

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

VOLUME No. 9 SUMMER 2/1977 DM 4.50



VHF COMMUNICATIONS

Published by:
Verlag UKW-BERICHTE · Hans J. Dohlus oHG · Jahnstraße 14 · D-8523 BAIERSDORF · Fed. Rep. of Germany · Telephones (0 91 33) 855, 856.

Publishers:
T. Bittan, H. Dohlus.

Editors:
Terry D. Bittan, G 3 JVQ / DJ 0 BQ, responsible for the text
Robert E. Lentz, DL 3 WR, responsible for the technical contents

Advertising manager:
T. Bittan.

VHF COMMUNICATIONS,
the international edition of the German publication UKW-BERICHTE, is a quarterly amateur radio magazine especially catering for the VHF/UHF/SHF technology. It is published in Spring, Summer, Autumn, and Winter. The subscription price is DM 16.00 or national equivalent per year. Individual copies are available at DM 4.50, or equivalent, each. Subscriptions, orders of individual copies, purchase of P.C. boards and advertised special components, advertisements and contributions to the magazine should be addressed to the national representative.

Printed in the Fed. Rep. of Germany by R. Reichenbach KG · Krelingstr.39 · 8500 Nuernberg

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VOLUME No. 9 SUMMER EDITION 2/1977

INTRODUCTION TO MICROWAVE TECHNIQUES AND A DESCRIPTION OF A 10 GHz TRANSCEIVER

by B. Heubusch, DC 5 CX · Dr.Ing. A.Hock, DC 0 MT · H. Knauf, DC 5 CY

Dr. Dain Evans's lecture on the Munich VHF-UHF Convention in 1976 led to a large amount of experimentation and activity on the 10 GHz band in Southern Germany. Microwave technology is extremely interesting, and is relatively easy to handle. Difficult metal work is usually not required. This article is to give an introduction to waveguide principles in the form of a constructional description of our 10 GHz transceiver, which we hope will help other amateurs to become active on the 10 GHz band (10.0 to 10.5 GHz, corresponding to approx. 3 cm).

1. FUNDAMENTALS

1.1. Waveguides

Virtually only waveguides can be used for transmitting energy in the frequency range of 10 GHz, since coaxial lines possess very high attenuation, and bring difficulties in the interface between different types of feeder. For various reasons, only rectangular waveguides of type WR 90 (1) are suitable. The operation of a waveguide can be explained as follows:

A waveguide usually consists of a rectangular tube with metallic (conductive) walls; also round, elliptical, as well as other shapes are possible. **Figure 1** shows a rectangular waveguide and indicates the most important dimensions. In order to be able to understand what happens in a waveguide, let us assume that a metal pin or a small metal ball is placed in the centre of a waveguide aperture and a DC-voltage is connected between this and the metallic surface of the waveguide **(see Figure 2)**. Electrical field lines will now be present in the waveguide which, for physical reasons, always leave the first surface and arrive at the second surface vertically. The voltage is continuously reduced along such a field line, until it becomes zero at the wall of the cavity. The electrical field strength also becomes weaker and weaker in the direction of the waveguide axis, which is indicated by the lines of equal potential (in other words equal voltage, or equi-potential lines) given in Figure 2. If an AC-voltage is connected between the metal ball and waveguide surface instead of the DC-voltage, the electrical field will vary together with the frequency. According to the magnetic flux laws, magnetic field lines will be generated that are similar to that shown in **Figure 3.**

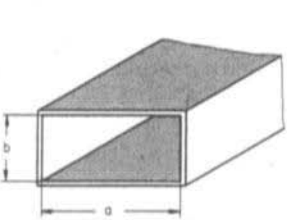

Fig. 1: Rectangular waveguide

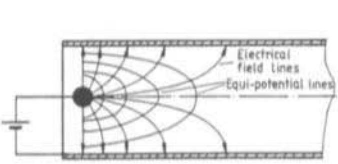

Fig. 2: Electrical field lines in the waveguide with DC-voltage

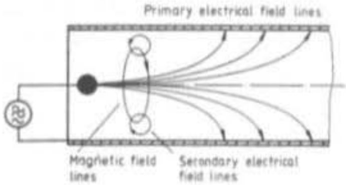

Fig. 3:
Electrical and magnetic fields in the waveguide with AC-voltage

The magnetic fields generated in this manner also alter with time, and additional electrical field lines will be generated according to the theory of induction, which means that the electrical field is concentrated in the axis of the waveguide, and is forced more and more into it. It will clearly be seen when comparing Figure 2 and 3. This effect becomes stronger and stronger on increasing the frequency, and the field will stretch further and further into the waveguide. However, it still represents a quasi-stationary field, that only converters reactive power, and is not able to transport energy. If the frequency is increased more and more, a point will finally be reached where this field reaches infinity. The frequency at this point is called the limit frequency of the waveguide. On increasing the frequency further, the field will be released from the energizing metal ball, and the wave will be passed along the waveguide and be able to transport energy. All types of waveguides have such a lower limit frequency, which is dependent on its shape and dimensions, and it is unable to transport waves below this frequency. It can therefore be assumed that a waveguide possesses something similar to a high-pass filter characteristic.

In practice, there are any number of waveforms in the waveguide, since there are very many types of waveguides, even more possibilities of input coupling, and also a large number of operating frequencies. In order to achieve clear operating conditions, it is therefore necessary for a matching waveguide to be used for each frequency range. For amateur radio applications, this will be a rectangular waveguide which is operated with the wave mode H_{10}. This is a very stable waveform which is shown in **Figure 4**.

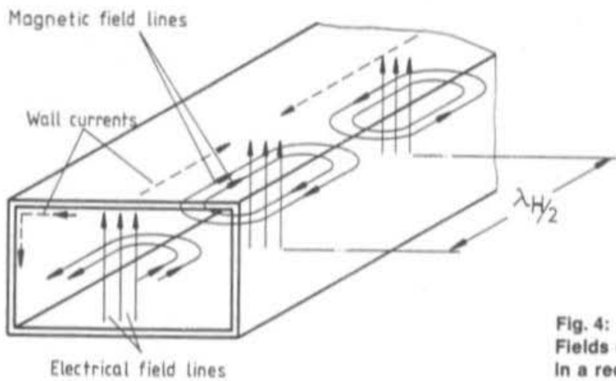

Fig. 4:
Fields of the H_{10}-wave in a rectangular waveguide

The electrical field lines run from one of the longest sides to the other, whereas the magnetic field lines orbit parallel to them. The limit wavelength is $\lambda = 2a$. In the range from $\lambda = 2a$ to approximately $\lambda = a$, the H_{10}-wave is the only waveform that can exist in this waveguide. This means that clear operating relationships exist.

If $a > \lambda$ is valid, it is possible for the H_{10}-wave to be converted to a different waveform at any interference in the waveguide (poor flange, filter, output coupling etc.). This will have an interference effect on the operation of the whole system. The H_{10}-wave in the waveguide has a wavelength of

$$\lambda_H = \frac{\lambda_0}{\sqrt{1-(\lambda_0/2a)^2}}$$

This means that it is greater than the wavelength in free space. The impedance of a waveguide is both frequency and waveform dependent. The following is valid for the H_{10}-wave :

$$Z_H = \frac{120\pi}{\sqrt{1-(\lambda_0/2a)^2}} = \frac{377}{\sqrt{1-(\lambda_0/2a)^2}} \quad \text{Ohm}$$

Since RF currents flow in the waveguide **(see Figure 4)**, the waveguide should be made from a flat, good conductive material. The individual waveguide segments are connected together with the aid of flanges, which must be clean and smooth, otherwise considerable losses will occur. It is important that they are flush-mounted for the same reasons. Such a flange is shown in **Figure 5**, and **Figure 6** gives the standard dimensions for the 10 GHz band.

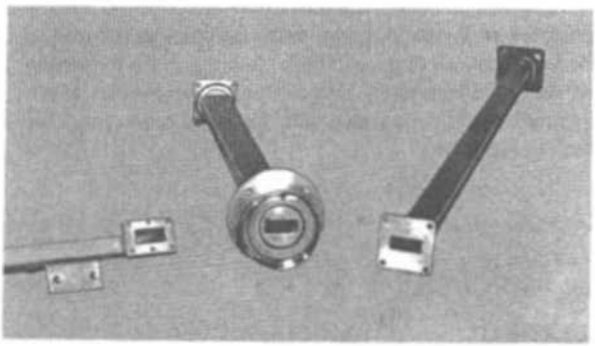

Fig. 5: Various types of waveguide with flange

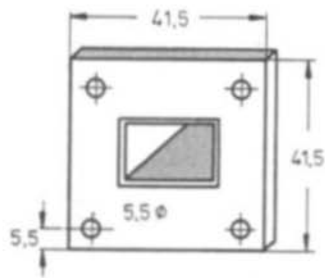

Fig. 6:
Standard flange for rectangular 10 GHz waveguides

1.2. CORNERS AND BENDS

No abrupt changes of direction should be made in the waveguide, since this would cause considerable reflections. For this reason, curved pieces of waveguide (so-called compensated corner pieces) are used. Attention must always be paid that the inner surface of the waveguide is smooth and that no residual solder, or cracks are present. **Figure 7** shows several types of corner pieces.

The energizing of a wave is achieved easily by exciting either the electrical wave capacitively, or the magnetic wave inductively. **Figure 8** shows such unterfaces. It is also possible to reverse the process so that energy can be taken from the waveguide in the same manner.

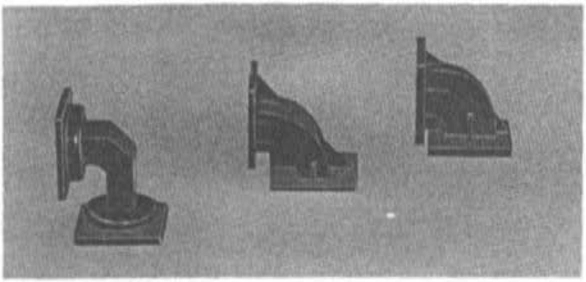

Fig. 7: Waveguide corner pieces

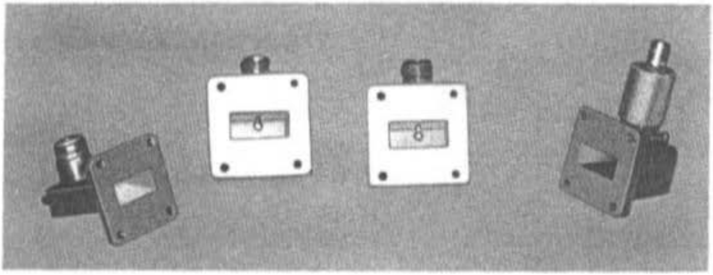

Fig. 8: Waveguide-coaxial interfaces for energizing the H_{10}-wave

Since the wave is propagated in both directions in the waveguide, it is necessary for a short circuit to be mounted at one end. This must be placed $\lambda_H/4$ from the energizing position so that the reflected wave at the short circuit returns at the correct phase to the energizing point. This is very important with respect to finding the most favorable power input or output coupling and represents a critical position. Very often this is to be found in the form of a short-circuit piston in order to obtain an exact matching or termination. The exciting point itself should provide a well matched termination, for instance, to the impedance of the coaxial cable. The matching can be made by varying the thickness of the energizing wire, or the diameter and insertion depth of the capacitive plate, as well as possibly by adding a small piece of low-loss plastic tube.

1.3. Horn Radiator

Finally a word regarding the transition from waveguide to free space (e.g. antennas). Since the impedance of the waveguide and free space do not coincide, strong reflections would result if the end of the waveguide was simply left open. For this reason, a slow transition is provided by extending the walls of the waveguide, which then leads to the horn radiator. The more gradually the waveguide is extended and the length of the transition, the lower will be the interfering reflection of the transition of the wave to free space, and the higher will be the gain. **Figure 9** shows various horn radiators and **Table 1** gives the practical dimensions for horn radiators constructed as shown in **Figure 10**.

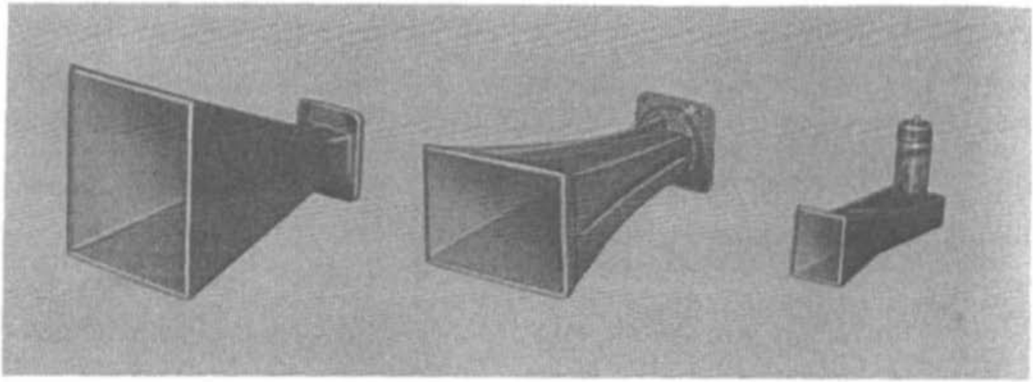

Fig. 9: Various types of horn radiators

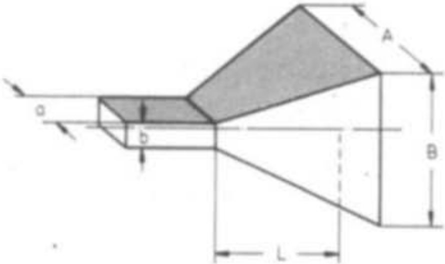

Fig. 10: Critical dimensions of horn radiators

Table 1:
Dimensions and attainable gain of horn radiators (f = 10.3 GHz)

Length of horn L cm	Width A cm	Width B cm	Gain dB
2.62	6.82	5.05	14
3.65	7.65	5.67	15
4.98	8.58	6.36	16
6.67	9.63	7.13	17
8.85	10.81	8.00	18
11.62	12.12	8.98	19
15.16	13.60	10.08	20
19.66	15.26	11.31	21
25.37	17.13	12.69	22
32.62	19.22	14.43	23
41.81	21.56	15.97	24
53.45	24.19	17.92	25

1.4. References to part 1

(1) R. Lentz: Designations of the microwave bands and waveguides
VHF COMMUNICATIONS 8, Edition 4/1976, Pages 232 - 233

(2) H. Meinke und F.W. Gundlach: Taschenbuch der Hochfrequenztechnik
Springer-Verlag Berlin 1968

This article is to be concluded in one of the next editions of VHF COMMUNICATIONS.

A COAXIAL-LINE POWER AMPLIFIER FOR 70 CM EQUIPPED WITH THE 4 CX 250 B

by W. Rahe, DC 8 NR

There are virtually only two families of tubes that allow higher outputs to be obtained on 70 cm inexpensively: The 2 C 39 and 4 X 150 / 4 CX 250 types.

Several 70 cm power amplifiers equipped with the tube 2 C 39 have been published in (1), and (2). This article is to describe a higher powered amplifier equipped with a 4 CX 250 B. A similar amplifier was described several years ago in (3/5). However, this amplifier possessed a number of disadvantages: the construction was rather difficult due to the use of many lathed parts, which is probably the reason why such power amplifiers have not been too popular in the past. In fact, a 4 CX 250 amplifier possesses with 14 to 16 dB a considerably higher gain than a 2 C 39 amplifier (10 to 11 dB). This means that approximately four times the output power will be provided by the 4 CX 250 amplifier for the same drive power. Since both tube types require forced-air cooling, it is only the power supply that is more extensive due to the higher voltages, larger transformer and the required screen grid voltage. The amount of mechanical construction is approximately the same for both types. The power amplifier shown in **Figure 1** can be constructed relatively easily; it possesses a number of special features which are to be described in more detail.

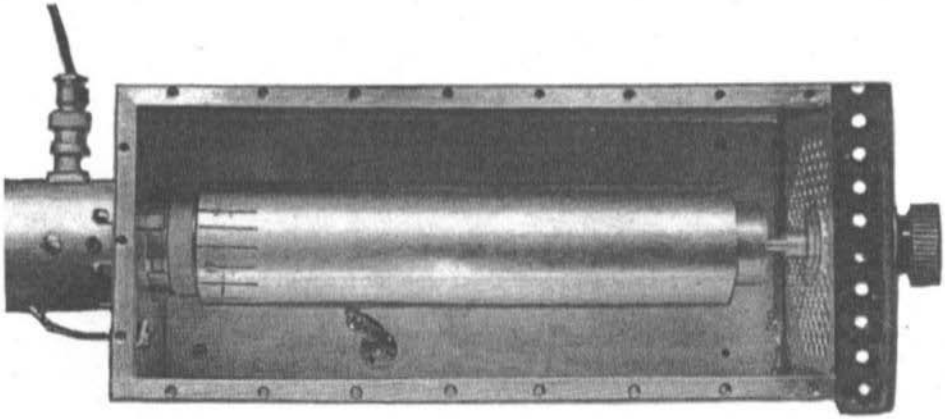

Fig. 1: A 70 cm power amplifier equipped with the tube 4 CX 250 B

1. SELECTION OF THE TUBE

There are several different types of the 4 X 150 A tubes according to the manufacturer. Normally, this tube is constructed in a metal-glass technology, but there are also versions with a ceramic base and with simple or double fins of the anode radiator. The socket connections are identical, and the main difference is in the heater voltage of 6 V or 26 V, and in the plate dissipation power of 150 W or 250 W.

In addition to this there is the type 4 X 150 G which possesses coaxial tube electrodes similar to the 2 C 39 in order to keep the connection inductances as low as possible.

Due to the metal-ceramic construction, a higher maximum plate temperature of 250°C is permissible for the tubes 4 CX 250 B/F/R (250 W plate dissipation), in contrast to 200°C for the glass types. This means that it is not necessary for them to be cooled so intensively. The limit values for the plate and screen grid voltage of 2000 V and 400 V respectively should, however, not be exceeded with either type. As can be seen in **Figure 2**, the gain of the 4 X 150 starts to fall at lower frequencies than with the 4 CX 250. **Figure 3** gives the maximum permissible input at the limit values, and the attainable output power as a function of frequency. A comparison of these tubes was given in (3). With the exception of the 4 CX 250 R, all types are available inexpensively on the surplus market.

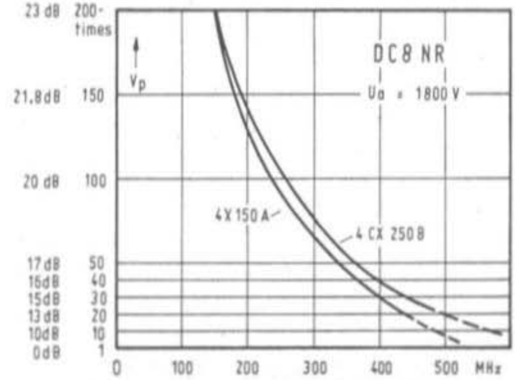

Fig. 2:
Power gain as a function of frequency.
Class AB$_1$, grounded cathode circuit

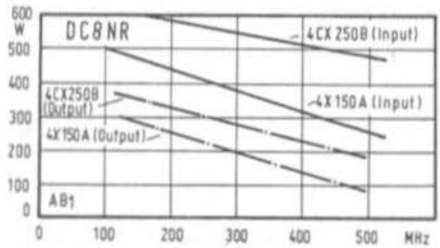

Fig. 3:
Input and output power as
a function of frequency

During the last few years, a whole series of more modern tubes has been developed in this power category, but their prices are usually too high for tnem to be used for amateur radio applications. Such a tube is the STC-tetrode 4 KC / 160 M with specifications similar to that of the 4 CX 250 B. This tube is contact cooled using an insulated, but good heat-conducting block of beryllium oxide mounted on a heat sink. This is also the case with the Eimac tube 8873. The two other tubes of this series 8874 and 8875 are forced-air cooled similar to the 4 CX 250 (6). The additional output capacitance of a contact cooled tube with beryllium oxide block is in the order of 6 to 10 pF. This must be taken into consideration on designing the anode circuit.

2. ELECTRICAL CONSTRUCTION

The tetrode tube operates in a grounded cathode circuit. The basic circuit diagram is given in **Figure 4**, as are the base connections of the tube. The 435 MHz drive power is fed via coaxial socket 1 and the input coupling L 1 to the λ/2 line circuit L 2. This is constructed in the form of a helical circuit. The choke Ch 1 feeds the negative bias voltage for the control grid (– 55 V for linear operation, or – 90 V for class C); this is made in the vicinity of the point of maximum current.

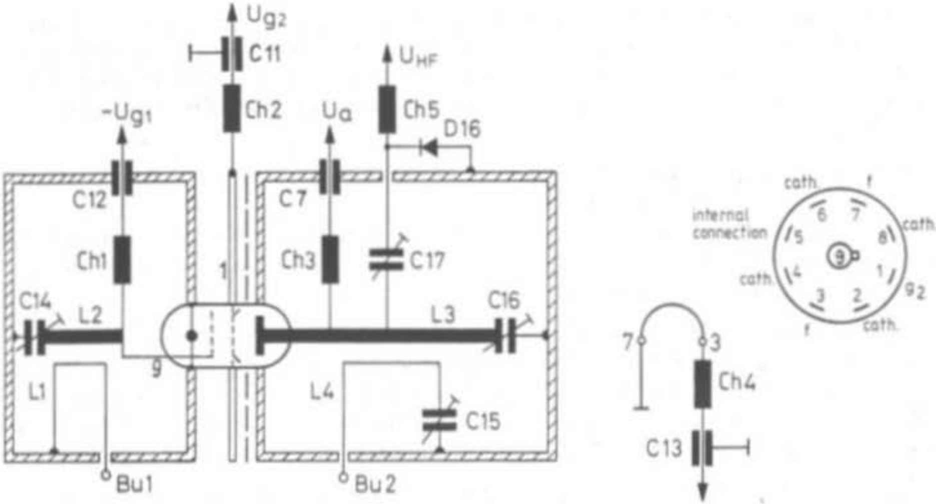

Fig. 4: Basic circuit of the 70 cm power amplifier

Due to the large capacitive shortening of the circuit (input capacitance 16.5 pF) this point is already within the tube. It is, of course, possible for the input coupling to be made capacitively via approximately 8 pF to the grid connection.

The screen grid is by-passed with effect to HF frequencies using the built-in by-pass capacitor C 17 provided in the tube socket. The required, stabilized screen grid voltage of 350 V required for class AB_1 linear operation is fed via choke Ch 2 to pin 1 of the socket.

The heater voltage of 6.0 V is fed to pin 3 of the tube socket via choke Ch 4. This voltage value may seem strange, since the heater voltage is usually reduced in the case of UHF operation in order to compensate for the unwanted heating effect caused by heating of the cathode by electrons not captured by the plate. In contrast to class C operation, this compensation is not required in the linear mode since this effect will be very small.

The tube is at the high voltage point of the plate circuit (L 3); the output coupling (L 4) and the plate voltage feed (Ch 3) are at the point of maximum current. Trimmer capacitor C 15 is provided to compensate for the reactive component of L 4. The coaxial anode circuit is also $\lambda/2$ long and is tuned to resonance with a home-made tubular trimmer (C 16) at the other end of the line from the tube (also at the point of maximum voltage). This type of construction saves the use of the by-pass capacitor that would be required with $\lambda/4$ circuits.

Due to the output capacitance of the tube (4.5 pF), trimmer C 16, and the inavoidable stray capacitances, the anode circuit is mechanically shorter than $\lambda/2$. Such an arrangement with a round inner conductor and a round, or square outer conductor completely surrounding the inner conductor is usually called a coaxial line circuit. The most favorable Q for the circuit, and thus the lowest loss will be provided at a ratio of D/d approx. 3.5. This diameter ratio corresponds to an impedance of 75 Ω. The higher the capacitive load of the circuit, the lower will be the most favorable impedance. The plate line circuit described here possesses an impedance of approximately 55 Ω.

The band width of coaxial line circuits decreases on increasing the line length in multiples of $\lambda/4$ if the impedance is kept constant. This means that the band width of a $\lambda/2$ circuit is only approximately half that of the band width of a $\lambda/4$ circuit. This means that power amplifiers equipped with $\lambda/2$ circuits may not be suitable for a wideband operation, such as for ATV.

3. MECHANICAL CONSTRUCTION

Figure 5 shows the compact, and uncomplicated construction of the power amplifier. The mechanical construction is extremely simple for a power amplifier of this class, since mainly already available tubular material is used.

The outer conductor of the plate coaxial line with the internal dimensions of 100 mm x 100 mm is firstly bent from 1 mm thick brass plate. Of course, it is also possible for this part to be soldered together from suitable metal plates. However, it is not recommended that a screwed construction is used at this point, since a poor electrical contact in the high-current paths would be extremely unfavorable. It is only the cover that is mounted in this manner.

This is followed by soldering the tube and metal plates into position; the plate with the 56 mm hole in the centre is for mounting the tube socket, and the other with the hole for the shaft of the capacitor C 16 and a large number of 6 mm dia. holes for exit of the cooling air. The other holes in the plate line circuit (cooling air input, socket mounting, cover mounting, plate voltage input, ceramic support for the inner conductor and output coupling) should be made as shown in Figure 5 and need not be described in detail. Since the output coupling is at low impedance (high current, low voltage) a spacing of 0.75 mm is sufficient for trimmer C 15.

The inner conductor of the coaxial line circuit comprises a 2 mm thick copper tube of 185 mm in length and 44 mm diameter. Brass tube can also be used, but is not so favorable due to its higher heat resistance. Aluminium tube should also be very suitable if it can be well clamped to the anode radiator (heat coefficient λ [W/grd x cm] : copper = 3.8; aluminium = 2.1; brass = 1.1).

Since the plate radiator of the 4 CX 250 B possesses a diameter of approximately 41.2 mm, it is necessary for the copper tube to be lathed to this diameter. Finally, the tube is provided with slots of 20 to 25 mm in length at this position. The 4 CX 250 B and the tube of the inner conductor should have a fit tightly. If necessary, a tubular clamp can also be used (see **Fig.7**).

In order to avoid lathing, already available brass tube of 44 mm / 1 mm was firstly used for the plate inner conductor. However, it was found that such a tube possesses considerable mechanical tensions which were noticeable as crackling sounds on heating. In addition to this, the stability and the heat conduction were inferior.

The plate circuit is supported at one end by the tube and at the other by a ceramic support. This simple construction allows the tube to be replaced quickly.

The plate trimmer C 16 consists of a readily available aluminium knob of 30 mm diameter and 15 mm in length **(Figure 6)**. It is screwed onto a 48 mm length of 6 mm brass rod that has been provided with a fine M 6 thread to a length of 30 mm. The threaded rod is placed through a guide for single-hole mounting that possesses an M 6 fine thread on the inside.

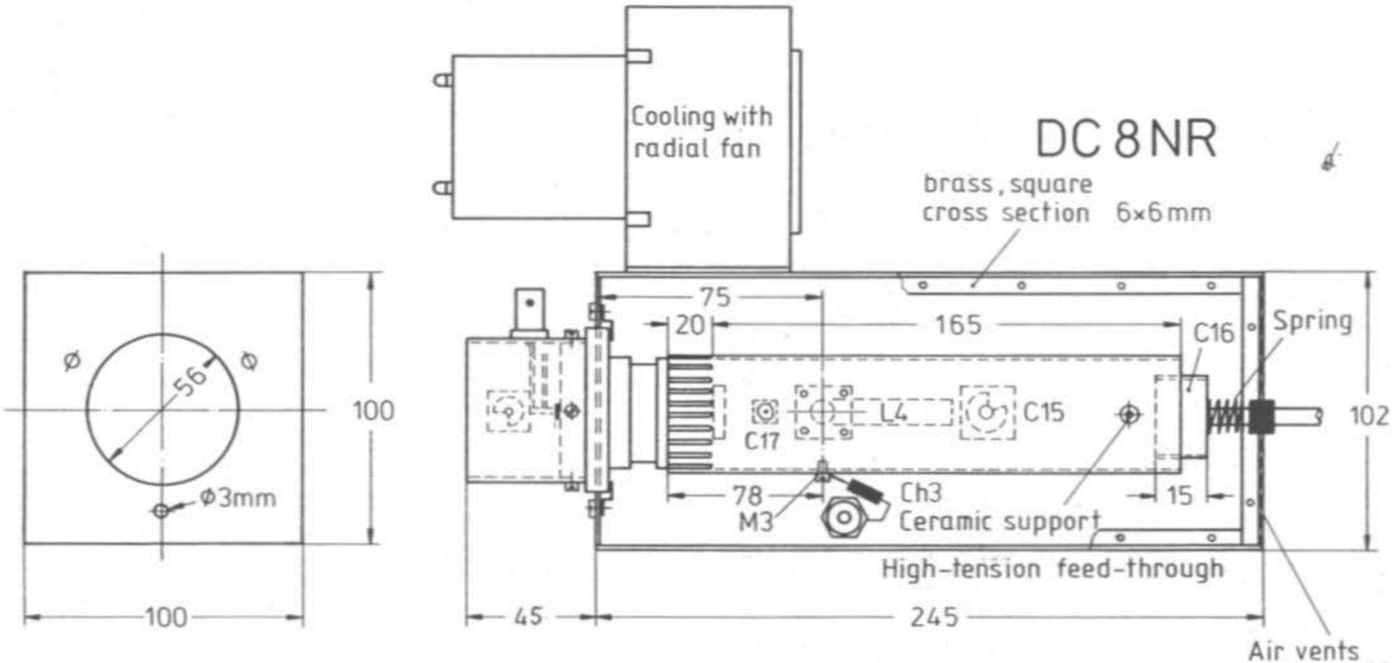

Fig. 5: Construction and dimensions of the 4 CX 250 PA for 435 MHz

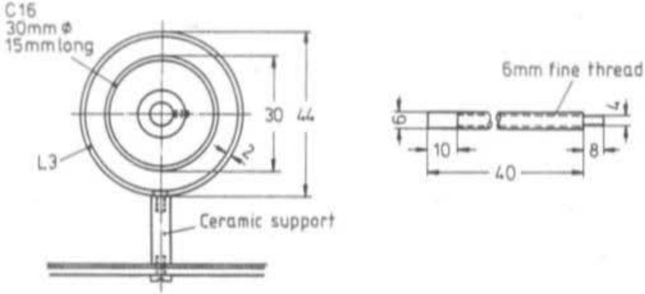

Fig. 6: Anode circuit tuning

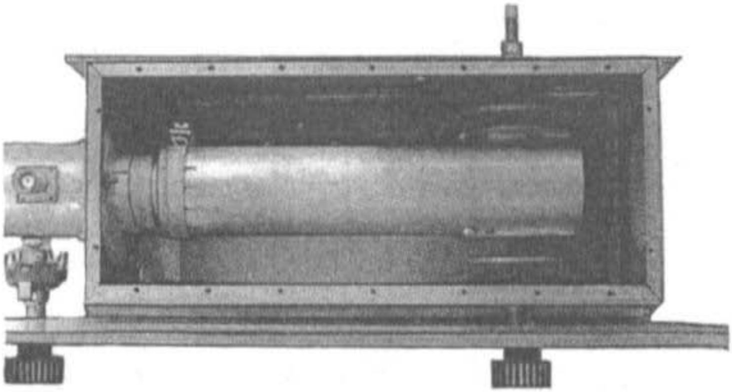

Fig. 7: Another prototype with a different plate circuit tuning

In order to obtain this, it is necessary for the diameter of 6 mm to be reduced to approximately 4.5 mm which can be done by soldering a suitable tube (6 mm / 0.75 mm) into place. In order to ensure, that the fine thread is not damaged by the mounting screws of the knob, the rod is lathed down to 4 mm diameter at one end for a length of 8 mm.

Of course, it is also possible for the tuning of the anode circuit to be made in a different manner, for instance, using two opposite plate capacitors as can be seen in Figure 7. The advantage of this is that the grid and anode circuit are accessible from the front panel. The disadvantage are the considerably larger dimensions of the power amplifier. Furthermore, the two plate capacitors must be tuned to the same capacitance values in order to ensure that a symmetric field prevails. Otherwise, it would not be possible to obtain the maximum output power.

The construction of the grid circuit and tuning is shown in **Figure 8**, which also shows the connections of the tube socket. The inner conductor (L 2) of the helical circuit is constructed from 2 mm thick, silver-plated copper wire of approximately 90 mm in length, and the coupling link L 1 from 1 mm diameter silver-plated copper wire. In order not to capacitively load the circuit more than necessary, the tuning is made with an air-spaced trimmer (C 14) having a low final capacitance (max. 2 pF). Since such trimmers are normally not available, the plates of a higher capacitance trimmer are removed until two stationary and one moving plate remain. The grid circuit can be aligned from the front panel using a flexible coupling and gear.

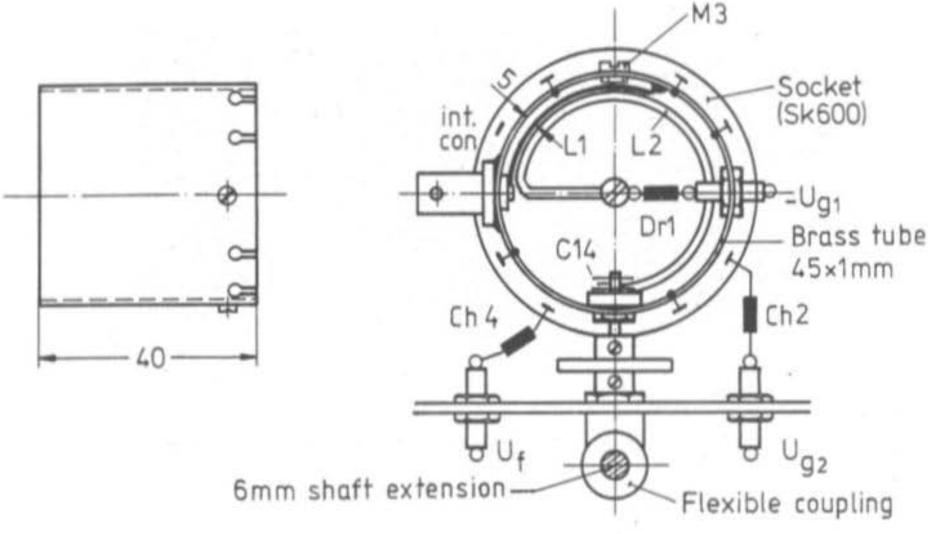

Fig. 8: The grid circuit

The outer conductor of the circuit is formed by a 40 mm length of 45 mm / 1 mm diameter tube that is placed over the tube socket and screwed with it at four points. In order to do this, M 3 nuts should be soldered into place on the socket over the holes provided. In order for the tube to be placed completely over the socket, it is necessary for slots to be made at the required positions (Figure 8). The five cathode connections of the tube are soldered using short pieces of wire to the inside of the socket as can be seen in Figure 8. **Figure 9** shows a photograph of the grid side of the power amplifier.

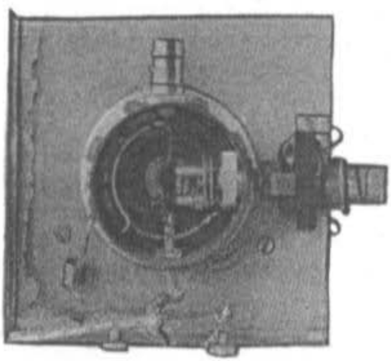

Fig. 9: Photograph of the grid circuit

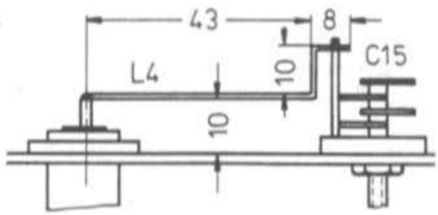

Fig. 10: Output coupling

The output coupling comprising L 4 and C 15 is shown in **Figures 10 and 11.**

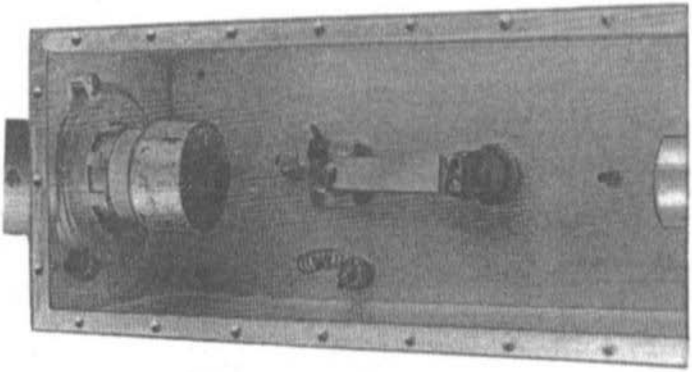

Fig. 11: View of the output coupling after removing the plate tube

4. COOLING

Tubes with the anode on the outside such as the 4 CX 250 B are designed for forced-air cooling. The cooling air is normally blown in from the grid side since the electrode connections also require cooling. It is necessary for the pin connections to be cooled even when only the heating is on. The cooling air then flows along the envelope of the tube with the aid of a ceramic chimney, flows through the plate cooling fins and escapes. A reduction in pressure results due to vortex and eddies. The 4 CX 250 B requires approximately 0.6 m^3/min. at the full plate dissipation power of 250 W and when using the matching socket and chimney.

A pressure of 1.6 mbar is required. The pressure drop can, however, be considerably higher due to an unfavorable air flow, so that the air flow is non-existent with an unsuitable fan.

Most axial fans provide sufficient quantities of air, even at low speed, and are sufficiently quiet. However, they are not able to operate against any noticeable pressure drop. Radial fans are better in this respect, but are louder, and are usually not powerful enough in the types usually used by radio amateurs (8). Experiments made, feeding the cooling air from a radial fan from the grid side, brought the expected negative result: after 10 s of continuous carrier, the output power dropped to more than 50 % of the commencement value and the solder on the plate capacitor at the end of the anode tube (Figure 7) melted. This means that the permissable anode temperature of 250°C was most certainly exceeded. Since no damage to the tube was noticed during the relatively long experimental process, this goes to show how robust the construction of the tube is. Due to its glass envelope, the 4 X 150 would not be able to stand such an overload condition.

In the case of the power amplifier described here, the cooling air is blown into the plate cavity in the vicinity of the tube anode using a relatively weak axial fan as shown in Figure 15. The air then flows without any considerable pressure drop along the plate tube so that sufficient heat is dissipated in this manner. A small portion of the cooling air is fed in this manner via the socket and the electrical connections and is exited via the grid circuit. A chimney should not be used here.

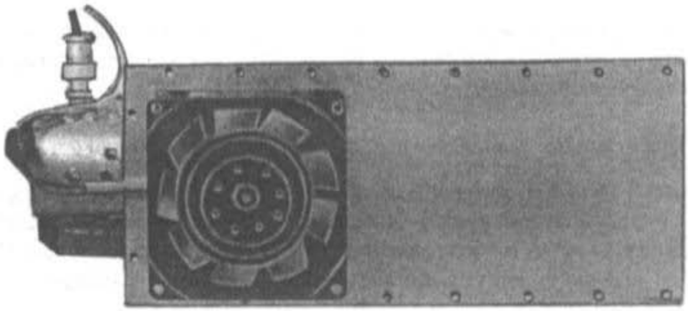

Fig. 15: Photograph of the power amplifier with fan

The most important point when using this concept is that a sufficiently large amount of air is moved and that hardly any pressure drop occurs on the large cooling surface of the plate tube. The shown axial fan with the dimensions 88 mm x 88 mm x 50 mm was found to be suitable for SSB operation up to power levels of approximately 500 W input. For higher power levels, a radial fan is more favorable which should be mounted as shown in Figure 5.

The air intakes of both types of fans should be kept free for a distance of at least 3 to 4 cm. The air intake of the anode chamber should be provided with the coarsest possible screening (spacing 5 to 6 mm). Brass netting can be soldered into place.

5. POWER SUPPLY

Figure 12 shows the circuit diagram of the simple power supply. The required high plate voltage of 1500 to 2000 V required in order to obtain the highest possible linearity is obtained with the aid of voltage doubling (Delon-circuit). This saves an expensive high-tension transformer; the output voltage is sufficiently stable. A series connection of cheap electrolytic capacitors replaces the large, and expensive metal paper capacitors (whose reliability is however greater). The resistors connected in parallel with the electrolytic capacitors compensate the leakage currents and discharge the capacitor chain on switching off. It is therefore possible for the whole power supply to be accommodated on a PC-board. **Fig. 13** shows this PC-board which has been designated DC 8 NR 007. Cheap rectifier diodes can be used; the two diode chains could, however, be replaced by two single diodes (inverse voltage 3 kV; I = 3 A). The dropper resistors R 7 to R 9 of the stabilizer chain generate a considerable heat. For this reason, they should be spaced somewhat from the PC-board. Capacitor C 7 is not to be found on the PC-board, but is connected in the vicinity of the high-tension feed-through.

In the linear mode, the tube requires a stabilized screen grid voltage of 350 V. If a failure of the anode voltage should occur, the screen grid with its maximum dissipation power of 12 W is endangered, since it will then work as plate and will attempt to accept the high anode current. The screen grid current can become negative at low anode currents. For this reason, it is favorable for the screen grid supply to be made with the aid of neon stabilizers which limit the available power to 11 W. This ensures that no overload condition can occur. The bias voltage source for the control grid can be at high impedance since no grid current will be drawn in the linear mode.

The meter switch S 3 connects all currents and voltages of interest to the meter I 1. The values of the shunt resistors R 10, R 11, R 14, and R 15 depend on the meter used, and should be recalculated if necessary.

The given values are valid for a meter having 60 μA, and an impedance of 2.5 kΩ.

The 25 Ω resistor R 19 in the power line is not absolutely necessary. It is only provided to ensure that the transient current peak on switching on does not actuate the power line fuse (10 A). Switch S 2 bridges the resistor after the electrolytic capacitors are fully charged after approximately 1 s, and the transformer will receive the full power line voltage.

After switching on, the heater, control grid and plate voltage remain connected to the tube. As soon as the auxiliary contact r 1 of the antenna relay is opened in the receive mode, the full negative grid bias voltage of 110 V will block the tube. If the driver is not keyed, this negative voltage can be used in the CW mode for keying.

The power amplifier can be switched to class C operation with the aid of two change-over switches that are not shown in Figure 12: The control grid bias voltage is increased from – 55 V to – 90 V, and the screen grid voltage reduced from 350 to 250 V. In this position, a higher plate efficiency (of more than 50 %) is possible, however, the gain will be lower. This mode can be used for FM and CW; however, it is easier to use the linear class AB^1 mode also for these modulation modes. When operating with lower plate voltages of less than 1500 V, it will be necessary for the screen grid voltage also to be reduced to 250 to 300 V.

6. COMPONENT DETAILS

Transformer Tr 1:	700 - 800 V, 600 mA (for 1700 - 2000 V DC under load) Secondary 2: 75 V / 20 mA Secondary 3: 6.0 V / 2.6 A
D 1 - D 14:	see text (1 N 4007 or similar)
D 15:	1 N 4007
D 16:	1 N 4148 or similar
C 1 - C 6:	100 μF / 550 V
C 7:	ceramic disk capacitor 470 pF / 3 kV
C 8, C 9:	25 μF / 350 V
C 11 - C 13:	500 pF / 500 V feedthrough capacitor for screw mounting
C 14:	0.4 - 2 pF (see text)
C 15:	15 pF (see text)
C 16:	0.3 - 6 pF (see text)
C 17:	10 x 10 mm mounted on insulated feedthrough spaced 20 mm from L 3
C 18:	220 pF ceramic disk capacitor

R 1 - R 6:	100 kΩ / 2 W	R 14:	6.0 kΩ / 0.5 W
R 7 - R 9:	8.2 kΩ / 11 W (better 15 W)	R 15:	60 Ω / 0.5 W
R 10:	0.6 kΩ wire-wound resistor (see text)	R 16:	1 kΩ trimmer pot.
R 11:	50 Ω trimmer potentiometer	R 17:	100 kΩ / 0.5 W
R 12:	2.7 kΩ / 2 W	R 18:	33 Ω / 0.5 W
R 13:	10 kΩ trimmer potentiometer	R 19:	25 Ω / min. 5 W

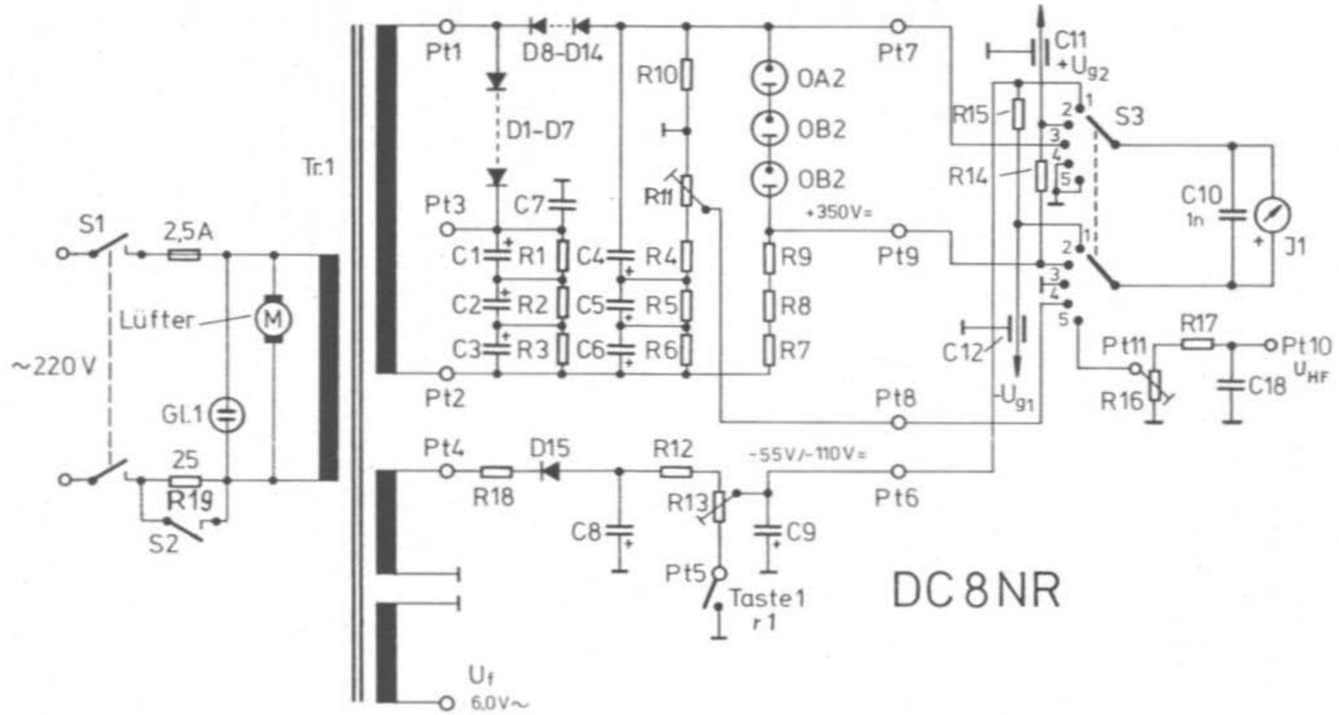

Fig. 12: Circuit diagram of the power supply with voltage doubling to obtain the high-tension voltage

All chokes are made from λ/4 length (17.2 cm) of enamelled copper wire wound in the form of a coil:

Ch 1, Ch 2,Ch 5:	0.3 mm enamelled copper wire, 3 mm diameter
Ch 3:	0.8 mm enamelled copper wire, 7 mm diameter
Ch 4:	0.8 mm enamelled copper wire, 4 mm diameter

Ceramic support, 28 mm long, with two metal caps with M 3 thread
Ceramic high-tension feedthrough
Aluminium knob
Flexible coupling
6 mm diameter shaft support
6 mm extension shaft
Gearing
Radial fan (Airflow 26 BTM)
Axial fan (Papst type 3050 with capacitor motor: Dimensions: 88 x 88 x 50 mm, 90 m^3/h, 220 V / 50 Hz)

7. CONNECTION AND ALIGNMENT

The power supply should firstly be tested. This is followed by checking the operation of the antenna relay. With relay contact r 1 closed, a voltage of approximately – 55 V should be aligned at Pt 6 with the aid of R 13. After this, the power amplifier can be connected to the operating voltages and connected to a load (antenna or dummy-load). Normally, it is recommended that the preliminary alignment of such a high-power amplifier to be made at reduced screen grid and plate voltage, and to increase these voltages after this preliminary alignment.

Although this method is to be recommended, practical difficulties take place in our application (power supply). This is also not absolutely necessary if the following procedure is carried out:

Switch on the power supply and allow a warm-up period of 1 minute; place switch S 2 into the anode current position. Switch on the power amplifier and adjust R 13 to obtain a quiescent plate current of 100 mA. Establish whether the linear amplifier is operating stabily: increase the anode current temporarily with R 13 to 200 mA and rotate the grid and plate circuit trimmers. No output power should be indicated, and the indicated plate current should not change.

Connect an exciter with an output power of approximately 3 to 4 W, and place the coupling trimmer C 15 to its minimum capacitance position. Align the grid circuit for maximum plate current. Bring the plate circuit to resonance with the aid of C 16, where a clear dip of the anode current should appear. Align the output coupling trimmer and the plate tuning alternately for maximum output power. The alignment is completed by carefully bending the coupling link L 1. The position and spacing of the output coupling link L 4 to the plate circuit as shown in Figure 10 should be most favorable. By the way, the stabilizer chain presents a good visual indication for the correct tuning of the power amplifier. The screen grid currents will fall noticeably at resonance.

After this, it is possible for the drive power to be increased. Approximately 7 to 8 W will be required for full drive according to the plate voltage which means that the 144 MHz - 432 MHz transverter with the EC 8020 in the power amplifier (9) is suitable. The control grid current can be used as criterion for the drive limit: The grid current should be zero in linear operation. Large intermodulation products will be generated when driving the amplifier into the grid current region, which will cause the signal band width to increase and the power amplifier will splatter. No neutralization has been found necessary with all the linear amplifiers constructed until now.

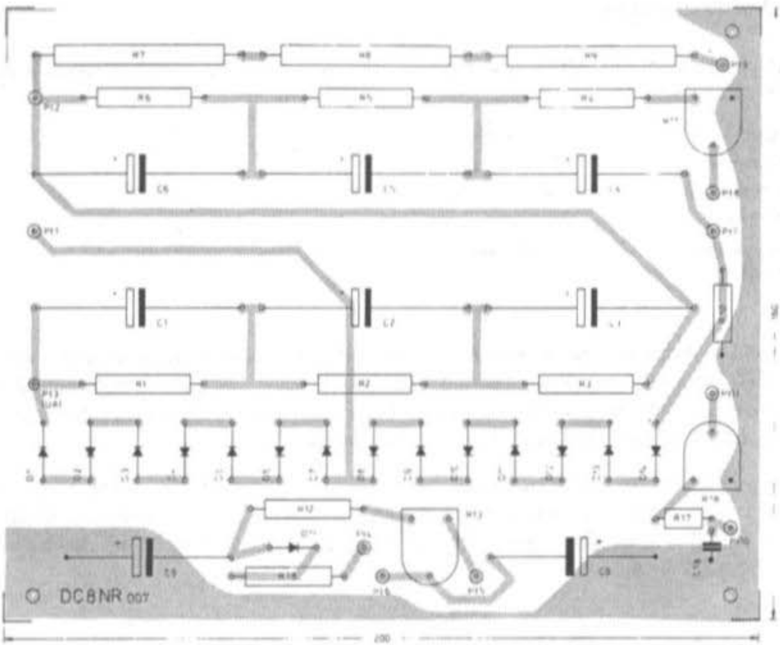

Fig. 13: PC-board DC 8 NR 007 for the power supply

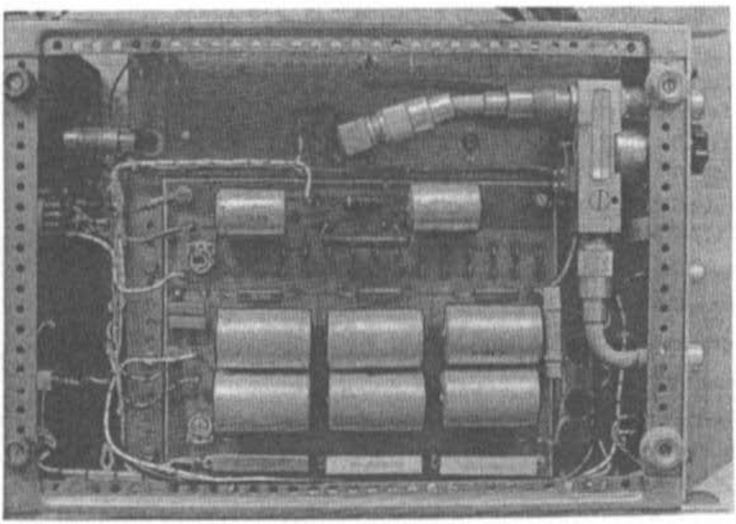

Fig. 14: Photograph of the power supply

8. MEASURED VALUES

The following measured values result as mean values of several different tubes operated at a plate voltage of 1700 V:

Power gain:	15 dB
Output power:	approx. 200 W
DC-current:	295 mA
Plate efficiency:	approx. 40 %
RF-drive power:	6.5 W
I_{G2}:	8 mA
I_{G1}:	0.5 mA

Higher power levels can be obtained by increasing the plate voltage and/or using the tube 4 CX 250 R. Measurements made before and after silver-plating the power amplifier showed no improvement, which is certainly due to the large, current flow areas. However, a silver-plated surface no doubt has advantages in the long run.

The third-order intermodulation rejection for the 4 CX 250 B in the linear mode and at full drive amounts to approximately 20 to 25 dB according to (10). In comparison, 35 dB are given in (6) for the 8874 at 30 MHz, and 1 kW input. Approximately 25 to 30 dB can be expected in the case of the 2 C 39.

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HOME-MADE FINGER STOCK

by J. Nilsson, SM 6 FHI

A tool for manufacturing finger stock is to be described that can be easily made at home. A typical application for such contact strips is shown in **Figure 1** where it is used as contact between the anode strip line circuit of a 70 cm amplifier and a 2 C 39 tube. The tool comprises a piece of right-angle iron profile and a modified saw blade which is usually placed into a vice as shown in **Figure 2**.

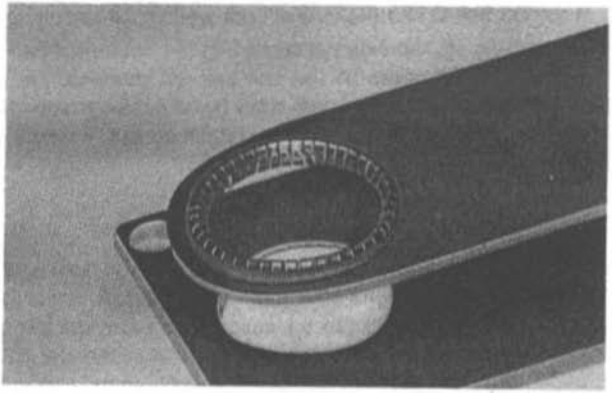


Fig. 1: Application for contact strips

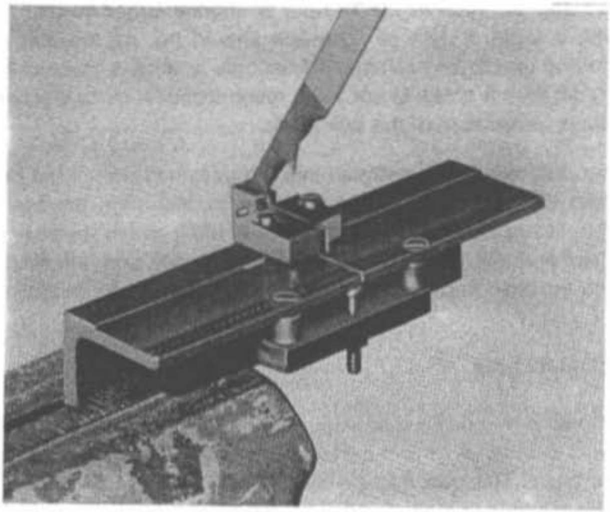


Fig. 2: Tool for manufacturing contact strips

1. MANUFACTURE OF THE TOOL

1.1. Base

The iron profile for the base of the tool can be of any size. Dimensions of 40 x 40 x 5 mm are favorable, and approximately a 150 mm long piece will be required. In the middle of the horizontal surface, a slot is sawn exactly perpendicular to the longitudinal axis. This must be made very carefully since the slot must have a constant width and may not be curved. If this slot cannot be made using a machine, it is advisable for it to be cut with a completely new saw blade of the same type used later for the cutting tool. The width of the slot should be only just as wide as that of the cutting tool. If the slot is wider than the tool itself, irregularities will be present at the cut edges and the tool will not work satisfactorily.

After this slot has been completed, the external surface of the horizontal part is smoothened. Attention should be paid that the edges of the slot are not rounded. The smoothing and polishing process can best be made using emery cloth (sand paper) wrapped around a file or using a sanding disc in conjunction with an electric drill. Of course, it would be better to use an automatic milling machine.

1.2. Bearing

The bearing onto which the modified saw blade is mounted is made from a 50 mm long piece of rectangular steel having the dimensions 10 x 8 mm or similar. A slot is sawn at one end that is as deep as the sawtooth is wide. As in the case of the base, the slot width must correspond to the thickness of the cutting tool. It is important that the slot is exactly perpendicular to the outer edge, in other words in the same direction as the slot in the base part. If this is not the case, the cutting tool will be under tension and can break.

A hole is now drilled in the slotted end for mounting the pivot. The diameter depends on the size of the hole in the saw blade that is to be used. In the case of a 4 mm hole, a 4 mm dia. bolt should be used and the hole in the bearing is slightly larger so that the bolt can be pushed through with a slight resistance. The bolt should not be too stiff since it will be necessary for the cutting tool to be exchanged from time to time. A piece of round hardened steel should be used as bolt. A screw is not to be recommended, since a groove will be worn into it quickly due to the movement of the saw blade.

Two holes of 4.2 mm diameter are now drilled into the unslotted part of the bearing so that it can be mounted onto the base portion. This is done by mounting the saw blade into the bearing with the teeth facing in upward direction and holding this in the base part so that the saw blade is in the slot and just visible on the inside. In this position, the two mounting holes should be marked on the base portion. 3.2 mm dia. holes should now be drilled and provided with an M 4 thread.

1.3. Guide for the Contact Strip

The base portion should now be provided with a guide for the contact strip. This can either be milled to a depth of 0.6 mm as shown in the photographs, or two 0.7 mm thick steel strips can be screwed into place. This second possibility is more favourable since it is possible for the guide strips to be made variable by providing adjustment slots. This allows the length of the cut in the contact strips to be varied, and simplifies the required sharpening of the cutting tool described later. Both parts of the guiding system should be plane and smooth and screwed securely to the base part.

1.4. Cutting Tool

As has been previously mentioned, a saw blade is used as cutting tool. A fast steel blade of the highest quality for cutting iron should be used, but need not be new, since the cutting teeth are not used. The saw blade is mounted with the teeth facing in an upward direction into the bearing which is already mounted on the base part, and adjusted so that its edge is just visible on the inside of the base.

The contours shown in **Figure 3** are then drawn onto the saw blade so that the steep lower edge of the cutting triangle is spaced 3 mm from the guide. This ensures that the contact strips are not out through completely, but 3 mm remains for mounting. The saw blade is removed again from the bearing and the required contours are sawn out. Only the minimum of material required should be removed in order to ensure that the resulting cutting tool is not too weak. It is now possible for the teeth of the saw blade to be removed and shortened to a handy length of say 100 mm.

The cutting tool is now mounted again and tested using a 0.2 mm thick piece of phosphor bronze. It is possible that a few corrections must be made. After it is working satisfactorily, the cutting edges of the tool are sharpened with the aid of a wet stone. The surface should be smooth and the edges should be sharp. A piece of plastic tube can be used as grip, or the saw blade can be wound with insulating tape.

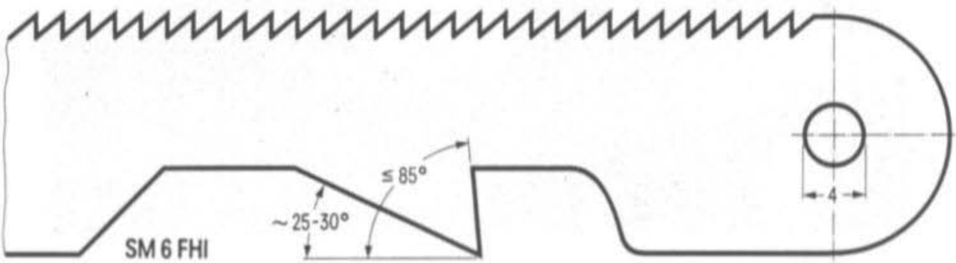


Fig. 3: A saw blade modified as cutting tool

1.5. Stop

In order to ensure that the contact strip is not accidentally cut through, and to ensure that the cuts are equally deep, a stop should be provided. The author uses an adjustable stop, that can be seen in Figure 2. A flat iron plate is mounted using spacers below the base plate and is provided with a M 4 thread, and a long screw which is then able to stop the cutting tool at the correct position. Of course, the stop can also be in the form of a large disc that is held with a screw and covers the slot from below.

1.6. Strip Holder

In order to ensure that the contact strip is held in position and is not lifted after each cutting process, a strip holder is provided. This holder is shaped like a tuning fork and comprises two pieces of square steel that have been sawn from the same material as the bearing. The special shape of the right hand half of the holder is shown clearly in **Figure 4**: The front of the right-hand piece is removed except for a 1 mm wide tongue. If the edge of each finger of the contact strip coincides with this tongue, this will ensure a constant spacing between each of the contact tongues, and will provide professional looking strips. The cuts are 0.6 mm wide (standard thickness of saw blades) and are spaced 1.6 mm from another.

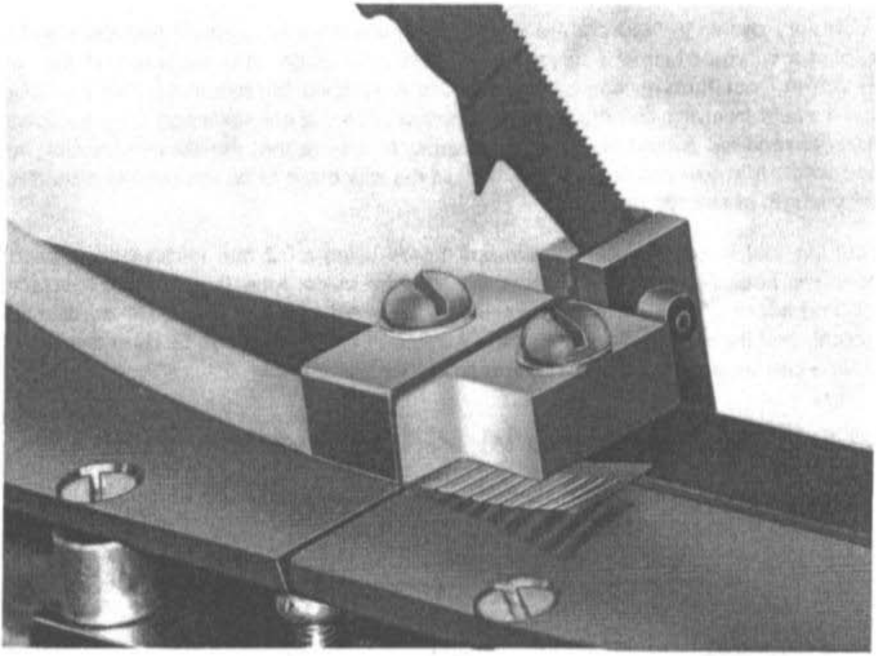


Fig. 4: Enlarged view of the strip holder

2. MANUFACTURE OF THE CONTACT STRIPS

The most suitable material for manufacturing the contact strips is 0.2 mm thick phosphor bronze plate that is available in strips of 150 mm in width. This is just the right size for the anode contacts of a 2 C 39. Longer contact strips are cut off in longitudinal direction from this metal plate. 10 mm wide strips are satisfactory, and 7 mm deep cuts can be provided. If the cuts are deeper, the remaining strip of 3 mm will not be sufficient and the contact strip will bend during the punching process. It in therefore more favorable to leave sufficient material during the cutting process, and remove this afterwards.

A burr will be present at the base of each cut, but this is inavoidable when using such a simple tool. The sharper the cutting tool, the smaller will be the burr. Of course, the burr can be flattened with the use of a hammer, or the burr can be bent in the direction of the solder side.

3. NOTES

The bronze plate can be easily cut with the aid of a normal pair of scissors. After punching, the contact strip should be bent in the form shown in Figure 1. It should be noted that this material easily breaks which means that sharp bending edges should be avoided.

It is now necessary for the contact strip to be bent in the form of a ring. This is made by pulling it around an edge that is not too sharp. In order to ensure that no sharp bents result, this process should not be made in one powerful pull, but the strip should be drawn several times over the edge so that it becomes rounder and rounder until the required radius is obtained. Do not forget to ensure that the burr is on the outside. Low-radius contact rings should be bent with the aid of a round pair of pliers and the narrow radius should also be obtained here by gradually obtaining the required radius. Of course, some practice is required at first, and it is advisable to firstly make somewhat more contact strips than actually required.

If the contact strips are not to be bent in a round form but used, for instance, for making an RF proof contact, a greater spacing can be selected between the individual contact fingers. For such applications, the author uses a spacing of approximately 5 mm.

By the way, the bronze plate should be soft soldered, since it will lose its spring characteristics if hard soldering is used.

The condition of the tool should be checked at regular intervals. The most important thing is a sharp edge of the cutting tool since burrs would otherwise result, and the punching process will not be easy, Clean cuts will only be provided when a good guiding of the actual cutting tool is provided through the contact strip holder and when maintaining the vertical arrangement between cutting tool and base. With good experience a 150 mm long strip is made in 2-3 minutes when using a spacing of 1.75 mm (approx. 85 steps).

AN ABSORPTION WAVEMETER FOR 70 MHz to 1350 MHz

by J. Dahms, DC 0 DA

It is often very difficult when constructing local oscillator chains and transmit mixers to check and align the various stages to the required, and not the unwanted frequencies. The dipmeters available on the market usually only have an upper frequency limit of 200 MHz, and the sensitivity at their upper frequency limit is usually poor. This article is to describe an absorption wavemeter designed by DJ 2 HF, which allows the frequency of low-level RF voltages to be measured in the frequency range from approximately 70 MHz to 950 MHz. The complete unit comprises a control unit (stabilized power supply, tuning potentiometer, meter) and a separate probe for each of the six frequency ranges. The probes can be remotely connected to the control unit using two screened cables similar to a tape recorder cable and can be of any required length. This means that it is possible for measurements to be made in existing equipment which would not be possible when using a dipmeter. A photograph of the control unit and two probes are shown in **Figure 1**.

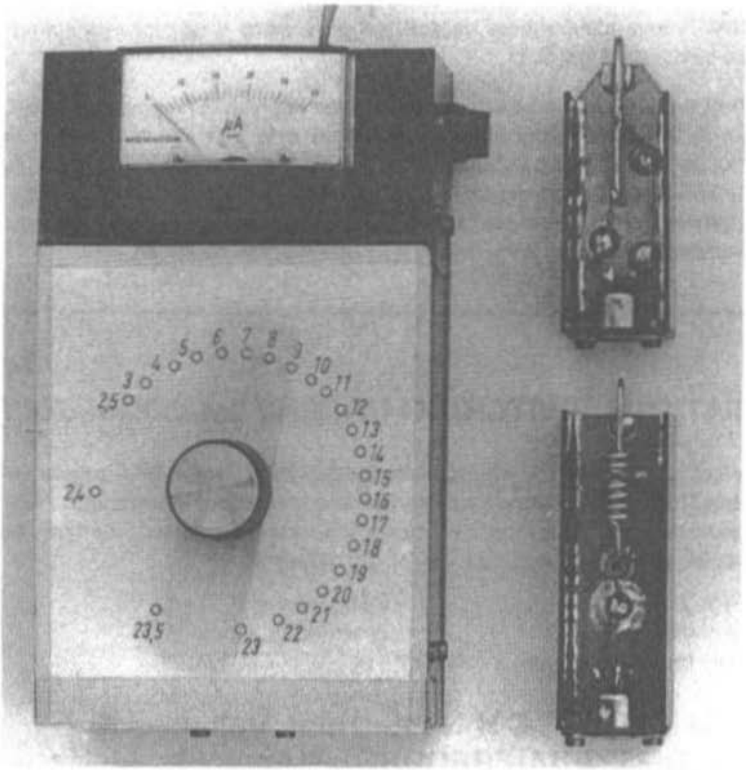


Fig. 1: An absorption wavemeter with measuring probes for 70 to 1350 MHz

1. CIRCUIT

Each probe is provided with a resonant circuit for the required frequency range, which is tuned with the aid of a varactor diode. The tuning voltage of 3 to 30 V is provided by the control unit. **Figure 2** shows a typical characteristic of a varactor diode BB 141. A portion of the RF-voltage present in the resonant circuit is rectified and fed to the meter via the interconnection cable. The meter will produce a reading under resonance conditions.

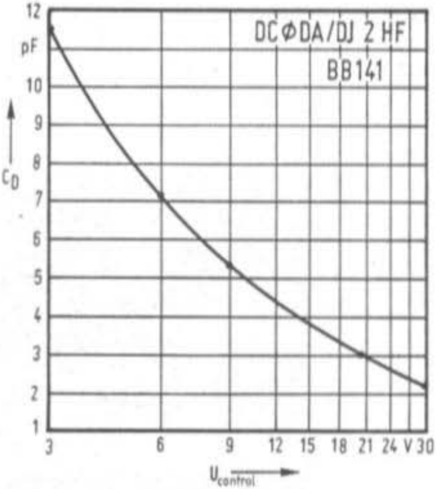


Fig. 2: Typical characteristic of the varactor diode BB 141

The circuit diagram of the power supply and basic circuit of probes A to E is given in **Figure 3**. A small power transformer, for instance from a transistor radio, can be used to provide the required tuning voltage.

Such transformers are usually very inexpensive. In order to obtain the required voltage of approximately 30 V, it is necessary for voltage doubling to be made when using such transformers. The power supply need only provide slightly more current than is necessary to operate the zener diode. It is important that the voltage is well filtered.

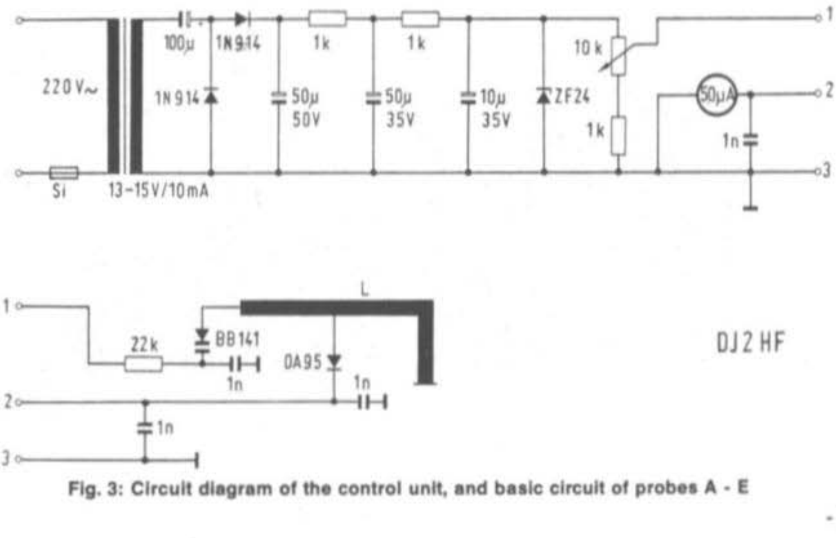


Fig. 3: Circuit diagram of the control unit, and basic circuit of probes A - E

The tuning voltage can be varied with the aid of a conventional carbon potentiometer having a linear characteristic. The meter for indication of the rectified RF-voltage should be as sensitive as possible (50 to 100 µA). Connections 1, 2, and 3 are fed to a three-pin tape recorder-type socket. Pin 1 provides the tuning voltage for the varactor diode, pin 2 provides the rectified RF-voltage from the diode probe, and pin 3 serves as common ground connection.

In the author's prototype, the small power supply, meter, potentiometer, and DIN socket are mounted in a home-made cabinet constructed from epoxy board **(Figure 4)**. The potentiometer shaft is provided with a knob and pointer, and the selected voltage is measured at connections 1 and 3 using a high-impedance (≧ 100 kΩ) voltmeter. The measured values are then traced in a calibration curve.

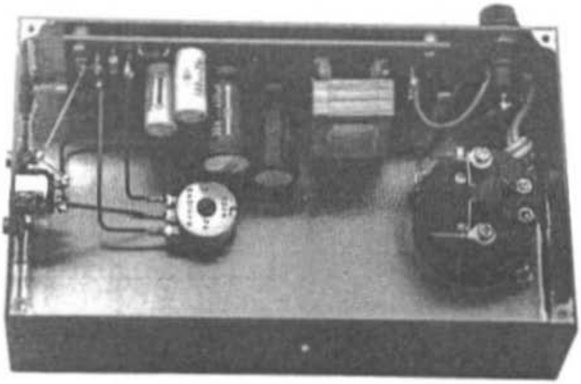


Fig. 4: Internal view of the control unit

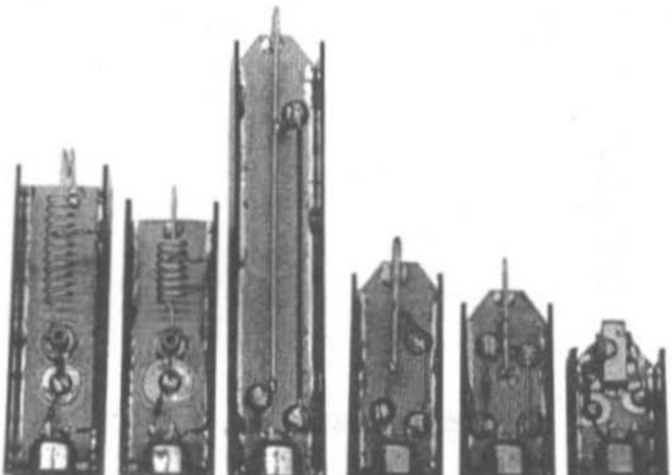


Fig. 5: The measuring probes

The measuring probes consist of a resonant circuit tuned with a varactor diode BB 141 to the required frequency, and a germanium diode which provides a DC-voltage when the frequency of the resonant circuit coincides with that of the frequency to be measured. This resonant condition is shown as an indication on the μA-meter. The construction and the electrical circuit of each of the separate measuring probes A to E is given in **Figures 5 to 10**. The side panels are made from single-coated PC-board material. All dimensions and component details are given in the diagrams. Ceramic capacitors without connection leads should be used for by-passing the diodes. The 0.5 pF capacitors of probes A and B are ceramic tubular trimmers without spindle, which are also used as support for inductance L.

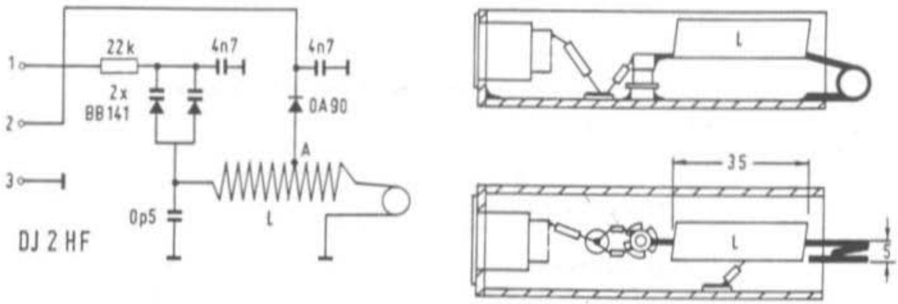


Fig. 6: Probe A for 70 to 120 MHz

L: 11 + 1.75 turns of 1.3 mm dia. (16 AWG) silver-plated copper wire wound on a 7 mm former

Tap: 4.5 turns from the cold end

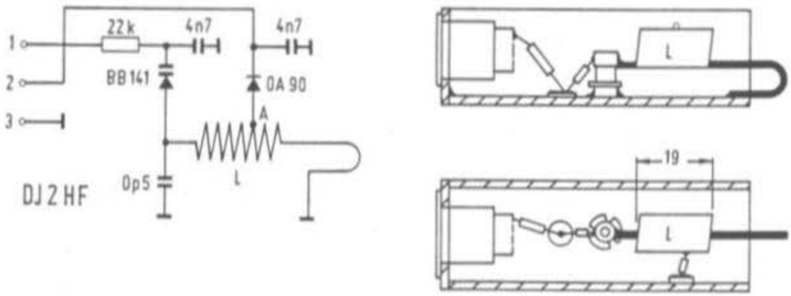


Fig. 7: Measuring probe B for 120 to 200 MHz

L: 7 turns of 1.3 mm dia. (16 AWG) silver-plated copper wire wound on a 7 mm former

Tap: 2.5 turns from the cold end

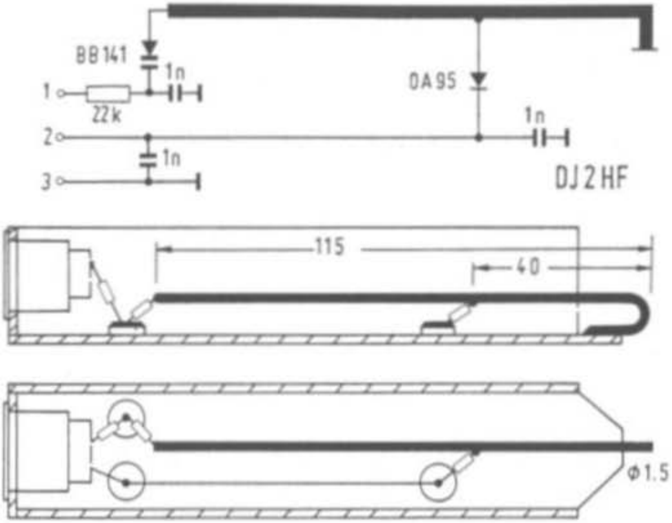


Fig. 8: Measuring probe C for 190 to 340 MHz

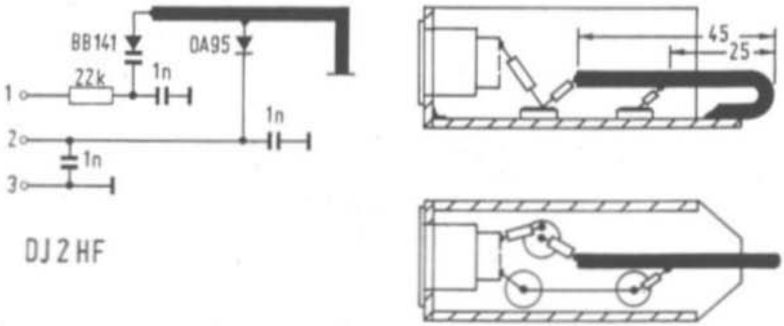


Fig. 9: Measuring probe D for 260 to 500 MHz

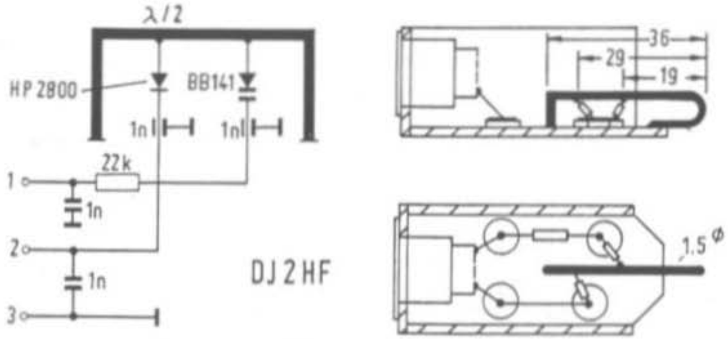


Fig. 10: Measuring probe E for 500 to 950 MHz

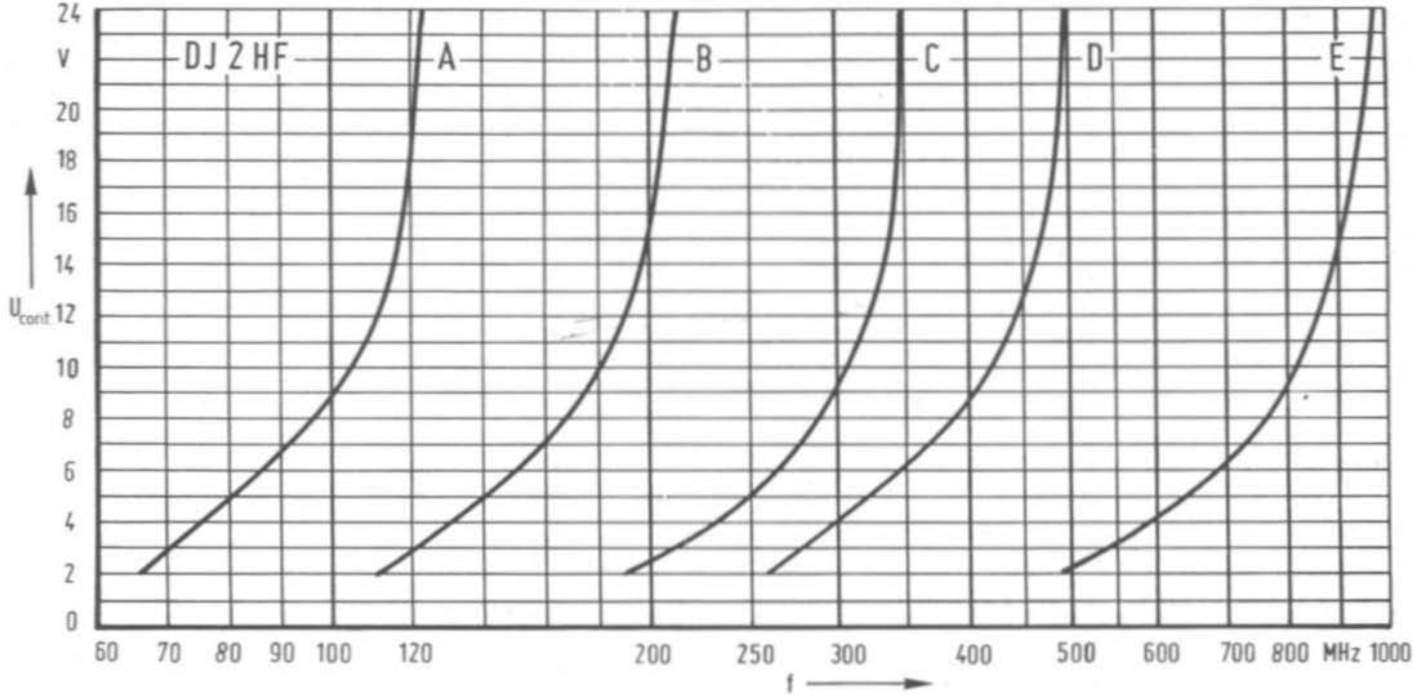


Fig. 11: Calibration curves for measuring probes A - E

The individual calibration curves are given in **Figure 11**. These were provided by DJ 2 HF for the measuring probes he designed. The author, and other amateurs manufactured probes from these diagrams, and were surprised at the good coincidence with the given curves.

This means that it is not absolutely necessary for the measuring probes to be calibrated on a measuring system. Of course, it is necessary for the given components and inductance dimensions to be maintained.

2. OPERATION

During frequency measurement, the probe should be loosely coupled to the test object so that the resonance condition lies in the last third of the μA-meter range. This will ensure that a very sharp resonance indication will be provided. If, for instance, a frequency of 404 MHz is to be measured on a transmit mixer from 28 MHz to 432 MHz, the coupling should not be too tight, otherwise the 404 MHz signal will be overloaded by the far higher level of the nominal frequency 432 MHz.

3. EXTENDING THE FREQUENCY RANGE TO 1350 MHz

The author developed a further measuring probe that is able to provide reliable resonance indications up to and including the 23 cm amateur band. This measuring probe F is given in **Figure 12**. A Schottky diode hp 2800 (or similar) is used in this probe as well as in probe E; this is to ensure sufficient sensitivity. Also in a similar manner to probe E, the resonant line is in the form of a λ/2 circuit which provides a better reproducibility and higher Q. In order to obtain the lowest possible commencement value of the tuning capacitance, it is necessary to connect a disk capacitor of 1.5 pF in series with the varactor diode BB 141. The calibration curve is given in **Figure 13**.

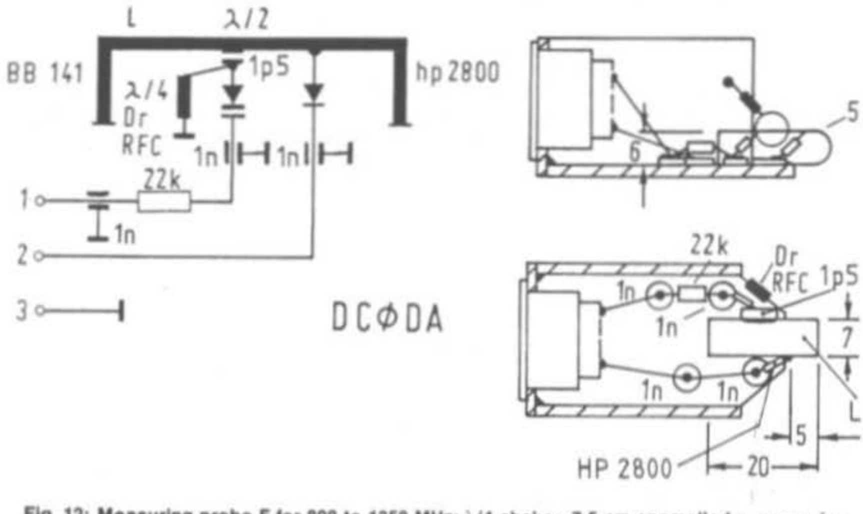


Fig. 12: Measuring probe F for 800 to 1350 MHz; λ/4 choke: 7.5 cm enamelled copper wire, approx. 0.4 mm dia. (26 AWG) wound on a 3.5 mm former. L: silver-plated copper strip

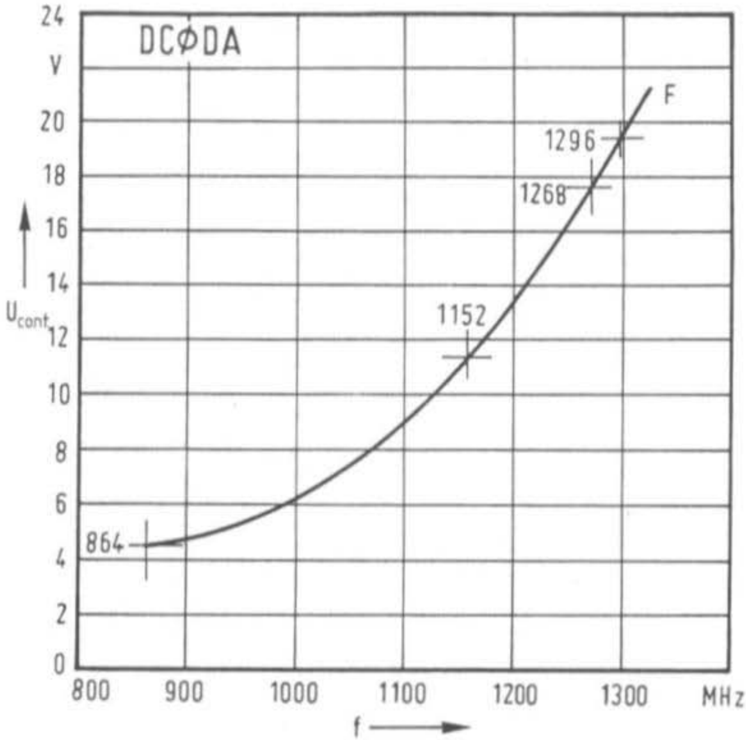


Fig. 13: Calibration curve for measuring probe F

ZENER DIODE NOISE IN OSCILLATOR AND MULTIPLIER CIRCUITS

by H.J. Franke, DK 1 PN

The local oscillator signal was examined on a spectrum analyzer during the construction of a 23 cm converter described by DJ 5 XA. The author was surprised at the high noise threshold of the signal. This article is to describe the experience gained in reducing the noise component.

Zener diodes with breakdown voltages of 5 V or more generate noise voltages due to the avalanche discharge (1), whose value is mainly dependent on the technology used during manufacture and from the operating point (e.g. the current flowing via the diode). If the zener diode is used directly or via a pass transistor to stabilize an oscillator, this noise will amplitude and frequency modulate the output signal. It is true that the modulation level will be very low, however, the FM component will be multiplied in the frequency multiplier chains together with the signal when used for UHF and SHF converters. The AM components will cause a phase modulation in the stages that operate in class C. The noise component can increase to such a level when using unsuitable zener diode circuits that it is not only measurable, but also audible.

Figure 1 shows the original voltage stabilizer circuit which feeds a crystal oscillator at 70.444 MHz. This frequency is multiplied to 1268 MHz (x 18). The zener diode is a BZY 85 / C 9 V 1, with a dropper resistor R 1 of 560 Ω. **Figure 2** shows a diagram of the output signal: With operating voltages of 12 V and more, the noise ratio amounts to – 55 dB. This noise ratio will fall to 30 dB when reducing the operating voltage to 9.5 V !

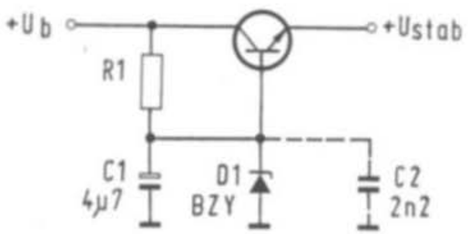


Fig. 1: Original voltage stabilizer circuit including first modification

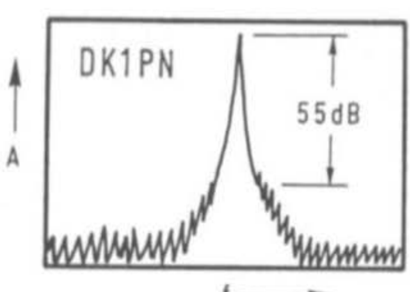


Fig. 2: Output frequency spectrum at 1268 MHz with a noise ratio of 55 dB

This poor value shows that the operating point of the zener diode should not be brought in the vicinity of the bend in the characteristic. This means that the dropper resistor R 1 should have a value so that sufficient current (several mA) flows through the zener diode even at the lowest operating voltage.

In order to increase the noise ratio, a ceramic disc capacitor of 2.2 nF was connected in parallel (shown as a dashed line in Figure 1). This increased the noise ratio from 55 to 60 dB.

This was followed by replacing the zener diode BZY 85 / C 9 V 1 (diffusion diode) for a planar diode BZX 85 / C 9 V 1, and the results were measured, using various values for the dropper resistor R 1 at U_b = 12 V and without parallel capacitor C 2:

R 1 (Ω)	Noise ratio	Zener diode current
1000	45 dB	2.9 mA
500	52 dB	5.8 mA
333	54 dB	8.7 mA

Surprisingly enough, the noise values for the modern diode were somewhat worse; however, the noise was found to be not so dense on the spectrum analyzer as when using a diode of the BZY series; on experimenting with an operating voltage of 9.5 V, it was found that the parallel capacitor of 2.2 nF was immediately able to improve the noise ratio from 30 dB to 60 dB.

The experience gained in this manner led to the design of the circuit given in **Figure 3** which was subsequently measured. The noise ratio of this circuit amounts to more than 70 dB ! Furthermore, a reduction of the operating voltage from 15 V down to 6 V does not have any effect on the noise ratio, even when voltage stabilization ceases.

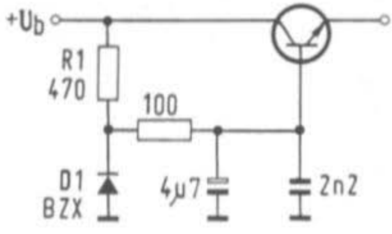


Fig. 3: Voltage stabilizer circuit with filtering of the noise voltage

No effect of the transistor type on the noise ratio could be observed when experimenting with transistors BC 107 and 2 N 918. It was also found that a ceramic capacitor connected between emitter and ground had no effect since the stabilizer circuit is very low-impedance at at this position.

REFERENCES

(1) Macek, Dr.O.: Z-Dioden, Eigenschaften und Anwendungen
Siemens, Technical Component Information, September 1976, Order-No. B 1593

STABILIZING THE OPERATING POINT OF TRANSISTORS WITH DIRECTLY GROUNDED EMITTER

by E. Schmitzer, DJ 4 BG

1. GENERAL

Several amplifier stages have been described for the UHF/SHF bands where the emitter of the transistor is directly grounded. This type of circuit brings several advantages at higher frequencies, however, it is necessary for special measures to be made to stabilize the operating point against temperature fluctuations. If this is not done – as was the case with the converter given in the references – operation will only be possible within room ambient temperature ranges. The following measured curves will show that it would be impossible to use it for portable or mobile operation.

There is an extremely simple method of stabilizing the operating point of such circuits, which should be generally known. The amount of circuitry required is virtually negligible. Some of the stages of the 13 cm converter described in (1) are to be used as example for the circuit technology and for the measurements.

2. CIRCUITS

The experimental circuits are given in **Figure 1: Figure 1 a** shows the DC-paths of the original circuit of the converter, **Figure 1 b** the modified circuit for the first RF-amplifier with an operating point of 9 V / 3 mA. **Figure 1 c** shows the modified circuit of the second RF-amplifier with an operating point of 6 V / 6 mA. A voltage feedback is now effective here: If the collector current increases with temperature, the collector-emitter voltage will drop due to the voltage drop across the DC-dropper resistor R_C, and thus will also reduce the base bias voltage. This means that the increasing current will be partly compensated. Prerequisite for this is that a considerable part of the overall operating voltage is dropped across the collector resistor. This results in a high DC residual gain and a sufficiently high feedback factor. **Figure 1 d** shows a recommended circuit for stabilizing the operating point of the mixer.

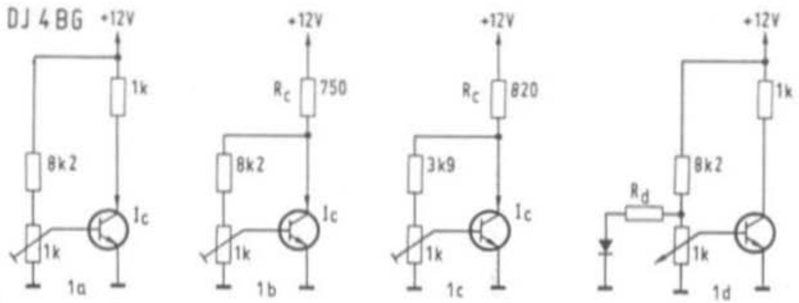


Fig. 1: Circuits: a) Original, b) 9 V / 3 mA stab., c) 6 V / 6 mA stab., d) Stabilized mixer

2.1. First RF-Stage

According to the information given in (1) the operating point of the first RF amplifier should be $U_{ce} = 9\,V / I_c = 3\,mA$. The temperature response of the original circuit is given in **Figure 2 a**. If one assumes that the circuit will still work satisfactorily with a deviation of ± 25 % from the nominal operating point, it will be seen that this is only valid in a temperature range between 16.5 and 23°C, if the nominal value is in the order of 20°C. Curve »d« is able to extend the usable temperature range to – 2 to + 39°C ! without requiring more components, and only by modifying the circuit to that shown in Figure 1 b. Even if one was able to allow a deviation of the collector current to a value of ± 50 %, the original circuit would only be able to operate between + 11.5 and + 25.5°C, whereas the modified circuit would be able to cover the range – 22 to + 60°C. If the more readily obtainable value 820 Ω is selected for R_C in Figure 1 b instead of the 750 Ω, a voltage of 8.7 V will result at $I_c = 3\,mA$, and thus provide a slightly better stabilization.

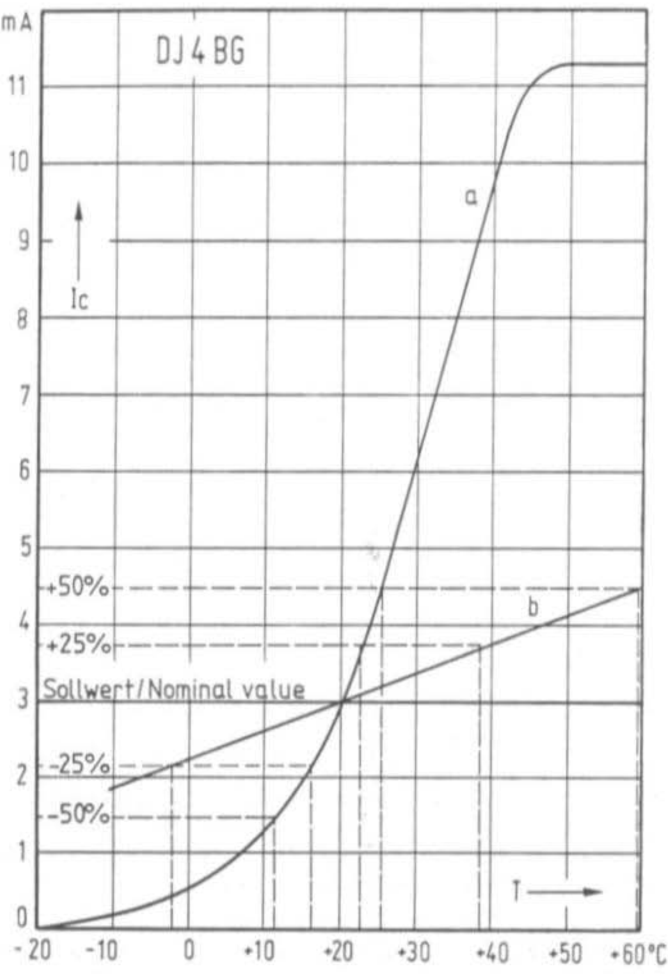


Fig. 2: Temperature response of the RF-stage
Original: a)
Stabilized: b)

2.2. Second RF-Amplifier

According to (1) the operating point of the second RF-stage should be U_{ce} = 6 V / I_c = 6 mA. The temperature response of the collector current of the original circuit is then as given in curve a of **Figure 3.** If a permissible deviation of ± 25 % is assumed, this results in a working range between + 14.5 and + 25.5°C. The stage will go into saturation at approximately 37°C, which means that the gain and thus the sensitivity will no longer be available. The use of the DC-feedback as shown in Figure 1 c results in a collector current response as shown by curve b in Figure 3. This results in a usable temperature range between – 30 and + 70°C !

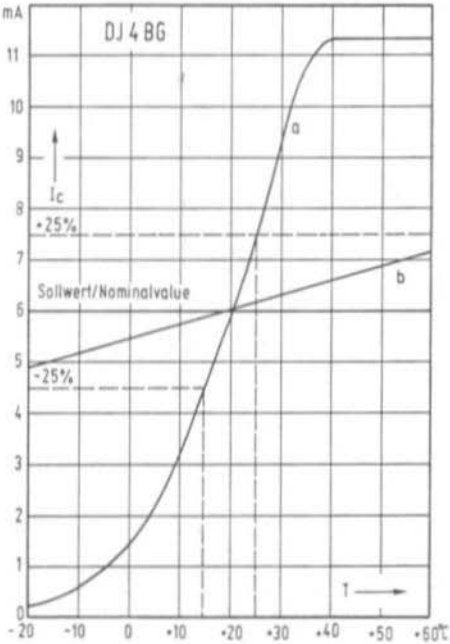


Fig. 3:
Temperature response of the second RF-stage

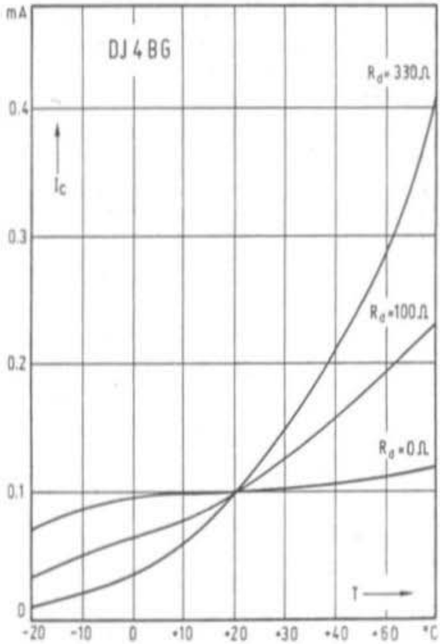


Fig. 4:
Temperature response of the mixer (Id)

2.3. Mixer

Since the mixer is to be aligned so that only a few µA flow without local oscillator signal, whereas several mA flow when this local oscillator signal is provided, the voltage feedback circuit will not be possible. A good stabilization of the collector current is possible by using an additional silicon diode (Figure 1 d). Any low-signal silicon diode such as BAW 76, 1 N 914, or 1 N 4148 is suitable. If it is seen that the diode provides too low a voltage when aligning this stage, in other words if the mixer cannot be opened, it will be necessary to provide a small dropper resistor shown in the diagram as R_d.

This resistor is then able to increase the DC-potential, but will deteriorate the temperature compensation noticeably as can be seen in the curves given in **Figure 4**.

Even though, the collector current is still far less temperature dependent than the original circuit without compensation. In addition to this, the mixer is driven better by the oscillator voltage so that correct operation is to be expected over the whole temperature range to be expected.

3. SUMMARY

It has been seen that the 13 cm transistor converter described by DC 0 DA can be modified easily, as can similar stages, so that the DC-operating points of the transistors can be stabilized and can be used over a wide ambient temperature range. Of course, there are more extensive circuits that can be used that provide an even better stabilization; however, the target of this article was to show how easy it is to stabilize the operating point of transistor circuits using a directly grounded emitter.

As can be seen in Figure 3 of (1), it is possible for the recommended modifications to be made on the DC 0 DA 13 cm converter without difficulties, and without affecting the RF-circuits.

4. REFERENCES

(1) J. Dahms: A Converter for the 13 cm Band Equipped with two Preamplifier Stages and an Active Mixer
VHF COMMUNICATIONS 8 (1976), Edition 4, Pages 194 - 201

THE 70 cm FM TRANSCEIVER »ULM 70«
Part 1: Introduction, Block Diagrams, Variations

by I. Sangmeister, DJ 7 OH · H.J. Franke, DK 1 PN · H. Bentivoglio, DJ 0 FW

1. CONCEPT

Required was an efficient, easy-to-construct 70 cm transceiver to forward 70 cm activity in the ULM area of West Germany. The basic requirements were as follows:

- Inexpensive: using a surplus 10.7 MHz crystal filter
- Stable, crystal-controlled operation for local communications
- Possibility of frequency variable operation, as long as this can be achieved at low expense
- Good sensitivity
- Output power in the order of 1 W so that it is possible to use a built-in accumulator
- Modular construction to simplify construction and modification of the unit
- Use of standard components

Of course, the following modules are not only suitable for use for local communication but can be extended as required to form even rather sophisticated equipment. Of the four versions, only versions A, B, and D are to be described in detail:

A: Two-channel using 2 transmit and 2 receive crystals
B: 16-channel unit with VXO using four crystals each in transmitter and receiver
C: 30-channel unit (using wider range VXO)
D: 100-channel unit equipped with a synthesizer

Full construction details are to be given for version B.
The synthesizer for version D will be described later.

Figures 1 a to 1 d show the simplified block diagrams of the four versions. The stages accomodated on a common board are designated using the same symbols in the corner of the block.

The normal version consists of two equally wide PC-boards having the dimensions 90 mm x 134 mm, and 90 mm x 120 mm. These boards are mounted back-to-back. (The receive board is accessible from above, and the transmit board from below). This means that both boards are easily accessible. Sufficient room is provided in the case for accommodation of a relatively large accumulator. Inspite of the use of normal-sized components not mounted vertically to the board, it is possible for the transceiver to be mounted in a case having the dimensions 140 mm x 185 mm x 67 mm high. There is also sufficient room for accommodation of a synthesizer above the receive board. The synthesizer board also has the dimensions of 90 mm x 134 mm.

The distribution of the stages onto the various PC-boards was mainly determined by the dimensions. The complete transmitter is accommodated on the shortest of the PC-boards, and sufficient room is provided for the whole of the audio stages. The voltage stabilizer and electronic transmit-receive switching for the operating voltages is also to be found on the transmit board.

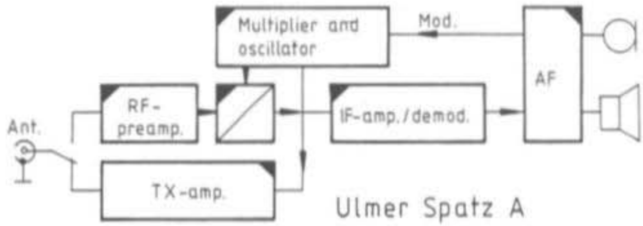


Fig. 1a: Block diagram of the simplest version

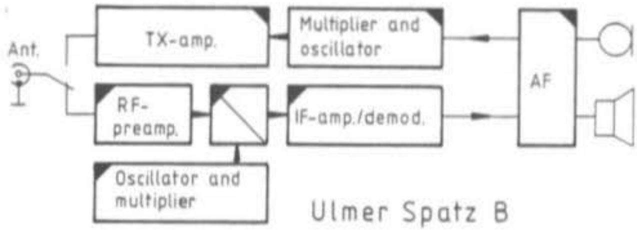


Fig. 1b: The 16-channel transceiver

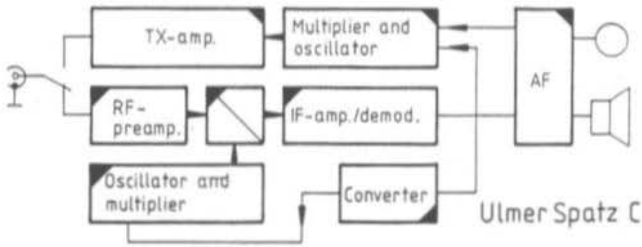


Fig. 1c: 30-channel transceiver with extended VXO

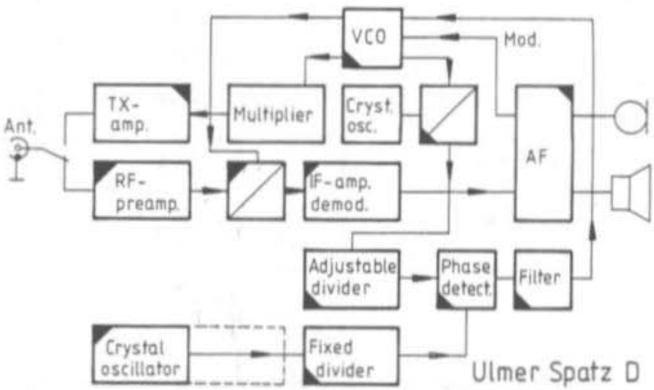


Fig. 1d: 100-channel synthesizer version

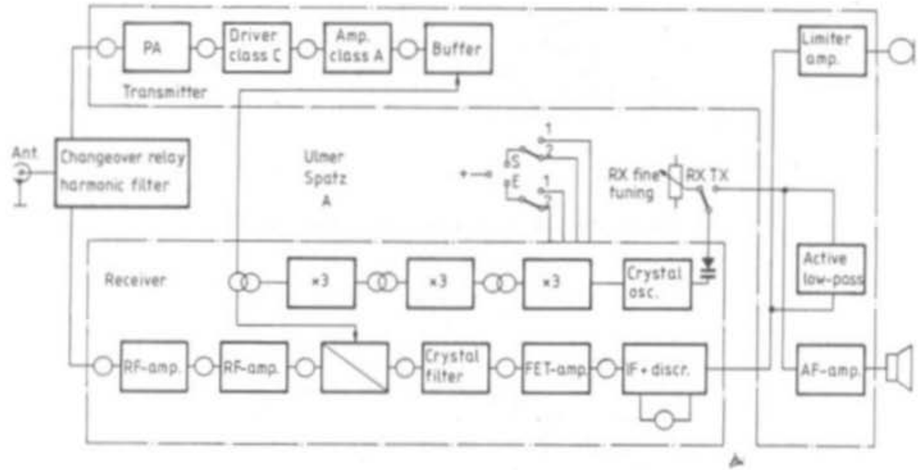


Fig. 2: Detailed block diagram of the simplest version »A«

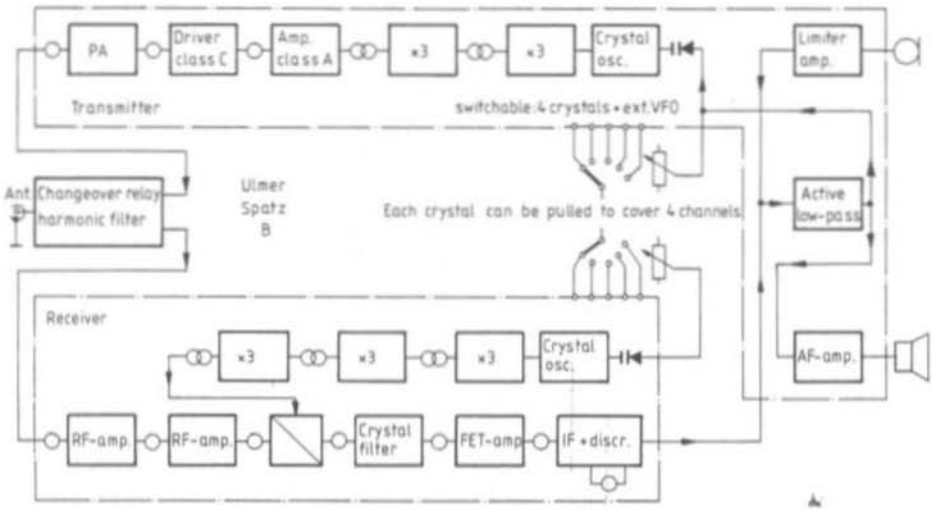


Fig. 3: Detailed block diagram of the 16-channel version (4 crystal VXO)

The antenna relay is part of the additional module »Harmonic Filter« which is mounted directly adjacent to the BNC socket on the rear panel. The longer board accommodates the whole receiver up to the 100 mV audio output. The stripline circuits of the RF-amplifier stages take up a large amount of the surface of this PC-board.

A more detailed block diagram of version B is shown in **Figure 3**; the previously mentioned distribution of the stages to the various modules can be seen clearly. The receiver line-up is as follows: Two-stage RF amplifier equipped with the well-proved stripline transistor AF 279, a mixer equipped with the inexpensive 1.2 GHz silicon transistor 2 N 5179. An intermediate frequency of 10.7 MHz was selected since the majority of surplus FM crystal filters are available for this frequency. The selection of this relatively low intermediate frequency for a 70 cm transceiver simplifies the rest of the circuitry: with the exception of the matching stage equipped with a dual-gate MOSFET, it is only necessary for the universal integrated IF circuit CA 3089 to be used. Of course, an intermediate frequency of 21.4 MHz would provide a better image rejection, however, it would be necessary for a dual-conversion system to be used to a lower second intermediate frequency in order to avoid an expensive 21.4 MHz crystal filter.

Provision is made for connecting an S-Meter, an adjustable squelch, and an external VFO. The VFO connection is electronically switched together with the crystals so that it is possible for the coaxial cable to remain connected.

The use of three frequency multipliers in the local oscillator chain of the receiver seems to be in contrast to the requirement for an inexpensive construction; however, this method is simpler, and not more expensive than when using a higher frequency oscillator. The crystals oscillate in a frequency range where they are least expensive. However, the separate oscillator frequency chain for the transmitter could be avoided. This results in the cheaper version »A« which uses a common frequency plan for transmitter and receiver, and is suitable for two-channel use. The detailed block diagram of this version is given in **Figure 2**. The same PC-boards are used as in version »B«, however, the oscillator and multiplier parts of the transmit board are not equipped. It should be noted, that the switching and pulling of the crystals, as well as the tuning of the multipliers is simpler when each board is completely equipped and can be aligned individually.

The three-stage transmitter provides an output power of approximately 1 W. All AF-stages are integrated in a four-stage operational amplifier and considerably simplified with respect to previous designs. Even though, the transceiver is provided with a clipper and active low-pass filter, as is to be expected from a modern transceiver.

A detailed block diagram of a transceiver equipped with a simple synthesizer for 100 channels is shown in **Figure 4**. CB-equipment has shown that synthesizers need not only be found in expensive equipment. Of course, some deterioration of the technical specifications could be accepted; for instance, the spectral purity of the VCO signal is deteriorated by tripling the frequency twice. The conventional arrangement using a VCO oscillating at the final frequency together with an ECL-divider would have led to an unacceptably high current drain. The current drain of the 48 MHz synthesizer only amounts to 28 mA at 12 V. The multiplier stages and the crystal oscillator require approximately 15 mA.

A special feature of the synthesizer is the use of the cheap C-MOS-IC CD 4017 which can be directly connected to a normal rotary switch. The phase detector is a sample-hold-type which means that the VCO signal is free of interfering noise, even without any special measures. The PC-board of the synthesizer is not so tightly packed as the other boards, and the component cost is somewhat lower.

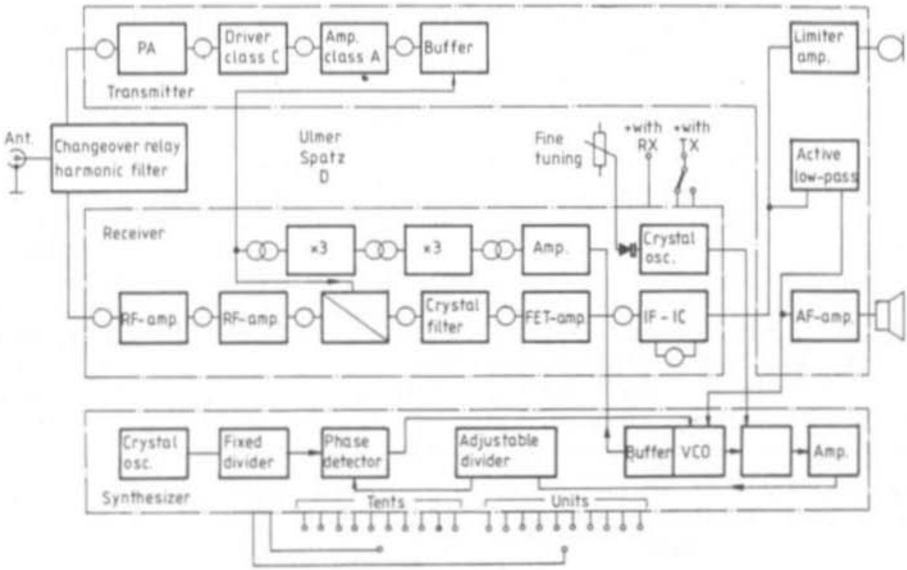


Fig. 4: Block diagram of the synthesizer version »D«

The construction details to be described in the next editions of VHF COMMUNICATIONS, do not only describe the PC-board modules, but also the overall transceiver.

Fig. 5: Photograph of the author's prototype

A SPECTRUM ANALYZER FOR AMATEUR APPLICATIONS

by E. Berberich, DL 8 ZX

Due to their versatility and wide range of applications, it would be difficult to imagine a professional laboratory or even a larger workshop without a spectrum analyzer. In the narrow band mode, it is possible to easily indicate the noise sidebands, modulation depth and characteristics, as well as drive-dependent distortions such as intermodulation etc. In the wideband mode, spurious and harmonic waves can be shown simultaneously over a wide frequency and level range. This allows both a quick and clear evaluation, and rapid alignment. In spite of the great difference in price, spectrum analyzers have become much more popular than the well-established selective voltmeters etc.

Spectrum analyzers are greatly respected by radio amateurs, since they allow a clear representation of all required and unwanted signals that are generated in a transmitter in the oscillator, and frequency multiplier stages. This means that those amateurs that have access to a spectrum analyzer have distinct advantages.

Home construction of a spectrum analyzer is most certainly possible if certain compromises are made with respect to frequency and dynamic range, and measuring accuracy. This article is to describe means of achieving this aim.

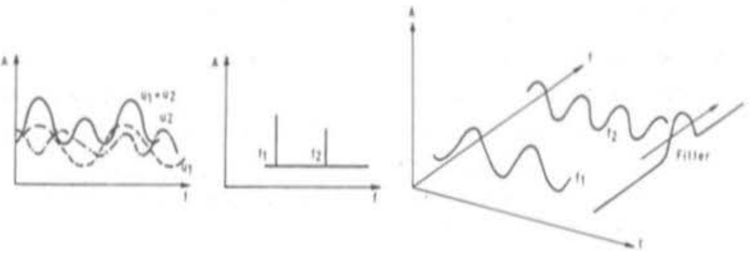


Fig. 1: Two sinusoidal voltages, shown in time-domain and frequency-domain, resp.

1. MEASURING PRINCIPLE

The amplitude of an electrical oscillation can be displayed on a spectrum analyzer in two ways: As a function of time (time domain measurement) using an oscilloscope, or as a function of frequency (frequency domain measurement). **Figure 1** shows how both types of display work. The simplest form of spectrum analyzer would be a resonant circuit with subsequent rectifier whose resonance can be tuned over the frequency range of interest. This would have to made in synchronous with the deflection voltage of the oscilloscope used for indication. Due to the low tuning range, low selectivity, and since selectivity and matching cannot be kept constant over the tuning range, this approach is not satisfactory and more extensive circuitry is required.

In principle, spectrum analyzers are swept frequency, superhet receivers with a highly selective IF circuit which is used as receive window. The amplitude is displayed on the screen of a CRT or drawn on a plotter.

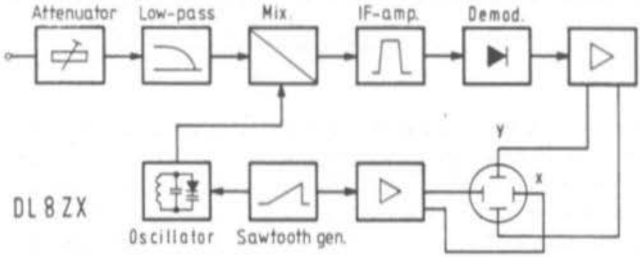


Fig. 2: Principle of a spectrum analyzer

2. PRINCIPLE OF OPERATION

Figure 2 shows the basic principle of a spectrum analyzer: Single-conversion superhet with calibrated attenuator and low-pass filter at the input. The local oscillator frequency is tuned across the range of interest using a sawtooth voltage. This voltage is also used for the X-deflection of the display unit which then runs synchronously. Such equipment is usually known as a panoramic receiver. As measuring unit, a spectrum analyzer should at least fulfil the following minimum requirements:

Calibrated display of the amplitude (Y-plane) in both a linear and logarithmic scale

Adjustable display width in the frequency axis (X-plane) with frequency markers.

Selectable filter bandwidth in the IF circuit to allow various resolutions.

Variable, calibrated attenuator in front of the measuring signal input which must have a defined, real impedance over the whole range of interest.

It is in the first point above that the spectrum analyzer differs greatly from a panoramic receiver, which usually only uses a controlled IF-amplifier with all its control time-constant problems. The uncontrolled, logarithmic IF amplifier forms the heart of this description.

The amplitude range displayed in the Y-plane amounts to approximately 100 dB in the case of modern professional spectrum analyzers. This dynamic range cannot usually be utilized during the display of frequency spectrums, since the intermodulation of the first mixer cannot normally be brought below 70 dB. However, the dynamic range is increased by use of the calibrated attenuator. The calibrated attenuator also allows any overload of the mixer to be seen: If, for instance, the attenuation is increased by 10 dB and the spectral line is decreased by more than 10 dB, then an overload condition is present. Any spectral lines caused by overload conditions within the measuring system itself will always be non-linear.

The required large frequency range of a spectrum analyzer makes it impossible to use a tuned circuit before the first mixer circuit. One way out is to place the first intermediate frequency above the highest input frequency (about twice the frequency), and to use a low-pass filter at the input to suppress spurious reception points. This leads to a clear spectrum display. The subsequent mixer uses a 50 Ω wideband ring mixer.

It is necessary to use multiple conversion to obtain the required selectivity when using such a high IF. Crystal filters can be used with advantage to obtain the main IF-selectivity. However, since each of the spectral lines runs through the passband range of the filter or filters, special attention must be paid to the transient behaviour. The narrower and steeper the filters slopes, the slower will have to be the sweep.

3. A SPECTRUM ANALYZER FOR AMATEUR APPLICATIONS

The following considerations were made during the conception of an amateur spectrum analyzer:

The main requirements are in the frequency range of 0.5 to 60 MHz and 120 to 180 MHz. Most of the amateur bands, oscillator and intermediate frequencies are within this range. The UHF bands can be converted to one of these two bands.

If an IF of 60 MHz is selected, it would be possible to cover 0 to 60 MHz (lower sideband) and 120 to 180 MHz (upper sideband) with just one local oscillator range of 60 to 120 MHz. The required sideband range is selected by placing a multistage highpass or lowpass filter in front of the input. The rejection of this filter determines the image rejection.

In order to use inexpensive 10.7 MHz crystal filters, the spectrum analyzer was designed as double conversion superhet **(Figure 3)**. When using a second IF of 10.7 MHz, both oscillator frequencies of 70.7 and 49.3 MHz would be possible. Since the later falls within a range of interest, 70.7 MHz was selected.

A 10.7 MHz FM crystal filter with a bandwidth of 15 kHz is used in narrow band IF circuit. If required, a narrower bandwidth can be obtained by placing an SSB crystal filter with a bandwidth of 2.1 kHz in series with the FM-filter. The increase in loss can be compensated for in an additional amplifier. The crystal filter is bypassed in the wideband mode.

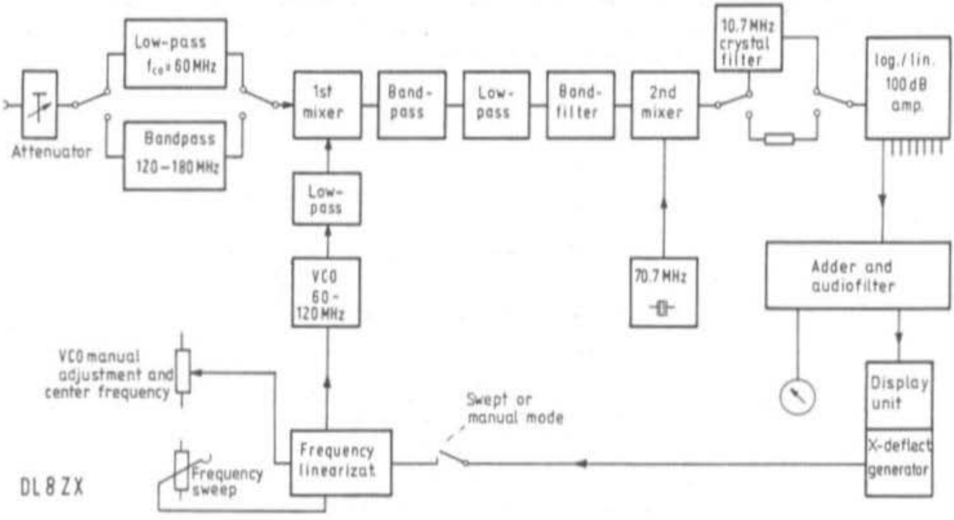


Fig. 3: Block diagram of an amateur spectrum analyzer

The resolution bandwidth is limited due to the residual FM of the local oscillator signal. In the case of very narrow band measurements, it would be possible to lock in the first local oscillator every 1 MHz and to sweep the second oscillator by ± 0.5 MHz.

Screened metal cans were used for the mechanical construction. This is necessary in order to obtain the full selectivity offered by the crystal filter. In this manner it was possible to eliminate all but one spurious reception point: twice the first IF (21.4 MHz) and even this is more than 100 dB down on the maximum level to be indicated. The heart of the spectrum analyzer is the logarithmic amplifier which is accommodated on a PC-board. This module is to be described in greater detail to allow home construction.

3.1. Circuit Description

Due to the complexity, the various modules of the spectrum analyzer are to be described individually. However, the display unit and its amplifiers, the sawtooth generator and the power supply are not to be discussed in detail.

3.1.1. Input Circuit

The input circuit comprises a calibrated attenuator (switchable or continuously variable), a lowpass filter with a cutoff-frequency of 60 MHz, a bandpass filter of 120-180 MHz, a two-pole changeover switch, and a wideband Schottky ring mixer. The circuit diagram of this module is shown in **Figure 4**. In contrast to the basic concept, a bandpass filter is used in order to suppress any interference from harmonics. It is important that all three connections of the ring mixer are terminated by a real impedance. The termination at the IF output of the mixer is at high impedance for the intermediate frequency. This is obtained by providing a parallel resonant circuit which isolates the actual IF from the 56 Ω terminating resistor. The gain of the required frequencies is therefore higher. The output coupling to the next module is in the form of a series resonance circuit and thus selective. Either two reed-relays or miniature relays RH-12 can be used for switching the input filter.

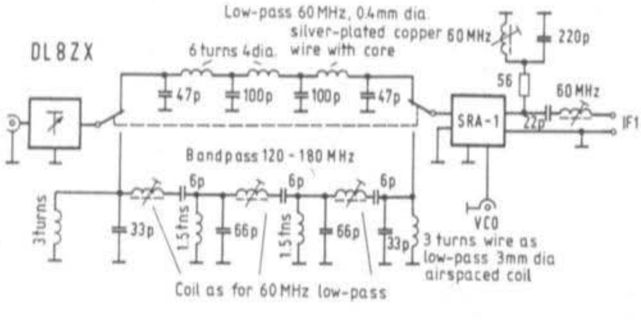


Fig. 4: Circuit diagram of the input filters and first mixer

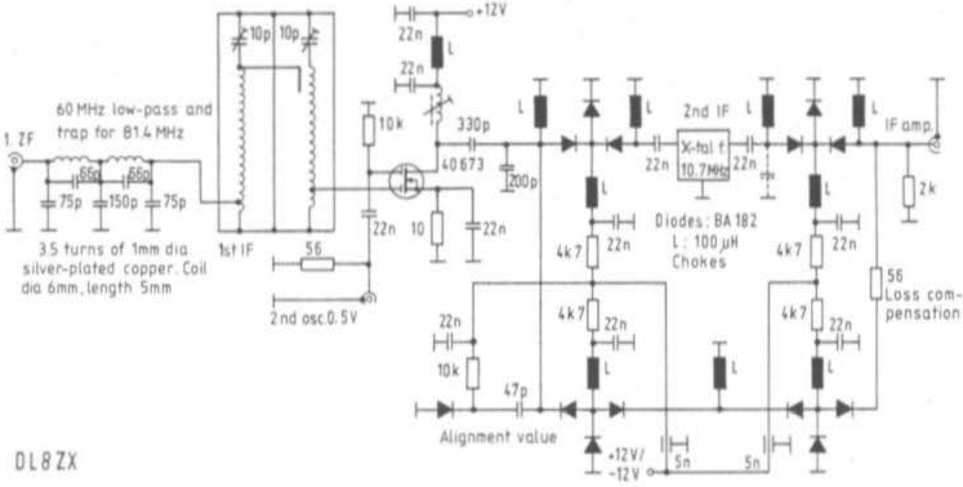


Fig. 5: Circuit diagram of the first IF amplifier, second mixer and filter switching

3.1.2. First Intermediate Frequency, Second Mixer, and Crystal Filter

The circuit diagram of this module is shown in **Figure 5**. The first IF of 60 MHz is fed via a lowpass filter with a cutoff frequency of slightly over 60 MHz. A trap is provided for the image frequency of 81.4 MHz (60 MHz + 2 x 2nd IF). This is formed by the resonance of the inductances with the 66 pF capacitors, which increase the suppression of this critical frequency to more than 40 dB. A helical bandpass filter is now provided, whose bandwidth is aligned to 1 MHz with the aid of the capacitive coupling. If the previously mentioned narrowband mode sweeping the second local oscillator is not required, it is possible for the helical filter to be aligned to a bandwidth of approximately 200 kHz. This will reduce the danger of intermodulation products being formed in the second mixer.

A dual-gate MOSFET is used as second mixer, whose high-impedance input allows a slight upward transformation of the first IF signal. The coaxial connection from the second oscillator is terminated with a 56 Ω resistor so that the cable length is not critical. This resistor also makes gate 2 low-impedance for the second IF in order to avoid feedback. The best intermodulation characteristics were obtained with the given oscillator level of 0.5 V (rms) and a source resistor of 10 Ω.

The second IF of 10.7 MHz is filtered out in the drain circuit and transformed to the required impedance for the crystal filter (in our case 2 kΩ). Either the crystal filter or another filter are switched into circuit using a diode switch. Only a wideband choke used to obtain the diode currents is shown in the circuit diagram; however, a multiple stage type with, for instance, a bandwidth of 100 kHz can be used. In the author's prototype, the selectivity of the 10-stage IF amplifier is used.

It is important, that the ultimate selectivity of 100 dB is not deteriorated by an unfavourable construction or by the diode switch. For this reason, the crystal filter should be accommodated in a screened metal box. The diodes are BA 182 types, which are used as switching diodes for range switching of Band III TV tuners. They are manufactured by almost all European semiconductor manufacturers. The DC for the diodes must not be fed via the crystal filter since the toroid transformer in the filter would produce non-linearities as soon as it goes into saturation. For this reason, approximately 100 µH chokes should be used in the connections to the diodes.

The difference in insertion loss between the different filters is compensated for using a series resistor (in our case 56 Ω). A switchable capacitor (in our case 4.7 pF) is used to ensure that the drain circuit is not detuned.

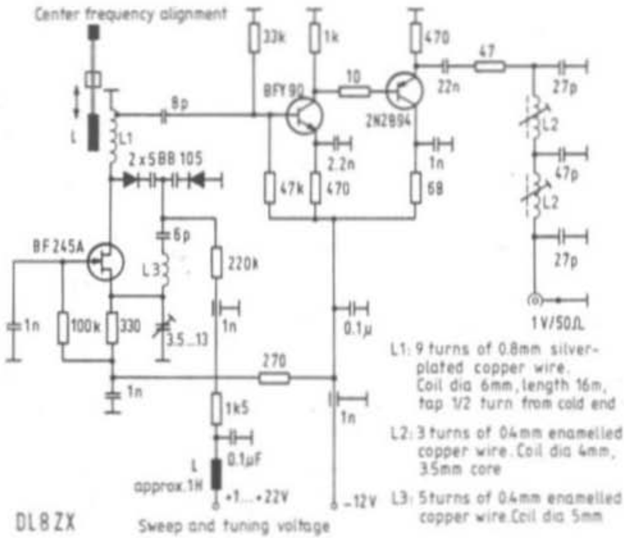


Fig. 6: The VCO for 60 to 120 MHz

3.1.3. First Local Oscillator

The circuit of the first local oscillator is shown in **Figure 6**. This oscillator is swept by the sawtooth deflection voltage from the display unit. In order to obtain the required tuning range of 1 : 2 (60 to 120 MHz), it is necessary for the oscillator circuit to be designed with the lowest possible commencement capacitance. It was found that this wide frequency range could be swept by parallel connection of ten UHF-tuning varactors BB 105. Frequency-dependent correction circuits were provided in the feedback path in order to keep the amplitude as constant as possible. A field-effect transistor in common gate configuration was found to be suitable both with respect to frequency stability and noise. It is important that all noise and hum voltages must suppressed, since a residual voltage of only 330 µV is required to cause a frequency variation of 1 kHz.

The actual oscillator is followed by two amplifier stages. The voltage amplifier stage is galvanically coupled to the subsequent impedance converter. A 10 Ω resistor is provided to suppress any tendency to UHF oscillation. The lowpass filter has an impedance of approximately 50 Ω and a cutoff frequency of somewhat over 120 MHz.

3.1.4. Second Local Oscillator

The second oscillator is crystal-controlled, but can be converted to a free-running LC-oscillator if the narrowband mode with swept second oscillator is required. A 70.7 MHz crystal is excited at its third overtone using a Butler-circuit equipped with FET's **(see Figure 7)**. This circuit can be used successfully upto 200 MHz. It is, however, necessary to neutralize the capacitance of the crystal holder, and LN is provided for this reason. The local-oscillator signal is fed to the second mixer via a buffer amplifier and bandpass filters. A voltage of 0.5 V at 50 Ω is available at this point. The bandpass filters are aligned for a bandwidth of 1 MHz.

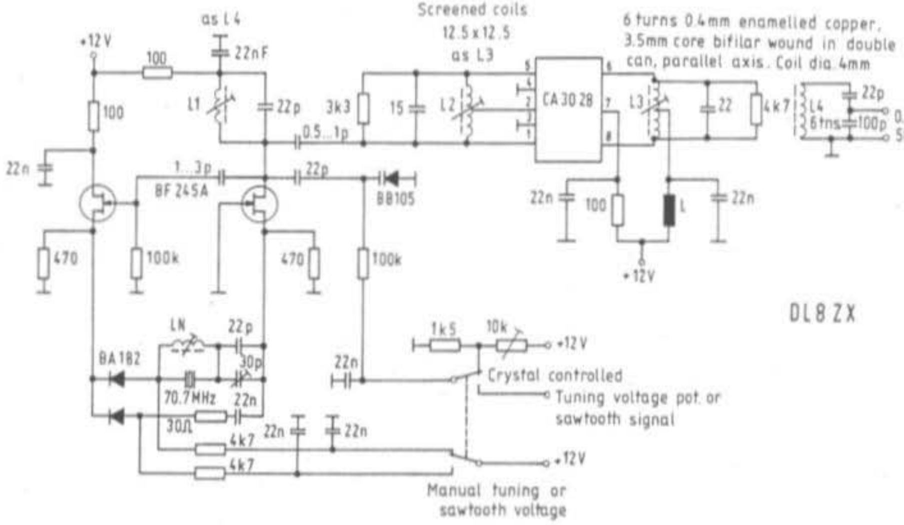


Fig. 7: The second oscillator (70.7 MHz)

In the narrowband mode, the series-resonance impedance of the crystal is replaced by a 30 Ω resistor which is switched into circuit using switching diodes. It is then possible to tune the drain circuit using a varactor diode. If the sawtooth voltage is too fast for measurements on narrowband IF filters with steep slopes, it is possible for this to be made manually with the aid of a potentiometer. Of course, in this case either a plotter or long-persistence CRT must be used.

3.1.5. Logarithmic IF Amplifier

The IF amplifier and its gain control is one of the most important modules, not only for the application in question. Nearly every amateur complains about the error of his S-meter. A logarithmic indication would be ideal, which could then be directly calibrated in dB. In many receivers, this is attempted in the AGC-circuit. In the case of dynamically-tuned equipment, such as spectrum-analyzers, this is difficult due to the transient effects of the control circuit.

These problems are avoided if a so-called successive-detection amplifier is used in the IF circuit. This principle is well-known in measuring technology. The author has developed a ten-stage amplifier for 10.7 MHz according to this principle **(see Figure 8)**. This amplifier comprises ten identical stages with a defined, adjustable gain. Each individual stage possesses a selective circuit for 10.7 MHz and an RF-detector circuit **(see Figure 9)**. The gain of each stage is adjusted to exactly 10 dB.

The logarithmic characteristics are obtained by addition of the rectified currents from the ten amplifier stages. If one assumes that no rectification takes place under no-signal conditions (excluding noise), the indication will be zero. If an RF-voltage is now fed to the circuit that is above the noise threshold (approx. 1 μV), the tenth stage will commence rectification. On increasing the input voltage, the rectified current will increase to approximately 10 μA where the stage goes into saturation and the current will remain constant.

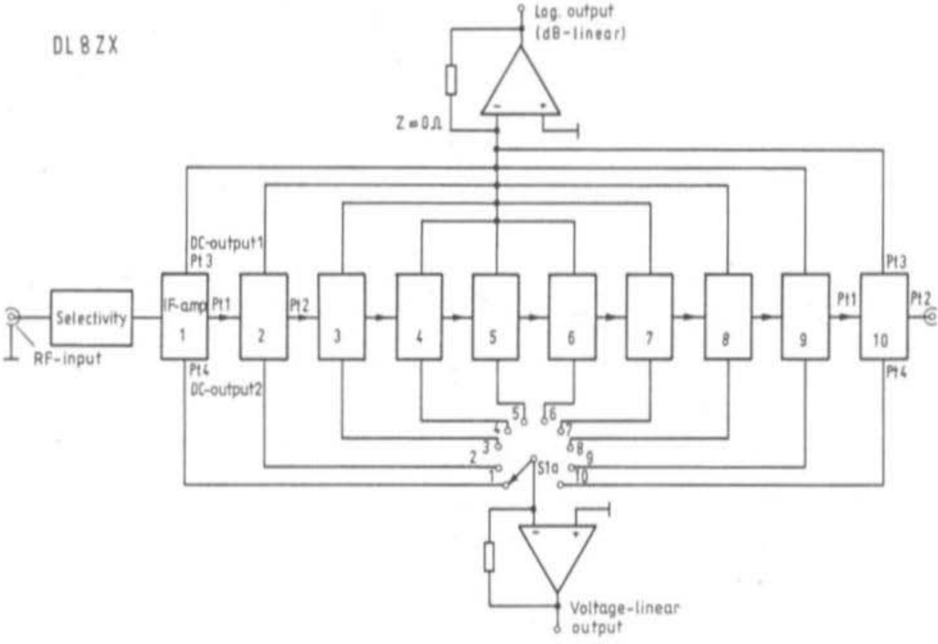


Fig. 8: Principle of the log./lin. amplifier with 100 dB dynamic range

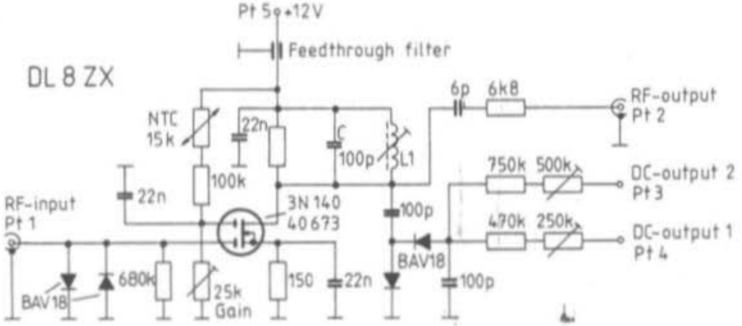


Fig. 9: A single IF amplifier stage (10.7 MHz)

In the meantime, the ninth stage will be generating a direct current which will increase on increasing the RF-voltage until the ninth stage also goes into saturation at approximately 10 µA. If the input voltage is increased in 10 dB steps upto 300 mV at the input Pt 1, each of the detectors will provide 10 µA. The ten individual currents are added in an operational amplifier; the resulting output signal is passed through an active filter for noise suppression and fed to the Y-plane amplifier of the display unit or plotter. A linear-dB scale can be used **(see Figure 10)**.

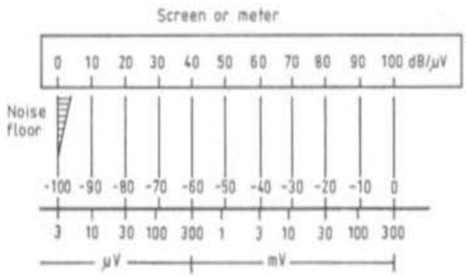


Fig. 10: Dynamic range of the IF amplifier

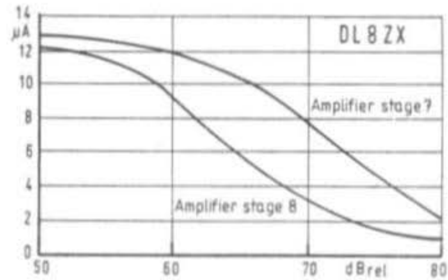


Fig. 11: Rectified currents of two individual IF stages

The successive detector amplifier possesses both a logarithmic and linear output, which allows either of these modes to be selected. In order to display the voltage-linear indication, the DC-voltage of the IF-stage still operating in its linear mode is selected with the aid of a rotary switch. To ensure that the linear outputs are always loaded, a second wafer is provided on the selector switch that shorts out all unused outputs. If this measure was not taken, amplitude errors would occur in the logarithmic circuit due to the load variations on the selector switch.

Approximately 20 dB are displayed linearly on the plotter or display unit. In order to obtain an exactly voltage-linear display, it is possible for the characteristic of the diode to be equalized. However, it is easier to make a corresponding scale. The original scale of a 100 µV meter can be used in the logarithmic mode. **Figure 11** shows the direct currents of two stages as a function of the RF-input voltage. These curves are typical for all 10 stages.

3.2. Construction and Alignment Details for the Logarithmic IF-Amplifier

The IF amplifier can be accommodated on the 260 mm by 60 mm single-coated PC-board DL 8 ZX 003. **Figure 12** shows part of this PC-board. The conductor lanes and components of each of the ten stages are identical. The RF-output of each stage is connected to the input (Pt 1) of the subsequent stage using a short wire bridge on the conductor side. The two DC-outputs for the logarithmic and linear outputs, and the operating voltages of each stage are fed via feedthrough capacitors that are soldered into position in the 30 mm high screening panels. **Figure 13** shows a photograph of part of the authors prototype.The PC-board is soldered into place about 10 mm from the bottom of the screening panel. Due to the construction and use of screened inductances, no screening panels are necessary between the individual stages. The gain of each stage can be adjusted with the aid of trimmer resistors; the trimmer resistors for the DC-outputs are not accommodated on this board.

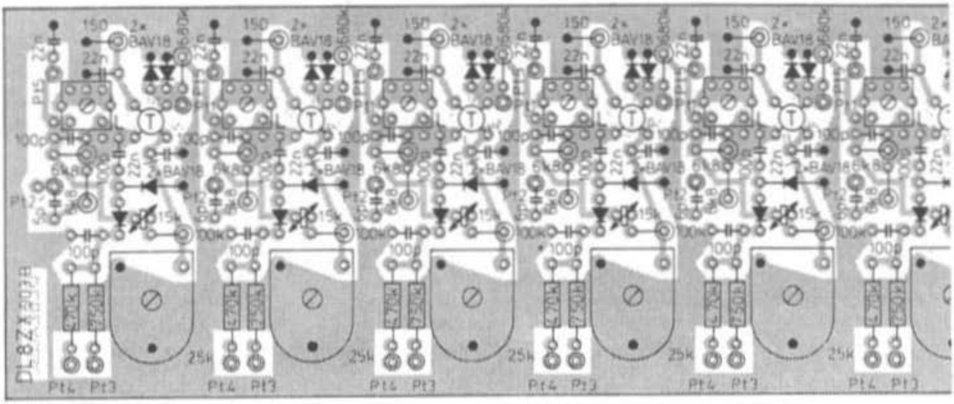


Fig. 12: PC-board with components location plan of the log./lin. IF amplifier DL 8 ZX 003

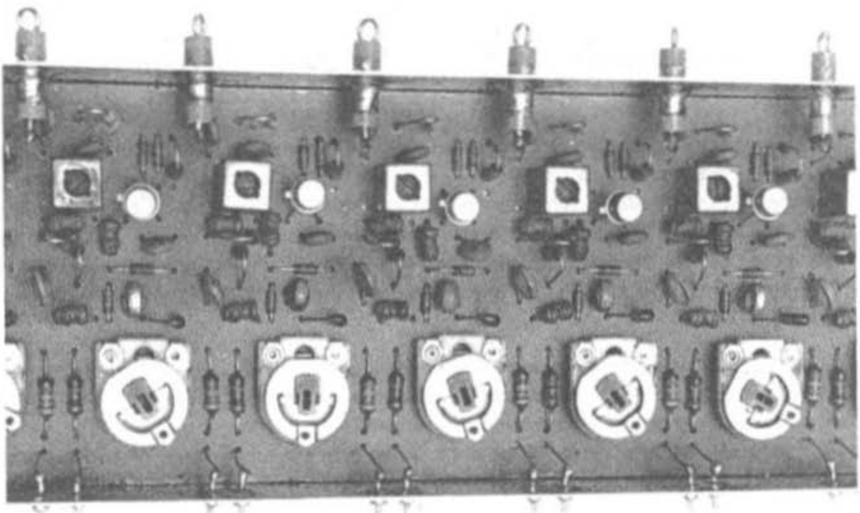


Fig. 13: Author's prototype of the log./lin. IF amplifier DL 8 ZX 003

3.2.1. Special Components

Transistors: 3 N 140, 40841, 40673 or other similar dual-gate MOSFET (RCA)

Diodes (4 pieces per stage): BAV 18 (ITT, TI, Philips), BAW 76 (Siemens) or 1 N 4151 (fast switching diode)

Thermistor: 15 kΩ (Philips)

Inductance: 2.1 μH (for 10.7 MHz and 100 pF)
Screening can 7.5 mm by 7.5 mm by 12 mm high

Trimmer potentiometers for gain adjustment: 25 kΩ, spacing 12.5/10 mm, preferably ceramic

All resistors: composite carbon types

All capacitors: ceramic disc types

Selector switch: Wafer 1: 26 positions
Wafer 2: 26 positions all shorted except one
(Type SB 50, ITT-Germany)

3.2.2. Alignment

The DC-outputs of the individual amplifier stages work into very low impedance consumers (virtually 0 Ω). For alignment it is necessary for these conditions to be simulated by grounding all logarithmic and linear outputs. The gain control trimmers (500 kΩ, 250 kΩ) are set to approximately mid range. The RF-input (Pt 1) is now terminated with 50 Ω and connected to a signal generator (or 10.7 MHz oscillator [RF-tight] via a calibrated attenuator).

1. Connect the operating voltage of 12 V. The current drain should be approximately 80 mA but can differ considerably. Adjust the gain controls so that approximately 1 V can be measured at G 2 (High impedance voltmeter).
2. Inject a signal of 300 mV (rms) at 10.7 MHz. The first stage should go into saturation at this level.
3. Connect a VTVM (digital voltmeter or oscilloscope with DC-input) with an input impedance of at least 10 MΩ to the 100 pF charge capacitor of the rectifier in the first stage. Align the resonant circuit for maximum reading. Adjust the trimmer potentiometer so that 10 V are present across the capacitor.
4. Reduce the RF-level by 10 dB (voltage divided by 3.16). Connect VTVM to second stage (100 pF capacitor) and repeat alignment as described in 3.
5. Repeat alignment as described in 3. and 4. for each of the ten stages. Repeat the alignment and then leave the trimmer potentiometers in this position.
6. Connect the logarithmic outputs to the inverted input of the operational amplifier, and the VTVM to the output of this op. amplifier.
7. Select a DC-voltage for the 0 dB point, for instance 10 V DC = Full scale for 300 mV input voltage.
8. Adjust the signal generator for 10 μV output.
9. Align the 500 kΩ trimmer resistor at the logarithmic output of stage 10 so that the meter indicates 1/10th of the voltage selected for 0 dB; 1 V in our example.
10. Increase the output voltage of the signal generator by 10 dB (to approx. 32 μV).
11. Align the trimmer at the log. output of stage 9 so that the meter indicates 2/10 th of the 0 dB voltage (e.g. 2 V).
12. Carry out alignment steps 10. and 11. for all stages.

13. Repeat this alignment from stage 8 onwards and check the linearity of the logarithmic scale. A dB-linear scale should result, and the error should be less than ± 1 dB/10 dB. At low input levels the errors will be greater due to the residual noise. The noise should not be greater than 4 dB.

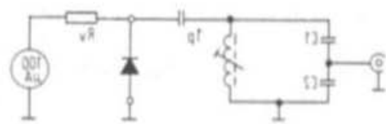
14. Connect the linear outputs to the two wafers of the rotary switch (selector and short-circuit wafers). Connect the wiper of the selector wafer to the input of the other op. amplifier (Z_{in} approx. 0 Ω), and connect the meter to this output.

15. Switch the selector switch from position 1 to 10 and change the level by 10 dB for each step. The indicated output voltage should not change. Any variations can be compensated using the 250 kΩ trimmers. Stages 9 and 10 will show noise components at the linear output.

The voltage linearity is, of course, limited due to the use of silicon diode rectifiers. However, the linearity should be sufficiently good within the 10 dB range.

Diodes having identical efficiency would improve the scale linearity over the various stages. The simple circuit shown in **Figure 14** is suitable for selection of suitable diodes. The ratio of C 1 to C 2 should amount to 1 : 10. The resonant circuit is aligned to the nominal frequency.

Fig. 14: Circuit for testing diodes



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(3) Kestler, J.: Matching Circuits for Schottky Ring Mixers
VHF COMMUNICATIONS 8, Edition 1/1976, Pages 13 - 18

A TRIANGULAR-WAVE GENERATOR

by H.J. Ehrke, DC 7 LE

A spectrum analyser for amateur applications is described by Eugen Berberich in this edition. A triangular-wave generator for feeding such a spectrum analyser, panoramic receiver or sweep generator is to be described in this article.

A simple circuit is to be described that generates the deflection voltage for the oscilloscope and control voltage for the voltage controlled oscillator (VCO) at the same time. An automatic deflection with the aid of a triangular voltage is to be provided. Since the voltage is symmetrical in contrast to that of a sawtooth voltage, any distortion caused by too high a deflection frequency will be immediately visible, e.g. as double passband curves or spectral lines.

1. CHARACTERISTICS

The deflection frequency, sweep deviation and center frequency are adjustable. In the manual deflection mode, it is possible for the frequency of the VCO to be varied manually, and the curve traced on the oscilloscope will be identical to that made in the automatic mode. If a frequency counter is connected to the output of the VCO, it is possible for passband curves and individual spectral lines to be measured exactly. This means that a frequency marker generator will no longer be required, and that frequency measurements will be more accurate.

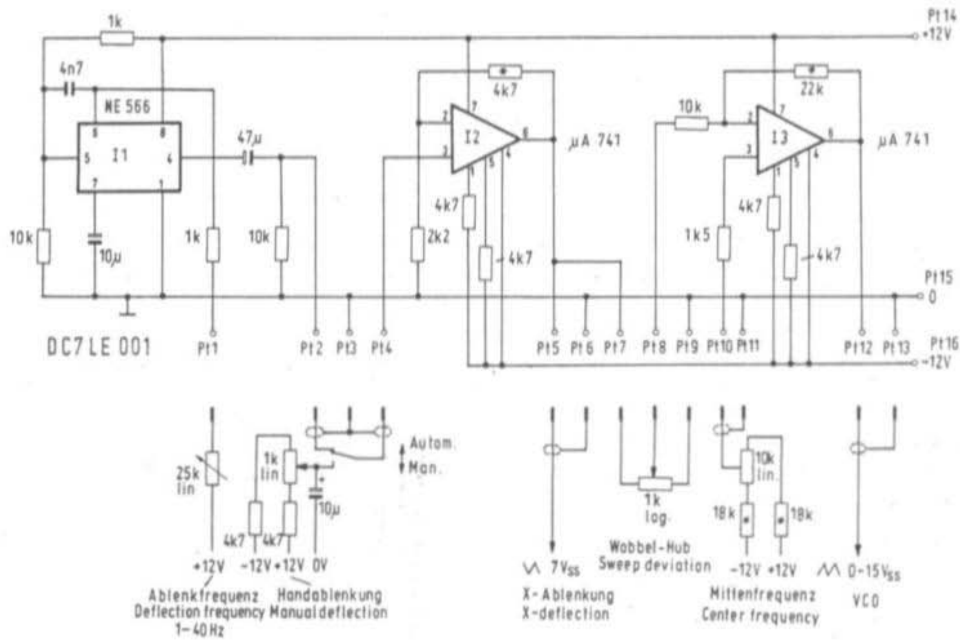

Fig. 1: Triangular-wave generator, deflection frequency, sweep deviation, and center frequency are all adjustable

2. CIRCUIT

The circuit diagram of the triangular-wave generator is given in **Figure 1**. A triangular wave voltage is generated in an integrated circuit NE 566, which is amplified in the two operational amplifiers I 2 and I 3 to the required output level. The resistance values given in the circuit are for orientation and are values suitable for use with a sweep generator and oscilloscope HAMEG HM 207 and a simple VCO. The resistors designated with »+« will probably have to be changed to suit the oscilloscope and VCO used.

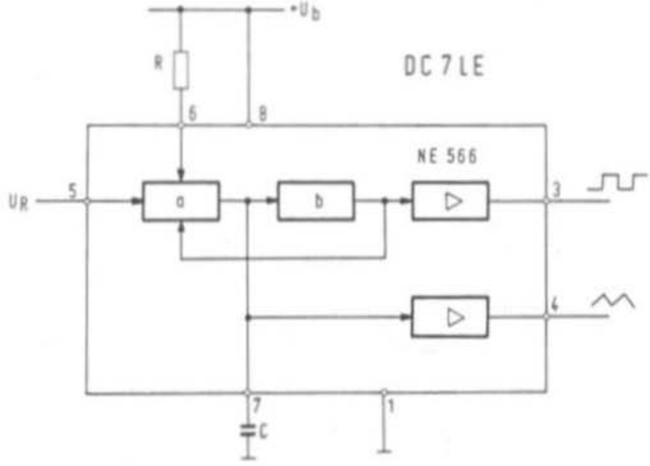

Fig. 2: Block diagram of the IC NE 566

2.1. Operation of the NE 566

The integrated circuit NE 566 manufactured by SIGNETICS is a voltage-controlled oscillator with a square-wave and triangular-wave output. The operation of this integrated circuit is to be explained with the aid of the block diagram given in **Figure 2**. The capacitor C is charged and discharged via resistor R and the voltage-controlled current source »a«. The charge voltage of this capacitor is present at the input of the Schmitt-trigger »b« which switches the current source (from charge to discharge and vice versa). Since the charge and discharge current are constant and equal, the curve of the voltage across C is linear and possesses a triangular characteristic. This voltage is fed via a buffer amplifier to output 4 and is used in the described circuit. The output (switching) voltage of the Schmitt-trigger is also available via a buffer at output 3.

The maximum frequency of the VCO is approximately 1 MHz, and is determined by $U_{control}$, R and C.

Further technical data and calculations, including information on using this integrated circuit for phase-locked loops are given in (1) and (2).

3. CONSTRUCTION

The described triangular generator is accomodated on a PC-board of 95 mm by 55 mm. The component locations and conductor lanes of this board, which has been designated DC 7 LE 001, is shown in **Figure 3**. The components and construction are not critical. The integrated circuit NE 566 is manufactured by SIGNETICS, whereas the well-known operational amplifier 741 is manufactured by a large number of companies; the Siemens designation is TBA 221 B.

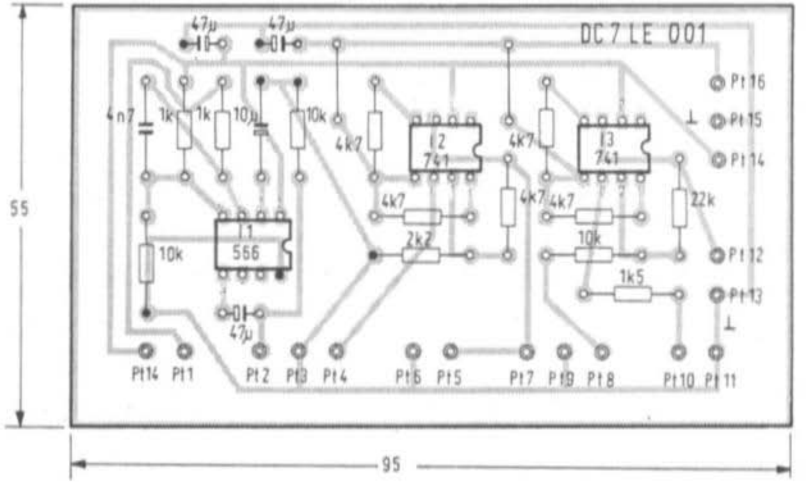

Fig. 3: PC-board DC 7 LE 001

4. REFERENCES

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(2) SIGNETICS Applications, 1974

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MATERIAL PRICE LIST OF EQUIPMENT
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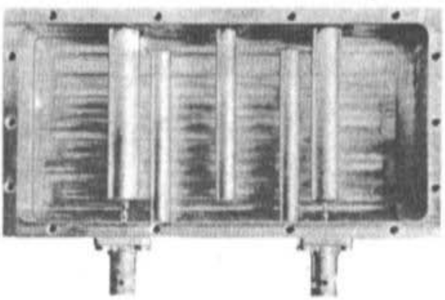
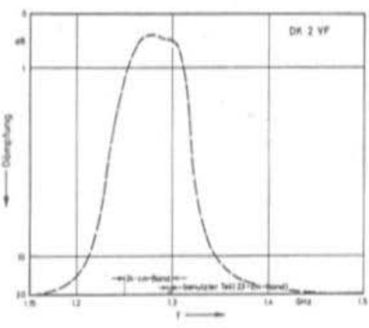
PC-board	DC 8 NR 007	(with printed plan)	DM 18.—
DL 8 ZX 003	**LOGARITHMIC IF-AMPLIFIER for SPECTRUM ANALYZERS**		**Ed. 2/1977**
PC-board	DL 8 ZX 003	(with printed plan)	DM 17.—
DC 7 LE 001	**TRIANGULAR WAVE GENERATOR**		**Ed. 2/1977**
PC-board	DC 7 LE 001	(with printed plan)	DM 10.—
Minikit	DC 7 LE 001	(3 IC's, 5 electrolytics, 1 ceramic cap.)	DM 27.—
Kit	**DC 7 LE 001**	**with above parts**	**DM 36.—**

INTERDIGITAL BANDPASS FILTER FOR 23 cm

Similar to that described in edition 3/71 VHF COMMUNICATIONS





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

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Single Band	HW-10	10	29.00	1000	750	1.25
Single Band	HW-15	15	21.20	500	750	1.25
Single Band	HW-20	20	14.25	250	750	1.85
Single Band	HW-40	40	7.05	100	450	1.85
Single Band	HW-80	80	3.75	50	350	1.85
Single Band	HW-880	80	3.75	60	350	2.50
HIGH PERFORMANCE MOBILE ANTENNAS						
Single Band	KW-20	20	14.25	250	2000	1.85
Single Band	KW-40	40	7.05	100	2000	1.85
Single Band	KW-80	80	3.75	60	2000	2.50
HELICAL DIPOLE ANTENNAS						
Single Band	HWD-20	20	14.25	250	750	3.70
Single Band	HWD-40	40	7.05	100	750	5.00
Single Band	HWD-80	80	3.75	60	350	5.00

Accessories: Mobile base HWM-1, glass-fiber, single-hole mounting (16 mm dia.).
Gutter mount RM for HWM-1

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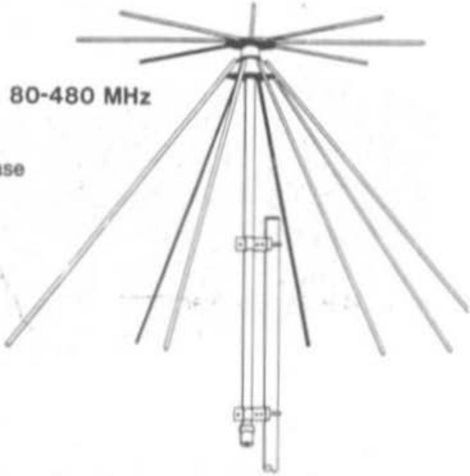
- Extremely sensitive double superhet
- For AC and battery operation (Connection cables provided)

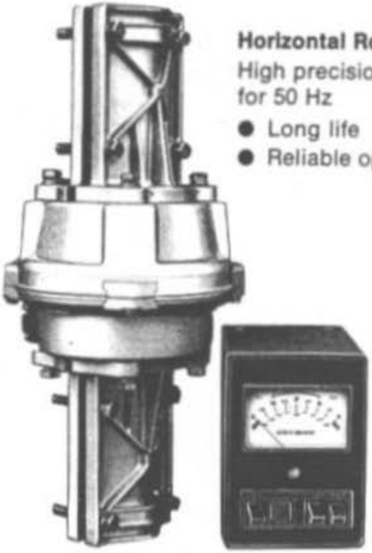
SPECIFICATIONS:

Frequency range:	108.000 - 135.975 MHz
Receive channels:	1120
Channel spacing:	25 kHz
Modulation mode:	AM (A 3)
Sensitivity:	$< 0.5\ \mu V$ / 20 dB (S + N)/N
Bandwidth:	15 kHz
Intermediate frequencies:	10.695 MHz / 455 kHz
Oscillators:	1st: PPL-synthesizer, 2nd: crystal
Spurious and image rejection:	Better than – 60 dB
Control time (AGC) :	0.1 - 0.5 s
Built-in effective noise blanker:	Especially designed for mobile use
Input connector:	50 Ω, SO 239
Temperature range:	– 20°C to + 60°C
Audio output:	> 1.5 W / < 5 % distortion
Built-in loudspeaker:	8 Ω, approx. 60 mm dia.
Operating voltages:	13.8 V ± 15 % / 220 VAC 50 Hz
DC-current drain:	0.8 A
Dimensions:	160 x 56 x 250 mm
Weight:	approx. 3 kg
Supplied accessories:	Power line cable, battery cable, telescope antenna, mobile mount.

WIDEBAND OMNIDIRECTIONAL DISCONE ANTENNA

- Frequency range: 80 - 480 MHz
- Gain: 3.4 dB / λ/4
- Impedance: 50 Ω
- Power rating: 500 W
- Polarisation: Vertical
- :Connection: SO 239 in water proof case
- VSWR: < 1.5 : 1
- Weight: 3 kg
- Dimensions: Height: 1.00 m
Diameter: 1.30 m
- Material: Aluminium
- Mounting: 2 mast clamps provided





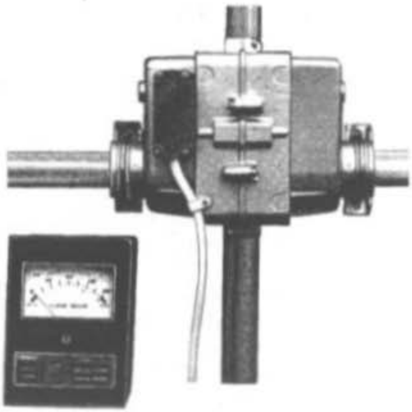
Horizontal Rotator KR 400
High precision rotator
for 50 Hz

- Long life
- Reliable operation

Type	KR 400
Load	200 kg
Brake torque	196 Nm *)
Torque	40 Nm *)
Mast diameter	38 - 63 mm
Speed (1 rev.)	60 s
Rotation angle	370°
Control cable	6 wires
Line voltage	220 V/50 Hz 50 VA
Dimensions (mm)	270 x 180 ∅
Weight	4.5 kg

*) 1 kpm ≙ 9.81 Nm
Both rotators designed for 50 Hz operation.

Vertical Rotator KR 500
Especially designed for
vertical tilting of antennas
for EME, OSCAR etc.



Type	KR 500
Load	ca. 400 kg
Brake torque	197 Nm *)
Torque	40 Nm *)
Horiz. tube diam.	32 - 43 mm
Mast diameter	38 - 63 mm
Speed (1 rev.)	74 s
Rotation angle	180° (+ 5°)
Control cable	6 wires
Line voltage	220 V/50 Hz 30 VA
Weight	4.5 kg

NEW SEMIPROFESSIONAL ANTENNAS FROM UKW-TECHNIK

Colinear Antennas for 2 m and 70 cm

Colinear C 5 / 2 m

No.Elements:	5 x λ/2
Gain:	4.8 dBd 8.55 dB / λ/4
Impedance:	50 Ω
Power Rating:	250 W
Length:	4.00 m
Weight:	3.2 kg
Wind Load at 160 km/h:	10.1 kg
Polarisation:	Vertical
Frequency:	144 - 148 MHz
Vertical Beamwidth:	24°
Termination:	50 Ω 'N' Type Socket
Shroud:	Glass-fibre
Mounting:	2 Type JBL 29/2 Steel Clamps
V.S.W.R.:	< 1.5 : 1

Colinear C 8 / 70 cm

No.Elements:	8 x λ/2
Gain:	7.8 dBd 11.55 dB / λ/4
Impedance:	50 Ω
Power Rating:	250 W
Length:	3.20 m
Weight:	3.5 kg
Wind Load at 160 km/h:	10.0 kg
Polarisation:	Vertical
Frequency:	430 - 440 MHz
Vertical Beamwidth:	12°
Termination:	50 Ω 'N' Type Socket
Shroud:	Glass-fibre
Mounting:	2 Type JBL 29/2 Steel Clamps
V.S.W.R.:	< 1.5 : 1

The following antennas are also available in semi-professional quality only from the UKW-TECHNIK outlets:

Parabeams PBM 10/2 m and PBM 14/2 m · Crossyagi 10 XY/2 m.

These antennas are suffixed »HP« and offer the following advantages over the standard models:

- 1/2" elements from seamless tubing
- Professional baluns with PBM 10/2 m HP and PBM 14/2 m HP
- Semiprofessional balun with 10 XY/2 m HP

The standard models of these antennas remain available.

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